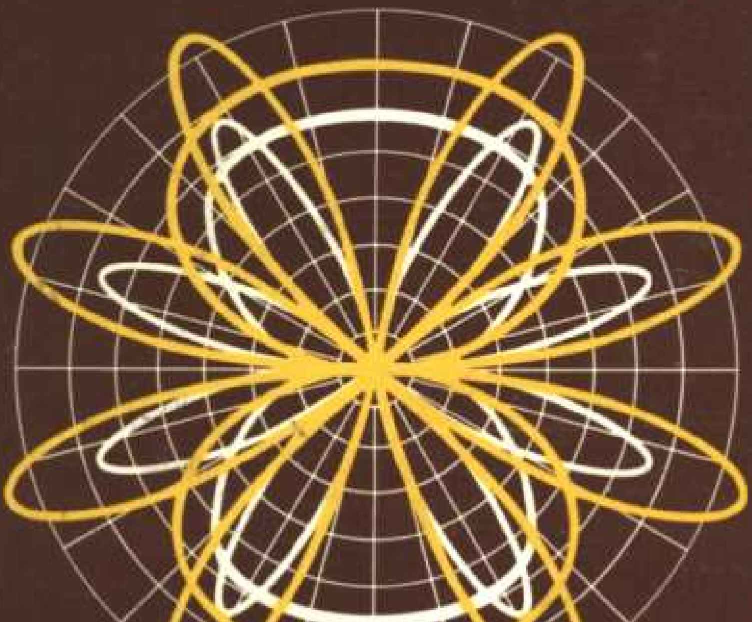
radio handbook

nineteenth edition

William I. Orr, W6SAI



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RADIO HANDBOOK

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PREFACE

At the turn of the Century farsighted experimenters were communicating by "wireless" over scores of miles using spark transmitters and magnetic detectors. Semiconductors and vacuum tubes were unknown to these daring men but the fundamental concepts of tuning, resonance and wave propagation were clearly understood.

Seventy years later the grandsons of these pioneers have instant, worldwide radio communication at their fingertips and can radio-command space probes traversing the outer reaches of the solar system.

The wireless experimenters of 1900 would be confounded by the vacuum-tube equipment of the "sixties" and dazzled by the solid-state techniques of the "seventies". But they would understand the underlying fundamentals of today's sophisticated equipment since it still obeys the natural laws of electricity as set forth by Ohm, Ampere, Faraday, Maxwell, Hertz, and others so long ago.

Thus radio communication is a continuing science and this latest edition of RADIO HANDBOOK reflects both basic fundamentals and the latest electronic techniques and practices.

Born in 1935 as a slim, paperback reference work, RADIO HANDBOOK has grown through 18 editions to its present position as the leading independent authority in the field of radio amateur h-f and vhf communication, faithfully covering more than three decades of development in the art of electronic communication.

Included in this 19th edition are expanded sections covering the latest advancements in communication electronics, circuit techniques, vhf solid-state power sources and uhf low-noise receiving devices. Of particular interest to the advanced amateur are the solid-state SSB receiver and exciter as well as the h-f broadband linear amplifiers and vhf f-m equipment.

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Contents

GLOSSARY OF TERMS

Symbol	Notation
A	Amperes (a-c, rms, or d-c)
A	Amplifier voltage gain
a	Amperes (peak)
ac, a-c, a.c.	Alternating current
a-m, a.m.	Amplitude modulation
C	Capacitance
c.f.m.	Cubic feet per minute
C_{gg}	Capacitance grid to ground
C_{gk}, C_{gp}, etc.	Tube capacitance between indicated electrodes
C_{in}	Input capacitance
C_k	Capacitance between cathode and ground
cm	Centimeter
C_N	Neutralizing capacitance
C_{out}	Output capacitance
C_{pg2}	Capacitance, plate to screen
cw, c.w. or c-w	Continuous wave
dB or db	Decibel
dc, d.c., d-c	Direct current
E	Voltage (a-c, rms, or d-c)
e	Peak voltage
E_b	Average plate voltage
e_b	Instantaneous plate voltage
$e_{b\,max}$	Peak plate voltage
$e_{b\,min}$	Minimum instantaneous plate voltage referenced to ground
e_{cmp}	Maximum positive grid voltage
E_{co}	Cutoff-bias voltage
E_{c1}	Average grid #1 voltage
E_{c2}	Average grid #2 voltage
E_{c3}	Average grid #3 voltage
e_{c1}	Instantaneous grid #1 voltage
e_{c2}	Instantaneous grid # 2 voltage
e_{c3}	Instantaneous grid #3 voltage
E_f	Filament voltage
e_g	Rms value of exciting voltage
e_p	Instantaneous plate voltage (a.c.) referenced to E_b
$e_{p\,max}$	Peak a-c plate voltage referenced to E_b
E_{sig}	Applied signal voltage (d-c)
e_{sig}	Applied signal voltage (a-c)
e_k	Instantaneous cathode voltage
$e_{k\,max}$	Peak cathode voltage
F	Farad
f	Frequency (in Hertz)
fil	Filament
G	Giga (10^9)
g, g_1, g_2, etc.	Grid (number to identify, starting from cathode)
$g_{2,4}$	Grids having common pin connection
GHz	Gigahertz (10^9 cycles per second)
G_m or S_m	Transconductance (grid-plate)
H	Henry
Hz	Hertz
i	Peak current
I	Current (a-c, rms, or d-c)
I_b	Average d-c plate current
$I_{b\,max}$	Peak signal d-c plate current
i_b	Instantaneous plate current
$i_{b\,max}$	Peak plate current
I_{bo}	Idling plate current
I_c	Average d-c grid current current
i_p	Instantaneous a-c plate current referred to I_b
$i_{p\,max}$	Peak a-c plate current referred to I_b
i_1 etc.	Fundamental component of r-f plate current
$i_{1\,max}$	Peak fundamental component of r-f plate current
I_1	Single tone d-c plate current
I_2 etc.	Two-tone, etc., d-c plate current
I_{c1}, $_{c2}$, etc.	Average grid #1, #2, etc. current
I_f	Filament current
i_{g1} i_{g2} etc.	Instantantous grid current
$i_{g1\,max}$, etc.	Peak grid current
I_k	Average cathode current
i_k	Instantaneous cathode current
$i_{k\,max}$	Peak cathode current
K	Cathode
k	Kilo(10^3)
kHz	Kilohertz
kV	Peak kilovolts
kVac	A-c kilovolts
kVdc	D-c kilovolts
kW	Kilowatts
λ	Wavelength
M	Mutual inductance
M	Mega (10^6)
m	Meter
m	One thousandth

Symbol	Notation
mm	Millimeter
mA or ma	Milliamperes
Meg or meg	Megohm
mH	Millihenry
MHz	Megahertz
Mu or μ	Amplification factor
mV or mv	Millivolts
MW	Megawatts
mW	Milliwatts
NF	Noise figure
N_p	Efficiency
p	Pico (10^{-12})
P_d	Average drive power
p_d	Peak drive power
P_{ft}	Average feedthrough power
p_{ft}	Peak feedthrough power
pF or pf	Picofarad
PEP	Peak envelope power
P_{g1}, P_{g2}, etc.	Power dissipation of respective grids
P_i	Power input (average)
p_i	Peak power input
P_o	Power output (average)
p_o	Peak power output
P_p	Plate dissipation
Q	Figure of merit
Q_L	Loaded Q
R	Resistance
r	Reflector
rf, r.f. or r-f	Radio frequency
R_g	Resistance in series with the grid.
r_g	Dynamic internal grid resistance
R_k	Resistance in series with the cathode
R_L	Load resistance
rms	Root mean square
R_p	Resistance in series with plate
r_p	Dynamic internal plate resistance
S_c or G_c	Conversion transconductance
S_m or G_m	Transconductance
SSB	Single sideband
SWR	Standing-wave ratio
T	Temperature (°C)
t	Time (seconds)
θ	Conduction angle
μ	Micro (10^{-6}) or amplification factor
μ	Amplification Factor
μA	Microampere
μmho	Micromho
μF or μfd	Microfarad
μH	Microhenry
μs	Microsecond
μV	Microvolt
μ_s	Grid-screen amplification factor
V	Volt(s), (a-c, rms, or d-c) or d.c.)
v	Peak volts
Vac	A-c volts
Vdc	D-c volts
VSWR	Voltage standing-wave ratio
W	Watts
Z	Impedance
Z_g	Grid impedance
Z_i	Input impedance
Z_k	Cathode impedance
Z_L	Load impedance
Z_o	Output impedance
Z_p	Impedance in plate circuit
Z_s	Screen bypass impedance

CHAPTER ONE

Introduction to Radio

The field of *radio* is a division of the much larger field of electronics. Radio itself is such a broad study that it is still further broken down into a number of smaller fields of which only short-wave or high-frequency radio is covered in this book. Specifically the field of communication on frequencies from 1.8 to 1296 MHz is taken as the subject matter for this work.

The largest group of persons interested in the subject of high-frequency communication is the more than 450,000 radio amateurs located in nearly all countries of the world. Strictly speaking, a *radio amateur* is anyone noncommercially interested in radio, but the term is ordinarily applied only to those hobbyists possessing transmitting equipment and a license to operate from the Government.

It was for the radio amateur, and particularly for the serious and more advanced amateur, that most of the equipment described in this book was developed. However, in each equipment group, simple items also are shown for the student or beginner. The design principles behind the equipment for high-frequency radio communication are of course the same whether the equipment is to be used for commercial, military, or amateur purposes. The principal differences lie in construction practices, and in the tolerances and safety factors placed on components.

With the increasing complexity of high-frequency communication, resulting primarily from increased utilization of the available spectrum, it becomes necessary to delve more deeply into the basic principles underlying radio communication, both from the standpoint of equipment design and operation and from the standpoint of signal propagation. Hence, it will be found that this edition of the RADIO HANDBOOK has been devoted in greater proportion to the teaching of the principles of equipment design and signal propagation. It is in response to requests from schools and agencies of the Department of Defense, in addition to persistent requests from the amateur radio fraternity, that coverage of these principles has been expanded.

1-1 Amateur Radio

Amateur radio is a fascinating hobby with many facets. So strong is the fascination offered by this hobby that many executives, engineers, and military and commercial operators enjoy amateur radio as an avocation, even though they are also engaged in the radio field commercially.

Amateurs have rendered much public service through furnishing communications to and from the outside world in cases where disaster has isolated an area by severing all wire communications. Amateurs have a proud record of heroism and service on such occasions. Many expeditions to remote places have been kept in touch with home by communication with amateur stations on the high frequencies. The amateur's fine record of performance with the "wireless" equipment of World War I has been surpassed by his outstanding service in World War II.

By the time peace came in the Pacific in the summer of 1945, many thousand amateur operators were serving in the Allied Armed Forces. They had supplied the Army, Navy, Marines, Coast Guard, Merchant Marine, Civil Service, war plants, and civilian defense organizations with *trained* personnel for radio, radar, wire, and visual communications and for teaching. The Veterans who came from these organizations, many of whom were radio amateurs, now are the backbone of our modern electronics industry. Their stature in the community emphasizes to the beginning radio amateur that his pastime is the gateway to a career in the expanding electronics industry and that amateur radio is indeed an impressive introduction to one of the most exciting fields of endeavor in this century.

1-2 Station and Operator Licenses

Every radio transmitting station in the United States (with the exception of certain low-power communication devices) must have a license from the Federal Government before being operated; some classes of stations must have a permit from the government even before being constructed. And every operator of a licensed transmitting station must have an operator's license before operating a transmitter. There are no exceptions. Similar laws apply in practically every major country.

Classes of Amateur Operator Licenses There are at present six classes of amateur operator licenses in the United States authorized by the Federal Communications Commission. These classes differ in many important respects, so each will be discussed briefly.

Novice Class—The Novice Class license is available to any U.S. citizen or national who has not previously held an amateur license of any class issued by any agency of the U.S. Government, military or civilian. The license is valid for a period of two years and is not renewable. However a former amateur licensee may apply for a new Novice Class license provided he has not held an amateur license for at least a period of one year prior to making application. The examination may be taken only by mail, under the direct supervision of an amateur holding a General Class license or higher, or a commercial radiotelegraph licensee. The examination consists of a code test in sending and receiving at a speed of 5 words per minute, plus a written examination on the rules and regulations essential to beginners operation, including sufficient elementary radio theory for the understanding of these rules. Restricted c-w privileges in segments of the 80-, 40-, 15-, and 2-meter amateur bands are currently available to the Novice licensee, whose transmitter is limited to crystal-controlled operation with an input power not exceeding 75 watts.

The receiving code test for the Novice Class license requires correct copy of five consecutive words of text counting five letters per word for a continuous period of at least one minute. Punctuation marks and numerals are included.

Technician Class—The Technician Class exists for the purpose of encouraging a greater interest in experimentation and development of the higher frequencies among experimenters and would-be radio amateurs. This Class of license is available to any U.S. Citizen or national. The examination is similar to that given for the General Class license, except that the code test in sending and receiving is at a speed of 5 words per minute.

The holder of a Technician Class license is accorded all authorized amateur privileges in all amateur bands above 220 MHz, and in portions of the 144-MHz and 50-MHz bands. This class of license may be taken only by mail, under the direct supervision of an amateur (21 years of age, or older) holding a General Class License, or higher, or a commercial radiotelegraph license. The license is valid for a period of five years, and may be renewed on proper application.

General Class—The General Class license is the standard radio amateur license and is available to any U.S. Citizen or national. The license is valid for a period of five years and is renewable on proper application. Applicants for the General Class license must take the examination before an FCC representative (with certain exceptions discussed under the Conditional Class license). The examination consists of a code test in sending and receiving at a code speed of 13 words per minute, plus a written examination in basic theory and regulations. It conveys all amateur privileges, *with the exceptions noted for the Advanced and Extra Class licenses.*

Conditional Class—The Conditional Class license is equivalent to the General Class license in the privileges accorded by its use. This license is issued to an applicant who: (1) lives more than 175 miles airline distance from the nearest point at which the FCC conducts examinations twice yearly, or oftener; (2) is unable to appear for examination because of physical disability to travel; (3) is unable to appear for examination because of military service; (4) is temporarily resident outside the United States, its territories, or possessions for a year or more. The Conditional Class license may be taken only by mail and is renewable.

Advanced Class—The Advanced Class license is equivalent to the old Class-A license and is available to any U.S. Citizen or national. The license is valid for a period of five years and is renewable on proper application. Applicants for the Advanced Class license must take the examination before an FCC representative. The examination consists of a general code test at 13 words per minute, questions covering general amateur practice and regulations involving radio operation, and technical questions covering intermediate-level radio theory and operation as applicable to modern amateur techniques, including, but not limited to, radiotelephony and radiotelegraphy. An applicant for the Advanced Class license will be given credit for that portion of the examination and the code test covered by the General Class license, if a valid license of that grade is held at the time of examination.

The Advanced Class license accords certain radiotelephone privileges in the amateur bands between 80 and 6 meters, which are unavailable to holders of lower-grade amateur licenses.

Amateur Extra Class—The Amateur Extra Class license is the highest-grade amateur license issued by the FCC and the recipient, on request, may receive a special diploma-type certificate from the District FCC Engineer-in-Charge. The license is valid for a period of five years and is renewable. Applicants for the Amateur Extra Class license must take the examination before an FCC representative. The examination consists of a code test in sending and receiving at a speed of 20 words per minute, a standard written examination in theory and regulations (credit will be given to holders of General and Advanced Class licenses for this requirement), and a written examination based on advanced radio theory and operation as applicable to modern amateur techniques, including, but not limited to, radiotelephony, radiotelegraphy, and transmissions of energy for measurements and observations applied to propagation, for the radio control of remote objects, and for similar experimental purposes. An applicant for the Amateur Extra Class license will be given credit for that portion of the examination covered by the General and Advanced Class licenses, if a valid license of either grade is held at the time of examination.

The Amateur Extra Class license accords certain radiotelephone and radiotelegraph privileges in the amateur bands between 80 and 6 meters, unavailable to holders of lower-grade licenses. In addition, the holder of an Amateur Extra Class license, licensed for 25 years or longer by the FCC prior to the date of the Amateur Extra Class license may request a two-letter call sign, in lieu of a three-letter call sign.

The Amateur Station License The station license authorizes the radio apparatus of the radio amateur for a particular address and designates the official call sign to be used. The license is a portion of the combined station-operator license normally issued to the radio amateur. Authorization is included for portable or mobile operation within the continental limits of the United States, its territories or possessions, on any amateur frequency authorized to the class of license granted the operator. If portable or mobile operation for a period of greater than 48 hours is contemplated, advance notice must be given to the FCC district in which operation will be conducted. The station license must be modified on a permanent change in address. The station license is customarily renewed with the operator license. Applications filed for amateur radio licenses (except that of a Novice Class) require a filing fee.

International Regulations The domestic regulatory pattern of the United States agrees with the international agreements established by the International Telecommunications Union and to which the United States is a signatory power. The frequency bands reserved for the Amateur Radio Service are included in the ITU frequency allocations table, as one of the services to which frequencies are made available. In the lower-frequency amateur bands, the international allocations provide for joint use of the bands by several services in addition to the amateur service in various areas of the world.

Article I of the ITU Radio Regulations defines the amateur service as: *"A service of self-training, intercommunication, and technical investigations carried on by amateurs, that is, by duly authorized persons interested in radio technique solely with a personal aim and without a pecuniary interest."* Within this concept, the U. S. radio regulations governing radio amateur licensing and regulation are formulated.

By reciprocal treaty, the United States now has a number of agreements with other countries permitting amateurs of one country to operate in the other. On the other hand, by international agreement, notification to the ITU may forbid international communications with radio amateurs of certain countries.

A comprehensive coverage of United States licensing procedure for radio amateurs and applicable rules and regulations may be found in *"The Radio Amateur's License Manual,"* published by the American Radio Relay League, Newington, Conn. 06111.

1-3 The Amateur Bands

Certain small segments of the radio-frequency spectrum between 1800 kHz and 22,000 MHz are reserved for operation of amateur radio stations. These segments are in general agreement throughout the world, although certain parts of different amateur bands may be used for other purposes in various geographic regions. In particular, the 40-meter amateur band is used legally (and illegally) for short-wave broadcasting by many countries in Europe, Africa and Asia. Parts of the 80-meter band are used for short distance marine work in Europe, and for broadcasting in South America. The amateur bands available to United States radio amateurs are:

160 Meters (1800 kHz—2000 kHz) The 160-meter band is divided into 25-kHz segments on a regional basis, with day and night power limitations, and is available for amateur use provided no interference is caused to the Loran (Long Range Navigation) stations operating in this band. This band is least affected by the 11-year solar sunspot cycle. The *maximum usable frequency* (MUF) even during the years of decreased sunspot activity does not usually drop below 4 MHz, therefore this band is not subject to the violent fluctuations found on the higher-frequency bands. DX contacts on this band are limited by the ionospheric absorption of radio signals, which is quite high. During winter nighttime hours the absorption is often of a low enough value to permit trans-oceanic contacts on this band. On rare occasions, contacts up to 10,000 miles have been made. As a usual rule, however, 160-meter amateur operation is confined to ground-wave contacts or single-skip contacts of 1000 miles or less. Popular before World War II, the 160-meter band is now only sparsely occupied since many areas of the country are blanketed by the megawatt pulses of the Loran chains.

80 Meters (3500 kHz—4000 kHz) The 80-meter band is the most popular amateur band in the continental United States for local "rag chewing" and traffic nets. During the years of minimum sunspot activity the ionospheric absorption on this band may be quite low, and long distance DX contacts are possible during the winter night hours. Daytime operation, in general, is limited to contacts of 500 miles or less. During the summer months, local static and high ionospheric absorption limit long distance contacts on this band. As the sunspot cycle advances and the MUF rises, increased ionospheric absorption will tend to degrade the long distance possibilities of this band. At the peak of the sunspot cycle, the 80-meter band becomes useful only for short-haul communication.

40 Meters (7000 kHz—7300 kHz) The 40-meter band is high enough in frequency to be severely affected by the 11-year sunspot cycle. During years of minimum solar activity, the MUF may drop below 7 MHz, and the band will become very erratic, with signals dropping completely out during the night hours. Ionospheric absorption of signals is not as large a problem on this band as it is on 80 and 160 meters. As the MUF gradually rises, the skip distance will increase on 40 meters, especially during the winter months. At the peak of the solar cycle, the daylight skip distance on 40 meters will be quite long, and stations within a distance of 500 miles or so of each other will not be able to hold communication. DX operation on the 40-meter band is considerably hampered by broadcasting stations, propaganda stations, and jamming transmitters. In Europe and Asia the band is in a chaotic state, and amateur operation in this region is severely hampered.

20 meters (14,000 kHz—14,350 kHz) At the present time, the 20-meter band is by far the most popular band for long-distance contacts. High enough in frequency to be almost obliterated at the bottom of the solar cycle, the band nevertheless provides good DX contacts during years of minimal sunspot activity. At the present time, the band is open to almost all parts of the world at some time during the year. During the summer months, the band is active until the late evening hours, but during the winter months the band is only good for a few hours during daylight. Extreme DX contacts are usually erratic, but the 20-meter band is the only band available for DX operation the year around during the bottom of the sunspot cycle. As the sunspot count increases and the MUF rises, the 20-meter band will become open for longer hours during the winter. The maximum skip distance increases, and DX contacts are possible over paths other than the Great Circle route. Signals can be heard via the "long path," 180 degrees opposite the Great Circle path. During daylight hours, absorption may become apparent on the 20-meter band, and all signals except very short skip may disappear. On the other hand, the band will be open for worldwide DX contacts all night long. The 20-meter band is very susceptible to "fadeouts" caused by solar disturbances, and all except local signals may completely disappear for periods of a few hours to a day or so.

15 Meters (21,000 kHz—21,450 kHz) This is a relatively new band for radio amateurs since it has only been available for amateur operation since 1952. It has characteristics similar to both the 20- and 10-meter amateur bands. During a period of low sunspot activity, the MUF will rarely rise as high as 15 meters, so this band will be "dead" for a large part of the sunspot cycle. During the next few years, 15-meter activity should be excellent, and the band should support extremely long DX contacts. The band will remain open 24 hours a day in Equatorial areas of the world.

Fifteen-meter operation may be hampered in some cases when neighbors possess older-model TV receivers having a 21-MHz i-f channel, which falls directly in the 15-meter band. The interference problem may be alleviated by retuning the i-f system to a frequency outside the amateur assignment.

10 Meters (28,000 kHz—29,700 kHz) During the peak of the sunspot cycle, the 10-meter band is without doubt the most popular amateur band. The combination of long skip and low ionospheric absorption make reliable DX contacts with low-powered equipment possible. The great width of the band (1700 kHz) provides room for a large number of amateurs. The long skip (1500 miles or so) prevents nearby amateurs from hearing each other, thus dropping the interference level. During the winter months, sporadic-E (short-skip) signals up to 1200 miles or so will be heard. The 10-meter band is poorest in the summer months, even during a sunspot maximum. Extremely long daylight skip is common on this band, and in years of high MUF the 10-meter band will support intercontinental DX contacts during daylight hours.

The second harmonic of stations operating in the 10-meter band falls directly into television channel 2, and the higher harmonics of 10-meter transmitters fall into the higher TV channels. This harmonic problem seriously curtailed amateur 10-meter operation during the late 40's. However, with new circuit techniques and the TVI precautionary measures stressed in this Handbook, 10-meter operation should cause little or no interference to nearby television receivers of modern design.

Six Meters (50 MHz—54 MHz) At the peak of the sunspot cycle, the MUF occasionally rises high enough to permit DX contacts up to 10,000 miles or so on 6 meters. Activity on this band during such a period is often quite high. Interest in this band wanes during a period of lesser solar activity, since contacts, as a rule, are restricted to short-skip work. The proximity of the 6-meter band to television channel 2 often causes interference problems to amateurs located in areas where channel 2 is active. As the sunspot cycle increases, activity on the 6-meter band will increase.

The VHF Bands (Two Meters and "Up") The vhf bands are the least affected by the vagaries of the sunspot cycle and Heaviside layer. Their predominant use is for reliable communication over distances of 150 miles or less. These bands are sparsely occupied in the rural sections of the United States, but are quite heavily congested in the urban areas of high population.

In recent years it has been found that vhf signals are propagated by other means than by line-of-sight transmission. "Scatter signals," Aurora reflection, and air-mass boundary bending are responsible for vhf communication up to 1200 miles or so. Weather conditions will often affect long-distance communication on the 2-meter band, and all the vhf bands are particularly sensitive to this condition.

In recent years the vhf bands have been used for experimental "moonbounce" (earth-moon-earth) transmissions and for repeater-satellite experiments (Project Oscar). The vhf bands hold great promise for serious experimenters as radio amateurs forge into the microwave region.

1-4 Starting Your Study

When you start to prepare yourself for the amateur examination you will find that the circuit diagrams, tube characteristic curves, and formulas appear confusing and difficult to understand. But after a few study sessions one becomes sufficiently familiar with the notation of the diagrams and the basic concepts of theory and operation so that the acquisition of further knowledge becomes easier and even fascinating.

Since it takes a considerable time to become proficient in sending and receiving code, it is a good idea to intersperse technical study sessions with periods of code practice. Many short code-practice sessions benfit one more than a small number of longer sessions. Alternating between one study and the other keeps the student from getting "stale" since each type of study serves as a sort of respite from the other.

When you have practiced the code long enough you will be able to follow the gist of the slower-sending stations. Many stations send very slowly when working other stations at great distances. Stations repeat their calls many times when calling other stations before contact is established, and one need not have achieved much code proficiency to make out their calls and thus determine their location.

The Code The applicant for any class of amateur operator license must be able to send and receive the Continental Code (sometimes called the International Morse Code). The speed required for the sending and receiving test may be either 5, 13, or 20 words per minute, depending on the class of license assuming an average of five characters to the word in each case. The sending and receiving tests run for five minutes, and one minute of errorless transmission or reception must be accomplished within the five-minute interval.

If the code test is failed, the applicant

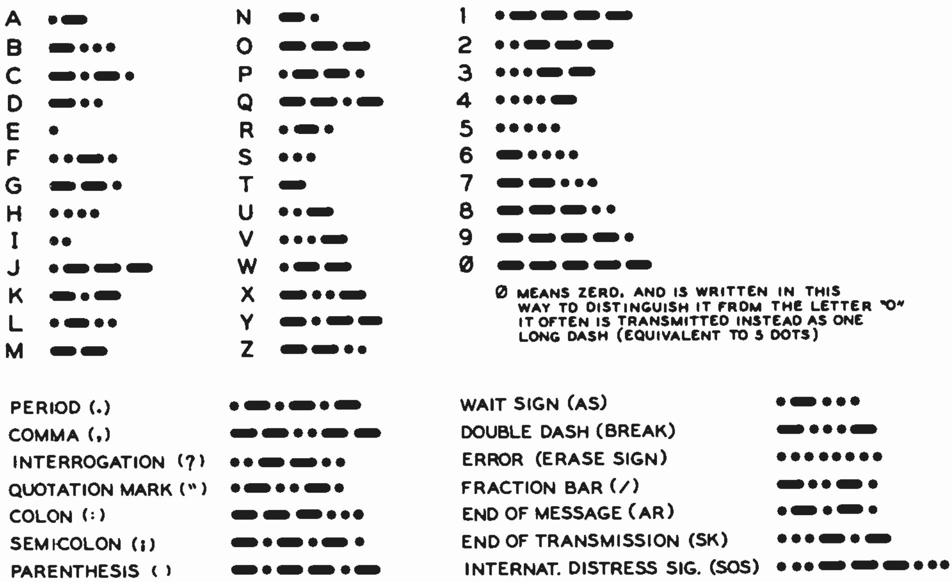


Figure 1

The Continental (or International Morse) Code is used for substantially all non-automatic radio communication. DO NOT memorize from the printed page; code is a language of SOUND, and must not be learned visually; learn by listening as explained in the text.

must wait at least one month before he may again appear for another test. Approximately 30% of amateur applicants fail to pass the test. It should be expected that nervousness and excitement will, at least to some degree, temporarily lower the applicant's code ability. The best insurance against this is to master the code at a little greater than the required speed under ordinary conditions. Then if you slow down a little due to nervousness during a test the result will not prove fatal.

Memorizing the Code There is no shortcut to code proficiency. To memorize the alphabet entails but a few evenings of diligent application, but considerable time is required to build up speed. The exact time required depends on the individual's ability and the regularity of practice.

While the speed of learning will naturally vary greatly with different individuals, about 70 hours of practice (no practice period to be over 30 minutes) will usually suffice to bring a speed of about 13 w.p.m.; 16 w.p.m. requires about 120 hours; 20 w.p.m., 175 hours.

Since code reading requires that individual letters be recognized instantly, any memorizing scheme which depends on orderly sequence, such as learning all "*dah*" letters and all "*dit*" letters in separate groups, is to be discouraged. Before beginning with a code practice set it is necessary to memorize the whole alphabet perfectly. A good plan is to study only two or three letters a day and to drill with those letters until they become part of your consciousness. Mentally translate each day's letters into their sound equivalent wherever they are seen, on signs, in papers, indoors and outdoors. Tackle two additional letters in the code chart each day, at the same time reviewing the characters already learned.

Avoid memorizing by routine. Be able to sound out any letter immediately without so much as hestitating to think about the letters preceding or following the one in question. Know *C*, for example, apart from the sequence *ABC*. Skip about among all the characters learned, and before very long sufficient letters will have been acquired to enable you to spell out simple words to yourself in "*dit dahs*." This is interesting

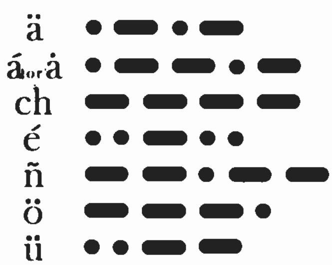


Figure 2

These code characters are used in languages other than English. They may occasionally be encountered so it is well to know them.

exercise, and for that reason it is good to memorize all the vowels first and the most common consonants next.

Actual code practice should start only when the entire alphabet, the numerals, period, comma, and question mark have been memorized so thoroughly that any one can be sounded without the slightest hesitation. Do not bother with other punctuation or miscellaneous signals until later.

Sound — Not Sight Each letter and figure *must* be memorized by its *sound* rather than its appearance. Code is a system of sound communication, the same as is the spoken word. The letter *A*, for example, is one short and one long sound in combination sounding like *dit dah*, and it must be remembered as such, and not as "dot dash."

Practice Time, patience, and regularity are required to learn the code properly. Do not expect to accomplish it within a few days.

Don't practice too long at one stretch; it does more harm than good. Thirty minutes at a time should be the limit.

Lack of regularity in practice is the most common cause of lack of progress. Irregular practice is very little better than no practice at all. Write down what you have heard; then forget it; *do not look back.* If your mind dwells even for an instant on a signal about which you have doubt, you will miss the next few characters while your attention is diverted.

While various automatic code machines, phonograph records, etc., will give you practice, by far the best practice is to obtain a study companion who is also interested in learning the code. When you have both memorized the alphabet you can start sending to each other. Practice with a key and oscillator or key and buzzer generally proves superior to all automatic equipment. Two such sets operated between two rooms are fine—or between your house and his will be just that much better. Avoid talking to your partner while practicing. If you must ask him a question, do it in code. It makes more interesting practice than confining yourself to random practice material.

When two co-learners have memorized the code and are ready to start sending to each other for practice, it is a good idea to enlist the aid of an experienced operator for the first practice session or two so that they will get an idea of how properly formed characters sound.

During the first practice period the speed should be such that substantially solid copy can be made without strain. Never mind if this is only two or three words per minute. In the next period the speed should be increased slightly to a point where nearly all of the characters can be caught only through conscious effort. When the student becomes proficient at this new speed, another slight increase may be made, progressing in this manner until a speed of about 16 words per minute is attained if the object is to pass the amateur 13-word per minute code test. The margin of 3 w.p.m. is recommended to overcome a possible excitement factor at examination time. Then when you take the test you don't have to worry about the "jitters" or an "off day."

Speed should not be increased to a new level until the student finally makes solid copy with ease for at least a five-minute period at the old level. How frequently increases of speed can be made depends on individual ability and the amount of practice. Each increase is apt to prove disconcerting, but remember "you are never learning when you are comfortable."

A number of amateurs are sending code practice on the air on schedule once or twice each week; excellent practice can be obtained after you have bought or constructed your receiver by taking advantage of these sessions.

If you live in a medium-size or large city, the chances are that there is an amateur-radio club in your vicinity which offers free code-practice lessons periodically.

Skill When you listen to someone speaking you do not consciously think how his words are spelled. This is also true when you read. In code you must train your ears to read code just as your eyes were trained in school to read printed matter. With enough practice you acquire skill, and from skill, speed. In other words, it becomes a *habit,* something which can be done without conscious effort. Conscious effort is fatal to speed; we can't think rapidly enough; a speed of 25 words a minute, which is a common one in commercial operations, means 125 characters per minute or more than two per second, which leaves no time for conscious thinking.

Perfect Formation of Characters When transmitting on the code practice set to your partner, concentrate on the *quality* of your sending, *not* on your speed. Your partner will appreciate it and he could not copy you if you speeded up anyhow.

If you want to get a reputation as having an excellent "fist" on the air, just remember that speed alone won't do the trick. Proper execution of your letters and spacing will make much more of an impression. Fortunately, as you get so that you can send evenly and accurately, your sending speed will automatically increase. Remember to try to see how *evenly* you can

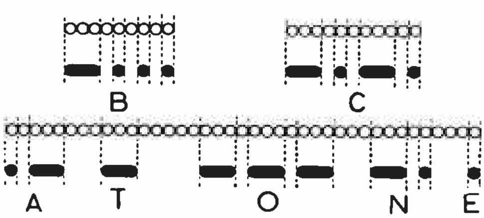


Figure 3

Diagram illustrating relative lengths of dashes and spaces referred to the duration of a dot. A dash is exactly equal in duration to three dots; spaces between parts of a letter equal one dot; those between letters, three dots; space between words, five dots. Note that a slight increase between two parts of a letter will make it sound like two letters.

send, and how *fast* you can *receive.* Concentrate on making signals properly with your key. Perfect formation of characters is paramount to everything else. Make every signal right no matter if you have to practice it hundreds or thousands of times. Never allow yourself to vary the slightest from perfect formation once you have learned it.

If possible, get a good operator to listen to your sending for a short time, asking him to criticize even the slightest imperfections.

Timing It is of the utmost importance to maintain uniform spacing in characters and combinations of characters. Lack of uniformity at this point probably causes beginners more trouble than any other single factor. Every dot, every dash, and every space must be correctly timed. In other words, accurate timing is absolutely essential to intelligibility, and timing of the spaces between the dots and dashes is just as important as the lengths of the dots and dashes themselves.

The characters are timed with the dot as a "yardstick." A standard dash is three times as long as a dot. The spacing between parts of the same letter is equal to one dot, the space between letters is equal to three dots, and that between words equal to five dots.

The rule for spacing between letters and words is not strictly observed when sending slower than about 10 words per minute for the benefit of someone learning the code and desiring receiving practice. When sending at, say, 5 w.p.m., the individual letters should be made the same as if the sending rate were about 10 w.p.m., except that the spacing between letters and words is greatly exaggerated. The reason for this is obvious. The letter *L,* for instance, will then sound exactly the same at 10 w.p.m. as at 5 w.p.m., and when the speed is increased above 5 w.p.m. the student will not have to become familiar with what may seem to him like a new sound, although it is in reality only a faster combination of dots and dashes. At the greater speed he will merely have to learn the identification of the *same* sound without taking as long to do so.

Be particularly careful of letters like *B.* Many beginners seem to have a tendency to leave a longer space after the dash than that which they place between succeeding dots, thus making it sound like *TS*. Similarly, make sure that you do not leave a longer space after the first dot in the letter *C* than you do between other parts of the same letter: otherwise it will sound like *NN*.

Sending vs. Receiving Once you have memorized the code thoroughly you should concentrate on increasing your *receiving* speed. True, if you have to practice with another newcomer who is learning the code with you, you will both have to do some sending. But don't attempt to practice *sending* just for the sake of increasing your sending *speed*.

When transmitting code to your partner so that he can practice, concentrate on the *quality* of your sending, not on your speed.

Because it is comparatively easy to learn to send rapidly, especially when no particular care is given to the quality of sending, many operators who have just received their licenses get on the air and send mediocre (or worse) code at 20 w.p.m. when they can barely receive good code at 13. Most old-timers remember their own period of initiation and are only too glad to be patient and considerate if you tell them that you are a newcomer. But the surest way to incur their scorn is to try to impress them with your "lightning speed," and then to request them to send more slowly when they come back at you at the same speed.

Stress your copying ability; never stress your sending ability. It should be obvious that if you try to send faster than you can receive, your ear will not recognize any mistakes which your hand may make.

Using the Key Figure 4 shows the proper position of the hand, fingers and wrist when manipulating a telegraph or radio key. The forearm should rest naturally on the desk. It is preferable that the key be placed far enough back from the edge of the table (about 18 inches) that the elbow can rest on the table. Otherwise, pressure of the table edge on the arm will tend to hinder the circulation of the blood and weaken the ulnar nerve at a point where it is close to the surface, which in turn will tend to increase fatigue considerably.

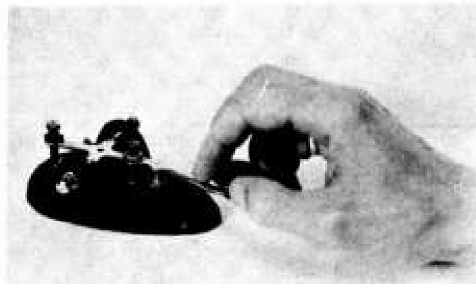


Figure 4

PROPER POSITION OF THE FINGERS FOR OPERATING A TELEGRAPH KEY

The fingers hold the knob and act as a cushion. The hand rests lightly on the key. The muscles of the forearm provide the power, the wrist acting as the fulcrum. The power should not come from the fingers, but rather from the forearm muscles.

The knob of the key is grasped lightly with the thumb along the edge; the index and third fingers rest on the top towards the front or far edge. The hand moves with a free up and down motion, the wrist acting as a fulcrum. The power must come entirely from the arm muscles. The third and index fingers will bend slightly during the sending but not because of deliberate effort to manipulate the finger muscles. Keep your finger muscles just tight enough to act as a cushion for the arm motion and let the slight movement of the fingers take care of itself. The key's spring is adjusted to the individual wrist and should be neither too stiff nor too loose. Use a moderately stiff tension at first and gradually lighten it as you become more proficient. The separation between the contacts must be the proper amount for the desired speed, being somewhat under 1/16 inch for slow speeds and slightly closer together (about 1/32 inch) for faster speeds. Avoid extremes in either direction.

Do not allow the muscles of arm, wrist or fingers to become tense. Send with a full, free arm movement. Avoid like the plague any finger motion other than the slight cushioning effect mentioned above.

Stick to the regular handkey for learning code. No other key is satisfactory for this purpose. Not until you have thoroughly mastered both sending and receiving at the maximum speed in which you are interested should you tackle any form of automatic or semiautomatic key such as the *Vibroplex* ("bug") or an electronic key.

Difficulties Should you experience difficulty in increasing your code speed after you have once memorized the characters, there is no reason to become discouraged. It is more difficult for some people to learn code than for others, but there is no justification for the contention sometimes made that "some people just can't learn the code." It is not a matter of intelligence; so don't feel ashamed if you seem to experience a little more than the usual difficulty in learning code. Your reaction time may be a little slower or your coordination not so good. If this is the case, remember *you can still learn the code.* You may never learn to send and receive at 40 w.p.m., but you can learn sufficient speed for all noncommercial purposes (and even for most commercial purposes) if you have patience, and refuse to be discouraged by the fact that others seem to pick it up more rapidly.

When the sending operator is sending just a bit too fast for you (the best speed for practice), you will occasionally miss a signal or a small group of them. When you do, leave a blank space; do not spend time futilely trying to recall it; dismiss it, and center attention on the next letter; otherwise you'll miss more. Do not ask the sender any questions until the transmission is finished.

To prevent guessing and get equal practice on the less common letters, depart occasionally from plain language material and use a jumble of letters in which the usually less commonly used letters predominate.

As mentioned before, many students put a greater space after the dash in the letter *B*, than between other parts of the same letter so it sounds like *TS*. *C*, *F*, *Q*, *V*, *X*, *Y*, and Z often give similar trouble. Make a list of words or arbitrary combinations in which these letters predominate and practice them, both sending and receiving until they no longer give you trouble. Stop everything else and stick to them. So long as these characters give you trouble you are not ready for anything else.

Follow the same procedure with letters which you may tend to confuse such as *F* and *L*, which are often confused by beginners. Keep at it until you *always* get them right without having to stop *even an instant* to think about it.

If you do not instantly recognize the sound of any character, you have not learned it; go back and practice your alphabet further. You should never have to omit writing down every signal you hear except when the transmission is too fast for you.

Write down what you hear, not what you think it should be. It is surprising how often the word which you guess will be wrong.

Copying Behind All good operators copy several words behind, that is, while one word is being received, they are writing down or typing, say the fourth or fifth previous word. At first this is very difficult, but after sufficient practice it will be found actually to be easier than copying close up. It also results in more accurate copy and enables the receiving operator to capitalize and punctuate copy as he goes along. It is not recommended that the beginner attempt to do this until he can send and receive accurately and with ease at a speed of at least 12 words a minute.

It requires a considerable amount of training to disassociate the action of the subconscious mind from the direction of the conscious mind. It may help some in obtaining this training to write down two columns of short words. Spell the first word in the first column out loud while writing down the first word in the second column. At first this will be a bit awkward, but you will rapidly gain facility with practice. Do the same with all the words, and then reverse columns.

Next try speaking aloud the words in the one column while writing those in the other column; then reverse columns.

After the foregoing can be done easily, try sending with your key the words in one column while spelling those in the other. It won't be easy at first, but it is well worth keeping after if you intend to develop any real code proficiency. Do *not* attempt to

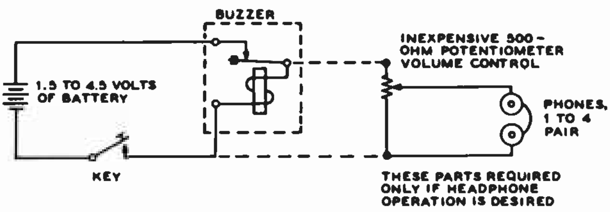


Figure 5

THE SIMPLEST CODE PRACTICE SET CONSISTS OF A KEY AND A BUZZER

The buzzer is adjusted to give a steady, high-pitched whine. If desired, the phones may be omitted, in which case the buzzer should be mounted firmly on a sounding board. Crystal, magnetic, or dynamic earphones may be used. Additional sets of phones should be connected in parallel, not in series.

catch up. There is a natural tendency to close up the gap, and you must train yourself to overcome this.

Next have your code companion send you a word either from a list or from straight text; do not write it down yet. Now have him send the next word; *after* receiving this second word, write down the first word. After receiving the third word, write the second word; and so on. Never mind how slowly you must go, even if it is only two or three words per minute. *Stay behind.*

It will probably take quite a number of practice sessions before you can do this with any facility. After it is relatively easy, then try staying two words behind; keep this up until it is easy. Then try three words, four words, and five words. The more you practice keeping received material in mind, the easier it will be to stay behind. It will be found easier at first to copy material with which one is fairly familiar, then gradually switch to less familiar material.

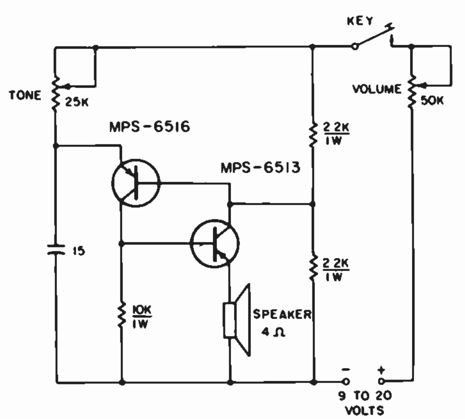


Figure 6

Two inexpensive "hobby"-type transistors and a 9-volt battery, plus a handful of parts make up a code-practice oscillator. Volume and tone are controlled by the potentiometers. Low-impedance earphones may be substituted for the speaker, if desired.

Automatic Code Machines The two practice sets which are described in this chapter are of most value when you have someone with whom to practice. Automatic code machines are not recommended to anyone who can possibly obtain a companion with whom to practice, someone who is also interested in learning the code. If you are unable to enlist a code partner and have to practice by yourself, the best way to get receiving practice is by the use of a tape machine (automatic code-sending machine) with several practice tapes. Or you can use a set of phonograph code-practice records. The records are of use only if you have a phonograph whose turntable speed is readily adjustable. The tape machine can be rented by the month for a reasonable fee.

Once you can copy about 10 w.p.m. you can also get receiving practice by listening to slow-sending stations on your receiver. Many amateur stations send slowly particularly when working far distant stations. When receiving conditions are particularly poor many commercial stations also send slowly, sometimes repeating every word. Until you can copy around 10 w.p.m. your receiver isn't much use, and either another operator or a machine or records is necessary for getting receiving practice after you have once memorized the code.

Code Practice Sets If you don't feel too foolish doing it, you can secure a measure of code practice with

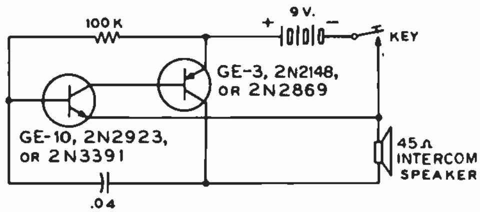


Figure 7

CODE-PRACTICE OSCILLATOR SUITABLE FOR SPEAKER OPERATION.

the help of a partner by sending "dit-dah" messages to each other while riding to work, eating lunch, etc. It is better, however, to use a buzzer or code-practice oscillator in conjunction with a regular telegraph key.

As a good key may be considered an investment it is wise to make a well-made key your first purchase. Regardless of what type code-practice set you use, you will need a key, and later on you will need one to key your transmitter. If you get a good key to begin with, you won't have to buy another one later.

The key should be rugged and have fairly heavy contacts. Not only will the key stand up better, but such a key will contribute to the "heavy" type of sending so desirable for radio work. Morse (telegraph) operators use a "light" style of sending and can send somewhat faster when using this light touch. But, in radio work static and interference are often present, and a slightly heavier dot is desirable. If you use a husky key, you will find yourself automatically sending in this manner.

To generate a tone simulating a code signal as heard on a receiver, either a mechanical buzzer or an audio oscillator may be used. Figure 5 shows a simple code-practice set using a buzzer which may be used directly simply by mounting the buzzer on a sounding board, or the buzzer may be used to feed from one to four pairs of conventional high-impedance phones.

An example of the audio-oscillator type of code-practice set is illustrated in figure 6. Two inexpensive "hobby-type transistors are used and the unit is powered by a 9-volt transistor radio battery. Low-impedance (500-ohm) earphones may be substituted for the speaker, if desired. The oscillator may be built up on a phenolic circuit board.

A code-practice oscillator that will drive a speaker to good room volume is shown in figure 7. Inexpensive entertainment-type transistors are used and any size permanent magnet speaker may be used. Mount the speaker on a large sounding board for best volume.

CHAPTER TWO

Direct-Current Circuits

All naturally occurring matter (excluding artifically produced radioactive substances) is made up of 92 fundamental constituents called *elements*. These elements can exist either in the free state such as iron, oxygen, carbon, copper, tungsten, and aluminum, or in chemical unions commonly called *compounds*. The smallest unit which still retains all the original characteristics of an element is the *atom*.

Combinations of atoms, or subdivisions of compounds, result in another fundamental unit, the *molecule*. The molecule is the smallest unit of any compound. All reactive elements when in the gaseous state also exist in the molecular form, made up of two or more atoms. The nonreactive gaseous elements helium, neon, argon, krypton, xenon, and radon are the only gaseous elements that ever exist in a stable monatomic state at ordinary temperatures.

2-1 The Atom

An atom is an extremely small unit of matter—there are literally billions of them making up so small a piece of material as a speck of dust. To understand the basic theory of electricity and hence of radio, we must go further and divide the atom into its main components, a positively charged *nucleus* and a cloud of negatively charged particles that surround the nucleus. These particles, swirling around the nucleus in elliptical orbits at an incredible rate of speed, are called *orbital electrons*.

It is on the behavior of these orbital electrons when freed from the atom, that depends the study of electricity and radio, as well as allied sciences. Actually it is possible to subdivide the nucleus of the atom into a dozen or so different particles, but this further subdivision can be left to quantum mechanics and atomic physics. As far as the study of electronics is concerned it is only necessary for the reader to think of the normal atom as being composed of a nucleus having a net positive charge that is exactly neutralized by the one or more orbital electrons surrounding it.

The atoms of different elements differ in respect to the charge on the positive nucleus and in the number of electrons revolving around this charge. They range all the way from hydrogen, having a net charge of one on the nucleus and one orbital electron, to uranium with a net charge of 92 on the nucleus and 92 orbital electrons. The number of orbital electrons is called the *atomic number* of the element.

Action of the Electrons From the foregoing it must not be thought that the electrons revolve in a haphazard manner around the nucleus. Rather, the electrons in an element having a large atomic number are grouped into rings having a definite number of electrons. The only atoms in which these rings are completely filled are those of the inert gases mentioned before; all other elements have one or more uncompleted rings of electrons. If the uncompleted ring is nearly empty, the element is *metallic* in character, being most metallic when there is only one electron in the outer ring. If the incomplete ring lacks only one or two electrons, the element is usually *nonmetallic*. Elements with a ring about half completed will exhibit both nonmetallic and metallic characteristics; carbon, silicon, germanium, and arsenic are examples. Such elements are called *semiconductors*.

In metallic elements these outer ring electrons are rather loosely held. Consequently, there is a continuous helter-skelter movement of these electrons and a continual shifting from one atom to another. The electrons which move about in a substance are called *free electrons*, and it is the ability of these electrons to drift from atom to atom which makes possible the *electric current*.

Conductors, Semiconductors, and Insulators If the free electrons are numerous and loosely held, the element is a good *conductor*. On the other hand, if there are few free electrons (as is the case when the electrons in an outer ring are tightly held), the element is a poor conductor. If there are virtually no free electrons, the element is a good *insulator*.

Materials having few free electrons are classed as semiconductors and exhibit conductivity approximately midway between that of good conductors and good insulators.

2-2 Fundamental Electrical Units and Relationships

Basic Electrical Dimensions, Units, and Symbols Electrical dimensions, units, and qualities are expressed as letters, combinations of letters, and other characters that may be used in place of the proper names for these characteristics. In addition, various prefixes are added to the symbols to indicate multiples or submultiples of units (Table 1).

The international system of fundamental units which covers mechanics, electricity, and magnetism is designated the *Rational MKS* (meter-kilogram-second) *System*. The *coulomb* is taken as a fourth fundamental unit.

Fundamental and Secondary Units Electrical measurements expressed in the MKS System are traceable to the *National Bureau of Standards* in the United States. Aside from the meter, kilogram, and second, the major electrical unit is the coulomb (Q), a unit of charge (6.28×10^{15} electron charges). The coulomb is defined as an *ampere-second*, or that steady current flowing through a solution of silver nitrate, which will deposit silver at the rate of 1.118×10^{-6} kilograms per second.

TABLE 1.

PREFIXES TO ELECTRICAL DIMENSIONS

PREFIX	MULTIPLE	SYMBOL
Giga-	10^9	G
Mega-	10^6	M
kilo-	10^3	k
deci-	10^{-1}	d
centi-	10^{-2}	c
milli-	10^{-3}	m
micro-	10^{-6}	μ or u
pico-	10^{-12}	p

Secondary, or *derived units*, are based on the above listed fundamental units. The rate of current flow is the *ampere* (*I*), whose dimensions are in coulombs per second. The unit of energy or work is the *joule* (*J*) whose dimensions are volts × coulombs. The unit of power is the *watt* (*W*), whose dimensions are joules per second. The electrical pressure that moves a coulomb of charge past a measuring point is the *volt* (*E* or *V*), whose dimensions are joules per coulomb.

The unit of opposition to current flow is the *ohm* (*R*), whose dimensions are volts per ampere. Two units express charge storage in a circuit. The first is the *farad* (*F*), a unit of capacitance whose dimensions are coulombs per volt. The second is the *henry* (*H*), a unit of inductance whose dimensions are volts per ampere-second. These and other electrical units are summarized in Table 2. Other complex quantities may be built up from these units.

Electromotive Force: Potential Difference The free electrons in a conductor move constantly about and

TABLE 2. ELECTRICAL UNITS

CHARACTERISTIC	SYMBOL	UNIT	DESCRIPTION
Charge	Q or q	coulomb	6.28×10^{18} electric charges
Voltage	E or e V or v	Volt	potential difference (joules per coulomb)
Current	I or i	Ampere	electrons in motion (coulombs per second)
Resistance	R or r	Ohm	electrical resistance (volts per ampere)
Conductance	G or g	mho	reciprocal of resistance
Energy	J	Joule	quantity of work (volts x coulombs)
Power	W	Watt	unit of power (joules per second)
Storage	F	Farad	unit of charge storage (coulombs per volt)
Storage	H	Henry	unit of inductance (volts per ampere-second)

change their position in a haphazard manner. To produce a drift of electrons, or *electric current*, along a wire it is necessary that there be a difference in "pressure" or *potential* between the two ends of the wire. This *potential difference* can be produced by connecting a source of *electrical potential* to the ends of the wire.

As will be explained later, there is an excess of electrons at the negative terminal of a battery and a deficiency of electrons at the positive terminal, due to chemical action. When the battery is connected to the wire, the deficient atoms at the positive terminal attract free electrons from the wire in order for the positive terminal to become neutral. The attracting of electrons continues through the wire, and finally the excess electrons at the negative terminal of the battery are attracted by the positively charged atoms at the end of the wire. Other sources of electrical potential (in addition to a battery) are: an electrical generator (dynamo), a thermocouple, an electrostatic generator (static machine), a photoelectric cell, and a crystal or piezoelectric generator.

Thus it is seen that a potential difference is the result of a difference in the number of electrons between the two (or more) points in question. The force or pressure due to a potential difference is termed the *electromotive* force, usually abbreviated *e.m.f.* or *E.M.F.* It is expressed in units called *volts*.

It should be noted that for there to be a potential difference between two bodies or points it is not necessary that one have a positive charge and the other a negative charge. If two bodies each have a negative charge, but one more negative than the other, the one with the lesser negative charge will act as though it were positively charged *with respect to the other body*. It is the *algebraic* potential difference that determines the force with which electrons are attracted or repulsed, the potential of the earth being taken as the zero reference point.

The Electric Current The flow of electric charges, either electrons, holes (see Chapter Four), or ions constitutes an *electric current*. The flow may be induced by the application of an electromotive force. This flow, or drift, is in addition to the irregular movements of the electrons. However, it must not be thought that each free electron travels from one end of the circuit to the other. On the contrary, each free electron travels only a short distance before colliding with an atom; this collision generally knocks off one or more electrons from the atom, which in turn move a short distance and collide with other atoms, knocking off other electrons. Thus, in the general drift of electrons along a wire carrying an electric current, each electron travels only a short distance and the excess of electrons at one end and the deficiency at the other are balanced by the source of the e.m.f. When this source is removed the state of normalcy returns; there is still the rapid interchange of free electrons between atoms, but there is no general trend or "net movement" in either one direction or the other—in other words, no current flows.

Current and Electron Flow Older textbooks speak of current flow as being from the positive terminal of the e.m.f. source through the conductor to the negative terminal. Nevertheless, it has long been an established fact that the current flow in a metallic conductor is the *electron* drift from the negative terminal of the source of voltage through the conductor to the positive terminal. The only exceptions to the electronic direction of flow occur in gaseous and electrolytic conductors where the flow of positive *ions* toward the cathode, or negative electrode, constitutes a positive flow in the opposite direction to the electron flow. (An ion is an atom, molecule, or particle which either lacks one or more electrons, or else has an excess of one or more electrons.)

In radio work the terms "electron flow" and "current" are becoming accepted as being synonymous, but the older terminology is still accepted in the electrical (industrial) field. Because of the confusion this sometimes causes, it is often safer to refer to the direction of electron flow rather than to the direction of the "current." Since electron flow consists actually of a passage of *negative* charges, current flow and *algebraic* electron flow do pass in the same direction.

Resistance The flow of current in a material depends on the ease with which electrons can be detached from the atoms of the material and on its molecular structure. In other words, the easier it is to detach electrons from the atoms the more free electrons there will be to contribute to the flow of current, and the fewer collisions that occur between free electrons and atoms the greater will be the total electron flow.

The opposition to a steady electron flow is called the *resistance* (R) of a material, and is one of its physical properties.

The unit of resistance is the *ohm*. Every substance has a *specific resistance*, usually expressed as *ohms per mil-foot*, which is determined by the material's molecular structure and temperature. A mil-foot is a piece of material one circular mil in area and one foot long. Another measure of resistivity frequently used is expressed in the units *microhms per centimeter cube*. The resistance of a uniform length of a given substance is directly proportional to its length and specific resistance, and inversely proportional to its cross-sectional area. A wire with a certain resistance for a given length will have twice as much resistance if the length of the wire is doubled. For a given length, doubling the cross-sectional area of the wire will *halve* the resistance, while doubling the *diameter* will reduce the resistance to *one fourth*. This is true since the cross-sectional area of a wire varies as the square of the diameter. The relationship between the resistance and the linear dimensions of a conductor may be expressed by the following equation:

TABLE 3. TABLE OF RESISTIVITY

Material	Resistivity in Ohms per Circular Mil-Foot	Temp. Coeff. of resistance per °C. at 20° C.
Aluminum	17	0.0049
Brass	45	0.003 to 0.007
Cadmium	46	0.0038
Chromium	16	0.00
Copper	10.4	0.0039
Iron	59	0.006
Silver	9.8	0.004
Zinc	36	0.0035
Nichrome	650	0.0002
Constantan	295	0.00001
Manganin	290	0.00001
Monel	255	0.0019

$$R = \frac{rl}{A}$$

where,

R equals resistance in ohms,
r equals resistivity in *ohms per mil-foot*,
l equals length of conductor in feet,
A equals cross-sectional area in circular mils.

For convenience, two larger units the *kilohm* (1000 ohms) and the *megohm* (1,000,000 ohms) are often used.

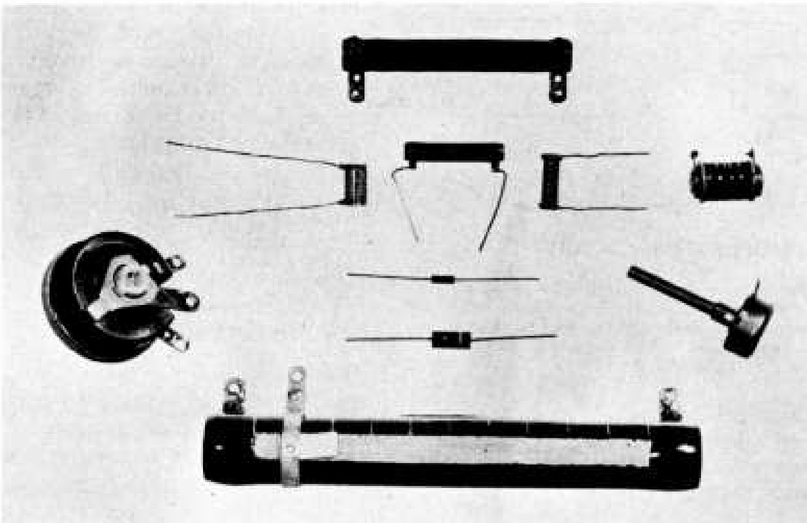


Figure 1

TYPICAL RESISTORS

Shown above are various types of resistors used in electronic circuits. The larger units are power resistors. On the left is a variable power resistor. Three precision-type resistors are shown in the center with two small composition resistors beneath them. At the right is a composition-type potentiometer, used for audio circuitry.

The resistance also depends on temperature, rising with an increase in temperature for most substances (including most metals), due to increased electron acceleration and hence a greater number of impacts between electrons and atoms. However, in the case of some substances such as carbon and glass the temperature coefficient is negative and the resistance decreases as the temperature increases.

Conductors and Insulators In the molecular structure of many materials such as glass, porcelain, and mica all electrons are tightly held within their orbits and there are comparatively few free electrons. This type of substance will conduct an electric current only with great difficulty and is known as an *insulator.* An insulator is said to have a high electrical *resistance.*

On the other hand, materials that have a large number of free electrons are known as *conductors.* Most metals (those elements which have only one or two electrons in their outer ring) are good conductors. Silver, copper, and aluminum, in that order, are the best of the common metals used as conductors and are said to have the greatest *conductivity*, or lowest resistance to the flow of an electric current.

Secondary Electrical Units These units are the *volt*, the *ampere*, and the *ohm*. They were mentioned in the preceding paragraphs, but were not completely defined in terms of fixed, known quantities.

The fundamental unit of *current*, or *rate of flow* of electricity is the ampere. A current of one ampere will deposit silver from a specified solution of silver nitrate at a rate of 1.118 milligrams per second.

The international standard for the ohm is the resistance offered by a uniform column of mercury at 0° C., 14.4521 grams in mass, of constant cross-sectional area and 106.300 centimeters in length. The expression *megohm* (1,000,000 ohms) is also sometimes used when speaking of very large values of resistance.

A volt is *the e.m.f. that will produce a current of one ampere through a resistance of one ohm*. The standard of electromotive force is the Weston cell which at 20° C. has a potential of 1.0183 volts across its terminals. This cell is used only for reference

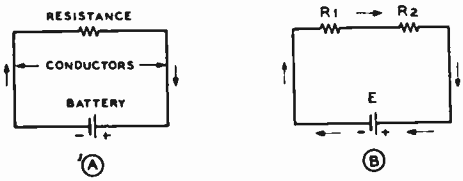


Figure 2

SIMPLE SERIES CIRCUITS

At (A) the battery is in series with a single resistor. At (B) the battery is in series with two resistors, the resistors themselves being in series. The arrows indicate the direction of electron flow.

purposes in a bridge circuit, since only an infinitesimal amount of current may be drawn from it without disturbing its characteristics.

Ohm's Law The relationship between the electromotive force (voltage), the flow of current (amperes), and the resistance which impedes the flow of current (ohms), is very clearly expressed in a simple but highly valuable law known as *Ohm's Law*. This law states that *the current in amperes is equal to the voltage in volts divided by the resistance in ohms*. Expressed as an equation:

$$I = \frac{E}{R}$$

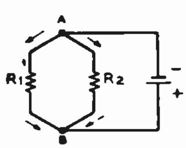


Figure 3
SIMPLE PARALLEL CIRCUIT

The two resistors R_1 and R_2 are said to be in parallel since the flow of current is offered two parallel paths. An electron leaving point A will pass either through R_1 or R_2, but not through both, to reach the positive terminal of the battery. If a large number of electrons are considered, the greater number will pass through whichever of the two resistors has the lower resistance.

If the voltage (E) and resistance (R) are known, the current (I) can be readily found. If the voltage and current are known, and the resistance is unknown, the resistance (R) is equal to $\frac{E}{I}$. When the voltage is the unknown quantity, it can be found by multiplying $I \times R$. These three equations are all secured from the original by simple transposition. The expressions are here repeated for quick reference:

$$I = \frac{E}{R} \quad R = \frac{E}{I} \quad E = IR$$

where,

I is the current in amperes,
R is the resistance in ohms,
E is the electromotive force in volts.

Taken in a broader sense, Ohm's Law expresses a *ratio* of voltage to current when the circuit resistance is known. This concept is important in transmission-line studies and antenna work.

Conductance Instead of speaking of the resistance of a circuit, the *conductance* may be referred to as a measure of the ease of current flow. Conductance is the reciprocal of resistance and is measured in *mhos* (ohms spelled backwards) and is designated by the letter *G*.

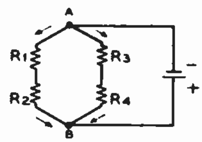
The relation between resistance and conductance is:

$$G = \frac{I}{R}, \; R = \frac{I}{G} \text{ or } I = EG$$

In electronics work, a small unit of conductance, which is equal to one-millionth of a mho, frequently is used. It is called a *micromho*.

Application of Ohm's Law All electrical circuits fall into one of three classes: *series circuits*, *parallel circuits*, and *series-parallel circuits*. A series circuit is one in which the current flows in a single continuous path and is of the same value at every point in the circuit (figure 2). In a parallel circuit there are two or more current paths between two points in the circuit, as shown in figure 3. Here the current divides at A, part going through R_1 and part through R_2, and combines at B to return to the battery. Figure 4 shows a series-parallel circuit. There are two paths between points A and B as in the parallel circuit, and in addition there are two resistances in series in each branch of the parallel combination. Two other examples of series-parallel ar-

Figure 4
SERIES-PARALLEL CIRCUIT



In this type of circuit the resistors are arranged in series groups, and these groups are then placed in parallel.

rangements appear in figure 5. The way in which the current splits to flow through the parallel branches is shown by the arrows.

In every circuit, each of the parts has some resistance: the batteries or generator, the connecting conductors, and the apparatus itself. Thus, if each part has some resistance, no matter how little, and a current is flowing through it, there will be a voltage drop across it. In other words, there will be a potential difference between the two ends of the circuit element in question. This drop in voltage is equal to the product of the current and the resistance hence it is called the *IR drop.*

Internal Resistance The source of voltage has an *internal* resistance, and when connected into a circuit so that current flows, there will be an *IR* drop in the source just as in every other part of the circuit. Thus, if the terminal voltage of the source could be measured in a way that would cause no current to flow, it would be found to be more than the voltage measured when a current flows by the amount of the *IR* drop in the source. The voltage measured with no current flowing is termed the *no load* voltage; that measured with current flowing is the *load* voltage. It is apparent that a voltage source having a low internal resistance is most desirable.

Resistances in Series The current flowing in a series circuit is equal to the voltage impressed divided by the *total* resistance across which the voltage is impressed. Since the same current flows through every part of the circuit, it is merely necessary to add all the individual resistances to obtain the total resistance. Expressed as a formula:

$$R_{Total} = R_1 + R_2 + R_3 + \ldots + R_N$$

Of course, if the resistances happened to be all the same value, the total resistance would be the resistance of one multiplied by the number of resistors in the circuit.

Resistances in Parallel Consider two resistors, one of 100 ohms and one of 10 ohms, connected in parallel as in figure 3, with a potential of 10 volts applied across each resistor, so the current through each can be easily calculated.

$$I = \frac{E}{R}$$

$E = 10$ volts
$R_1 = 100$ ohms
$$I_1 = \frac{10}{100} = 0.1 \text{ ampere}$$

$E = 10$ volts
$R_2 = 10$ ohms
$$I_2 = \frac{10}{10} = 1.0 \text{ ampere}$$

$$\text{Total current} = I_1 + I_2 = 1.1 \text{ ampere}$$

Until it divides at A, the entire current of 1.1 amperes is flowing through the conductor from the battery to A, and again from B through the conductor to the battery. Since this is more current than flows through the smaller resistor it is evident that the resistance of the parallel combination must be less than 10 ohms, the resistance of the smaller resistor. We can find this value by applying Ohm's Law.

$$R_T = \frac{E}{I}$$

$E = 10$ volts
$I = 1.1$ amperes
$$R_T = \frac{10}{1.1} = 9.09 \text{ ohms}$$

The resistance of the parallel combination is 9.09 ohms.

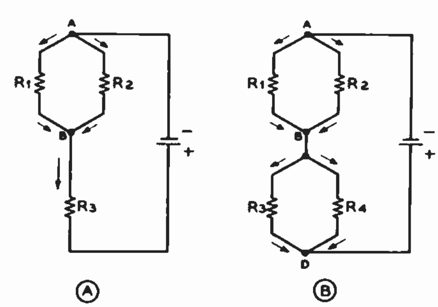


Figure 5

OTHER COMMON SERIES-PARALLEL CIRCUITS

Mathematically, we can derive a simple formula for finding the effective resistance of two resistors connected in parallel.

$$R_T = \frac{R_1 \times R_2}{R_1 + R_2}$$

where,

R_T is the unknown resistance,
R_1 is the resistance of the first resistor,
R_2 is the resistance of the second resistor.

If the effective value required is known, and it is desired to connect one unknown resistor in parallel with one of known value, the following transposition of the above formula will simplify the problem of obtaining the unknown value:

$$R_2 = \frac{R_1 \times R_T}{R_1 - R_T}$$

where,

R_T is the effective value required,
R_1 is the known resistor,
R_2 is the value of the unknown resistance necessary to give R_T when in parallel with R_1.

The resultant value of placing a number of unlike resistors in parallel is equal to the reciprocal of the sum of the reciprocals of the various resistors. This can be expressed as:

$$R_T = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \ldots \frac{1}{R_n}}$$

The effective value of placing any number of unlike resistors in parallel can be determined from the above formula. However, it is commonly used only when there are three or more resistors under consideration, since the simplified formula given before is more convenient when only two resistors are being used.

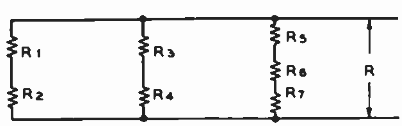


Figure 6

ANOTHER TYPE OF SERIES-PARALLEL CIRCUIT

From the above, it also follows that when two or more resistors of the same value are placed in parallel, the effective resistance of the paralleled resistors is equal to the value of one of the resistors divided by the number of resistors in parallel.

The effective value of resistance of two or more resistors connected in parallel is *always* less than the value of the lowest resistance in the combination. It is well to bear this simple rule in mind, as it will assist greatly in approximating the value of paralleled resistors.

Resistors in Series-Parallel To find the total resistance of several resistors connected in series-parallel, it is usually easiest to apply either the formula for series resistors or the parallel resistor formula first, in order to reduce the original arrangement to a simpler one. For instance, in figure 4 the series resistors should be added in each branch, then there will be but two resistors in parallel to be calculated. Similarly in figure 6, although here there will be three parallel resistors after adding the series resistors in each branch. In figure 6B the paralleled resistors should be reduced to the equivalent series value, and then the series resistance value can be added.

Resistances in series-parallel can be solved by combining the series and parallel formulas into one similar to the following (refer to figure 6):

$$R_T = \frac{1}{\frac{1}{R_1 + R_2} + \frac{1}{R_3 + R_4} + \frac{1}{R_5 + R_6 + R_7}}.$$

Voltage Dividers A *voltage divider* is exactly what its name implies: a resistor or a series of resistors connected across a source of voltage from which various lesser values of voltage may be obtained by connection to various points along the resistor.

A voltage divider serves a most useful purpose in a radio receiver, transmitter or amplifier, because it offers a simple means of obtaining plate, screen, and bias voltages of different values from a common power supply source. It may also be used to obtain very low voltages of the order of .01 to .001 volt with a high degree of accuracy, even though a means of measuring such voltages is lacking. The procedure for making these measurements can best be given in the following example.

Assume that an accurately calibrated voltmeter reading from 0 to 150 volts is available, and that the source of voltage is exactly 100 volts. This 100 volts is then impressed through a resistance of exactly 1000 ohms. It will then be found that the voltage along various points on the resistor, with respect to the grounded end, is exactly proportional to the resistance at that point. From Ohm's Law, the current would be 0.1 ampere; this current remains unchanged since the original value of resistance (1000 ohms) and the voltage source (100 volts) are unchanged. Thus, at a 500-ohm point on the resistor (half its entire resistance), the voltage will likewise be halved or reduced to 50 volts.

The equation ($E = I \times R$) gives the proof: $E = 500 \times 0.1 = 50$. At the point of 250 ohms on the resistor, the voltage will be one-fourth the total value, or 25 volts ($E = 250 \times 0.1 = 25$). Continuing with this process, a point can be found where the resistance measures exactly 1 ohm and where the voltage equals 0.1 volt. It is, therefore, obvious that if the original source of voltage and the resistance can be measured, it is a simple matter to predetermine the voltage at any point along the resistor, provided that the current remains constant, and provided that no current is taken from the tap-on point unless this current is taken into consideration.

Voltage-Divider Calculations Proper design of a voltage divider for any type of radio equipment is a relatively simple matter. The first consideration is the amount of "bleeder current" to be drawn. In addition, it is also necessary that the desired voltage and the exact current at each tap on the voltage divider be known.

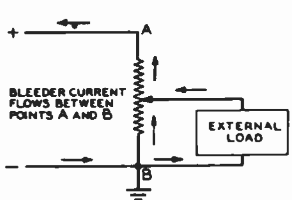


Figure 7

SIMPLE VOLTAGE-DIVIDER CIRCUIT

The arrows indicate the manner in which the current flow divides between the voltage divider itself and the external load circuit.

Figure 7 illustrates the flow of current in a simple voltage-divider and load circuit. The light arrows indicate the flow of bleeder current, while the heavy arrows indicate the flow of the load current. The design of a combined bleeder resistor and voltage divider, such as is commonly used in radio equipment, is illustrated in the following example:

A power supply delivers 300 volts and is conservatively rated to supply all needed current for the receiver and still allow a bleeder current of 10 milliamperes. The following voltages are wanted: 75 volts at 2 milliamperes for the detector tube, 100 volts at 5 milliamperes for the screens of the tubes, and 250 volts at 20 milliamperes for the plates of the tubes. The required voltage drop across R_1 is 75 volts, across R_2 25 volts, across R_3 150 volts, and across R_4 it is 50 volts. These values are shown in the diagram of figure 8. The respective current values are also indicated. Apply Ohm's Law:

$$R_1 = \frac{E}{I} = \frac{75}{.01} = 7500 \text{ ohms}$$

$$R_2 = \frac{E}{I} = \frac{25}{.012} = 2083 \text{ ohms}$$

$$R_3 = \frac{E}{I} = \frac{150}{.017} = 8823 \text{ ohms}$$

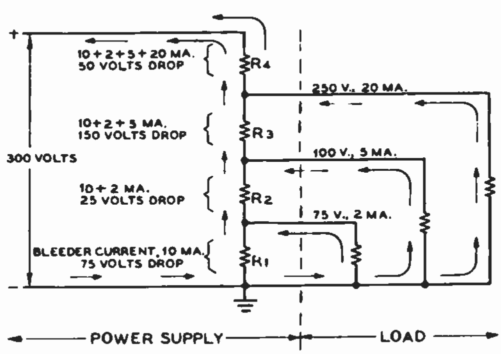


Figure 8

MORE COMPLEX VOLTAGE DIVIDER

The method for computing the values of the resistors is discussed in the accompanying text.

$$R_4 = \frac{E}{I} = \frac{50}{.037} = 1351 \text{ ohms}$$

$$R_{Total} = 7500 + 2083 + 8823 + 1351 = 19{,}757 \text{ ohms}$$

A 20,000-ohm resistor with three sliding taps will be the approximately correct size, and would ordinarily be used because of the difficulty in securing four separate resistors of the exact odd values indicated, and because no adjustment would be possible to compensate for any slight error in estimating the probable currents through the various taps.

When the sliders on the resistor once are set to the proper point, as in the above example, the voltages will remain constant at the values shown as long as the current remains a constant value.

Disadvantages of Voltage Dividers One of the serious disadvantages of the voltage divider becomes evident when the current drawn from one of the taps changes. It is obvious that the voltage drops are interdependent and, in turn, the individual drops are in proportion to the current which flows through the respective sections of the divider resistor. The only remedy lies in providing a heavy steady bleeder current in order to make the individual currents so small a part of the total current that any change in current will result in only a slight change in voltage. This can seldom be realized in practice because of the excessive values of bleeder current which would be required.

Kirchhoff's Laws Ohm's Law is all that is necessary to calculate the values in simple circuits, such as the preceding examples; but in more complex problems, involving several loops, or more than one voltage in the same closed circuit, the use of *Kirchhoff's laws* will greatly simplify the calculations. These laws are merely rules for applying Ohm's Law.

Kirchhoff's first law is concerned with net current to a point in a circuit and states that:

At any point in a circuit the current flowing toward the point is equal to the current flowing away from the point.

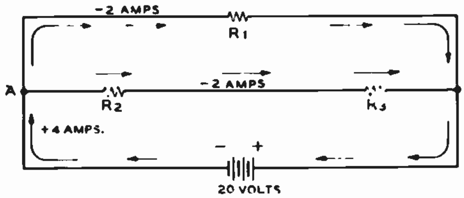


Figure 9

ILLUSTRATING KIRCHHOFF'S FIRST LAW

The current flowing toward point "A" is equal to the current flowing away from point "A."

Stated in another way: if currents flowing to the point are considered positive, and those flowing from the point are considered negative, the sum of all currents flowing toward and away from the point — taking signs into account — is equal to zero. Such a sum is known as an *algebraic sum;* such that the law can be stated thus: *The algebraic sum of all currents entering and leaving a point is zero.*

Figure 9 illustrates this first law. If the effective resistance of the network of resistors is 5 ohms, it can be seen that 4 amperes flow toward point A, and 2 amperes flow away through the two 5-ohm resistors in series. The remaining 2 amperes flow away through the 10-ohm resistor. Thus, there are 4 amperes flowing to point A and 4 amperes flowing away from the point. If R_T is the effective resistance of the network (5 ohms), $R_1 = 10$ ohms, $R_2 = 5$ ohms, $R_3 = 5$ ohms, and $E = 20$ volts, we can set up the following equation:

$$\frac{E}{R_T} - \frac{E}{R_1} - \frac{E}{R_2 + R_3} = 0$$

$$\frac{20}{5} - \frac{20}{10} - \frac{20}{5+5} = 0$$

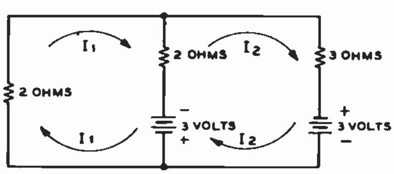
$$4 - 2 - 2 = 0$$

Kirchhoff's second law is concerned with net voltage drop around a closed loop in a circuit and states that:

In any closed path or loop in a circuit the sum of the IR drops must equal the sum of the applied e.m.f.'s.

The second law also may be conveniently stated in terms of an algebraic sum as: *The* *algebraic sum of all voltage drops around a closed path or loop in a circuit is zero.* The applied e.m.f.'s (voltages) are considered positive, while IR drops taken in the direction of current flow (including the internal drop of the sources of voltage) are considered negative.

Figure 10 shows an example of the application of Kirchhoff's laws to a comparatively simple circuit consisting of three resistors and two batteries. First assume an arbitrary direction of current flow in each closed loop of the circuit, drawing an arrow to indicate the assumed direction of current flow. Then equate the sum of all IR drops plus battery drops around each loop to zero. You will need one equation for each unknown to be determined. Then solve the equations for the unknown currents in the general manner indicated in figure 10. If the answer comes out positive the direction of current flow you originally assumed was correct. If the answer comes out negative, the current flow is in the opposite direction to the arrow



1. SET VOLTAGE DROPS AROUND EACH LOOP EQUAL TO ZERO.

$I_1 2_{(\text{OHMS})} + 2(I_1 - I_2) + 3 = 0$ (FIRST LOOP)

$-6 + 2(I_2 - I_1) + 3 I_2 = 0$ (SECOND LOOP)

2. SIMPLIFY

$2I_1 + 2I_1 - 2I_2 + 3 = 0$ $2I_2 - 2I_1 + 3I_2 - 6 = 0$

$\frac{4I_1 + 3}{2} = I_2$ $5I_2 - 2I_1 - 6 = 0$

$\frac{2I_1 + 6}{5} = I_2$

3. EQUATE

$\frac{4I_1 + 3}{2} = \frac{2I_1 + 6}{5}$

4. SIMPLIFY

$20I_1 + 15 = 4I_1 + 12$

$I_1 = -\frac{3}{16}$ AMPERE

5. RE-SUBSTITUTE

$I_2 = \frac{-\frac{12}{16} + 3}{2} = \frac{2\frac{1}{4}}{2} = 1\frac{1}{8}$ AMPERE

Figure 10

ILLUSTRATING KIRCHHOFF'S SECOND LAW

The voltage drop around any closed loop in a network is equal to zero.

which was drawn originally. This is illustrated in the example of figure 10, where the direction of flow of I_1 *is* opposite to the direction assumed in the sketch.

Power in Resistive Circuits In order to cause electrons to flow through a conductor, constituting a current flow, it is necessary to apply an electromotive force (voltage) across the circuit. Less power is expended in creating a small current flow through a given resistance than in creating a large one; so it is necessary to have a unit of power as a reference.

The unit of electrical power is the *watt*, which is the rate of energy consumption when an e.m.f. of 1 volt forces a current of 1 ampere through a circuit. The power in a resistive circuit is equal to the product of the voltage applied across, and the current flowing in, a given circuit. Hence: P (watts) $= E$ (volts) $\times$ I (amperes).

Since it is often convenient to express power in terms of the resistance of the circuit and the current flowing through it, a substitution of IR for E ($E = IR$) in the above formula gives: $P = IR \times I$ or $P = I^2R$. In terms of voltage and resistance, $P = E^2/R$. Here, $I = E/R$ and when this is substituted for I the original formula becomes $P = E \times E/R$, or $P = E^2/R$. To repeat these three expressions:

$$P = EI,\ P = I^2R, \text{ and } P = E^2/R$$

where,

P is the power in watts,
E is the electromotive force in volts, and
I is the current in amperes.

To apply the above equations to a typical problem: The voltage drop across a cathode resistor in a power amplifier stage is 50 volts; the plate current flowing through the resistor is 150 milliamperes. The number of watts the resistor will be required to dissipate is found from the formula: $P = EI$, or $50 \times .150 = 7.5$ watts (.150 ampere is equal to 150 milliamperes). From the foregoing it is seen that a 7.5-watt resistor will safely carry the required current, yet a 10- or 20-watt resistor would ordinarily be used to provide a safety factor.

In another problem, the conditions being similar to those above, but with the resistance ($R = 333\frac{1}{3}$ ohms), and current being the *known* factors, the solution is obtained as follows: $P = I^2R = .0225 \times 333.33 = 7.5$. If only the voltage and resistance are known, $P = E^2/R = 2500/333.33 = 7.5$ watts. It is seen that all three equations give the same results; the selection of the particular equation depends only on the known factors.

Power, Energy and Work It is important to remember that power (expressed in watts, horsepower, etc.), represents the *rate* of energy consumption or the *rate* of doing work. But when we pay our electric bill to the power company we have purchased a specific *amount* of *energy* or *work* expressed in the common units of *kilowatt-hours*. Thus *rate* of energy consumption (watts or kilowatts) multiplied by *time* (seconds, minutes, or hours) gives us total energy or work. Other units of energy are the watt-second, BTU, calorie, erg, and joule.

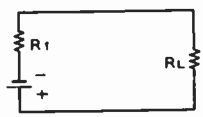


Figure 11

MATCHING OF RESISTANCES

To deliver the greatest amount of power to the load, the load resistance R_L should be equal to the internal resistance of the battery R_I.

Heating Effect Heat is generated when a source of voltage causes a current to flow through a resistor (or, for that matter, through any conductor). As explained earlier, this is due to the fact that heat is given off when free electrons collide with the atoms of the material. More heat is generated in high-resistance materials than in those of low resistance, since the free electrons must strike the atoms harder to knock off other electrons. As the heating effect is a function of the current flowing and the resistance of the circuit, the power expended in heat is given by the second formula: $P = I^2R$.

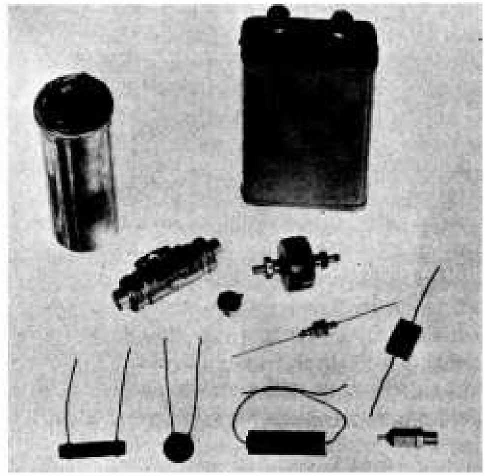


Figure 12

TYPICAL FIXED CAPACITORS

The two large units are high value filter capacitors. Shown beneath these are various types of bypass capacitors for r-f and audio application.

2-3 Electrostatics and Capacitors

Electrical energy can be stored in an *electrostatic field*. A device capable of storing energy in such a field is called a *capacitor* (in earlier usage the term *condenser* was frequently used but the IEEE standards call for the use of capacitor instead of condenser) and is said to have a certain *capacitance*. The *energy* stored in an electrostatic field is expressed in *joules* (watt-seconds) and is equal to $CE^2/2$, where C is the capacitance in *farads* (a unit of capacitance to be discussed) and E is the potential in volts. The *charge* is equal to CE, the charge being expressed in coulombs.

Capacitance and Capacitors Two metallic plates separated from each other by a thin layer of insulating material (called a *dielectric*, in this case) becomes a *capacitor*. When a source of d-c potential is momentarily applied across these plates, they may be said to become charged. If the same two plates are then joined together momentarily by means of a switch, the capacitor will *discharge*.

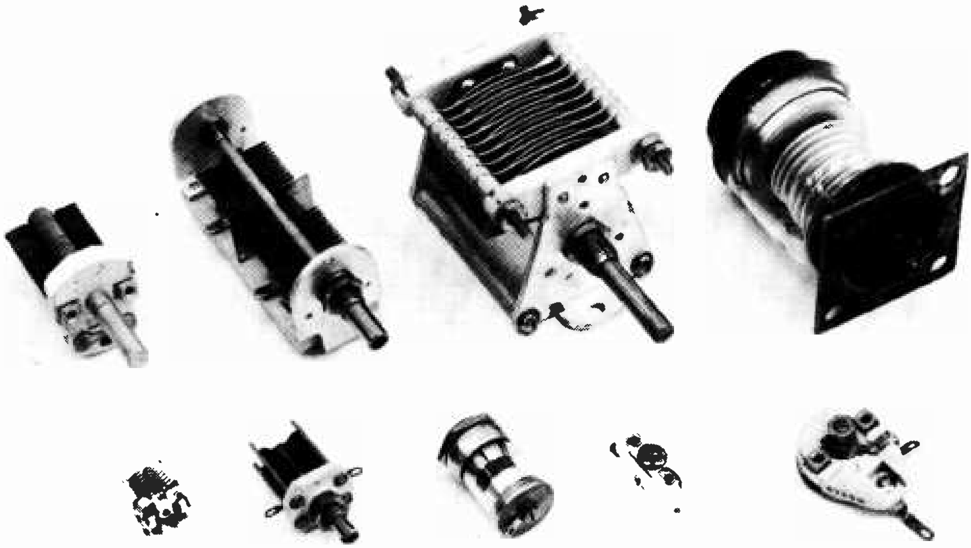


Figure 13

At top left are three variable air capacitors intended for hf/vhf use. At the right is a small variable vacuum capacitor intended for high-voltage service. Across the bottom are (left to right): two subminiature variable split-stator capacitors, a precision "plunger" capacitor, a compression mica capacitor, and a variable ceramic trimming capacitor.

When the potential was first applied, electrons immediately flowed from one plate to the other through the battery or such source of d-c potential as was applied to the capacitor plates. However, the circuit from plate to plate in the capacitor was *incomplete* (the two plates being separated by an insulator) and thus the electron flow ceased, meanwhile establishing a shortage of electrons on one plate and a surplus of electrons on the other.

Remember that when a deficiency of electrons exists at one end of a conductor, there is always a tendency for the electrons to move about in such a manner as to re-establish a state of balance. In the case of the capacitor herein discussed, the surplus quantity of electrons on one of the capacitor plates cannot move to the other plate because the circuit has been broken; that is, the battery or d-c potential was removed. This leaves the capacitor in a *charged* condition; the capacitor plate with the electron *deficiency* is *positively* charged, the other plate being *negative*.

In this condition, a considerable stress exists in the insulating material (dielectric) which separates the two capacitor plates, due to the mutual attraction of two unlike potentials on the plates. This stress is known known as *electrostatic* energy, as contrasted with *electromagnetic* energy in the case of an inductor. This charge can also be called *potential energy* because it is capable of performing work when the charge is released through an external circuit. The charge is proportional to the voltage but the energy is proportional to the voltage squared, as shown in the following analogy.

The charge represents a definite amount of electricity, or a given number of electrons. The potential energy possessed by these electrons depends not only on their number, but also on their potential or voltage.

Compare the electrons to water, and two capacitors to standpipes, a 1-μfd capacitor to a standpipe having a cross section of 1 square inch and a 2-μfd capacitor to a standpipe having a cross section of 2 square inches. The charge will represent a given volume of water, as the "charge" simply indicates a certain number of electrons. Suppose the quantity of water is equal to 5 gallons.

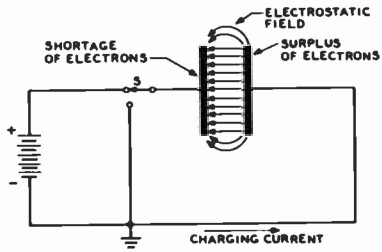


Figure 14

SIMPLE CAPACITOR

Illustrating the imaginary lines of force representing the paths along which the repelling force of the electrons would act on a free electron located between the two capacitor plates.

Now the potential energy, or capacity for doing work, of the 5 gallons of water will be twice as great when confined to the 1 sq. in. standpipe as when confined to the 2 sq. in. standpipe. Yet the volume of water or "charge" is the same in either case.

Likewise a 1-μfd capacitor charged to 1000 volts possesses twice as much potential energy as does a 2-μfd capacitor charged to 500 volts, though the *charge* (expressed in *coulombs:* $Q = CE$) is the same in either case.

TABLE 4. TABLE OF DIELECTRIC MATERIALS

Material	Dielectric Constant 10 MHz	Power Factor 10 MHz	Softening Point Fahrenheit
Aniline-Formaldehyde Resin	3.4	0.004	260°
Barium Titianate	1200	1.0	—
Castor Oil	4.67		
Cellulose Acetate	3.7	0.04	180°
Glass, Window	6-8	Poor	2000°
Glass, Pyrex	4.5	0.02	
Kel-F Fluorothene	2.5	0.6	—
Methyl-Methacrylate-Lucite	2.6	0.007	160°
Mica	5.4	0.0003	
Mycalex Mykroy	7.0	0.002	650°
Phenol-Formaldehyde, Low-Loss Yellow	5.0	0.015	270°
Phenol-Formaldehyde Black Bakelite	5.5	0.03	350°
Porcelain	7.0	0.005	2800°
Polyethylene	2.25	0.0003	220°
Polystyrene	2.55	0.0002	175°
Quartz, Fused	4.2	0.0002	2600°
Rubber Hard-Ebonite	2.8	0.007	150°
Steatite	6.1	0.003	2700°
Sulfur	3.8	0.003	236°
Teflon	2.1	.0006	—
Titanium Dioxide	100-175	0.0006	2700°
Transformer Oil	2.2	0.003	
Urea-Formaldehyde	5.0	0.05	260°
Vinyl Resins	4.0	0.02	200°
Wood, Maple	4.4	Poor	

The Unit of Capacitance: The Farad If the external circuit of the two capacitor plates is completed by joining the terminals together with a piece of wire, the electrons will rush immediately from one plate to the other through the external circuit and establish a state of equilibrium. This latter phenomenon explains the *discharge* of a capacitor. The amount of stored energy in a charged capacitor is dependent on the charging potential, as well as a factor which takes into account the *size* of the plates, *dielectric thickness, nature* of the dielectric, and the *number* of plates. This factor, which is determined by the foregoing, is called the *capacitance* of a capacitor and is expressed in *farads.*

The farad is such a large unit of capacitance that it is rarely used in radio calculations, and the following more practical units have, therefore, been chosen.

1 *microfarad* = 1/1,000,000 *farad, or* .000001 *farad, or* 10^{-6} *farad.*

1 *micromicrofarad or picofarad* = 1/1,000,000 *microfarad, or* .000001 *microfarad, or* 10^{-6} *microfarad.*

1 *micromicrofarad or picofarad* = *one-millionth of one-millionth of a farad, or* 10^{-12} *farad.*

If the capacitance is to be expressed in *microfarads* in the equation given for *energy storage,* the factor C would then have to be divided by 1,000,000, thus:

$$\text{Stored energy in joules} = \frac{C \times E^2}{2 \times 1{,}000{,}000}$$

This storage of energy in a capacitor is one of its very important properties, particularly in those capacitors which are used in power-supply filter circuits.

Dielectric Materials Although any substance which has the characteristics of a good insulator may be used as a dielectric material, commercially manufactured capacitors make use of dielectric materials which have been selected because their characteristics are particularly suited to the job at hand. Air is a very good dielectric material, but an air-spaced capacitor does not have a high capacitance since the dielectric constant of air is only slightly greater than one. A group of other commonly used dielectric materials is listed in Table 4.

Certain materials, such as *bakelite, lucite,* and other plastics dissipate considerable energy when used as capacitor dielectrics. This energy loss is expressed in terms of the *power factor* of the capacitor, which represents the portion of the input volt-amperes lost in the dielectric material. Other materials including air, polystyrene and quartz have a very low power factor.

The new *ceramic* dielectrics such as *steatite* (talc) and titanium dioxide products are especially suited for high-frequency and high-temperature operation. Ceramics based on titanium dioxide have an unusually high dielectric constant combined with a low power factor. The temperature coefficient with respect to capacitance of units made with this material depends on the mixture of oxides, and coefficients ranging from zero to over −700 parts per million per degree Centigrade may be obtained in commercial production.

Mycalex is a composition of minute mica particles and lead-borate glass, mixed and fired at a relatively low temperature. It is hard and brittle, but can be drilled or machined when water is used as the cutting lubricant.

Mica dielectric capacitors have a very low power factor and extremely high voltage breakdown per unit of thickness. A mica and copperfoil "sandwich" is formed under pressure to obtain the desired capacity value. The effect of temperature on the pressures in the "sandwich" causes the capacitance of the usual mica capacitor to have large, noncyclic variations. If the copper electrodes are plated directly on the mica sheets, the temperature coefficient can be stabilized at about 20 parts per million per degree Centigrade. A process of this type is used in the manufacture of "silver mica" capacitors.

Paper dielectric capacitors consist of strips of aluminum foil insulated from each other by a thin layer of paper, the whole assembly being wrapped in a circular bundle. The cost of such a capacitor is low, the capacitance is high in proportion to the size and weight, and the power factor is good. The life of such a capacitor is dependent on the moisture penetration of the paper dielectric, and on the level of the applied d-c voltage.

Air-dielectric capacitors are used in transmitting and receiving circuits, principally where a variable capacitor of high resetability is required. The dielectric strength is high, though somewhat less at radio frequencies than at 60 Hz. In addition, corona discharge at high frequencies will

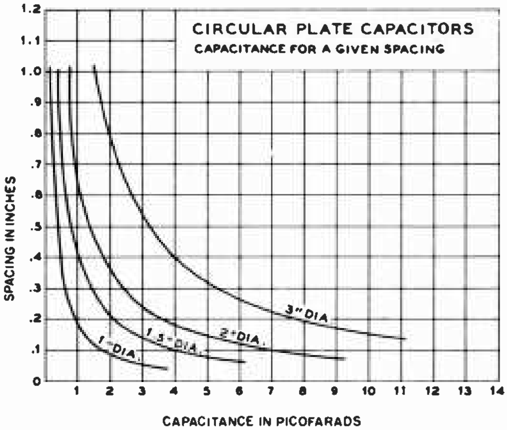


CHART 1

Through the use of this chart it is possible to determine the required plate diameter (with the necessary spacing established by peak voltage considerations) for a circular-plate neutralizing capacitor. The capacitance given is for a dielectric of air and the spacing given is between adjacent faces of the two plates.

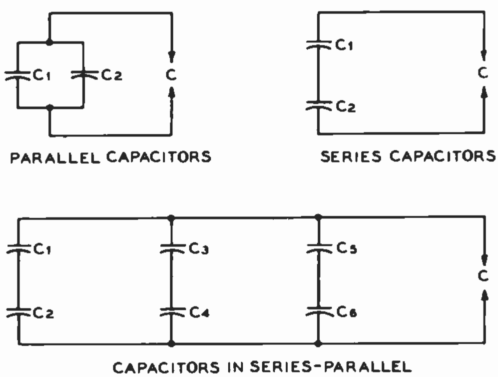


Figure 15

CAPACITORS IN SERIES, PARALLEL, AND SERIES-PARALLEL

cause ionization of the air dielectric causing an increase in power loss. Dielectric strength may be increased by increasing the air pressure, as is done in hermetically sealed radar units. In some units, dry nitrogen gas may be used in place of air to provide a higher dielectric strength than that of air.

Likewise, the dielectric strength of an "air" capacitor may be increased by placing the unit in a vacuum chamber to prevent ionization of the dielectric.

The temperature coefficient of a variable air-dielectric capacitor varies widely and is often noncyclic. Such things as differential expansion of various parts of the capacitor, changes in internal stresses, and different temperature coefficients of various parts contribute to these variances.

Dielectric Constant The capacitance of a capacitor is determined by the thickness and nature of the dielectric material between plates. Certain materials offer a greater capacitance than others, depending on their physical makeup and chemical constitution. This property is expressed by a constant K, called the *dielectric constant*. ($K = 1$ for air.)

Dielectric Breakdown If the charge becomes too great for a given thickness of a certain dielectric, the capacitor will break down, i.e., the dielectric will puncture. It is for this reason that capacitors are rated in the manner of the amount of voltage they will safely withstand as well as the capacitance in microfarads. This rating is commonly expressed as the *d-c working voltage* (*DCWV*).

Calculation of Capacitance The capacitance of two parallel plates may be determined with good accuracy by the following formula:

$$C = 0.2248 \times K \times \frac{A}{t}$$

where,

C equals capacitance in picofarads,
K equals dielectric constant of spacing material,
A equals area of dielectric in square inches,
t equals thickness of dielectric in inches.

This formula indicates that the capacitance is *directly* proportional to the area of the plates and *inversely* proportional to the thickness of the dielectric (spacing between the plates). This simply means that when the area of the plate is doubled, the spacing between plates remaining constant, the capacitance will be doubled. Also, if the area of the plates remains constant, and the plate spacing is doubled the capacitance will be reduced to half.

The above equation also shows that capacitance is directly proportional to the dielectric constant of the spacing material. An air-spaced capacitor that has a capacitance of 100 pf in air would have a capacitance of 467 pf when immersed in castor oil, because the dielectric constant of castor oil is 4.67 times as great as the dielectric constant of air.

Where the area of the plate is definitely set, when it is desired to know the spacing needed to secure a required capacitance,

$$t = \frac{A \times 0.2248 \times K}{C}$$

where all units are expressed just as in the preceding formula. This formula is not confined to capacitors having only square or rectangular plates, but also applies when the plates are circular in shape. The only change will be the calculation of the *area* of such circular plates; this area can be computed by squaring the *radius* of the plate, then multiplying by 3.1416, or "pi." Expressed as an equation:

$$A = 3.1416 \times r^2$$

where,
r equals radius in inches.

The capacitance of a multiplate capacitor can be calculated by taking the capacitance of one section and multiplying this by the number of dielectric spaces. In such cases, however, the formula gives no consideration to the effects of edge capacitance; so the capacitance as calculated will not be entirely accurate. These additional capacitances will be but a small part of the effective total capacitance, particularly when the plates are reasonably large and thin, and the final result will, therefore, be within practical limits of accuracy.

Capacitors in Parallel and in Series Equations for calculating capacitances of capacitors in *parallel* connections are the same as those for resistors in *series.*

$$C_T = C_1 + C_2 + \ldots + C_n$$

Capacitors in *series* connection are calculated in the same manner as are resistors in *parallel* connection.

The formulas are repeated: (1) For two or more capacitors of *unequal* capacitance in series:

$$C_T = \cfrac{1}{\cfrac{1}{C_1} + \cfrac{1}{C_2} + \cfrac{1}{C_3}}$$

or,
$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

(2) *Two* capacitors of *unequal* capacitance in series:

$$C_T = \frac{C_1 \times C_2}{C_1 + C_2}$$

(3) Three capacitors of *equal* capacitance in series:

$$C_T = \frac{C_1}{3}$$

where,

C_1 is the common capacitance.

(4) Three or more capacitors of *equal* capacitance in series.

$$C_T = \frac{\text{Value of common capacitance}}{\text{Number of capacitors in series}}$$

(5) Six capacitors in series-parallel:

$$C_T = \cfrac{1}{\cfrac{1}{C_1} + \cfrac{1}{C_2}} + \cfrac{1}{\cfrac{1}{C_3} + \cfrac{1}{C_4}} + \cfrac{1}{\cfrac{1}{C_5} + \cfrac{1}{C_6}}$$

Capacitors in A-C and D-C Circuits When a capacitor is connected into a direct-current circuit, it will block the d.c., or stop the flow of current. Beyond the initial movement of electrons during the period when the capacitor is being charged, there will be no flow of current because the circuit is effectively broken by the dielectric of the capacitor.

Strictly speaking, a very small current may actually flow because the dielectric of the capacitor may not be a perfect insulator. This minute current flow is the leakage current previously referred to and is dependent on the internal d-c resistance of the capacitor. This leakage current is usually quite noticeable in most types of electrolytic capacitors.

When an alternating current is applied to a capacitor, the capacitor will charge and discharge a certain number of times per second in accordance with the frequency of the alternating voltage. The electron flow in the charge and discharge of a capacitor when an a-c potential is applied constitutes an alternating current, in effect. It is for this reason that a capacitor will pass an alternating current yet offer practically infinite opposition to a direct current. These two properties are repeatedly in evidence in a radio circuit.

Voltage Rating of Capacitors in Series Any good paper-dielectric filter capacitor has such a high internal resistance (indicating a good dielectric) that the exact resistance will vary considerably from capacitor to capacitor even though they are made by the same manufacturer and are of the same rating. Thus, when 1000 volts d. c. are connected across two 1-μfd 500-volt capacitors in series, the chances are that the voltage will divide unevenly; one capacitor will receive more than 500 volts and the other less than 500 volts.

Voltage Equalizing Resistors By connecting a half-megohm 1-watt carbon resistor across each capacitor, the voltage will be equalized because the resistors act as a voltage divider,

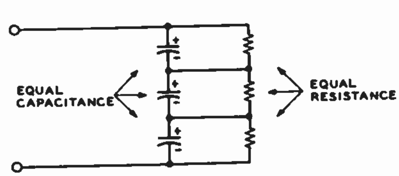


Figure 16

SHOWING THE USE OF VOLTAGE EQUALIZING RESISTORS ACROSS CAPACITORS CONNECTED IN SERIES

and the internal resistances of the capacitors are so much higher (many megohms) that they have but little effect in disturbing the voltage divider balance.

Carbon resistors of the inexpensive type are not particularly accurate (not being designed for precision service); therefore it is advisable to check several on an accurate ohmmeter to find two that are as close as possible in resistance. The exact resistance is unimportant, just so it is the same for the two resistors used.

Capacitors in Series on A.C. When two capacitors are connected in series, *alternating* voltage pays no heed to the relatively high internal resistance of each capacitor, but divides across the capacitors in inverse proportion to the *capacitance*. Because, in addition to the d-c voltage across a capacitor in a filter or audio amplifier circuit there is usually an a-c or a-f voltage component, it is inadvisable to series-connect capacitors of unequal capacitance even if dividers are provided to keep the d-c voltages within the ratings of the individual capacitors.

For instance, if a 500-volt 1-μfd capacitor is used in series with a 4-μfd 500-volt capacitor across a 250-volt a-c supply, the 1-μfd capacitor will have 200 a-c volts across it and the 4-μfd capacitor only 50 volts. An equalizing divider, to do any good in this case, would have to be of very low resistance because of the comparatively low impedance of the capacitors to alternating current. Such a divider would draw excessive current and be impracticable.

The safest rule to follow is to use only capacitors of the same capacitance and voltage rating and to install matched high-resistance proportioning resistors across the various capacitors to equalize the d-c voltage drop across each capacitor. This holds regardless of how many capacitors are series-connected.

Electrolytic Capacitors *Electrolytic capacitors* use a very thin film of oxide as the dielectric, and are polarized; that is, they have a positive and a negative terminal which must be properly connected in a circuit; otherwise, the oxide will break down and the capacitor will overheat. The unit then will no longer be of service. When electrolytic capacitors are connected in series, the positive terminal is always connected to the positive lead of the power supply; the negative terminal of the capacitor connects to the *positive* terminal of the *next* capacitor in the series combination. The method of connection for electrolytic capacitors in series is shown in figure 16. Electrolytic capacitors have very low cost per microfarad of capacitance, but also have a large power factor and high leakage; both dependent on applied voltage, temperature, and the age of the capacitor. The modern electrolytic capacitor uses a dry paste electrolyte embedded in a gauze or paper dielectric. Aluminum foil and the dielectric are wrapped in a circular bundle and are mounted in a cardboard or metal box. Etched electrodes may be employed to increase the effective anode area, and the total capacitance of the unit.

The capacitance of an electrolytic capacitor is affected by the applied voltage, the usage of the capacitor, the temperature and the humidity of the environment. The capacitance usually drops with the aging of the unit. The leakage current and power factor increase with age. At high frequencies the power factor becomes so poor that the electrolytic capacitor acts as a series resistance rather than as a capacitance.

2-4 Magnetism and Electromagnetism

The common bar or horseshoe magnet is familiar to most people. The magnetic field which surrounds it causes the magnet to attract other magnetic materials, such as iron nails or tacks. Exactly the same kind of magnetic field is set up around any conductor carrying a current, but the field exists only while the current is flowing.

Magnetic Fields Before a potential, or voltage, is applied to a conductor there is no external field, because there is no general movement of the electrons in one direction. However, the electrons do progressively move along the conductor when an e.m.f. is applied, the direction of motion depending on the polarity of the e.m.f. Since each electron has an electric field about it, the flow of electrons causes these fields to build up into a resultant external field which acts in a plane at right angles to the direction in which the current is flowing. This field is known as the *magnetic field*.

The magnetic field around a current-carrying conductor is illustrated in figure 17. The direction of this magnetic field depends entirely on the direction of electron drift or current flow in the conductor. When the flow is toward the observer, the field about the conductor is clockwise; when the flow is away from the observer, the field is counterclockwise. This is easily remembered if the left hand is clenched, with the thumb outstretched and pointing in the direction of electron flow. The fingers then indicate the direction of the magnetic field around the conductor.

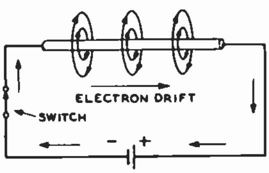


Figure 17

LEFT-HAND RULE

Showing the direction of the magnetic lines of force produced around a conductor carrying an electric current.

Each electron adds its field to the total external magnetic field, so that the greater the number of electrons moving along the conductor, the stronger will be the resulting field.

One of the fundamental laws of magnetism is that *like poles repel one another and unlike poles attract one another*. This is true of current-carrying conductors as well as of permanent magnets. Thus, if two conductors are placed side by side and the current in each is flowing in the same direction, the magnetic fields will also be in the same direction and will combine to form a larger and stronger field. If the current flow in adjacent conductors is in opposite directions, the magnetic fields oppose each other and tend to cancel.

The magnetic field around a conductor may be considerably increased in strength by winding the wire into a coil. The field around each wire then combines with those of the adjacent turns to form a total field through the coil which is concentrated along the axis of the coil and behaves externally in a way similar to the field of a bar magnet.

If the left hand is held so that the thumb is outstretched and parallel to the axis of a coil, with the fingers curled to indicate the direction of electron flow around the turns of the coil, the thumb then points in the direction of the north pole of the magnetic field.

The Magnetic Circuit In the magnetic circuit, the units which correspond to current, voltage, and resistance in the electrical circuit are *flux*, *magnetomotive force*, and *reluctance*.

Flux; Flux Density As a current is made up of a drift of electrons, so is a magnetic field made up of lines of force, and the total number of lines of force in a given magnetic circuit is termed the *flux*. The flux depends on the material, cross section, and length of the magnetic circuit, and it varies directly as the current flowing in the circuit. The unit of flux is the *maxwell*, and the symbol is the Greek letter ϕ (phi).

Flux density is the number of lines of force per unit area. It is expressed in *gauss* if the unit of area is the square centimeter (1 gauss = 1 line of force per square centimeter), or in *lines per square inch*. The symbol for flux density is B if it is expressed in gauss, or B if expressed in lines per sq. in.

Magnetomotive Force The force which produces a flux in a magnetic circuit is called *magnetomotive force*. It is abbreviated m.m.f. and is designated by the letter F. The unit of magnetomotive force is the *gilbert*, which is equivalent to $1.26 \times NI$, where N is the number of turns and I is the current flowing in the circuit in amperes.

The m.m.f. necessary to produce a given flux density is stated in gilberts per centimeter (oersteds) (H), or in ampere-turns per inch (H).

Reluctance Magnetic reluctance corresponds to electrical resistance, and is the property of a material that opposes the creation of a magnetic flux in the material. It is expressed in *rels*, and the symbol is the letter R. A material has a reluctance of 1 rel when an m.m.f. of 1 ampere-turn (NI) generates a flux of 1 line of force in it. Combinations of reluctances are treated the same as resistances in finding the total effective reluctance. The *specific reluctance* of any substance is its reluctance per unit volume.

Except for iron and its alloys, most common materials have a specific reluctance very nearly the same as that of a vacuum, which, for all practical purposes, may be considered the same as the specific reluctance of air.

Ohm's Law for Magnetic Circuits The relations between flux, magnetomotive force, and reluctance are exactly the same as the relations between current, voltage, and resistance in the electrical circuit. These can be stated as follows:

$$\phi = \frac{F}{R} \qquad R = \frac{F}{\phi} \qquad F = \phi R$$

where,

ϕ equals flux, F equals m.m.f.,
R equals reluctance.

Permeability *Permeability* expresses the ease with which a magnetic field may be set up in a material as compared with the effort required in the case of air. Iron, for example, has a permeability of around 2000 times that of air, which means that a given amount of magnetizing effort produced in an iron core by a current flowing through a coil of wire will produce 2000 times the *flux* density that the same magnetizing effect would produce in air. It may be expressed by the ratio B/H or B/H. In other words,

$$\mu = \frac{B}{H} \qquad \text{or} \qquad \mu = \frac{\mathrm{B}}{\mathrm{H}}$$

where μ is the permeability, B is the flux density in gausses, B is the flux density in lines per square inch, H is the m.m.f. in gilberts per centimeter (oersteds), and H is the m.m.f. in ampere-turns per inch. These relations may also be stated as follows:

$$H = \frac{B}{\mu} \text{ or } \mathrm{H} = \frac{\mathrm{B}}{\mu}, \text{ and } B = H\mu \text{ or } \mathrm{B} = \mathrm{H}\mu$$

It can be seen from the foregoing that permeability is inversely proportional to the specific reluctance of a material.

Saturation Permeability is similar to *electric conductivity*. This is, however, one important difference: the permeability of magnetic materials is not independent of the magnetic current (flux) flowing through it, although electrical conductivity is substantially independent of the electric current in a wire. When the flux density of a magnetic conductor has been increased to the *saturation point*, a further increase in the magnetizing force will not produce a corresponding increase in flux density.

B-H Curve To simplify magnetic circuit calculations, a magnetization curve may be drawn for a given unit of material. Such a curve is termed a B-H curve, and may be determined by experiment. When the current in an iron-core coil is first applied, the relation between the winding current and the core flux is shown at A-B in figure 18. If the current is then reduced to zero, reversed, brought back again to zero and reversed to the original direction, the flux passes through a typical hysteresis loop as shown.

Residual Magnetism; Retentivity The magnetism remaining in a material after the magnetizing force is removed is called *residual magnetism. Retentivity* is the property which causes a magnetic material to have residual magnetism after having been magnetized.

Hysteresis; Coercive Force *Hysteresis* is the characteristic of a magnetic system which causes a loss of power due to the fact that a negative magnetizing force must be applied to reduce the residual magnetism to zero. This negative force is termed *coercive force*. By "negative" magnetizing force is meant one which is of the opposite polarity with respect to the original magnetizing force. Hysteresis loss is apparent in

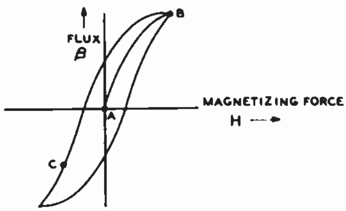


Figure 18

TYPICAL HYSTERESIS LOOP
(B-H CURVE = A-B)

Showing relationship between the current in the winding of an iron-core inductor and the core flux. A direct current flowing through the inductance brings the magnetic state of the core to some point on the hysteresis loop, such as C.

transformers and chokes by the heating of the core.

Inductance If the switch shown in figure 17 is opened and closed, a pulsating direct current will be produced. When it is first closed, the current does not instantaneously rise to its maximum value, but builds up to it. While it is building up, the magnetic field is expanding around the conductor. Of course, this happens in a small fraction of a second. If the switch is then opened, the current stops and the magnetic field contracts quickly. This expanding and contracting field will induce a current in any other conductor that is part of a continuous circuit which it cuts. Such a field can be obtained in the way just mentioned by means of a vibrator interruptor, or by applying a.c. to the circuit in place of the battery. Varying the resistance of the circuit will also produce the same effect. This inducing of a current in a conductor due to a varying current in another conductor not in actual contact is called *electromagnetic induction.*

Self-inductance If an alternating current flows through a coil the varying magnetic field around each turn cuts itself and the adjacent turn and *induces a voltage in the coil of opposite polarity to the applied e.m.f.* The amount of induced voltage depends on the number of turns in the coil, the current flowing in the coil, and the number of lines of force threading the coil. The voltage so induced is known as a *counter e.m.f.* or *back e.m.f.*, and the effect is termed *self-induction.* When the applied voltage is building up, the counter e.m.f. opposes the rise; when the applied voltage is decreasing, the counter e.m.f. is of the same polarity and tends to maintain the current. Thus, it can be seen that self-inductance tends to prevent any change in the current in the circuit.

The storage of energy in a magnetic field is expressed in *joules* and is equal to $(LI^2)/2$. (A joule is equal to 1 watt-second. L is defined immediately following.)

The Unit of Inductance: The Henry Inductance is usually denoted by the letter L, and is expressed in *henrys.* A coil has an inductance of 1 henry when a voltage of 1 volt is induced by a current change of 1 ampere per second. The henry, while commonly used in audio-frequency circuits, is too large for reference to inductance coils, such as those used in radio-frequency circuits; *millihenry* or *microhenry* is more commonly used, in the following manner:

1 *henry* = 1000 *millihenrys*, or 10^3 *millihenrys.*

1 *millihenry* = 1/1000 *henry*, .001 *henry*, or 10^{-3} *henry.*

1 *microhenry* = 1/1,000,000 *henry*, .000001 *henry*, or 10^{-6} *henry.*

1 *microhenry* = 1/1000 *millihenry*, .001, or 10^{-3} *millihenry.*

1000 *microhenrys* = 1 *millihenry.*

Mutual Inductance When one coil is near another, a varying current in one will produce a varying magnetic field which cuts the turns of the other coil, inducing a current in it. This induced current is also varying, and will therefore induce another current in the first coil. This reaction between two coupled circuits is called *mutual inductance*, and can be calculated and expressed in henrys. The symbol for mutual inductance is M. Two circuits thus joined are said to be *inductively coupled.*

The magnitude of the mutual inductance

depends on the shape and size of the two circuits, their positions and distances apart, and the permeability of the medium. The extent to which two inductors are coupled is expressed by a relation known as *coefficient of coupling* (k). This is the ratio of the mutual inductance actually present to the maximum possible value.

Thus, when k is 1, the coils have the maximum quantity mutual induction.

The mutual inductance of two coils can be formulated in terms of the individual inductances and the coefficient of coupling:

$$M = k\sqrt{L_1 \times L_2}$$

For example, the mutual inductance of two coils, each with an inductance of 10 henrys and a coupling coefficient of 0.8 is:

$$M = 0.8\sqrt{10 \times 10} = 0.8 \times 10 = 8$$

The formula for mutual inductance is $L = L_1 + L_2 + 2M$ when the coils are poled so that their fields add. When they are poled so that their fields buck, then $L = L_1 + L_2 - 2M$ (figure 19).

Inductors in Parallel Inductors in parallel are combined exactly as are resistors in parallel, provided that they are far enough apart so that the mutual inductance is entirely negligible.

Inductors in Series Inductors in series are additive, just as are resistors in series, again provided that no mutual inductance exists. In this case, the total inductance L is:

$$L = L_1 + L_2 + \ldots, \text{etc.}$$

Where mutual inductance does exist:

$$L = L_1 + L_2 + 2M$$

where,

M is the mutual inductance.

This latter expression assumes that the coils are connected in such a way that all flux linkages are in the same direction, i.e., additive. If this is not the case and the mutual linkages *subtract* from the self-linkages, the following formula holds:

$$L = L_1 + L_2 - 2M$$

where,

M is the mutual inductance.

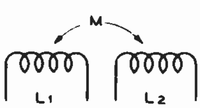


Figure 19

MUTUAL INDUCTANCE

The quantity M represents the mutual inductance between the two coils L_1 and L_2.

Core Material Ordinary magnetic cores cannot be used for radio frequencies because the *eddy current and hysteresis losses* in the core material become enormous as the frequency is increased. The principal use for conventional magnetic cores is in the audio-frequency range below approximately 15,000 Hertz, whereas at very low frequencies (50 to 60 Hertz) their use is mandatory if an appreciable value of inductance is desired.

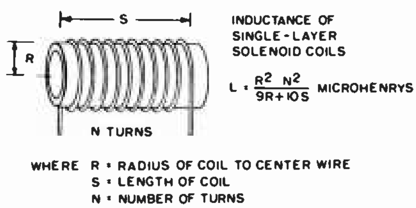


Figure 20

FORMULA FOR CALCULATING INDUCTANCE

Through the use of the equation and the sketch shown above the inductance of single-layer solenoid coils can be calculated with an accuracy of about one percent for the types of coils normally used in the hf and vhf range.

An air-core inductor of only 1 henry inductance would be quite large in size, yet values as high as 500 henrys are commonly available in small iron-core chokes. The inductance of a coil with a magnetic core will vary with the amount of current (both a-c and d-c) which passes through the coil. For this reason, iron-core chokes that are used in power supplies have a certain inductance rating at a *predetermined value of direct current.*

The permeability of air does not change with flux density; so the inductance of iron-core coils often is made less dependent on flux density by making part of the magnetic path air, instead of utilizing a closed loop of iron. This incorporation of an *air gap* is necessary in many applications of iron-core coils, particularly where the coil carries a considerable d-c component. Because the permeability of air is so much lower than that of iron, the air gap need comprise only a small fraction of the magnetic circuit in order to provide a substantial proportion of the total reluctance.

Iron-Core Inductors at Radio Frequencies Iron-core inductors may be used at radio frequencies if the iron is in a very finely divided form, as in the case of the powdered-iron cores used in some types of r-f coils and i-f transformers. These cores are made of extremely small particles of iron. The particles are treated with an insulating material so that each particle will be insulated from the others, and the treated powder is molded with a binder into cores. Eddy current losses are greatly reduced, with the result that these special iron cores are entirely practical in circuits which operate up to 100 MHz in frequency.

Ferrite-Core Inductors *Ferrite* materials provide high permeability and low core loss characteristics well into the vhf range. Commonly used ferrites are made primarily of iron oxide with a trace of manganese sintered at a high temperature into a ceramic-like material. Ferrite may take the form of a very small *ferrite bead* slipped over a wire to form a simple r-f choke, a *ferrite rod,* or a more complex *ferrite cup-core* assembly providing very high Q in a small volume. In addition, ferrite cores are available as *toroids* or standard E-I combinations (see Chapter 3).

Toroid Inductors The toroid winding provides a closed magnetic field allowing a large value of inductance per winding turn combined with minimum magnetic field outside the winding. Powdered-iron and ferrite toroids are available for hf and vhf operation. The powdered-iron toroids are commonly made of a carbonyl material and coils can be wound on such forms or cores showing Q values of several hundred. Ferrite toroids have a high value of permeability and provide coils having a greater inductance value for a given number of turns than a comparable powdered-iron unit.

Ferrite and powdered-iron cores of all types are widely used in bandswitching and broadband r-f transformers for both transmitters and receivers.

2-5 RC and RL Transients

A voltage divider may be constructed as shown in figure 21. Kirchhoff's and Ohm's Laws hold for such a divider. This circuit is known as an *RC circuit.*

Time Constant-RC and RL Circuits When switch S in figure 21 is placed in position 1, a voltmeter across capacitor C will indicate the manner in which the capacitor will become charged through the resistor R from battery B. If relatively large values are used for R and C, and if a vacuum-tube voltmeter which

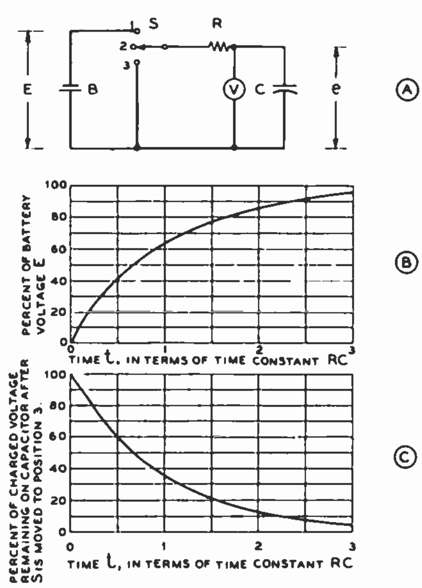


Figure 21

TIME CONSTANT OF AN RC CIRCUIT

Shown at (A) is the circuit upon which is based the curves of (B) and (C). (B) shows the rate at which capacitor C will charge from the instant at which switch S is placed in position 1. (C) shows the discharge curve of capacitor C from the instant at which switch S is placed in position 3.

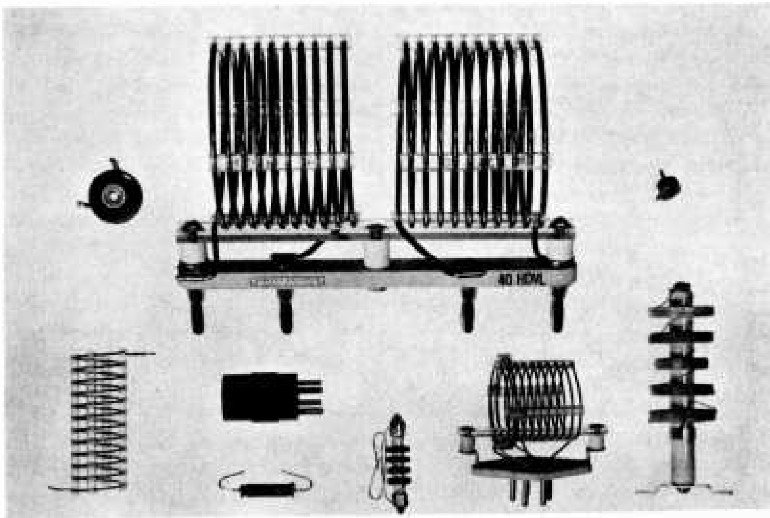


Figure 22

TYPICAL INDUCTANCES

The large inductance is a 1000-watt transmitting coil. To the right and left of this coil are small r-f chokes. Several varieties of low power capability coils are shown below, along with various types of r-f chokes intended for high-frequency operation.

draws negligible current is used to measure the voltage (e), the rate of charge of the capacitor may actually be plotted with the aid of a stop watch.

Voltage Gradient It will be found that the voltage (e) will begin to rise rapidly from zero the instant the switch is closed. Then, as the capacitor begins to charge, the rate of change of voltage across the capacitor will be found to decrease, the charging taking place more and more slowly as capacitor voltage e approaches battery voltage E. Actually, it will be found that in any given interval a constant percentage of the remaining difference between e and E will be delivered to the capacitor as an increase in voltage. A voltage which changes in this manner is said to increase *logarithmically*, or follows an *exponential* curve.

Time Constant A mathematical analysis of the charging of a capacitor in this manner would show that the relationship between battery voltage E and the voltage across the capacitor (e) could be expressed in the following manner:

$$e = E\,(1 - \varepsilon^{-t/RC})$$

where e,E,R, and C have the values discussed above, $\varepsilon = 2.716$ (the base of Naperian or natural logarithms), and t represents the time which has elapsed since the closing of

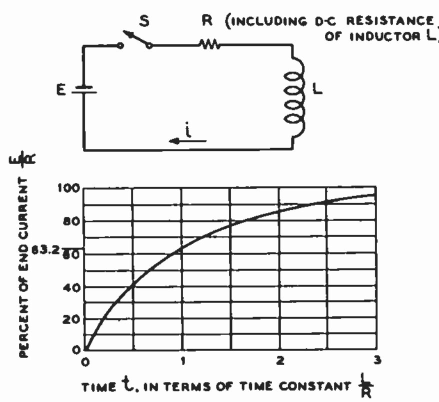


Figure 23

TIME CONSTANT OF AN RL CIRCUIT

Note that the time constant for the increase in current through an R L circuit is identical to the rate of increase in voltage across the capacitor in an R C circuit.

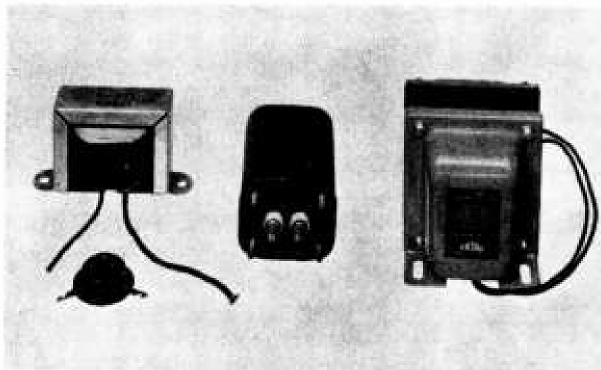


Figure 24

TYPICAL IRON-CORE INDUCTANCES

At the right is an upright mounting filter choke intended for use in low-powered transmitters and audio equipment. At the center is a hermetically sealed inductance for use under poor environmental conditions. To the left is an inexpensive receiving-type choke, with a small iron-core r-f choke directly in front of it.

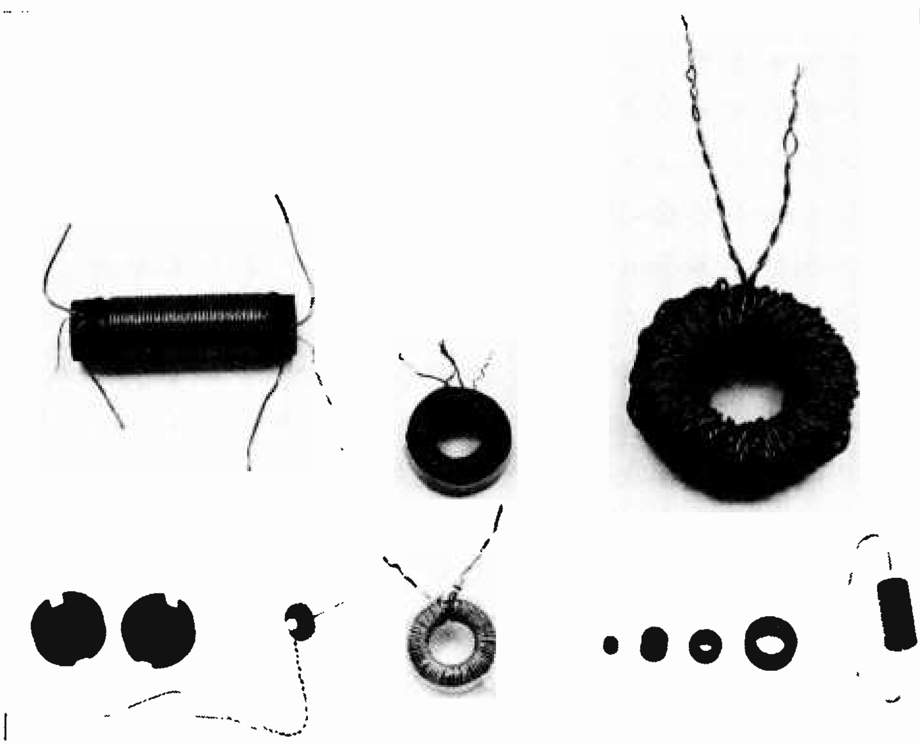


Figure 25

At top left is a trifilar (three-winding) filament choke wound on a ferrite rod. To the right are two toroid inductors with bifilar windings on ferrite cores. At the lower left is a ferrite cup-core assembly, with two miniature ferrite toroid inductors at the center. To the lower right are typical miniature ferrite toroid cores and an encapsulated ferrite-core r-f choke.

the switch. With t expressed in seconds, R and C may be expressed in farads and ohms, or R and C may be expressed in microfarads and megohms. The product RC is called the *time constant* of the circuit, and is expressed in seconds. As an example, if R is one megohm and C is one microfarad, the time constant RC will be equal to the product of the two, or one second.

When the elapsed time (t) is equal to the time constant of the RC network under consideration, the exponent of ε becomes -1. Now ε^{-1} is equal to $1/\varepsilon$, or $1/2.716$, which is 0.368. The quantity $(1-0.368)$ then is equal to 0.632. Expressed as percentage, the above means that the voltage across the capacitor will have increased to 63.2 per cent of the battery voltage in an interval equal to the time constant or RC product of the circuit. Then, during the next period equal to the time constant of the RC combination, the voltage across the capacitor will have risen to 63.2 per cent of the remaining difference in voltage, or 86.5 per cent of the applied voltage (E).

RL Circuit In the case of a series combination of a resistor and an inductor, as shown in figure 23, the current through the combination follows a very similar law to that given above for the voltage appearing across the capacitor in an RC series circuit. The equation for the current through the combination is:

$$i = \frac{E}{R}\ (1-\varepsilon^{-t/RL})$$

where i represents the current at any instant through the series circuit, E represents the applied voltage, and R represents the total resistance of the resistor and the d-c resistance of the inductor in series. Thus the time constant of the RL circuit is L/R, with R expressed in ohms and L expressed in henrys.

Voltage Decay When the switch in figure 21 is moved to position 3 after the capacitor has been charged, the capacitor voltage will drop in the manner shown in figure 21-C. In this case the voltage across the capacitor will decrease to 36.8 percent of the initial voltage (will make 63.2 per cent of the total drop) in a period of time equal to the time constant of the RC circuit.

CHAPTER THREE

Alternating-Current Circuits

The previous chapter has been devoted to a discussion of circuits and circuit elements upon which is impressed a current consisting of a flow of electrons in one direction. This type of unidirectional current flow is called direct current (abbreviated *d-c* or *d.c.*). Equally as important in radio and communications work and power practice is a type of current whose direction of electron flow reverses periodically. The reversal of flow may take place at a low rate, in the case of power systems, or it may take place millions of times per second, in the case of communications frequencies. This type of current flow is called *alternating current* (abbreviated *a-c* or *a.c.*).

3-1 Alternating Current

Frequency of an Alternating Current An alternating current is one whose amplitude of current flow periodically rises from zero to a maximum in one direction, decreases to zero, changes its direction, rises to maximum in the opposite direction, and decreases to zero again. This complete process, starting from zero, passing through two maximums in opposite directions, and returning to zero again, is called a *cycle.* The number of times per second that a current passes through the complete cycle is called the *frequency* (*f*) of the current. One and one-quarter cycles of an alternating current wave are illustrated diagrammatically in figure 1.

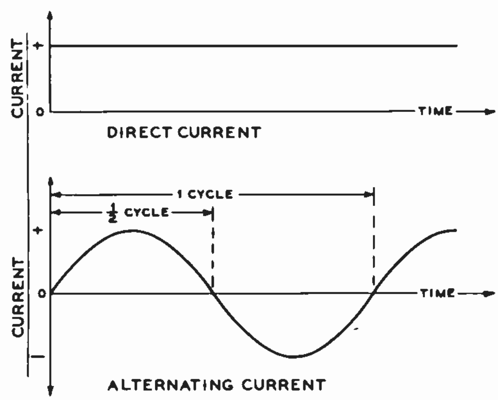


Figure 1

ALTERNATING CURRENT AND DIRECT CURRENT

Graphical comparison between unidirectional (direct) current and alternating current as plotted against time.

Frequency Spectrum At present the usable frequency range for alternating electrical currents extends over the electromagnetic spectrum from about 15 cycles per second to perhaps 30,000,000,000 cycles per second. It is obviously cumbersome to use a frequency designation in c.p.s. for enormously high frequencies, so three common units which are multiples of one cycle per second were established and are still used by many engineers.

These units have been:

(1) The kilocycle (kc), 1000 c.p.s.
(2) The megacycle (Mc), 1,000,000 c.p.s. or 1000 kc.
(3) The kilomegacycle (kMc), 1,000,000,000 c.p.s. or 1000 Mc.

Used for some time in other countries, and recently adopted by the U. S. National Bureau of Standards, IEEE, and many other American organizations, the *Hertz* is the present unit of frequency measurement.

One *Hertz* is precisely defined as *one cycle per second* and is not to be confused with any other time base. Hertz is abbreviated as *Hz* (no period). The standard metric prefixes for *kilo*, *mega*, *giga*, etc. are used with the basic unit. Since "m" denotes "milli," capital "M" is used for mega, and small "k" is kilo. Thus megacycle becomes *megahertz* (MHz), kilocycle is *kilohertz* (kHz), etc.

The frequencies falling between about 15 and 20,000 hertz are called *audio* frequencies (abbreviated *a.f.*), since these frequencies are audible to the human ear when converted from electrical to acoustical signals by a speaker or headphone. Frequencies in the vicinity of 60 Hz also are called *power* frequencies, since they are commonly used to distribute electrical power to the consumer.

The frequencies falling between 3,000 c.p.s. (3 kHz) and 300 GHz are commonly called *radio* frequencies (abbreviated *r.f.*), since they are commonly used in radio communication and allied arts. The radio-frequency spectrum is often arbitrarily classified into eight frequency bands, each one of which is ten times as high in frequency as the one just below it in the spectrum. The present spectrum, with classifications, is given in Table I.

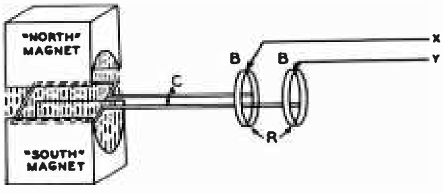


Figure 2

THE ALTERNATOR

Semi-schematic representation of the simplest form of the alternator.

TABLE 1.

FREQUENCY CLASSIFICATION

FREQUENCY	CLASSIFICATION	DESIGNATION
3 to 30 kHz	Very-low frequency	VLF
30 to 300 kHz	Low frequency	LF
300 to 3000 kHz	Medium frequency	MF
3 to 30 MHz	High frequency	HF
30 to 300 MHz	Very-high frequency	VHF
300 to 3000 MHz	Ultrahigh frequency	UHF
3 to 30 GHz	Superhigh frequency	SHF
30 to 300 GHz	Extremely high frequency	EHF

Generation of Alternating Current Faraday discovered that if a conductor which forms part of a closed circuit is moved through a magnetic field so as to cut across the lines of force, a current will flow in the conductor. He also discovered that, when a conductor in a second closed circuit is brought near the first conductor and the current in the first one is varied, a current will flow in the second conductor. This effect is known as *induction*, and the currents so generated are *induced currents*. In the latter case it is the lines of force which are moving and cutting the second conductor, due to the varying current strength in the first conductor.

A current is induced in a conductor if there is a relative motion between the conductor and a magnetic field, its direction of flow depending on the direction of the relative motion between the conductor and the field, and its strength depends on the intensity of the field, the rate of cutting lines of force, and the number of turns in the conductor.

Alternators A machine that generates an alternating current is called an *alternator* or *a-c* generator. Such a machine in its basic form is shown in figure 2. It consists of two permanent magnets, the opposite poles of which face each other and are machined so that they have a common radius. Between these two poles (north and south) a substantially constant magnetic field exists. If a conductor in the form of a loop (C) is suspended so that it can be

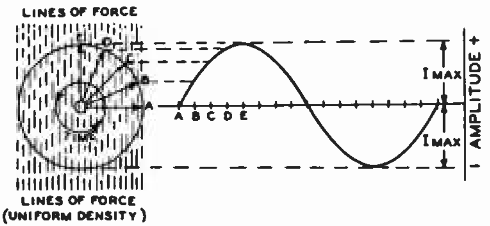


Figure 3

OUTPUT OF THE ALTERNATOR

Graph showing sine-wave output current of the alternator of figure 2.

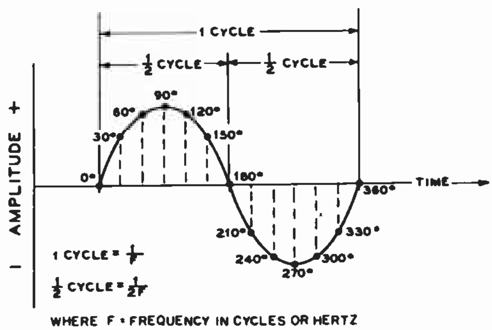


Figure 4

THE SINE WAVE

Illustrating one cycle of a sine wave. One complete cycle of alternation is broken up into 360 degrees. Then one-half cycle is 180 degrees, one-quarter cycle is 90 degrees, and so on down to the smallest division of the wave. A cosine wave has a shape identical to a sine wave but is shifted 90 degrees in phase — in other words the wave begins at full amplitude, the 90-degree point comes at zero amplitude, the 180-degree point comes at full amplitude in the opposite direction of current flow, etc.

freely rotated between the two poles, and if the opposite ends of conductor C are brought to collector rings, there will be a flow of alternating current when conductor C is rotated. This current flows out through the collector rings (R) and brushes (B) to the external circuit (X-Y).

The field intensity between the two pole pieces is substantially constant over the entire area of the pole face. However, when the conductor is moving parallel to the lines of force at the top or bottom of the pole faces, no lines are being cut. As the conductor moves on across the pole face it cuts more and more lines of force for each unit distance of travel, until it is cutting the maximum number of lines when opposite the center of the pole. Therefore, zero current is induced in the conductor at the instant it is midway between the two poles, and maximum current is induced when it is opposite the center of the pole face. After the conductor has rotated through 180° it can be seen that its position with respect to the pole pieces will be exactly opposite to that when it started. Hence, the second 180° of rotation will produce an alternation of current in the opposite direction to that of the first alternation.

The current does *not* increase directly as the angle of rotation, but rather as the *sine* of the angle; hence, such a current has the mathematical form of a *sine wave.* Although most electrical machinery does not produce a strictly pure sine curve, the departures are usually so slight that the assumption can be regarded as fact for most practical purposes. All that has been said in the foregoing paragraphs concerning alternating current also is applicable to alternating voltage.

The rotating arrow to the left in figure 3 represents a conductor rotating in a constant magnetic field of uniform density. The arrow also can be taken as a *vector* representing the strength of the magnetic field. This means that the length of the arrow is determined by the strength of the field (number of lines of force), which is constant. Now if the arrow is rotating at a constant rate (that is, with constant *angular velocity*), then the voltage developed across the conductor will be proportional to the rate at which it is cutting lines of force, which rate is proportional to the vertical distance between the tip of the arrow and the horizontal base line.

If EO is taken as unity, or a voltage of 1, then the voltage (vertical distance from tip of arrow to the horizontal base line) at point C for instance may be determined simply by referring to a table of sines and looking up the sine of the angle which the arrow makes with the horizontal.

When the arrow has traveled from point A to point E, it has traveled 90 degrees or one quarter cycle. The other three quadrants are not shown because their complementary or mirror relationship to the first quadrant is obvious.

It is important to note that time units are represented by *degrees* or quadrants. The fact that AB, BC, CD, and DE are equal chords (forming equal quadrants) simply means that the arrow (conductor or vector) is traveling at a constant speed, because these points on the radius represent the passage of equal units of time.

The whole picture can be represented in another way, and its derivation from the foregoing is shown in figure 3. The time base is represented by a straight line rather than by angular rotation. Points A, B, C, etc., represent the same units of time as before. When the voltage corresponding to each point is projected to the corresponding time unit, the familiar *sine curve* is the result.

The frequency of the generated voltage is proportional to the speed of rotation of the alternator, and to the number of magnetic poles in the field. Alternators may be built to produce radio frequencies up to 30 kHz, and some such machines are still used for low-frequency communication purposes. By means of multiple windings, three-phase output may be obtained from large industrial alternators.

Radian Notation From figure 1 we see that the value of an a-c wave varies continuously. It is often of importance to know the amplitude of the wave in terms of the total amplitude at any instant or at any time within the cycle. To be able to establish the instant in question we must be able to divide the cycle into parts. We could divide the cycle into eighths, hundredths, or any other ratio that suited our fancy. However, it is much more convenient mathematically to divide the cycle either into *electrical degrees* (360° represent one cycle) or into *radians*. A radian is an arc of a circle equal to the radius of the circle; hence there are 2π radians per cycle—or per circle (since there are π diameters per circumference, there are 2π radii).

Both radian notation and electrical-degree notation are used in discussions of alternating-current circuits. However, trigonometric tables are much more readily available in terms of degrees than radians, so the following simple conversions are useful.

$$2\pi \text{ radians} = 1 \text{ cycle} = 360^\circ$$

$$\pi \text{ radians} = 1/2 \text{ cycle} = 180^\circ$$

$$\frac{\pi}{2} \text{ radians} = 1/4 \text{ cycle} = 90^\circ$$

$$\frac{\pi}{3} \text{ radians} = 1/6 \text{ cycle} = 60^\circ$$

$$\frac{\pi}{4} \text{ radians} = 1/8 \text{ cycle} = 45^\circ$$

$$1 \text{ radian} = \frac{1}{2\pi} \text{cycle} = 57.3^\circ$$

When the conductor in the simple alternator of figure 2 has made one complete revolution it has generated one cycle and has rotated through 2π radians. The expression $2\pi f$ then represents the number of radians in one cycle multiplied by the number of cycles per second (the frequency) of the alternating voltage or current. The expression then represents the number of radians per second through which the conductor has rotated. Hence $2\pi f$ represents the angular velocity of the rotating conductor, or of the rotating vector, which represents any alternating current or voltage, expressed in radians per second.

In technical literature the expression $2\pi f$ is often replaced by ω, the lower-case Greek letter *omega*. Velocity multiplied by time gives the distance travelled, so $2\pi ft$ (or ωt) represents the angular distance through which the rotating conductor or the rotating vector has travelled since the reference time $t = 0$. In the case of a sine wave the reference time $t = 0$ represents the instant when the voltage or the current, whichever is under discussion, also is equal to zero.

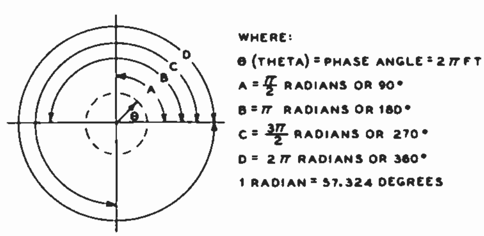


Figure 5

ILLUSTRATING RADIAN NOTATION

The radian is a unit of phase angle, equal to 57.324 degrees. It is commonly used in mathematical relationships involving phase angles since such relationships are simplified when radian notation is used.

Instantaneous Value of Voltage or Current The instantaneous voltage or current is proportional to the sine of the angle through which the rotating vector has travelled since reference time $t = 0$. Hence, when the peak value of the a-c wave amplitude (either voltage or current amplitude) is known, and the angle through which the rotating vector has travelled is established, the amplitude of the wave at this instant can be determined through use of the following expression:

$$e = E_{max} \sin 2\pi f t$$

where,

e equals the instantaneous voltage,
E_{max} equals maximum peak value of voltage,
f equals frequency in hertz,
t equals period of time which has elapsed since $t = 0$ (expressed as a fraction of one second).

The instantaneous current can be found from the same expression by substituting i for e and I_{max} for E_{max}.

It is often easier to visualize the process of determining the instantaneous amplitude by ignoring the frequency and considering only one cycle of the a-c wave. In this case, for a sine wave, the expression becomes:

$$e = E_{max} \sin \theta$$

where θ represents the angle through which the vector has rotated since time (and amplitude) were zero. As examples:

when $\theta = 30°$

$\sin \theta = 0.5$

so $e = 0.5\ E_{max}$

. . . .

when $\theta = 60°$

$\sin \theta = 0.866$

so $e = 0.866\ E_{max}$

. . . .

when $\theta = 90°$

$\sin \theta = 1.0$

so $e = E_{max}$

. . . .

when $\theta = 1$ radian

$\sin \theta = 0.8415$

so $e = 0.8415\ E_{max}$

Effective Value of an Alternating Current The instantaneous value of an alternating current or voltage varies continuously throughout the cycle, so some value of an a-c wave must be chosen to establish a relationship between the effectiveness of an a-c and a d-c voltage or current. The heating value of an alternating current has been chosen to establish the reference between the *effective* values of a.c. and d.c. Thus *an alternating current will have an effective value of 1 ampere when it produces the same heat in a resistor as does 1 ampere of direct current.*

The effective value is derived by taking the instantaneous values of current over a cycle of alternating current, squaring these values, taking an average of the squares, and then taking the square root of the average. By this procedure, the effective value becomes known as the *root mean square*, or rms, value. This is the value that is read on a-c voltmeters and a-c ammeters. The rms value is 70.7 percent of the peak or maximum instantaneous value (for sine waves only) and is expressed as follows:

E_{eff} or $E_{rms} = 0.707 \times E_{max}$, or

I_{eff} or $I_{rms} = 0.707 \times I_{max}$

The following relations are extremely useful in radio and power work:

$E_{rms} = 0.707 \times E_{max}$, and

$E_{max} = 1.414 \times E_{rms}$

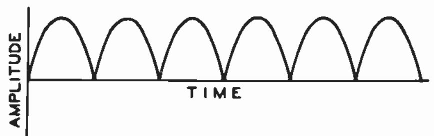


Figure 6

FULL-WAVE RECTIFIED SINE WAVE

Waveform obtained at the output of a full-wave rectifier being fed with a sine wave and having 100 per cent rectification efficiency. Each pulse has the same shape as one-half cycle of a sine wave. This type of current is known as pulsating direct current.

Rectified Alternating Current or Pulsating Direct Current If an alternating current is passed through a rectifier, it emerges in the form of a current of *varying amplitude* which flows in *one* direction only. Such a current is known as *rectified a.c.* or *pulsating d.c.* A typical wave form of a pulsating direct current as would be obtained from the output of a full-wave rectifier is shown in figure 6.

Measuring instruments designed for d-c operation will not read the peak or instantaneous maximum value of the pulsating d-c output from the rectifier; they will read only the *average value*. This can be explained by assuming that it could be possible to cut off some of the peaks of the waves, using the cutoff portions to fill in the spaces that are open, thereby obtaining an *average* d-c value. A milliammeter and voltmeter connected to the adjoining circuit, or across the output of the rectifier, will read this average value. It is related to *peak* value by the following expression:

$$E_{avg} = 0.636 \times E_{max}$$

It is thus seen that the average value is 63.6 percent of the peak value.

Relationship Between Peak, RMS, or Effective, and Average Values To summarize the three most significant values of an a-c sine wave: the peak value is equal to 1.41 times the rms or effective, and the rms value is equal to 0.707 times the peak value; the average value of a full-wave rectified a-c wave is 0.636 times the peak value, and the average value of a rectified wave is equal to 0.9 times the rms value.

$$\text{rms} = 0.707 \times \text{peak}$$

$$\text{average} = 0.636 \times \text{peak}$$

$$\text{average} = 0.9 \times \text{rms}$$

$$\text{rms} = 1.11 \times \text{average}$$

$$\text{peak} = 1.414 \times \text{rms}$$

$$\text{peak} = 1.57 \times \text{average}$$

Applying Ohm's Law to Alternating Current Ohm's Law applies equally to direct or alternating current, *provided* the circuits under consideration are purely resistive, that is, circuits which have neither inductance (coils) nor capacitance (capacitors). Problems which involve tube filaments, dropping resistors, electric lamps, heaters or similar resistive devices can be solved with Ohm's Law, regardless of whether the current is direct or alternating. When a capacitor or coil is made a part of the circuit, a property common to either, called *reactance*, must be taken into consideration. Ohm's Law still applies to a-c circuits containing reactance, but additional considerations are involved; these will be discussed in a later paragraph.

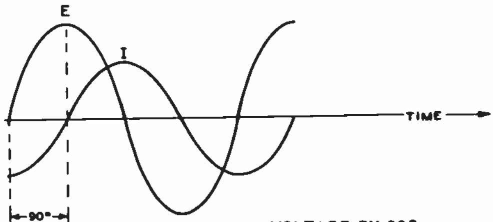


Figure 7

LAGGING PHASE ANGLE

Showing the manner in which the current lags the voltage in an a-c circuit containing pure inductance only. The lag is equal to one-quarter cycle or 90 degrees.

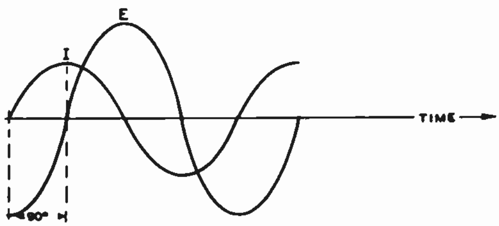
Inductive Reactance As was stated in Chapter Two, when a changing current flows through an inductor a back- or counterelectromotive force is developed, opposing any change in the initial current. This property of an inductor causes it to offer opposition or *impedance* to a change in current. The measure of impedance offered by an inductor to an alternating current of a given frequency is known as its *inductive reactance*. This is expressed as X_L and is shown in figure 7.

$$X_L = 2\pi f L$$

where,

X_L equals inductive reactance expressed in ohms,
π equals 3.1416 ($2\pi = 6.283$),
f equals frequency in Hertz,
L equals inductance in henrys.

Inductive Reactance at Radio Frequencies It is very often necessary to compute inductive reactance at radio frequencies. The same formula may be used, but to make it less cumbersome the inductance is expressed in *millihenrys* and the frequency in *kilohertz*. For higher frequencies and smaller values of inductance, frequency is expressed in *megahertz* and inductance in *microhenrys*. The basic equation need not be changed, since the multiplying factors for inductance and frequency appear in numerator and denominator, and hence are cancelled out. However, it is not possible in the same equation to express L in millihenrys and f in Hertz without conversion factors.



CURRENT LEADING VOLTAGE BY 90°
(CIRCUIT CONTAINING PURE CAPACITANCE ONLY)

Figure 8

LEADING PHASE ANGLE

Showing the manner in which the current leads the voltage in an a-c circuit containing pure capacitance only. The lead is equal to one-quarter cycle or 90 degrees.

Capacitive Reactance It has been explained that inductive reactance is the measure of the ability of an inductor to offer impedance to the flow of an alternating current. Capacitors have a similar property although in this case the opposition is to any change in the voltage across the capacitor. This property is called *capacitive reactance* and is expressed as follows:

$$X_C = \frac{1}{2\pi fC}$$

where,

X_C equals capacitive reactance in ohms,
π equals 3.1416,
f equals frequency in Hertz,
C equals capacitance in farads.

Capacitive Reactance at Radio Frequencies Here again, as in the case of inductive reactance, the units of capacitance and frequency can be

TABLE 2. Quantities, Units, and Symbols

Symbol	Quantity	Unit	Abbreviation
f	Frequency	hertz	Hz
λ	Wavelength	meter	M
X_L	Inductive Reactance	ohm	Ω
X_C	Capacitive Reactance	ohm	Ω
Q	Figure of merit	reactance / resistance	—
Z	Impedance	ohm	Ω

e = instantaneous value of voltage
E_{max} = peak value of voltage
i = instantaneous value of current
I_{max} = peak value of current
θ = phase angle, expressed in degrees
E_{eff} or E_{rms} = effective or rms value of voltage
I_{eff} or I_{rms} = effective or rms value of current
j = vector operator (90° rotation)

converted into smaller units for practical problems encountered in radio work. The equation may be written:

$$X_C = \frac{1{,}000{,}000}{2\pi fC}$$

where,

f equals frequency in megahertz,
C equals capacitance in picofarads.

In the audio range it is often convenient to express frequency (f) in *Hertz* and capacitance (C) in *microfarads*, in which event the same formula applies.

Phase When an alternating current flows through a purely resistive circuit, it will be found that the current will go through maximum and minimum in perfect step with the voltage. In this case the current is said to be in step, or *in phase* with the voltage. For this reason, Ohm's Law will apply equally well for *a.c.* or *d.c.* where pure resistances are concerned, provided that the same values of the wave (either peak or rms) for both voltage and current are used in the calculations.

However, in calculations involving alternating currents the voltage and current are not necessarily in phase. The current through the circuit may lag behind the voltage, in which case the current is said to have *lagging* phase. Lagging phase is caused by inductive reactance. If the current reaches its maximum value ahead of the voltage (figure 8) the current is said to have a *leading* phase. A leading phase angle is caused by capacitive reactance.

In an electrical circuit containing reactance only, the current will either lead or lag the voltage by 90°. If the circuit contains inductive reactance only, the current will lag the voltage by 90°. If only capacitive reactance is in the circuit, the current will lead the voltage by 90°.

Reactances in Combination Inductive and capacitive reactance have exactly opposite effects on the phase relation between current and voltage in a circuit. Hence when they are used in combination their effects tend to neutralize. The combined effect of a capacitive and an inductive reactance is often called the *net reactance* of a circuit. The net reactance (X) is found by subtracting the capacitive reactance from the inductive reactance ($X = X_L - X_C$).

The result of such a combination of pure reactances may be either positive, in which case the positive reactance is greater so that the net reactance is inductive, or it may be negative in which case the capacitive reactance is greater so that the net reactance is capacitive. The net reactance may also be zero in which case the circuit is said to be *resonant*. The condition of resonance will be discussed in a later section. Note that inductive reactance is always taken as being positive while capacitive reactance is always taken as being negative.

Impedance; Circuits Containing Reactance and Resistance Pure reactances introduce a phase angle of 90° between voltage and current; pure resistance introduces no phase shift between voltage and current. Hence we cannot add a reactance and a resistance directly. When a reactance and a resistance are used in combination the resulting phase angle of current flow with respect to the impressed voltage lies somewhere between plus or minus 90° and 0° depending on the relative magnitudes of the reactance and the resistance.

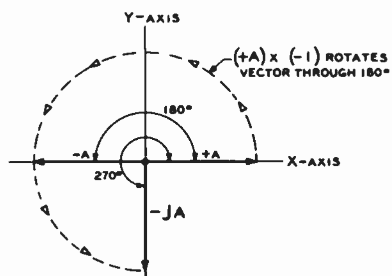


Figure 9

Operation on the vector (+A) by the quantity (−1) causes vector to rotate through 180 degrees.

The term *impedance* is a general term which can be applied to any electrical entity which impedes the flow of current. Hence the term may be used to designate a resistance, a pure reactance, or a complex combination of both reactance and resistance. The designation for impedance is Z. An impedance must be defined in such a manner that both its magnitude and its phase angle are established. The designation may be accomplished in either of two ways—one of which is convertible into the other by simple mathematical operations.

The j Operator The first method of designating an impedance is actually to specify both the resistive and the reactive component in the form $R + jX$. In this form R represents the resistive component in ohms and X represents the reactive component. The j merely means that the X component is reactive and thus cannot be added directly to the R component. Plus jX means that the reactance is positive or inductive, while if minus jX were given it would mean that the reactive component was negative or capacitive.

In figure 9 we have a vector $(+A)$ lying along the positive X-axis of the usual X-Y coordinate system. If this vector is multiplied by the quantity (-1), it becomes $(-A)$ and its position now lies along the X-axis in the negative direction. The *operator* (-1) has caused the vector to rotate through an angle of 180 degrees. Since (-1) is equal to $(\sqrt{-1} \times \sqrt{-1})$, the same result may be obtained by operating on the vector with the operator $(\sqrt{-1} \times \sqrt{-1})$. However if the vector is operated on but once by the operator $(\sqrt{-1})$, it is caused to rotate only 90 degrees (figure 10). Thus the operator $(\sqrt{-1})$ rotates a vector by 90 degrees. For convenience, this operator is called the *j operator*. In like fashion, the operator $(-j)$ rotates the vector of figure 9 through an angle of 270 degrees, so that the resulting vector $(-jA)$ falls on the $(-Y)$ axis of the coordinate system.

Polar Notation The second method of representing an impedance is to specify its absolute magnitude and the phase angle of current with respect to voltage, in the form $Z \angle \theta$. Figure 11 shows graphically the relationship between the two common ways of representing an impedance.

The construction of figure 11 is called an *impedance diagram*. Through the use of such a diagram we can add graphically a resistance and a reactance to obtain a value for the resulting impedance in the scalar form. With zero at the origin, resistances are plotted to the right, positive values of reactance (inductive) in the upward direction, and negative values of reactance (capacitive) in the downward direction.

Note that the resistance and reactance are drawn as the two sides of a right triangle, with the hypotenuse representing the resulting impedance. Hence it is possible to determine mathematically the value of a resultant impedance through the familiar right-triangle relationship—*the square of the hypotenuse is equal to the sum of the squares of the other two sides:*

$$Z^2 = R^2 + X^2$$

or,

$$|Z| = \sqrt{R^2 + X^2}$$

Note also that the angle θ included between R and Z can be determined from any of the following trigonometric relationships:

$$\sin \theta = \frac{X}{|Z|}$$

$$\cos \theta = \frac{R}{|Z|}$$

$$\tan \theta = \frac{X}{R}$$

One common problem is that of determining the *scalar magnitude* of the impedance, $|Z|$, and the phase angle θ, when resistance and reactance are known; hence, of converting from the $Z = R + jX$ to the $|Z| \angle \theta$ form. In this case we use two of the expressions just given:

$$|Z| = \sqrt{R^2 + X^2}$$

$$\tan \theta = \frac{X}{R}, \text{ (or } \theta = \tan^{-1}\frac{X}{R}\text{)}$$

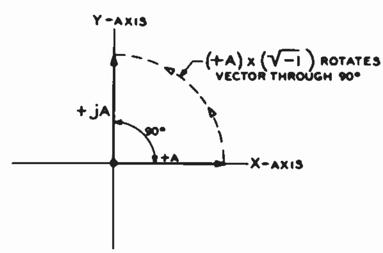


Figure 10

Operation on the vector (+A) by the quantity (j) causes vector to rotate through 90 degrees.

The inverse problem, that of converting from the $|Z| \angle \theta$ to the $R + jX$ form is done with the following relationships, both of which are obtainable by simple division from the trigonometric expressions just given for determining the angle θ:

$$R = |Z| \cos \theta$$

$$jX = |Z|\, j \sin \theta$$

By simple addition these two expressions may be combined to give the relationship between the two most common methods of indicating an impedance:

$$R + jX = |Z|\,(\cos \theta + j \sin \theta)$$

In the case of impedance, resistance, or reactance, the unit of measurement is the ohm; hence, the ohm may be thought of as a unit of *opposition to current flow*, without reference to the relative phase angle between the applied voltage and the current which flows.

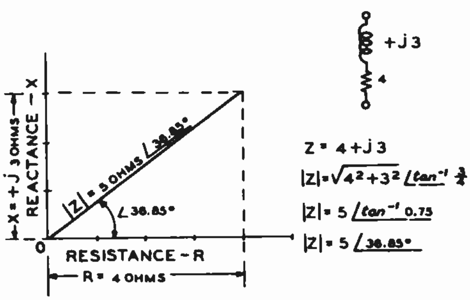


Figure 11

THE IMPEDANCE TRIANGLE

Showing the graphical construction of a triangle for obtaining the net (scalar) impedance resulting from the connection of a resistance and a reactance in series. Shown also alongside is the alternative mathematical procedure for obtaining the values associated with the triangle.

Further, since both capacitive and inductive reactance are functions of frequency, impedance will vary with frequency. Figure 12 shows the manner in which $|Z|$ will vary with frequency in an *RL* series circuit and in an *RC* series circuit.

Series RLC Circuits In a series circuit containing *R*, *L*, and *C*, the impedance is determined as discussed before except that the reactive component in the expressions defines the net reactance—that is, the difference between X_L and X_C. Hence $(X_L - X_C)$ may be substituted for X in the equations. Thus:

$$|Z| = \sqrt{R^2 + (X_L - X_C)^2}$$

$$\theta = \tan^{-1} \frac{(X_L - X_C)}{R}$$

A series *RLC* circuit thus may present an impedance which is capacitively reactive if the net reactance is capacitive, inductively reactive if the net reactance is inductive, or resistive if the capacitive and inductive reactances are equal.

Addition of Complex Quantities The addition of complex quantities (for example, impedances in series) is quite simple if the quantities are in the rectangular form. If they are in the polar form they only can be added graphically, unless they are converted to the rectangular form by the relationships previously given. As an example of the addition of complex quantities in the rectangular form, the equation for the addition impedance is:

$$(R_1 + jX_1) + (R_2 + jX_2) = (R_1 + R_2) + j(X_1 + X_2)$$

For example if we wish to add the impedances $(10 + j50)$ and $(20 - j30)$ we obtain:

$$\begin{aligned}(10 + j50) + (20 - j30) &= (10 + 20) + j[50 + (-30)] \\ &= 30 + j(50 - 30) \\ &= 30 + j20\end{aligned}$$

Multiplication and Division of Complex Quantities It is often necessary in solving certain types of circuits to multiply or divide two complex quantities. It is a much simpler mathematical operation to multiply or divide complex quantities if they are expressed in the polar form. Hence if they are given in the rectangular form they should be converted to the polar form before multiplication or division is begun. Then the multiplication is accomplished by multiplying the $|Z|$ terms together and *adding* algebraically the $\angle \theta$ terms, as:

$$(|Z_1| \angle\theta_1)\ (|Z_2| \angle\theta_2) = |Z_1|\,|Z_2|\ (\angle\theta_1 + \angle\theta_2)$$

For example, suppose that the two impedances $|20| \angle 43°$ and $|32| \angle -23°$ are to be multiplied. Then:

$$(|20| \angle 43°)\ (|32| \angle -23°) = |20 \cdot 32|\ (\angle 43° + \angle -23°) = 640 \angle 20°$$

Division is accomplished by dividing the denominator into the numerator, and *subtracting* the angle of the denominator from that of the numerator, as:

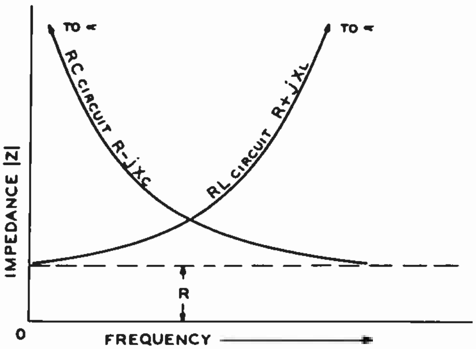


Figure 12

IMPEDANCE—FREQUENCY GRAPH FOR RL AND RC CIRCUITS

The impedance of an RC circuit approaches infinity as the frequency approaches zero (d.c.), while the impedance of a series RL circuit approaches infinity as the frequency approaches infinity. The impedance of an RC circuit approaches the impedance of the series resistor as the frequency approaches infinity, while the impedance of a series RL circuit approaches the resistance as the frequency approaches zero.

$$\frac{|Z_1| \angle\theta_1}{|Z_2| \angle\theta_2} = \frac{|Z_1|}{|Z_2|}(\angle\theta_1 - \angle\theta_2)$$

For example, suppose that an impedance of $|50| \angle 67°$ is to be divided by an impedance of $|10| \angle 45°$. Then:

$$\frac{|50| \angle 67°}{|10| \angle 45°} = \frac{|50|}{|10|}(\angle 67° - \angle 45°) = |5|(\angle 22°)$$

Ohm's Law for Complex Quantities The simple form of Ohm's Law used for d-c circuits may be stated in a more general form for application to a-c circuits involving either complex quantities or simple resistive elements. The form is:

$$I = \frac{E}{Z}$$

in which, in the general case, I, E, and Z are complex (vector) quantities. In the simple case where the impedance is a pure resistance with an a-c voltage applied, the equation simplifies to the familiar $I = E/R$. In any case the applied voltage may be expressed either as peak, rms, or average; the resulting current always will be in the same type used to define the voltage.

In the more general case vector algebra must be used to solve the equation. And, since either division or multiplication is involved, the complex quantities should be expressed in the polar form. As an example, take the case of the series circuit shown in figure 13 with 100 volts applied. The impedance of the series circuit can best be obtained first in the rectangular form, as:

$$200 + j(100 - 300) = 200 - j200$$

Now, to obtain the current we must convert this impedance to the polar form.

$$|Z| = \sqrt{200^2 + (-200)^2} = \sqrt{40{,}000 + 40{,}000} = \sqrt{80{,}000} = 282\ \Omega$$

$$\theta = \tan^{-1}\frac{X}{R} = \tan^{-1}\frac{-200}{200} = \tan^{-1}(-1) = -45°$$

Therefore, $Z = 282 \angle -45°$

Note that in a series circuit the resulting impedance takes the sign of the largest reactance in the series combination.

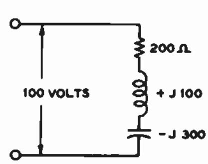


Figure 13

SERIES RLC CIRCUIT

Where a slide rule is being used to make the computations, the impedance may be found without any addition or subtraction operations by finding the angle θ first, and then using the trigonometric equation below for obtaining the impedance. Thus:

$$\theta = \tan^{-1}\frac{X}{R} = \tan^{-1}\frac{-200}{200} = \tan^{-1}(-1)$$
$$= -45°.$$

Then, $\quad Z$ equals $\dfrac{R}{\cos\theta}$

and $\quad \cos -45° = 0.707$

$$|Z| = \frac{200}{0.707} = 282 \text{ ohms}$$

Since the applied voltage will be the reference for the currents and voltages within the circuit, we may define it as having a zero phase angle: $E = 100\ \angle 0°$. Then:

$$I = \frac{100\ \angle 0°}{282\ \angle -45°} = 0.354\ \angle 0° - (-45°)$$
$$= 0.354\ \angle 45° \text{ amperes}$$

This same current must flow through all three elements of the circuit, since they are in series and the current through one must already have passed through the other two. Hence the voltage drop across the resistor (whose phase angle of course is 0°) is:

$$E = IR$$
$$E = (0.354\ \angle 45°)\ (200\ \angle 0°)$$
$$= 70.8\ \angle 45° \text{ volts}$$

The voltage drop across the inductive reactance is:

$$E = IX_L$$
$$E = (0.354\ \angle 45°)\ (100\ \angle 90°)$$
$$= 35.4\ \angle 135° \text{ volts}$$

Similarly, the voltage drop across the capacitive reactance is:

$$E = IX_C$$
$$E = (0.354\ \angle 45°)\ (300\ \angle -90°)$$
$$= 106.2\ \angle -45°$$

Note that the voltage drop across the capacitive reactance is greater than the supply voltage. This condition often occurs in a series *RLC* circuit, and is explained by the fact that the drop across the capacitive reactance is cancelled to a lesser or greater extent by the drop across the inductive reactance.

It is often desirable in a problem such as the above to check the validity of the answer by adding vectorially the voltage drops across the components of the series circuit to make sure that they add up to the supply voltage — or to use the terminology of Kirchhoff's Second Law, to make sure that the voltage drops across all elements of the circuit, including the source taken as negative, is equal to zero.

In the general case of the addition of a number of voltage vectors in series it is best to resolve the voltages into their in-phase and out-of-phase components with respect to the supply voltage. Then these components may be added directly. Hence:

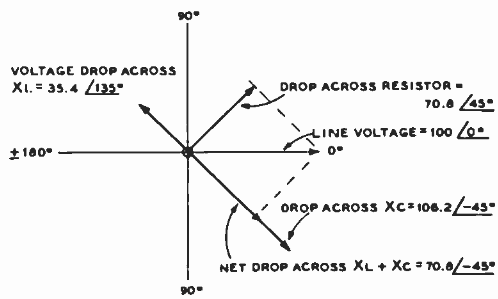


Figure 14

Graphical construction of the voltage drops associated with the series RLC circuit of figure 13.

$$E_R = 70.8\ \angle 45°$$
$$= 70.8\ (\cos 45° + j \sin 45°)$$
$$= 70.8\ (0.707 + j0.707)$$
$$= 50 + j50$$

. . . .

$$E_L = 35.4\ \angle 135°$$
$$= 35.4\ (\cos 135° + j \sin 135°)$$
$$= 35.4\ (-0.707 + j0.707)$$
$$= -25 + j25$$

. . . .

$$E_C = 106.2\ \angle 45°$$

$= 106.2\ (\cos -45° + j \sin -45°)$

$= 106.2\ (0.707\ - j0.707)$

$= 75\ - j75$

. . . .

$E_R + E_L + E_C = (50 + j50) + (-25 + j25) + (75 - j75)$

$= (50 - 25 + 75) + j(50 + 25 - 75)$

$E_R + E_L + E_C = 100 + j0$

$= 100\ \angle 0°,$

which is equal to the supply voltage.

Checking by Construction on the Complex Plane It is frequently desirable to check computations involving complex quantities by constructing vectors representing the quantities on the complex plane. Fig. 14 shows such a construction for the quantities of the problem just completed. Note that the answer to the problem may be checked by constructing a parallelogram with the voltage drop across the resistor as one side and the net voltage drop across the capacitor plus the inductor (these may be added algebraically as they are 180° out of phase) as the adjacent side. The vector sum of these two voltages, which is represented by the diagonal of the parallelogram, is equal to the supply voltage of 100 volts at zero phase angle.

Resistance and Reactance in Parallel In a series circuit, such as just discussed, the current through all the elements which go to make up the series circuit is the same. But the voltage drops across each of the components are, in general, different from one another. Conversely, in a parallel *RLC* or *RX* circuit the voltage is, obviously, the same across each of the elements. But the currents through each of the elements are usually different.

There are many ways of solving a problem involving paralleled resistance and reactance; several of these ways will be described. In general, it may be said that the impedance of a number of elements in parallel is solved using the same relations as are used for solving resistors in parallel, except that complex quantities are employed. The basic relation is:

$$\frac{1}{Z_{Total}} = \frac{1}{Z_1} + \frac{1}{Z_2} + \frac{1}{Z_3} + \ldots\ldots,$$

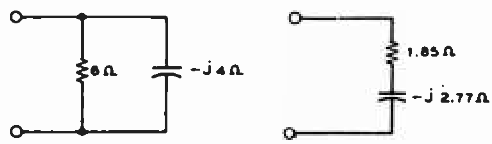
or when only two impedances are involved:

$$Z_{Total} = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

As an example, using the two-impedance relation, take the simple case, illustrated in figure 15, of a resistance of 6 ohms in parallel with a capacitive reactance of 4 ohms. To simplify the first step in the computation it is best to put the impedances in the polar form for the numerator, since multiplication is involved, and in the rectangular form for the addition in the denominator.

$$Z_{Total} = \frac{(6\ \angle 0°)\ (4\ \angle -90°)}{6 - j4}$$

$$= \frac{24\ \angle -90°}{6 - j4}$$



PARALLEL CIRCUIT   EQUIVALENT SERIES CIRCUIT

Figure 15

THE EQUIVALENT SERIES CIRCUIT

Showing a parallel RC circuit and the equivalent series RC circuit which represents the same net impedance as the parallel circuit.

Then the denominator is changed to the polar form for the division operation:

$$\theta = \tan^{-1}\frac{-4}{6} = \tan^{-1} -0.667 = -33.7°$$

$$|Z| = \frac{6}{\cos -33.7°} = \frac{6}{0.832} = 7.21 \text{ ohms}$$

$$6 - j4 = 7.21 \angle -33.7°$$

Then:

$$Z_{Total} = \frac{24\ \angle -90°}{7.21\ \angle -33.7°} = 3.33\ \angle -56.3°$$

$= 3.33\ (\cos - 56.3^\circ + j \sin - 56.3^\circ)$

$= 3.33\ [0.5548 + j\ (-0.832)]$

$= 1.85 - j\ 2.77$

Equivalent Series Circuit Through the series of operations in the previous paragraph we have converted a circuit composed of two impedances in parallel into an *equivalent series circuit* composed of impedances in series. An equivalent series circuit is one which, as far as the terminals are concerned, acts identically to the original parallel circuit; the current through the circuit and the power dissipation of the resistive elements are the same for a given voltage at the specified frequency.

We can check the equivalent series circuit of figure 15 with respect to the original circuit by assuming that one volt a-c (at the frequency where the capacitive reactance in the parallel circuit is 4 ohms) is applied to the terminals of both the series and parallel circuits.

In the parallel circuit the current through the resistor will be 1/6 ampere (0.166 amp) while the current through the capacitor will be j 1/4 ampere ($+ j$ 0.25 amp). The total current will be the sum of these two currents, or $0.166 + j\ 0.25$ amp. Adding these vectorially we obtain:

$$|I| = \sqrt{0.166^2 + 0.25^2} = \sqrt{0.09}$$
$$= 0.3 \text{ amp.}$$

The dissipation in the resistor will be $1^2/6 = 0.166$ watts.

In the case of the equivalent series circuit the current will be:

$$|I| = \frac{E}{|Z|} = \frac{1}{3.33} = 0.3 \text{ amp}$$

And the dissipation in the resistor will be:

$$W = I^2R = 0.3^2 \times 1.85$$
$$= 0.09 \times 1.85$$
$$= 0.166 \text{ watts}$$

So we see that the equivalent series circuit checks exactly with the original parallel circuit.

Parallel RLC Circuits In solving a more complicated circuit made up of more than two impedances in parallel we may elect to use either of two methods of solution. These methods are called the *admittance* method and the *assumed-voltage* method. However, the two methods are equivalent since both use the sum-of-reciprocals equation:

$$\frac{1}{Z_{Total}} = \frac{1}{Z_1} + \frac{1}{Z_2} + \frac{1}{Z_3} \ldots\ldots$$

In the admittance method we use the relation $Y = 1/Z$, where $Y = G + jB$; Y is called the *admittance*, defined above, G is the *conductance* or R/Z^2 and B is the *susceptance or* $-X/Z^2$. Then $Y_{total} = 1/Z_{total} = Y_1 + Y_2 + Y_3 \ldots$. In the assumed-voltage method we multiply both sides of the equation above by E, the assumed voltage, and add the currents, as:

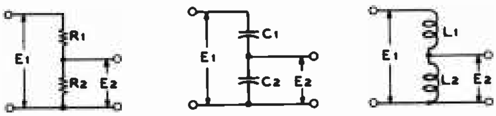
$$\frac{E}{Z_{Total}} = \frac{E}{Z_1} + \frac{E}{Z_2} + \frac{E}{Z_3} \ldots = I_{Z_1} + I_{Z_2} + I_{Z_3} \ldots$$

Then the impedance of the parallel combination may be determined from the relation:

$$Z_{Total} = E/I_{Z\,Total}$$

A-C Voltage Dividers Voltage dividers for use with alternating current are quite similar to d-c voltage dividers. However, since capacitors and inductors as well as resistors oppose the flow of a-c current, voltage dividers for alternating voltages may take any of the configurations shown in figure 16.

Since the impedances within each divider are of the same type, the output voltage is



$E_2 = E_1 \frac{R_2}{R_1+R_2}$ $E_2 = E_1 \frac{XC_2}{XC_1+XC_2}$ $E_2 = E_1 \frac{L_2}{L_1+L_2}$

$E_2 = E_1 \frac{C_1}{C_1+C_2}$

Ⓐ Ⓑ Ⓒ

Figure 16

SIMPLE A-C VOLTAGE DIVIDERS

in phase with the input voltage. By using combinations of different types of impedances, the phase angle of the output may be shifted in relation to the input phase angle at the same time the amplitude is reduced. Several dividers of this type are shown in figure 17. Note that the ratio of output voltage is equal to the ratio of the output impedance to the total divider impedance. This relationship is true only if negligible current is drawn by a load on the output terminals.

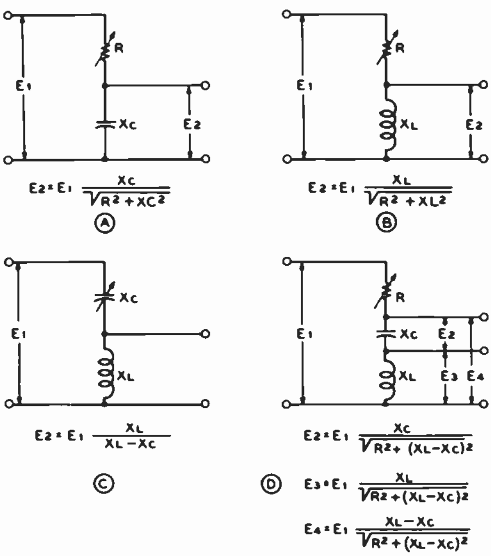


Figure 17

COMPLEX A-C VOLTAGE DIVIDERS

3-2 Resonant Circuits

A series circuit such as shown in figure 18 is said to be in *resonance* when the applied frequency is such that the capacitive reactance is exactly balanced by the inductive reactance. At this frequency the two reactances will cancel in their effects, and the impedance of the circuit will be at a minimum so that maximum current will flow. In fact, as shown in figure 19 the net impedance of a series circuit at resonance is equal to the resistance which remains in the circuit after the reactances have been cancelled.

Resonant Frequency Some resistance is always present in a circuit because it is possessed in some degree by both the inductor and the capacitor. If the frequency of the alternator E is varied from nearly zero to some high frequency, there will be one particular frequency at which the inductive reactance and capacitive reactance will be equal. This is known as the *resonant frequency*, and in a series circuit it is the frequency at which the circuit current will be a maximum. Such series-resonant circuits are chiefly used when it is desirable to allow a certain frequency to pass through the circuit (low impedance to this frequency), while at the same time the circuit is made to offer considerable opposition to currents of other frequencies.

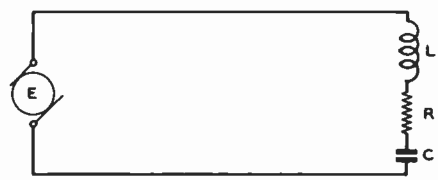


Figure 18

SERIES-RESONANT CIRCUIT

If the values of inductance and capacitance both are fixed, there will be only one resonant frequency.

If both the inductance and capacitance are made variable, the circuit may then be changed or *tuned*, so that a number of combinations of inductance and capacitance can resonate at the same frequency. This can be more easily understood when one considers that inductive reactance and capacitive reactance change in opposite directions as the frequency is varied. For example, if the frequency were to remain constant and the values of inductance and capacitance were then changed, the following combinations would have equal *reactance:*

Frequency is constant at 60 Hz.

L is expressed in henrys.

C is expressed in microfarads (.000001 farad.)

L	X_L	C	X_C
.265	100	26.5	100
2.65	1000	2.65	1000
26.5	10,000	.265	10,000
265.00	100,000	.0265	100,000
2,650.00	1,000,000	.00265	1,000,000

Frequency of Resonance From the formula for resonance, $2\pi fL = 1/2\pi fC$, the resonant frequency is determined:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

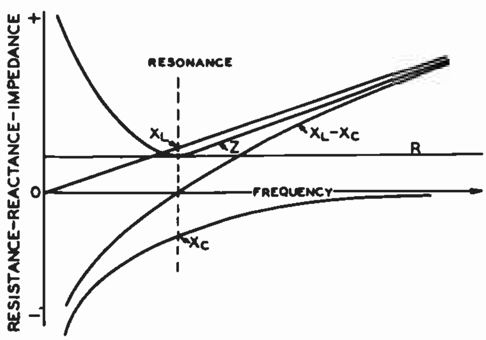


Figure 19

IMPEDANCE OF A SERIES-RESONANT CIRCUIT

Showing the variation in reactance of the separate elements and in the net impedance of a series resonant circuit (such as figure 18) with changing frequency. The vertical line is drawn at the point of resonance ($X_L - X_C = 0$) in the series circuit.

where,

f equals frequency in hertz,
L equals inductance in henrys,
C equals capacitance in farads.

It is more convenient to express L and C in smaller units, especially in making radio-frequency calculations; f can also be expressed in MHz or kHz. A very useful group of such formulas is:

$$f^2 = \frac{25,330}{LC} \text{ or } L = \frac{25,330}{f^2C} \text{ or } C = \frac{25,330}{f^2L}$$

where,

L equals inductance in microhenrys,
f equals frequency in MHz,
C equals capacitance in picofarads.

Impedance of Series Resonant Circuits The impedance across the terminals of a series-resonant circuit (figure 18) is:

$$Z = \sqrt{r^2 + (X_L - X_C)^2}$$

where,

Z equals impedance in ohms,
r equals resistance in ohms,
X_C equals capacitive reactance in ohms,
X_L equals inductive reactance in ohms.

From this equation, it can be seen that the impedance is equal to the vector sum of the circuit resistance and the *difference* between the two reactances. Since at the resonant frequency X_L equals X_C, the difference between them (figure 19) is zero, so that at resonance the impedance is simply equal to the resistance of the circuit; therefore, because the resistance of most normal radio-frequency circuits is of a very low order, the impedance is also low.

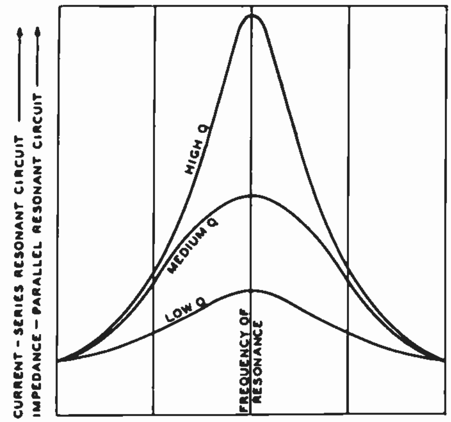


Figure 20

RESONANCE CURVE

Showing the increase in impedance at resonance for a parallel-resonant circuit, and similarly, the increase in current at resonance for a series-resonant circuit. The sharpness of resonance is determined by the Q of the circuit, as illustrated by a comparison between the three curves.

At frequencies higher and lower than the resonant frequency, the difference between the reactances will be a definite quantity and will add with the resistance to make the impedance higher and higher as the circuit is tuned off the resonant frequency.

If X_C should be greater than X_L, then the term $(X_L - X_C)$ will give a negative number. However, when the difference is squared the product is always positive. This means that the smaller reactance is subtracted from the larger, regardless of whether it be capacitive or inductive, and the difference is squared.

Current and Voltage in Series-Resonant Circuits Formulas for calculating currents and voltages in a series-resonant circuit are similar to those of Ohm's Law.

$$I = \frac{E}{Z} \quad E = IZ$$

The complete equations are:

$$I = \frac{E}{\sqrt{r^2 + (X_L - X_C)^2}}$$

$$E = I\sqrt{r^2 + (X_L - X_C)^2}$$

Inspection of the above formulas will show the following to apply to series-resonant circuits: When the impedance is low, the current will be high; conversely, when the impedance is high, the current will be low.

Since it is known that the impedance will be very low at the resonant frequency, it follows that the current will be a maximum at this point. If a graph is plotted of the current against the frequency either side of resonance, the resultant curve becomes what is known as a *resonance curve*. Such a curve is shown in figure 20, the frequency being plotted against *current* in the series-resonant circuit.

Several factors will have an effect on the shape of this resonance curve, of which resistance and L-to-C ratio are the important considerations. The lower curves in figure 20 show the effect of adding increasing values of resistance to the circuit. It will be seen that the peaks become less and less prominent as the resistance is increased; thus, it can be said that the *selectivity* of the circuit is thereby *decreased*. Selectivity in this case can be defined as the ability of a circuit to discriminate against frequencies adjacent to (both above and below) the resonant frequency.

Voltage Across Coil and Capacitor in Series Circuit Because the a-c or r-f voltage across a coil and capacitor is proportional to the reactance (for a given current), the actual voltages across the coil and across the capacitor may be many times greater than the *terminal* voltage of the circuit. At resonance, the voltage across the coil (or the capacitor) is Q times the applied voltage. Since the Q (or *merit factor*) of a series circuit can be in the neighborhood of 100 or more, the voltage across the capacitor, for example, may be high enough to cause flashover, even though the applied voltage is of a value considerably below that at which the capacitor is rated.

Circuit Q — Sharpness of Resonance An extremely important property of a capacitor or an inductor is its *factor-of-merit*, more generally called its Q. It is this factor, Q, which primarily determines the sharpness of resonance of a tuned circuit. This factor can be expressed as the ratio of the reactance to the resistance, as follows:

$$Q = \frac{2\pi f L}{R}$$

where,

R equals total resistance.

Skin Effect The actual resistance in a wire or an inductor can be far greater than the d-c value when the coil is used in a radio-frequency circuit; this is because the current does not travel through the entire cross section of the conductor, but has a tendency to travel closer and closer to the surface of the wire as the frequency is increased. This is known as the *skin effect*.

The actual current carrying portion of the wire is decreased as a result of the skin effect so that the ratio of a-c to d-c resistance of the wire, called the *resistance ratio*, is increased. The resistance ratio of wires to be used at frequencies below about 500 kHz may be materially reduced through the use of *litz* wire. Litz wire, of the type commonly used to wind the coils of 455-kHz i-f transformers, may consist of 3 to 10 strands of insulated wire, about No. 40 in size, with the individual strands connected together only at the ends of the coils.

Variation of Q with Frequency Examination of the equation for determining Q might give rise to the thought that even though the resistance of an inductor increases with frequency, the inductive reactance does likewise, so that the Q might be a constant. Actually, however, it works out in practice that the Q of an inductor will reach a relatively broad maximum at some particular frequency. Hence, coils normally are designed in such a manner that the peak in their curve of Q versus frequency will occur at the normal operating frequency of the coil in the circuit for which it is designed.

The Q of a capacitor ordinarily is much higher than that of the best coil. Therefore, it usually is the merit of the coil that limits the over-all Q of the circuit.

At audio frequencies the core losses in an iron-core inductor greatly reduce the Q from the value that would be obtained simply by dividing the reactance by the resistance. Obviously the core losses also represent circuit resistance, just as though the loss occurred in the wire itself.

Parallel Resonance In radio circuits, parallel resonance (more correctly termed *antiresonance*) is more frequently encountered than series resonance; in fact, it is the basic foundation of receiver and transmitter circuit operation. A circuit is shown in figure 21.

The "Tank" Circuit In this circuit, as contrasted with a circuit for series resonance, *L* (inductance) and *C* (capacitance) are connected in *parallel*, yet the *combination* can be considered to be in series with the remainder of the circuit. This combination of *L* and *C*, in conjunction with *R*, the resistance which is principally included in *L*, is sometimes called a *tank circuit* because it effectively functions as a

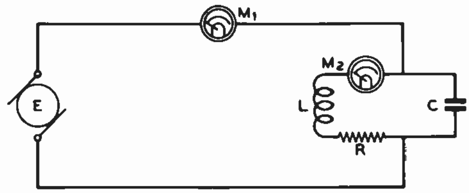


Figure 21

PARALLEL-RESONANT CIRCUIT

The inductance L and capacitance C comprise the reactive elements of the parallel-resonant (antiresonant) tank circuit, and the resistance R indicates the sum of the r-f resistance of the coil and capacitor, plus the resistance coupled into the circuit from the external load. In most cases the tuning capacitor has much lower r-f resistance than the coil and can therefore be ignored in comparison with the coil resistance and the coupled-in resistance. The instrument M_1 indicates the "line current" which keeps the circuit in a state of oscillation—this current is the same as the fundamental component of the plate current of a class-C amplifier which might be feeding the tank circuit. The instrument M_2 indicates the "tank current" which is equal to the line current multiplied by the operating Q of the tank circuit.

storage tank when incorporated in vacuum-tube circuits.

Contrasted with series resonance, there are two kinds of current which must be considered in a parallel-resonant circuit: (1) the *line current*, as read on the indicating meter M_1, (2) the *circulating current* which flows within the parallel *LCR* portion of the circuit. See figure 21.

At the resonant frequency, the line current (as read on the meter M_1) will drop to a very low value although the circulating current in the *LC* circuit may be quite large. It is interesting to note that the parallel-resonant circuit acts in a distinctly opposite manner to that of a series-resonant circuit, in which the current is at a maximum and the impedance is minimum at resonance. It is for this reason that in a parallel-resonant circuit the principal consideration is one of impedance rather than current. It is also significant that the *impedance* curve for *parallel* circuits is very nearly indentical to that of the *current* curve for *series* resonance. The impedance at resonance is expressed as:

$$Z = \frac{(2\pi f L)^2}{R}$$

where,

Z equals impedance in ohms,
L equals inductance in henrys,
f equals frequency in hertz,
R equals resistance in ohms.

Or, impedance can be expressed as a function of Q as:

$$Z = 2\pi fLQ$$

showing that the impedance of a circuit is directly proportional to its effective Q at resonance.

The curves illustrated in figure 20 can be applied to parallel resonance. Reference to the curve will show that the effect of adding resistance to the circuit will result in both a broadening out and lowering of the peak of the curve. Since the voltage of the circuit is directly proportional to the impedance, and since it is this voltage that is applied to the grid of the vacuum tube in a detector or amplifier circuit, the impedance curve must have a sharp peak in order for the circuit to be *selective*. If the curve is broad-topped in shape, both the desired signal and the interfering signals at close proximity to resonance will give nearly equal voltages on the grid of the tube, and the circuit will then be *nonselective* ; that is , it will tune broadly.

Effect of L/C Ratio in Parallel Circuits In order that the highest possible voltage can be developed across a parallel-resonant circuit, the impedance of this circuit must be very high. The impedance will be greater with conventional coils of limited Q when the ratio of inductance to capacitance is great, that is, when L is large as compared with C. When the resistance of the circuit is very low, X_L will equal X_C at maximum impedance. There are innumerable ratios of L and C that will have *equal* reactance, at a given resonant frequency, exactly as in the case in a series-resonant circuit.

In practice, where a certain value of inductance is tuned by a variable capacitance over a fairly wide range in frequency, the L/C ratio will be small at the lowest-frequency end and large at the high-frequency end. The circuit, therefore, will have unequal gain and selectivity at the two ends of the band of frequencies which is being tuned. Increasing the Q of the circuit (lowering the resistance) will obviously increase *both* the selectivity and gain.

Circulating Tank Current at Resonance The Q of a circuit has a definite bearing on the circulating tank current at resonance. This tank current is very nearly the value of the line current multiplied by the effective circuit Q. For example: an r-f line current of 0.050 ampere, with a circuit Q of 100, will give a circulating tank current of approximately 5 amperes. From this it can be seen that both the inductor and the connecting wires in a circuit with a high Q must be of very low resistance, particularly in the case of high-power transmitters, if heat losses are to be held to a minimum.

Because the voltage across the tank at resonance is determined by the Q, it is possible to develop very high peak voltages across a high-Q tank with but little line current.

Effect of Coupling on Impedance If a parallel-resonant circuit is coupled to another circuit, such as an antenna output circuit, the impedance and the effective Q of the parallel circuit is decreased as the coupling becomes closer. The effect of closer (tighter) coupling is the same as though an actual resistance were added in series with the parallel tank circuit. The resistance thus coupled into the tank circuit can be considered as being *reflected* from the output or load circuit to the driver circuit.

The behavior of *coupled circuits* depends largely on the amount of coupling, as shown in figure 22. The coupled currrent in the secondary circuit is small, varying with frequency, being maximum at the resonant frequency of the circuit. As the coupling is increased between the two circuits, the secondary resonance curve becomes broader and the resonant amplitude increases, until the reflected resistance is equal to the primary resistance. This point is called the *critical coupling point*. With greater coupling, the secondary resonance curve becomes broader and develops double resonance humps, which become more pronounced and farther apart in frequency as the coupling between the two circuits is increased.

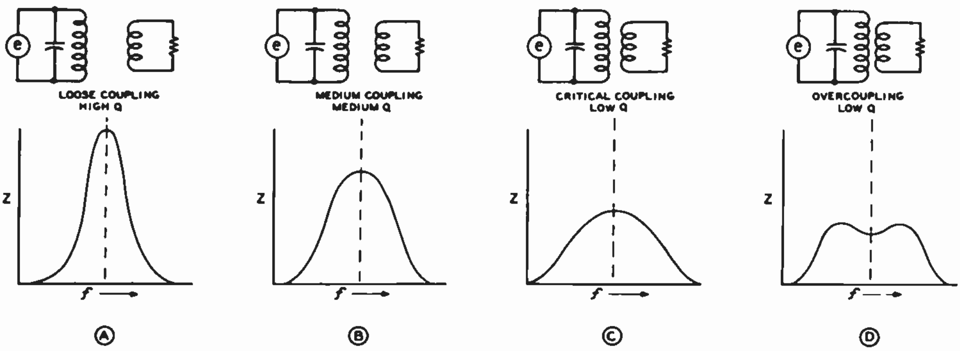


Figure 22
EFFECT OF COUPLING ON CIRCUIT IMPEDANCE AND Q

Tank-Circuit Flywheel Effect When the plate circuit of a class-B or class-C operated tube is connected to a parallel-resonant circuit tuned to the same frequency as the exciting voltage for the amplifier, the plate current serves to maintain this L/C circuit in a state of oscillation.

The plate current is supplied in short pulses which do not begin to resemble a sine wave, even though the grid may be excited by a sine-wave voltage. These spurts of plate current are converted into a sine wave in the plate tank circuit by virtue of the Q or *flywheel effect* of the tank.

If a tank did not have some resistance losses, it would, when given a "kick" with a single pulse, continue to oscillate indefinitely. With a moderate amount of resistance or "friction" in the circuit the tank will still have inertia, and continue to oscillate with decreasing amplitude for a time after being given a "kick." With such a circuit, almost pure sine-wave voltage will be developed across the tank circuit even though power is supplied to the tank in short pulses or spurts, so long as the spurts are evenly spaced with respect to time and have a frequency that is the same as the resonant frequency of the tank.

Another way to visualize the action of the tank is to recall that a resonant tank with moderate Q will discriminate strongly against harmonics of the resonant frequency. The distorted plate current pulse in a class-C amplifier contains not only the fundamental frequency (that of the grid excitation voltage) but also higher harmonics. As the tank offers low impedance to the harmonics and high impedance to the fundamental (being resonant to the latter), only the fundamental — a sine-wave voltage — appears across the tank circuit in substantial magnitude.

Loaded and Unloaded Q Confusion sometimes exists as to the relationship between the unloaded and the loaded Q of the tank circuit in the plate of an r-f power amplifier. In the normal case the loaded Q of the tank circuit is determined by such factors as the operating conditions of the amplifier, bandwidth of the signal to be emitted, permissible level of harmonic radiation, and such factors. The normal value of *loaded* Q for an r-f amplifier used for communications service is from perhaps 6 to 20. The *unloaded* Q of the tank circuit determines the efficiency of the output circuit and is determined by the losses in the tank coil, its leads and plugs and jacks if any, and by the losses in the tank capacitor which ordinarily are very low. The unloaded Q of a good quality large diameter tank coil in the high-frequency range may be as high as 500 to 800, and values greater than 300 are quite common.

Tank-Circuit Efficiency Since the unloaded Q of a tank circuit is determined by the minimum losses in the tank, while the loaded Q is determined by useful loading of the tank circuit from the external load in addition to the internal losses in the tank circuit, the relationship between the two Q values determines the operating efficiency of the tank circuit. Expressed in the form of an equation, the loaded efficiency of a tank circuit is:

$$\text{Tank efficiency} = \left(1 - \frac{Q_l}{Q_u}\right) \times 100$$

where,

Q_u equals unloaded Q of the tank circuit,
Q_l equals loaded Q of the tank circuit.

As an example, if the unloaded Q of the tank circuit for a class-C r-f power amplifier is 400, and the external load is coupled to the tank circuit by an amount such that the loaded Q is 20, the tank-circuit efficiency will be: eff. = (1 − 20/400) × 100, or (1 − 0.05) × 100, or 95 per cent. Hence 5 per cent of the power output of the class-C amplifier will be lost as heat in the tank circuit and the remaining 95 per cent will be delivered to the load.

3-3 Nonsinusoidal Waves and Transients

Pure sine waves, discussed previously, are basic wave shapes. Waves of many different and complex shapes are used in electronics, particularly square waves, sawtooth waves, and peaked waves.

Wave Composition Any periodic wave (one that repeats itself in definite time intervals) is composed of sine waves of different frequencies and amplitudes, added together. The sine wave which has the same frequency as the complex, periodic wave is called the *fundamental*. The frequencies higher than the fundamental are

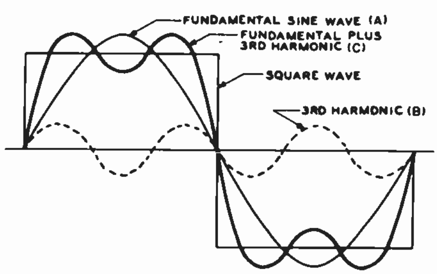


Figure 23

COMPOSITE WAVE—FUNDAMENTAL PLUS THIRD HARMONIC

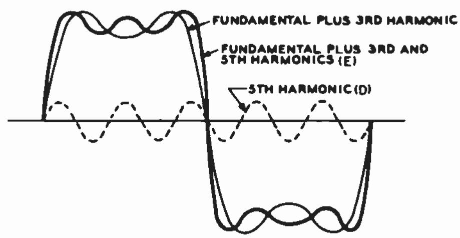


Figure 24

THIRD-HARMONIC WAVE PLUS FIFTH HARMONIC

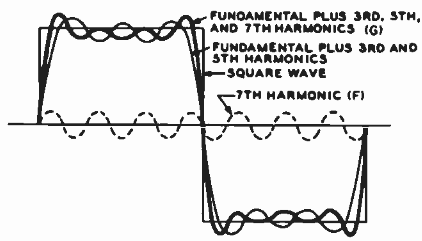


Figure 25

RESULTANT WAVE, COMPOSED OF FUNDAMENTAL, THIRD, FIFTH, AND SEVENTH HARMONICS

called *harmonics*, and are always a whole number of times higher than the fundamental. For example, the frequency twice as high as the fundamental is called the *second harmonic*.

The Square Wave Figure 23 compares a square wave with a sine wave (A) of the same frequency. If another sine wave (B) of smaller amplitude, but three times the frequency of A, called the third harmonic, is added to A, the resultant wave (C) more nearly approaches the desired square wave.

This resultant curve (figure 24) is added to a fifth-harmonic curve (D), and the sides of the resulting curve (E) are steeper than before. This new curve is shown in figure 25 after a 7th-harmonic component has been added to it, making the sides of the composite wave even steeper. Addition of more higher odd harmonics will bring the resultant wave nearer and nearer to the desired square-wave shape. The square wave will be achieved if an infinite number of odd harmonics are added to the original sine wave.

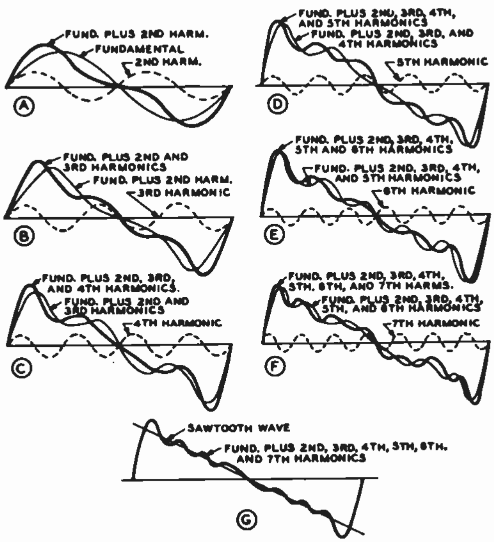


Figure 26

COMPOSITION OF A SAWTOOTH WAVE

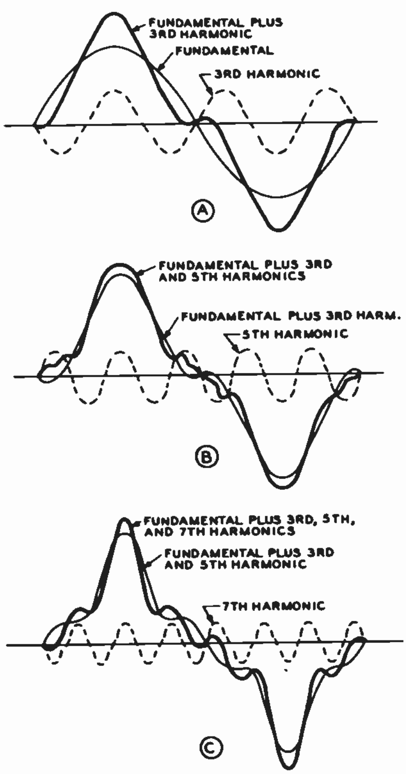


Figure 27

COMPOSITION OF A PEAKED WAVE

The Sawtooth Wave In the same fashion, a *sawtooth wave* is made up of different sine waves (figure 26). The addition of all harmonics, odd and even, produces the sawtooth waveform.

The Peaked Wave Figure 27 shows the composition of a *peaked wave*. Note how the addition of each successive harmonic makes the peak of the resultant higher, and the sides steeper.

Other Waveforms The three preceding examples show how a complex periodic wave is composed of a fundamental wave and different harmonics. The shape of the resultant wave depends on the harmonics that are added, their relative amplitudes, and relative phase relationships. In general, the steeper the sides of the waveform, the more harmonics it contains.

A-C Transient Circuits If an a-c voltage is substituted for the d-c input voltage in the *RC* transient circuits discussed in Chapter 2, the same principles may be applied in the analysis of the transient behavior. An *RC* coupling circuit is designed to have a long time constant with respect to the lowest frequency it must pass. Such a circuit is shown in figure 28. If a nonsinusoidal voltage is to be passed unchanged through the coupling circuit, the time constant must be long with respect to the period of the lowest frequency contained in the voltage wave.

RC Differentiator and Integrator An *RC* voltage divider that is designed to distort the input waveform is known as a *differentiator* or *integrator*, depending on the locations of the output taps. The output from a differentiator is taken across the resistance, while the output from an integrator is taken across the capacitor. Such circuits will change the shape of any complex a-c waveform that is impressed on them. This distortion is a function of the value of the time constant of the circuit as compared to the period of the waveform. Neither a differentiator nor an integrator can change the shape of a pure sine wave, they will merely shift the phase of the wave (figure 29). The differentiator output is a sine wave leading the input wave, and the integrator output is a sine wave which lags the input wave. The sum of the two outputs at any instant equals the instantaneous input voltage.

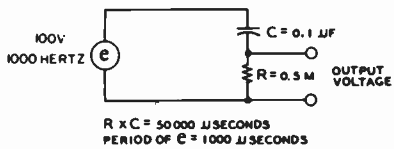


Figure 28

RC COUPLING CIRCUIT WITH LONG TIME CONSTANT

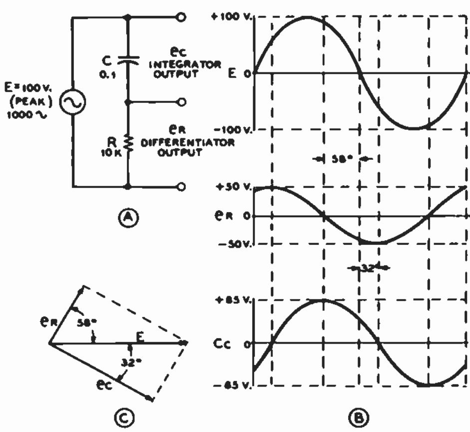


Figure 29

RC DIFFERENTIATOR AND INTEGRATOR ACTION ON A SINE WAVE

Square-Wave Input If a square-wave voltage is impressed on the circuit of figure 30, a square-wave voltage output may be obtained across the integrating capacitor if the time constant of the circuit allows the capacitor to become fully charged. In this particular case, the capacitor never fully charges, and as a result the output of the integrator has a smaller amplitude than the input. The differentiator output has a maximum value greater than the input amplitude, since the voltage left on the capacitor from the previous half wave will add to the input voltage. Such a circuit, when used as a differentiator, is often called a *peaker*. Peaks of twice the input amplitude may be produced.

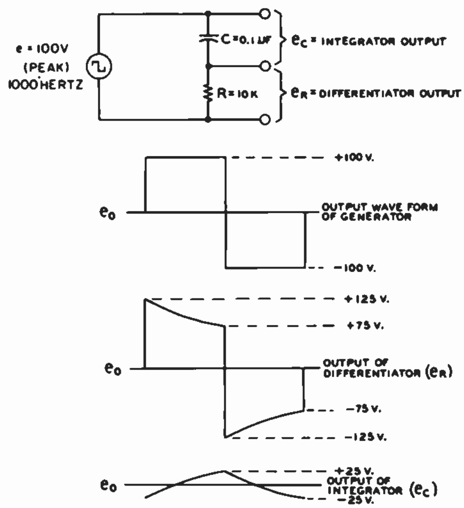


Figure 30

RC DIFFERENTIATOR AND INTEGRATOR ACTION ON A SQUARE WAVE

Sawtooth-Wave Input If a back-to-back sawtooth voltage is applied to an *RC* circuit having a time constant one-sixth the period of the input voltage, the result is shown in figure 31. The capacitor voltage will closely follow the input voltage, if the time constant is short, and the integrator output closely resembles the input. The amplitude is slightly reduced and there is a slight phase lag. Since the voltage across the capacitor is increasing at a constant rate, the charging and discharging current is constant. The output voltage of the differentiator, therefore, is constant during each half of the sawtooth input.

Miscellaneous Inputs Various voltage waveforms other than those represented here may be applied to short-time-constant *RC* circuits for the purpose of producing across the resistor an output voltage with an amplitude *proportional to the rate of change* of the input signal. The shorter the *RC* time constant is made with respect to the period of the input wave, the more nearly the voltage across the capacitor conforms to the input voltage. Thus, the differentiator output becomes of particular importance in very short-time-constant *RC*

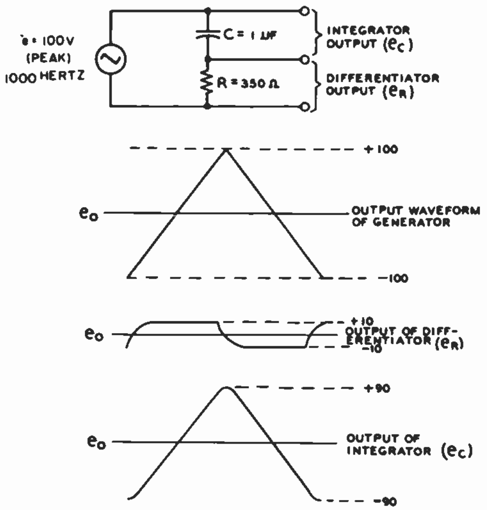


Figure 31

RC DIFFERENTIATOR AND INTEGRATOR ACTION ON A SAWTOOTH WAVE

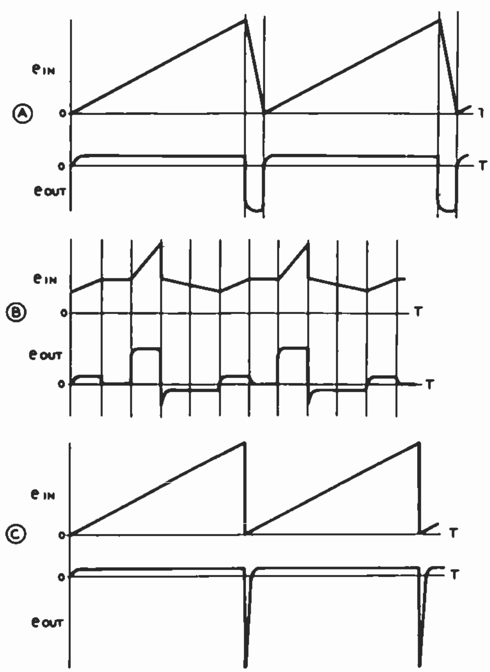


Figure 32

Differentiator outputs of short-time-constant RC circuits for various input voltage waveshapes. The output voltage is proportional to the rate of change of the input voltage.

circuits. Differentiator outputs for various types of input waves are shown in figure 32.

Square-Wave Test for Audio Equipment The application of a square-wave input signal to audio equipment, and the observation of the reproduced output signal on an oscilloscope will provide a quick and accurate check of the over-all operation of audio equipment.

Low-frequency and high-frequency response, as well as transient response can be examined easily.

If the amplifier is deficient in low-frequency response, the flat top of the square wave will be canted, as in figure 33. If the high-frequency response is inferior, the rise time of the output wave will be retarded (figure 34).

An amplifier with a limited high- and low-frequency response will turn the square wave into the approximation of a sawtooth wave (figure 35).

3-4 Transformers

When two coils are placed in such inductive relation to each other that the lines of force from one cut across the turns of the other inducing a current, the combination can be called a *transformer*. The name is derived from the fact that energy is transformed from one winding to another. The inductance in which the original flux is produced is called the *primary;* the inductance which *receives* the induced current is called the *secondary*. In a radio-receiver power transformer, for example, the coil through which the 120-volt a.c. passes is the *primary*, and the coil from which a higher or lower voltage than the a-c line potential is obtained is the *secondary*.

Transformers can have either air or magnetic cores, depending on the frequencies at which they are to be operated. The reader should thoroughly impress on his mind the fact that current can be transferred from one circuit to another *only* if the primary current is changing or alternating. From this it can be seen that a power transformer cannot possibly function as such when the primary is supplied with nonpulsating d-c.

A power transformer usually has a magnetic core which consists of laminations of

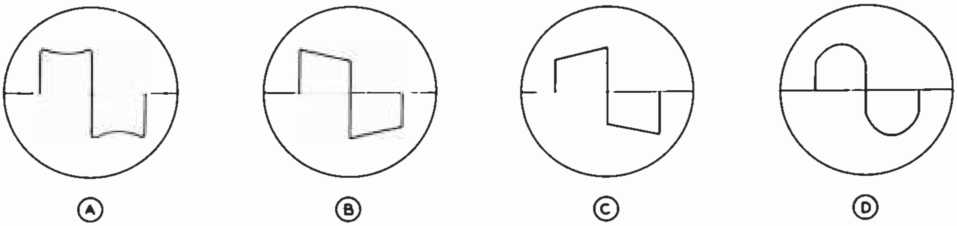


Figure 33

Amplifier deficient in low-frequency response will distort square wave applied to the input circuit, as shown. A 60-Hz square wave may be used.
A: Drop in gain at low frequencies
B: Leading phase shift at low frequencies
C: Lagging phase shift at low frequencies
D: Accentuated low-frequency gain

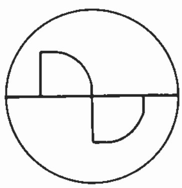


Figure 34

Output waveshape of amplifier having deficiency in high-frequency response. Tested with 10-kHz square wave.

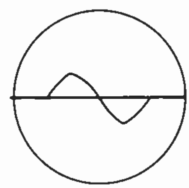


Figure 35

Output waveshape of amplifier having limited low-frequency and high-frequency response. Tested with 1 kHz square wave.

iron, built up into a square or rectangular form, with a center opening or window. The secondary windings may be several in number, each perhaps delivering a different voltage. The secondary voltages will be proportional to the turns ratio and the primary voltage.

Types of Transformers Transformers are used in alternating-current circuits to transfer power at one voltage and impedance to another circuit at another voltage and impedance. There are three main classifications of transformers: those made for use in power-frequency circuits, those made for audio-frequency applications, and those made for radio frequencies.

The Transformation Ratio In a perfect transformer all the magnetic flux lines produced by the primary winding link crosses every turn of the secondary winding (figure 36). For such a transformer, the ratio of the primary and secondary voltages is exactly the same as the ratio of the number of turns in the two windings:

$$\frac{N_P}{N_S} = \frac{E_P}{E_S}$$

where,

N_P equals number of turns in the primary,
N_S equals number of turns in the secondary,
E_P equals voltage across the primary,
E_S equals voltage across the secondary.

In practice, the transformation ratio of a transformer is somewhat less than the turns ratio, since unity coupling does not exist between the primary and secondary windings.

Ampere Turns (NI) The current that flows in the secondary winding as a result of the induced voltage must produce a flux which exactly equals the primary flux. The magnetizing force of a coil is expressed as the product of the number of turns in the coil times the current flowing in it:

$$N_P \times I_P = N_S \times I_S, \text{ or } \frac{N_P}{N_S} = \frac{I_S}{I_P}$$

where,

I_P equals primary current,
I_S equals secondary current.

It can be seen from this expression that when the voltage is stepped up, the current is stepped down, and vice versa.

Leakage Reactance Since unity coupling does not exist in a practical transformer, part of the flux passing from the primary circuit to the secondary circuit follows a magnetic circuit acted on by the primary only. The same is true of the secondary flux. These leakage fluxes cause *leakage reactance* in the transformer, and tend to cause the transformer to have poor voltage regulation. To reduce such leakage reactance, the primary and secondary windings should be in close proximity to each other. The more expensive transformers have interleaved windings to reduce inherent leakage reactance.

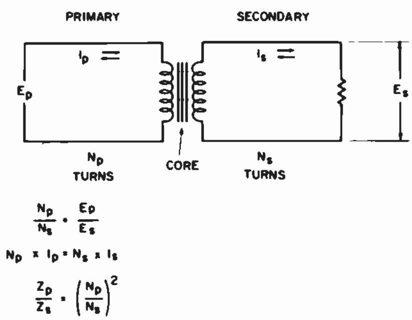


Figure 36

THE LOW-FREQUENCY TRANSFORMER

Power is transformed from the primary to the secondary winding by means of the varying magnetic field. The voltage induced in the secondary for a given primary voltage is proportional to the ratio of secondary to primary turns. The impedance transformation is proportional to the square of the primary to secondary turns ratio.

Impedance Transformation In the ideal transformer, the impedance of the secondary load is reflected back into the primary winding in the following relationship:

$$Z_P = N^2 Z_S, \text{ or } N = \sqrt{Z_P/Z_S}$$

where,

Z_P equals reflected primary impedance,

N equals turns ratio of transformer,

Z_S equals impedance of secondary load.

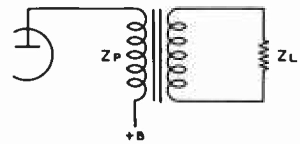


Figure 37

IMPEDANCE-MATCHING TRANSFORMER

The reflected impedance Z_p varies directly in proportion to the secondary load Z_L and directly in proportion to the square of the primary-to-secondary turns ratio.

Thus any specific load connected to the secondary terminals of the transformer will be transformed to a different specific value appearing across the primary terminals of the transformer. By the proper choice of turns ratio, any reasonable value of secondary load impedance may be "reflected" into the primary winding of the transformer to produce the desired transformer primary impedance. The phase angle of the primary "reflected" impedance will be the same as the phase angle of the load impedance. A capacitive secondary load will be presented to the transformer source as a capacitance, a resistive load will present a resistive "reflection" to the primary source. Thus the primary source "sees" a transformer load entirely dependent on the secondary load impedance and the turns ratio of the transformer (figure 37).

The Auto-transformer The type of transformer in figure 38, when wound with heavy wire over an iron core, is a common device in primary power circuits for the purpose of increasing or decreasing the line voltage. In effect, it is merely a continuous winding with taps taken at various points along the winding, the input voltage being applied to the bottom and also to one tap on the winding. If the output is taken from this same tap, the voltage ratio will be 1 to 1; i.e., the input voltage will be the same as the output voltage. On the other hand, if the output tap is moved down toward the common terminal, there will be a stepdown in the turns ratio with a consequent stepdown in voltage. The initial setting of the middle input tap is chosen so that the number of turns will have sufficient

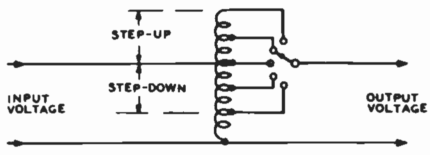


Figure 38

THE AUTOTRANSFORMER

Schematic diagram of an autotransformer showing the method of connecting it to the line and to the load. When only a small amount of step up or step down is required, the autotransformer may be much smaller physically than would be a transformer with a separate secondary winding. Continuously variable autotransformers (Variac and Powerstat) are widely used commercially.

reactance to keep the no-load primary current at a reasonable low value.

3-5 Electric Filters

There are many applications where it is desirable to pass a d-c component without passing a superimposed a-c component, or to pass all frequencies above or below a certain frequency while rejecting or attenuating all others, or to pass only a certain band or bands of frequencies while attenuating all others.

All of these things can be done by suitable combinations of inductance, capacitance, and resistance. However, as whole books have been devoted to nothing but *electric filters*, it can be appreciated that it is possible only to touch on them superficially in a general-coverage book.

Filter Operation A filter acts by virtue of its property of offering very high impedance to the undesired frequencies, while offering but little impedance to the desired frequencies. This will also apply to d. c. with a superimposed a-c component, as d. c. can be considered as an alternating current of zero frequency so far as filter discussion goes.

Basic Filters Filters are divided into four classes, descriptive of the frequency bands which they are designed to transmit: high-pass, low-pass, bandpass, and band-elimination. Each of these classes of filters is made up of elementary filter sections called *L sections* which consist of a series element (Z_A) and a parallel element (Z_B) as illustrated in figure 39. A definite number of *L* sections may be combined into basic filter sections, called *T networks* or *π networks*, also shown in figure 39. Both the *T* and *π* networks may be divided in two to form *half-sections*.

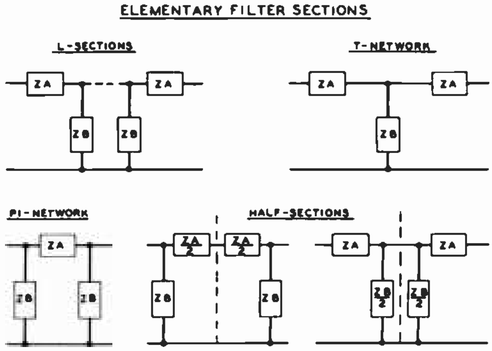


Figure 39

Complex filters may be made up from these basic filter sections.

Filter Sections The most common filter section is one in which the two impedances Z_A and Z_B are so related that their arithmetical product is a constant: $Z_A \times Z_B = k^2$ at all frequencies. This type of filter section is called a *constant-k section.*

A section having a sharper cutoff frequency than a constant-*k* section, but less attenuation at frequencies far removed from cutoff is the *m-derived section*, so called because the shunt or series element is resonated with a reactance of the opposite sign. If the complementary reactance is added to the series arm, the section is said to be *shunt derived;* if added to the shunt arm, *series derived*. Each impedance of the *m*-derived section is related to a corresponding impedance in the constant-*k* section by some factor which is a function of the constant *m*. In turn, *m* is a function of the ratio between the cutoff frequency and the frequency of infinite attenuation, and will have some value between zero and one. As the value of *m* approaches zero, the sharpness of cutoff increases, but the less will be the attenuation at several times cutoff frequency. A value of 0.6 may be used for *m* in most applications. The "notch" frequency is determined by the resonant frequency of

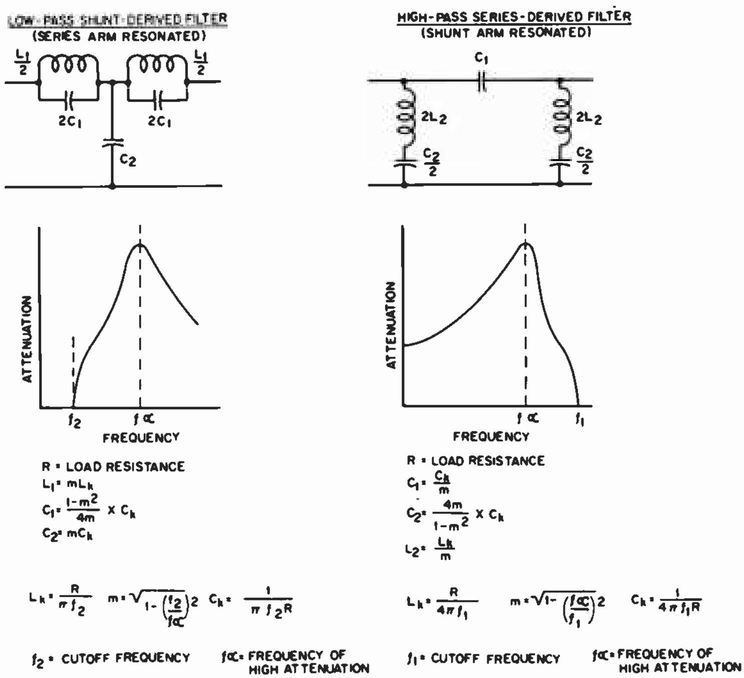


Figure 40

TYPICAL LOW-PASS AND HIGH-PASS FILTERS, ILLUSTRATING SHUNT AND SERIES DERIVATIONS

the tuned filter element. The amount of attenuation obtained at the "notch" when a derived section is used is determined by the effective Q of the resonant arm (figure 40).

Filter Assembly Constant-*k* sections and *m*-derived sections may be cascaded to obtain the combined characteristics of sharp cutoff and good remote frequency attenuation. Such a filter is known as a *composite* filter. The amount of attenuation will depend on the number of filter sections used, and the shape of the transmission curve depends on the type of filter sections used. All filters have some *insertion loss.* This attenuation is usually uniform to all frequencies within the passband. The insertion loss varies with the type of filter, the Q of the components, and the type of termination employed.

Filter Design Techniques Electric wave filters have long been used in some amateur stations in the audio channel to reduce the transmission of unwanted high frequencies and hence to reduce the bandwidth occupied by a radiophone signal. The effectiveness of a properly designed and properly used filter circuit in reducing QRM and sideband splatter should not be underestimated.

In recent years, high-frequency filters have become commonplace in TVI reduction. High-pass type filters are placed before the input stage of television receivers to reject the fundamental signal of low-frequency transmitters. Low-pass filters are used in the output circuits of low-frequency transmitters to prevent harmonics of the transmitter from being radiated in the television channels.

The chart of figure 41 gives design data and procedure on the π section type of filter. The *m*-derived sections with an *m* of 0.6 will be found to be most satisfactory as the input section (or half-section) of the usual filter since the input impedance of such a section is most constant over the passband of the filter section.

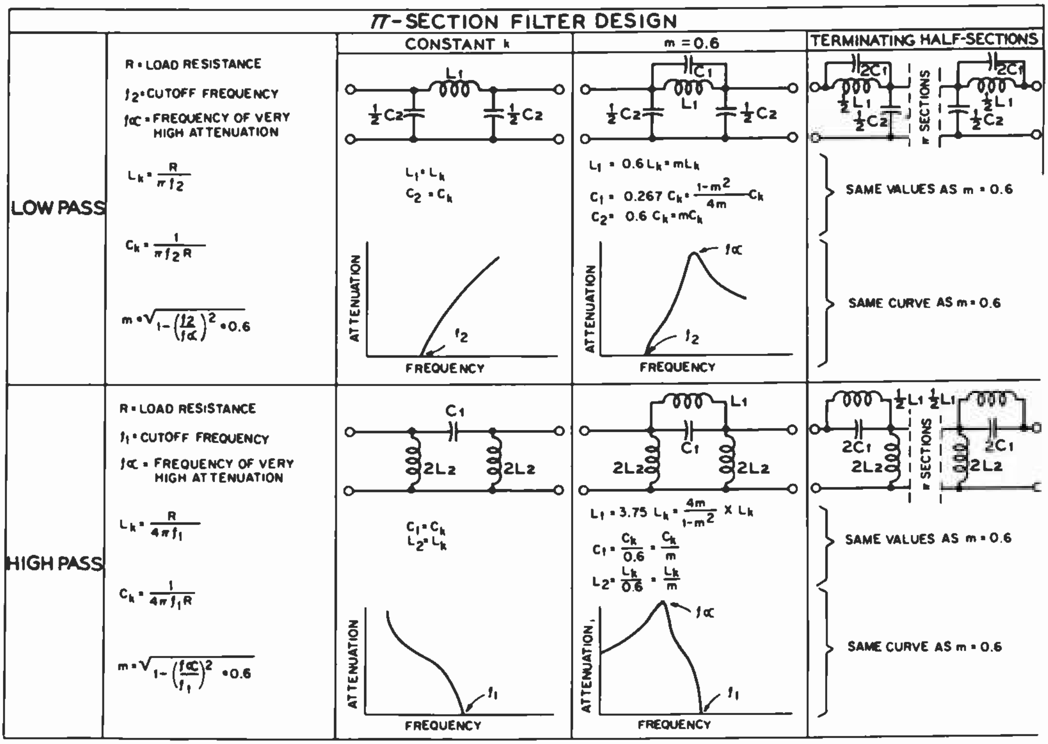


Figure 41

Through the use of the curves and equations which accompany the diagrams in the illustration above it is possible to determine the correct values of inductance and capacitance for the usual types of pi-section filters.

Simple filters may use either L, T, or π sections. Since the π section is the more commonly used type, figure 41 gives design data and characteristics for this type of filter.

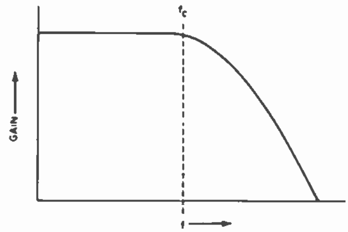
3-6 Low-Pass Filter Nomographs

The Constant-K Filter The low-pass constant-k filter has a passband from d.c. to the cutoff frequency (f_c). Beyond this frequency, the signal is attenuated as shown in figure 42.

Pi and T configurations for constant-k filters are shown in the illustration, with appropriate design formulas. The nomograph (#1 of figure 43) provides a graphical solution to these equations. The values of L and C can be determined by aligning a straightedge from f_c on the left-hand scale to $R(L)$ or $R(C)$, respectively, on the right-hand scale. The values of L and C are found where the straightedge intersects the center scales. *(Nomograph by Applebaum, reprinted with permission from the March, 1967 issue of EDN Magazine, Rogers Publishing Co., Englewood, Colo.)*

The Series m-Derived Filter The low-pass, m-derived filter has a passband from d.c. to the cutoff frequency, f_c. Beyond this frequency, the signal is attenuated considerably to $f\,\infty$, as shown in figure 44.

The T section configuration used in series m-derived filters is shown in the nomograph of figure 44, with the appropriate design formulas. The correct value of m is found by the use of nomograph # 2 of figure 45. No units are given for f_c and f_∞ since any frequency may be used provided that both scales use the same units. The value



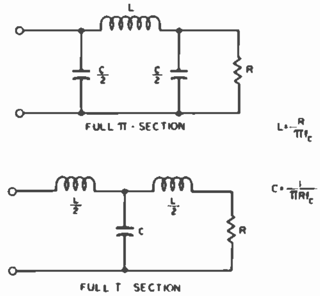


Figure 42

CONSTANT-k FILTER AND LOW-FREQUENCY BANDPASS

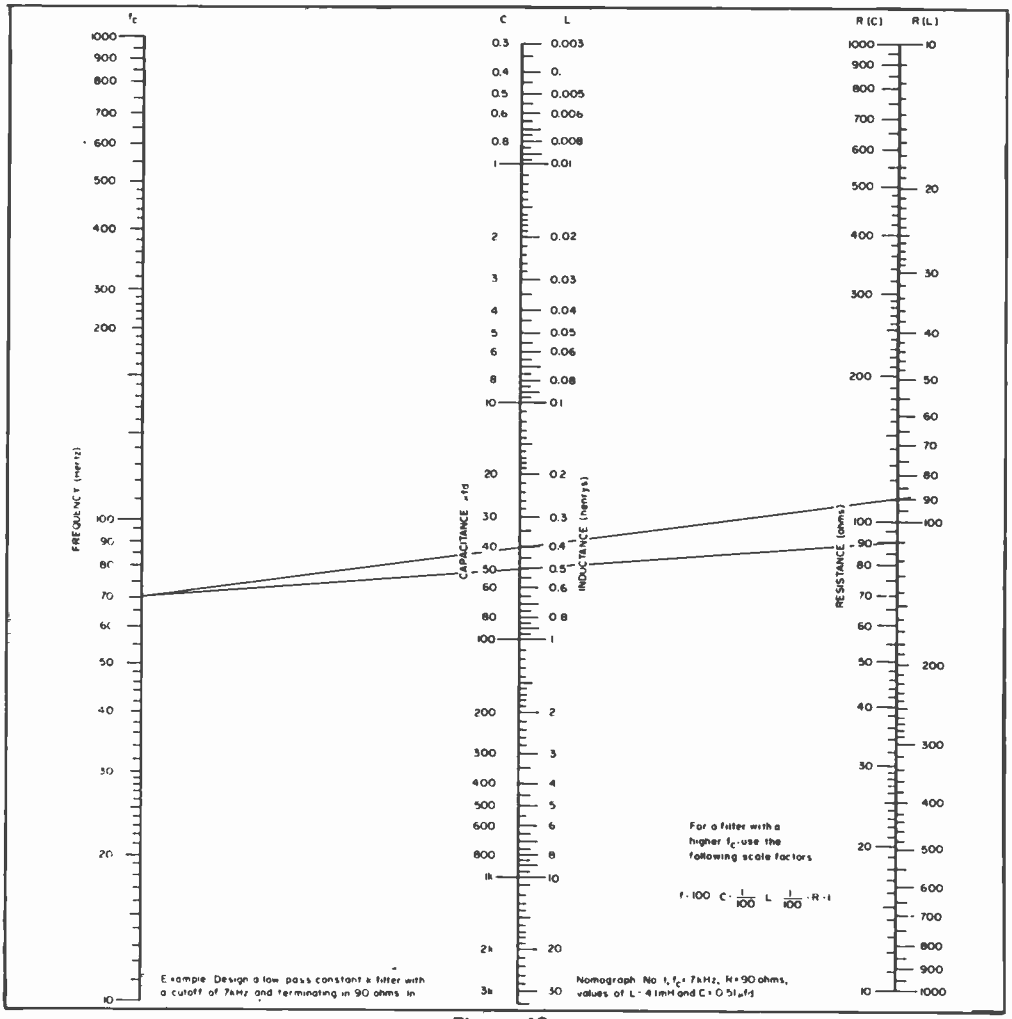
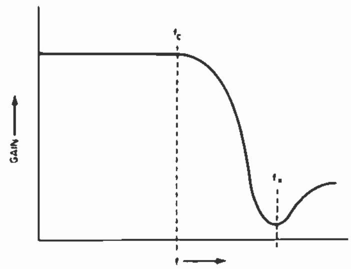


Figure 43

NOMOGRAPH #1, CONSTANT-k FILTER

The filter termination value (R) is used on separate scales (right-hand) for determination of C and L. An example is shown for R = 90 ohms.



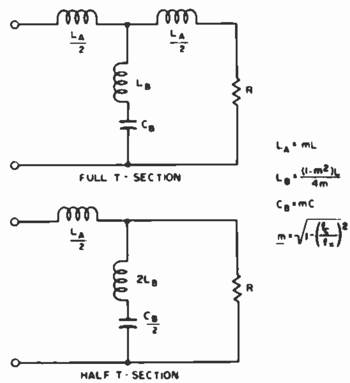


Figure 44

SERIES *m*-DERIVED FILTER AND LOW-FREQUENCY BANDPASS

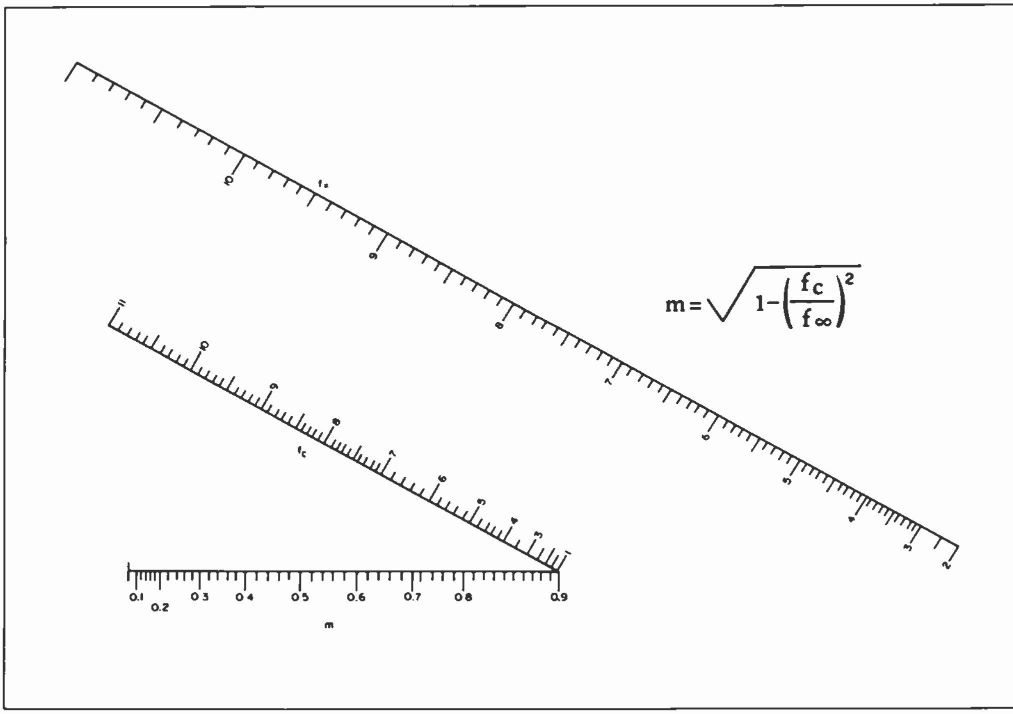


Figure 45

NOMOGRAPH #2. FILTER CONSTANT *m* IS DETERMINED FROM f_∞ and f_c

of m is determined by aligning a straightedge from the value of f_∞ on its scale through the value of f_c on its scale. The value of m is found where the straightedge intersects the horizontal m scale.

The values of filter components L_A, L_B, and C_B are found with the aid of nomograph #3 of figure 46. Note that L_A, and C_B are found by using the left-hand scales, and L_B and C_A are found by using the right-hand scales.

By extending a straightedge from either L or C to the value of m (as found in figure 45) on their appropriate scales, L_A, L_B, and C_B are found where this line intersects the center scale. Any units may be used for L or C provided the same units are used for C_B or L_A and L_B, respectively. *(Nomo-*

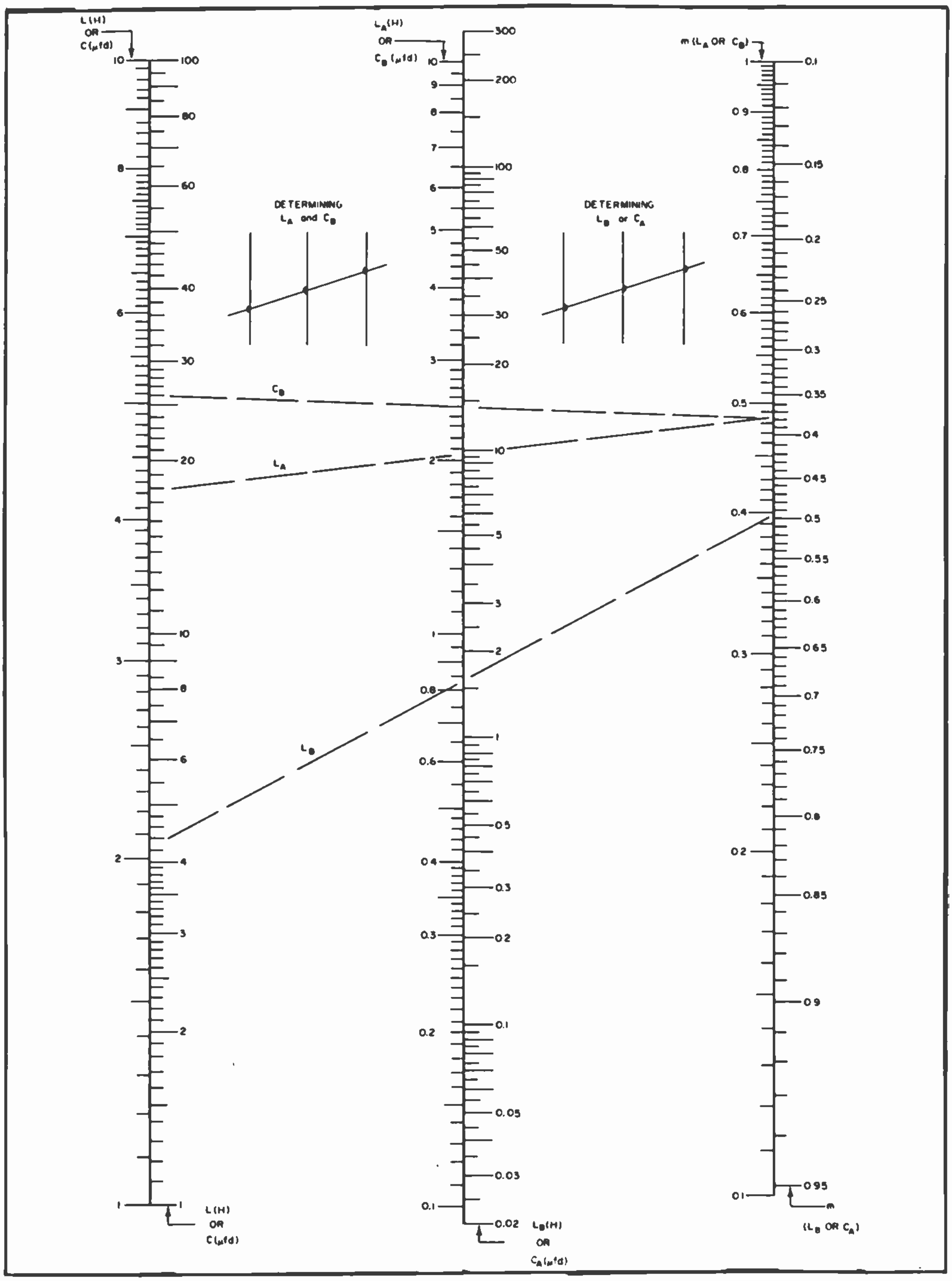


Figure 46

NOMOGRAPH #3. L_A AND C_B ARE DETERMINED USING L AND C (NOMOGRAPH #1) AND *m* (NOMOGRAPH #2.) ALL NUMBERS ARE FOUND WITH LEFT SIDE OF SCALES. L_B AND C_A ARE DETERMINED IN THE SAME MANNER, USING RIGHT SIDE OF SCALES.

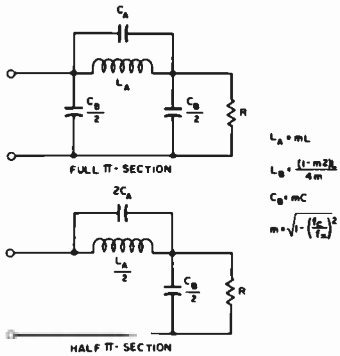


Figure 47

SHUNT m-DERIVED FILTER PI SECTIONS

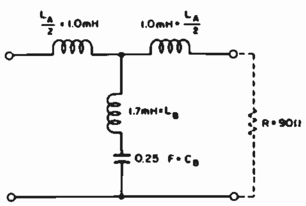


Figure 48

SERIES m-DERIVED FILTER DESIGNED FOR $f_c = 7$ kHz, $f_{\infty} = 8$ kHz AND $R = 90$ OHMS

graph by Applebaum, reprinted with permission from the April, 1697 issue of EDN magazine, Rogers Publishing Co., Englewood, Colo.)

The Shunt, m-Derived Filter The pi section for the shunt, m-derived filter is shown in figure 47, using the nomenclature shown. The values of these components are found by using nomographs #2 and #3 of figures 45 and 46, just as with the series, m-derived filter design.

Example: Design a low-pass, series m-derived filter with a cutoff frequency of 7 kHz, a maximum attenuation at 8 kHz, and terminating in 90 ohms. Using the nomograph of figure 45, m is determined to be 0.485.

On the nomograph (#1 of figure 42) using f_c = 7 kHz and R = 90 on both $R(C)$ and $R(L)$ scales, the value of L and C are determined to be: L_2 = 0.0042 henry

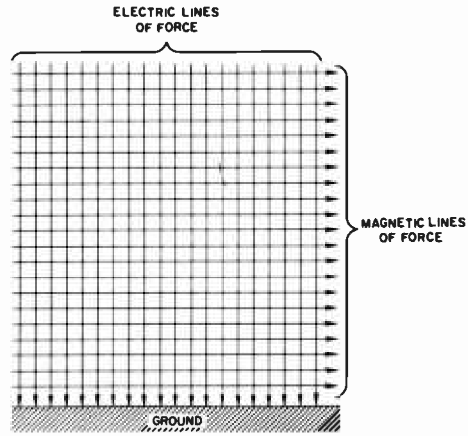


Figure 49

THE ELECTROMAGNETIC FIELD

A time-varying electromagnetic field may be propagated through empty space. The wave may be considered to be made up of interrelated electric and magnetic fields at right angles to each other and lying in a plane. In this illustration, the electric (E) and magnetic (H) fields are transverse to the direction of propagation (out of the page).

and C_2 = 0.51 μfd. Therefore, on nomograph #3 of figure 46, L_A = 0.002 henry (2mH), C_B = 0.25 μfd and L_B = 0.0017 henry (1.7 mH). The final filter design is shown in figure 48.

3-7 Modern Filter Design

The traditional image-parameter filter (discussed in the previous section) has largely been superseded by the recently developed *elliptic-function (Chebishev)* filter design. This technique is well suited to computer programming which stores a file of precalculated and cataloged designs normalized to a cutoff frequency of one Hz and terminations of one ohm. The catalog may be readily adapted to a specific use by scaling the normalized parameters to the cutoff frequency and terminating resistance desired. Filters designed by this new technique provide superior performance with less components than equivalent filters designed by the image-parameter scheme. A catalog of synthesis systems may be found in *Simplified Modern Filter Design*, by Geffe, published by

John F. Ryder Publisher, Inc., New York, a division of Hayden Publishing Co., Inc.

3-8 The Electromagnetic Field

A time-varying *electromagnetic field* may be propagated through empty space at the velocity of light. Such a moving field is commonly called an *electromagnetic wave*. The wave may be considered to be made up of interrelated electric and magnetic fields at right angles to each other and lying in a plane, as indicated in figure 49. The wave, in addition to being propagated through space, may be reflected or refracted at the boundary between two types of media. The drawing illustrates a plane wave, with the electric (E) and magnetic (H) fields transverse to the direction of propagation (out of page).

The abstract concept of an electromagnetic wave traveling through space is difficult to comprehend without the assistance of mathematical proof. Viewed from the concept of electron flow in a conductor, there is no suggestion of energy radiation into space. A set of relationships termed *Maxwell's equations* form the basic tools for the analysis of most electromagnetic-wave problems. The equations picture an interplay of energy between electric and magnetic fields in free space which is self-maintained, with the energy radiating outward from the point of origin. The equations express the continuous nature of electric and magnetic fields and define how changes in one field bring about changes in the other field. The interplay of energy between the fields, moreover, produce displacement waves traveling with the velocity of light. The compound disturbance thus created is described in Maxwell's equations, which were first proven in fact by Hertz, who generated an electromagnetic (radio) wave in 1888, fifteen years after Maxwell predicted its existence.

A complete discussion of electromagnetic fields and Maxwell's equations may be found in *Electromagnetics*, by John D. Kraus, McGraw-Hill Book Co., New York.

CHAPTER FOUR

Semiconductor Devices

Part I—Diodes and Bipolar Devices

One of the earliest detection devices used in radio was the galena crystal, a crude example of a *semiconductor*. More modern examples of semiconductors are the selenium and silicon rectifiers, the germanium diode, and numerous varieties of the transistor. All of these devices offer the interesting property of greater resistance to the flow of electrical current in one direction than in the opposite direction. Typical conduction curves for some semiconductors are shown in figure 1. The *transistor*, a three-terminal device, moreover, offers current amplification and may be used for a wide variety of control functions including amplification, oscillation, and frequency conversion.

Semiconductors have important advantages over other types of electron devices. They are very small, light and require no filament voltage. In addition, they consume very little power, are rugged, and can be made impervious to many harsh environmental conditions. Transistors are capable of usable amplification into the uhf region and provide hundreds of watts of power capacity at the lower frequencies.

Common transistors are current-operated devices whereas vacuum tubes are voltage-operated devices so that direct comparisons between the two may prove to be misleading, however economic competition exists between the two devices and the inexpensive and compact transistor is rapidly taking over some of the functions previously reserved for the more expensive vacuum tube.

4-1 Atomic Structure of Germanium and Silicon

Since the mechanism of conduction of a semiconductor device is different from that of a vacuum tube, it is well to briefly review the atomic structure of various materials used in the manufacture of transistors and diodes.

It was stated in an earlier chapter that the electrons in an element having a large atomic number are conveniently pictured as being grouped into rings, each ring having a definite number of electrons. Atoms in which these rings are completely filled are termed *inert gases*, of which helium and argon are examples. All other elements have one or more incomplete rings of electrons. If the incomplete ring is loosely bound, the electrons may be easily removed, the element is called *metallic*, and is a conductor of electric current. Copper and iron are examples of conductors. If the incomplete ring is tightly bound, with only a few electrons missing, the element is called *nonmetallic*, and is an insulator (nonconductor) to electric current. A group of elements, of which germanium, gallium, and silicon are examples, fall between these two sharply defined groups and exhibit both metallic and nonmetallic characteristics. Pure germanium or silicon may be considered to be a good insulator. The addition of certain impurities in carefully controlled amounts to the pure

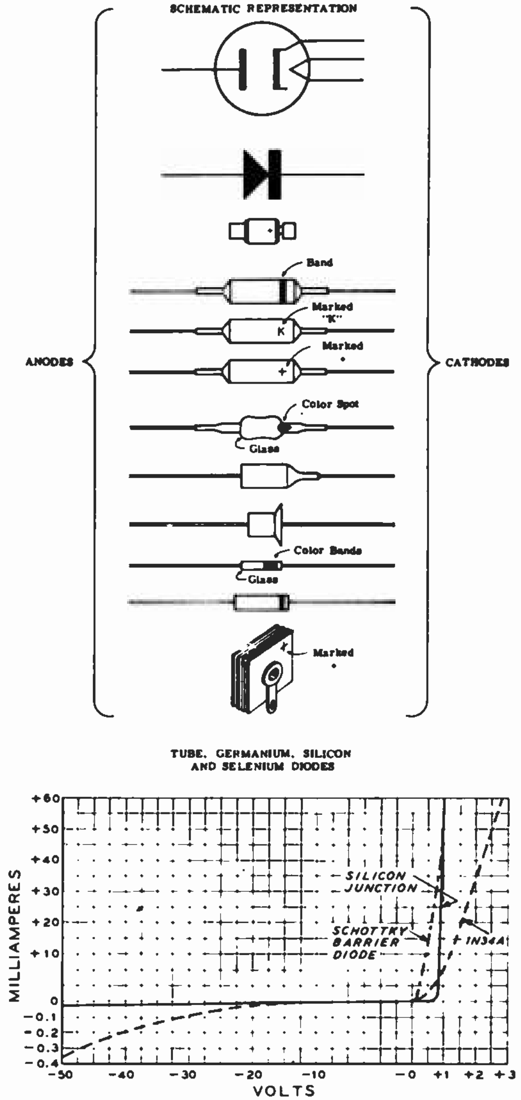


Figure 1

DIODE CHARACTERISTICS AND CODING

The semiconductor diode offers greater resistance to the flow of current in one direction than in the opposite direction. Note expansion of negative current and positive voltage scales. Diode coding is shown above, with notations usually placed on cathode (positive) end of unit.

element will alter the conductivity of the material. In addition, the choice of the impurity can change the direction of conductivity through the element, some impurities increasing conductivity to positive potentials and others increasing conductivity to negative potentials. Early transistors were mainly made of germanium but most modern transistors possessing power capability are made of silicon. Experimental transistors are being made of gallium arsenide which combines some of the desirable features of both germanium and silicon.

Both germanium and silicon may be "grown" in a diamond lattice crystal configuration, the atoms being held together by bonds involving a shared pair of electrons (figure 2). Electrical conduction within the crystal takes place when a bond is broken, or when the lattice structure is altered to obtain an excess electron by the addition of an impurity. When the impurity is added, it may have more or less loosely held electrons than the original atom, thus allowing an electron to become available for conduction, or creating a vacancy, or *hole*, in the shared electron bond. The presence of a hole encourages the flow of electrons and may be considered to have a positive charge, since it represents the absence of an electron. The hole behaves, then, as if it were an electron, but it does not exist outside the crystal.

4-2 Mechanism of Conduction

There exist in semiconductors both negatively charged electrons and absence of electrons in the lattice (holes), which behave as though they had a positive electrical charge equal in magnitude to the negative charge on the electron. These electrons and holes drift in an electrical field with a velocity which is proportional to the field itself:

$$V_{dh} = \mu_h E$$

where,

V_{dh} equals drift velocity of hole,
E equals magnitude of electric field,
μ_h equals mobility of hole.

In an electric field the holes will drift in a direction opposite to that of the electron and with about one-half the velocity, since the hole mobility is about one-half the electron mobility. A sample of a semiconductor,

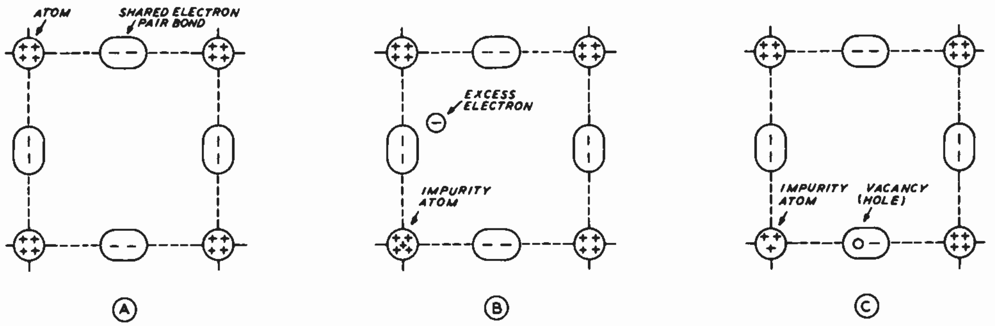


Figure 2

SEMICONDUCTOR CRYSTAL LATTICE

Silicon and germanium lattice configuration made up of atoms held by bonds involving a shared pair of electrons. Conduction takes place when bond is altered to provide excess electron (B) or to create electron vacancy or conducting "hole" (C).

such as germanium or silicon, which is both chemically pure and mechanically perfect will contain in it approximately equal numbers of holes and electrons and is called an *intrinsic* semiconductor. The intrinsic resistivity of the semiconductor depends strongly on the temperature, being about 50 ohm/cm for germanium at room temperature. The intrinsic resistivity of silicon is about 65,000 ohm/cm at the same temperature.

If, in the growing of the semiconductor crystal, a small amount of an impurity, such as phosphorous is included in the crystal, each atom of the impurity contributes one free electron. This electron is available for conduction. The crystal is said to be *doped* and has become electron-conducting in nature and is called *N (negative)-type* silicon. The impurities which contribute electrons are called *donors*. N-type silicon has better conductivity than pure silicon in one direction, and a continuous stream of electrons will flow through the crystal in this direction as long as an external potential of the correct polarity is applied across the crystal.

Other impurities, such as boron add one hole to the semiconducting crystal by accepting one electron for each atom of impurity, thus creating additional holes in the semiconducting crystal. The material is now said to be hole-conducting, or *P (positive)-type* silicon. The impurities which create holes are called *acceptors*. P-type silicon has better conductivity than pure silicon in one direction. This direction is opposite to that of the N-type material. Either the N-type or the P-type silicon is called *extrinsic* conducting type. The doped materials have lower resistivities than the pure materials, and doped semiconductor material in the resistivity range of .01 to 10 ohm/cm is normally used in the production of transistors.

The electrons and holes are called *carriers;* the electrons are termed majority carriers, and the holes are called minority carriers.

4-3 The PN Junction

The semiconductor diode is a *PN junction,* or *junction diode* having the general electrical characteristic of figure 1 and the electrical configuration of figure 3. The anode of the junction diode is always positive type

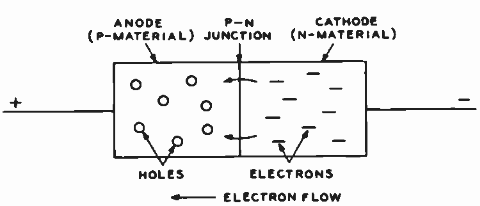


Figure 3

PN JUNCTION DIODE

P-type and N-type materials form junction diode. Current flows when P anode is positive with respect to the N cathode (forward bias). Electrons and holes are termed carriers, with holes behaving as though they have a positive charge.

(P) material while the cathode is always negative–type (N) material. Current flow occurs when the P-anode is positive with respect to the N-cathode. This state is termed *forward bias.* Blocking occurs when the P-anode is negative with respect to the N-cathode. This is termed *reverse bias.* When no external voltage is applied to the PN junction, the energy barrier created at the junction prevents diffusion of carriers across the junction. Application of a positive potential to the P-anode effectively reduces the energy barrier, and application of a negative potential increases the energy barrier, limiting current flow through the junction.

In the forward-bias region shown in figure 1, current rises rapidly as the voltage is increased, whereas in the reverse-bias region current is much lower. The junction, in other words is a high-resistance element in the reverse-bias direction and a low-resistance element in the forward-bias direction.

Junction diodes are rated in terms of average and peak-inverse voltage in a given environment, much in the same manner as thermionic rectifiers. Unlike the latter, however, a small *leakage current* will flow in the reverse-biased junction diode because of a few hole-electron pairs thermally generated in the junction. As the applied inverse voltage is increased, a potential will be reached at which the leakage current rises abruptly at an *avalanche voltage* point. An increase in inverse voltage above this value can result in the flow of a large reverse current and the possible destruction of the diode.

Maximum permissible forward current in the junction diode is limited by the voltage drop across the diode and the heat-dissipation capability of the diode structure. Power diodes are often attached to the chassis of the equipment by means of a *heat-sink* to remove excess heat from the small junction.

Silicon diode rectifiers exhibit a forward voltage drop of 0.4 to 0.8 volts, depending on the junction temperature and the impurity concentration of the junction. The forward voltage drop is not constant, increasing directly as the forward current increases. Internal power loss in the diode increases as the square of the current and thus increases rapidly at high current and temperature levels.

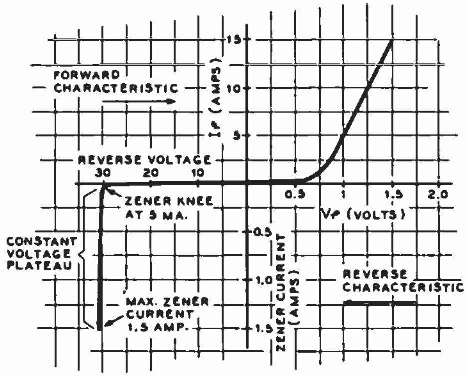


Figure 4

ZENER-DIODE CHARACTERISTIC CURVE

Between zener knee and point of maximum current, the zener voltage is essentially constant at 30 volts. Units are available with zener voltages from approximately 4 to 200.

After a period of conduction, a silicon rectifier requires a finite time interval to elapse before it may return to the reverse-bias condition. This *reverse recovery time* imposes an upper limit on the frequency at which a silicon rectifier may be used. Operation at a frequency above this limit results in overheating of the junction and possible destruction of the diode because of the power loss during the period of recovery.

The Zener Diode The *zener diode* (reference diode) is a PN junction that can be used as a constant-voltage reference, or as a control element. It is a silicon element operated in the reverse-bias avalanche breakdown region (figure 4). The break from nonconductance to conductance is very sharp and at applied voltages greater than the breakdown point, the voltage drop across the diode junction becomes essentially constant for a relatively wide range of currents. This is the *zener control region.* Zener diodes are available in ratings to 50 watts, with zener voltages ranging from approximately 4 volts to 200 volts.

Thermal dissipation is obtained by mounting the zener diode to a heat sink composed of a large area of metal having free access to ambient air.

The zener diode has no ignition potential as does a gas regulator tube, thus eliminating the problems of relaxation oscillation and high firing potential, two ailments of the gas tube. Furthermore, the zener regulator or combinations can be obtained for almost any voltage or power range, while the gas tubes are limited to specific voltages and restricted current ranges.

Actually, only the zener diode having a voltage rating below approximately 6.8 volts is really operating in the zener region. A higher voltage zener diode displays its constant voltage characteristic by virtue of the *avalanche effect*, which has a very sharp knee (figure 4). A diode for a voltage below 6.8 operates in the true zener region and is characterized by a relatively soft knee.

Avalanche and zener modes of breakdown have quite different temperature characteristics and breakdown diodes that regulate in the 5.6- to 6.2 volt region often combine some of each mechanism of breakdown and have a voltage versus temperature characteristic which is nearly flat. Many of the very stable *reference diodes* are rated at 6.2 volts. Since the avalanche diode (breakdown voltage higher than 6.8 volts) displays a positive voltage-temperature slope, it is possible to temperature-compensate it with one or more series forward-biased silicon diodes (D_1) as shown in figure 5. The 1N935 series (9 volt) is apparently of this sort, since the voltage is not 6.2 or some integer multiple thereof.

Silicon epitaxial transistors may also be used as zener diodes, if the current requirement is not too large. Most small, modern, silicon signal transistors have a V_{BEO} (back emitter-base breakdown voltage) between 3

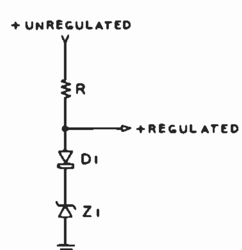


Figure 5
TEMPERATURE-COMPENSATED ZENER DIODE

and 5 volts. If the base and emitter leads are used as a zener diode, the breakdown will occur at a volt or so in excess of the V_{BEO} rating. Figure 6 shows NPN and PNP transistors used in this fashion. For safety, no more than one quarter the rated power dissipation of the transistor should be used when the device is operated this way.

All types of zener diodes are a potential source of noise, although some types are worse than others. If circuit noise is critical, the zener diode should be bypassed with a low-inductance capacitor. This noise can be evident at any frequency, and in the worst cases it may be necessary to use LC decoupling circuits between the diode and highly sensitive r-f circuits.

Junction Capacitance The PN junction possesses capacitance as the result of opposite electric charges existing on the sides of the barrier. Junction capacitance may change with applied voltage, as shown in figure 7.

A voltage-variable capacitor (*varactor* or *varicap*) is generally made of a silicon junction having a special impurity concentration to enhance the capacitance variation and to minimize series resistance losses.

The varicap and the varactor are fundamentally the same type of device, the former used in tuning resonant circuits electrically and the latter used in parametric amplifiers and frequency multipliers. Both devices have

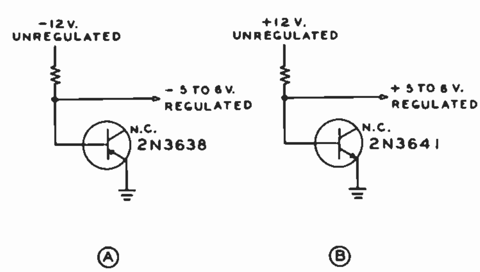


Figure 6
SMALL-SIGNAL SILICON TRANSISTOR USED AS ZENER DIODE

been designed to give a high-Q capacitance vs. voltage relationship at radio frequencies.

The circuit of figure 8A shows a varicap used to electrically tune a resonant circuit. This form of tuning is restricted to circuits which have a very small r-f voltage across them, such as in receiver r-f amplifier stages. Any appreciable a-c voltage (compared to the d-c control voltage across the device)

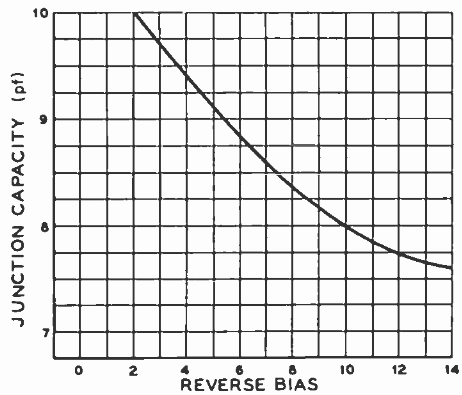


Figure 7

JUNCTION CAPACITANCE VARIATION WITH RESPECT TO REVERSE VOLTAGE

will swing its capacitance at the r-f rate, causing circuit nonlinearity and possible crossmodulation of incoming signals. This nonlinearity may be overcome by using two varicap devices as shown in figure 8B. In this case, the a-c component increases the capacitance of one varicap while decreasing that of the other. This tuning method may be used in circuits having relatively high r-f voltages without the danger of nonlinearity.

The Varactor The varactor frequency multiplier (also called the *parametric multiplier*) is a useful vhf/uhf multiplier which requires no d-c input power. The input power consists only of the fundamental-frequency signal to be multiplied and typically 50% to 70% of that r-f power is recovered at the output of the multiplier unit. Since the efficiency of a varactor multiplier drops as the square of the multiple (n), such devices are not usually used for values of n greater than five.

Examples of varactor multipliers are shown in figure 9. There are usually a number of *idlers* (series-resonant circuits) in a varactor multiplier. In general, there will be n-2 idlers. These idlers are high-Q selective short circuits which reflect undesired harmonics back into the nonlinear capacitance diode.

An interesting development in multiplier diodes is the *step-recovery* diode. Like the varactor, this device is a frequency multiplier requiring no d-c input. The important difference between the step-recovery diode and the varactor is that the former is deliberately driven into forward conduction by the fundamental drive voltage. In addition, the step-recovery diode multiplier requires no idler circuits and has an output efficiency that falls off only as $1/n$. A "times-ten" frequency multiplier could then approach 10% efficiency, as compared to a varactor multiplier whose efficiency would be in the neighborhood of 1%. A typical step-recovery multiplier is shown in figure 10. Diode multipliers are capable of providing output powers of over 25 watts at 1 GHz, and several watts at 5 GHz. Experimental devices have been used for frequency multiplication at frequencies over 20 GHz, with power capabilities in the milliwatt region.

Point-Contact Diodes A rectifying junction can be made of a metal "whisker" touching a very small semiconductor die. When properly assembled,

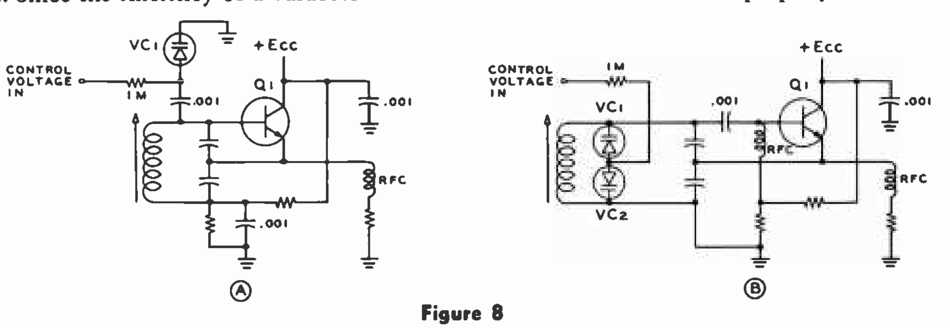


Figure 8

VOLTAGE VARIABLE CAPACITORS

A—Single varicap used to tune resonant circuit
B—Back-to-back varicaps provide increased tuning range with improved linearity

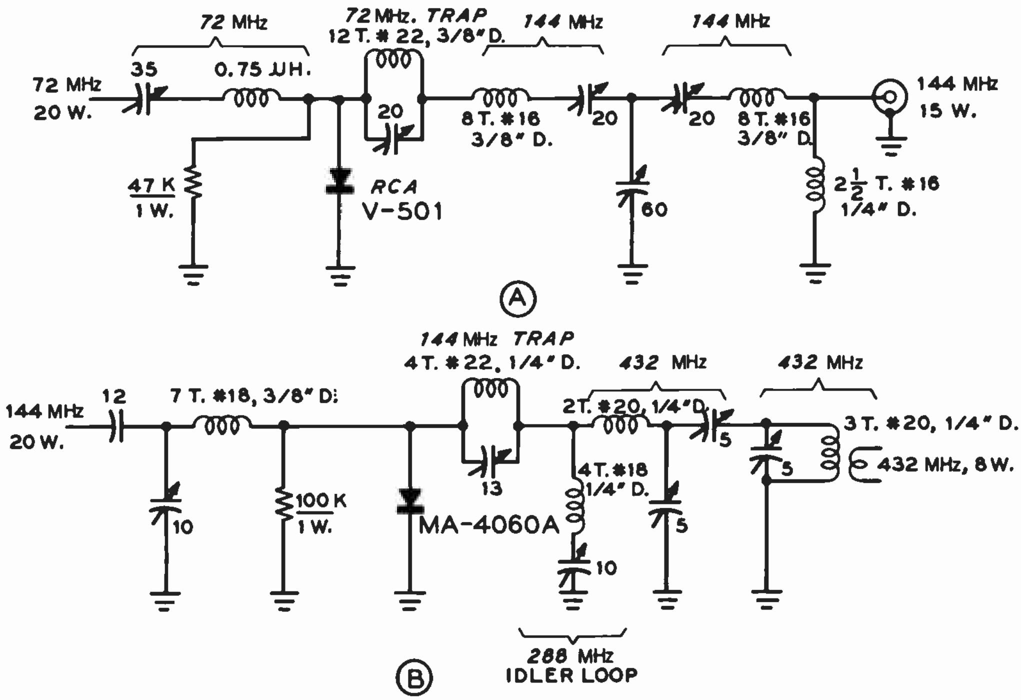


Figure 9

BASIC VARACTOR DOUBLING AND TRIPLING CIRCUITS

If "step-recovery" diode is used, idler loop may be omitted.

the die injects electrons into the metal. The contact area exhibits extremely low capacitance and *point-contact* diodes are widely used as uhf mixers, having noise figures ranging up to 5 db at 3 GHz. The 1N21-1N26 series devices are examples of silicon point-contact diodes made for microwave radar mixer use. However, it is the inexpensive germanium point-contact diode which is now universally used for r-f detection. The germanium device is still quite useful for, a h-f, low-voltage-drop diode but the new gold bonded germanium point-contact diode has a lower forward voltage drop than the older design.

The *Schottky-barrier* or *hot-carrier diode* is similar to the silicon point-contact diode, with the metal-to-silicon interface made by metal deposition on silicon. This device behaves like a silicon point-contact diode, having a lower forward voltage drop than an equivalent silicon unit, good high-frequency response, and a lower noise figure.

Other Diode Devices *Impatt*, *Trapatt*, and *Gunn* diodes are used to produce r-f directly from d-c when used in microwave cavities. The PIN diode is useful as an attenuator or switch at radio frequencies. This is a PN junction with a layer of undoped (intrinsic) silicon between

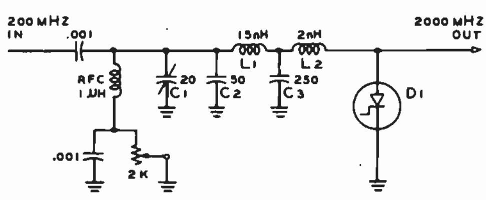


Figure 10

STEP-RECOVERY FREQUENCY MULTIPLIER

Step-recovery diode is used as multiplier. No idler circuits are required, such as used with varactor.

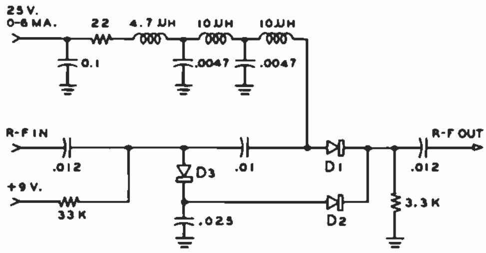


Figure 11

PIN DIODE USED AS R-F ATTENUATOR OR SWITCH

Diode D_1 appears resistive to frequencies whose period is shorter than "carrier" lifetime. Control voltage varies r-f attenuation of diode.

the P and N regions. Because of the neutral intrinsic layer, the charge carriers in the diode are relatively slow; that is, they have a long carrier lifetime. If this lifetime is long compared to the radio frequency impressed on the device, the diode appears resistive to that frequency. Since PIN diodes appear resistive to frequencies whose period is shorter than their carrier lifetime, these diodes can be used as attenuators and switches. An example of such an electrically variable PIN diode attenuator is shown in figure 11.

4-4 Diode Power Devices

Semiconductor devices have ratings which are based on thermal considerations similar to other electronic devices. The majority of power lost in semiconductors is lost internally and within a very small volume of the device. Heat generated by these losses must flow outward to some form of *heat exchanger* in order to hold junction temperature to a reasonable degree. The largest amount of heat flows out through the case and mounting stud of the semiconductor and thence through the heat exchanger into the air. The heat exchanger (or *heat sink*) must be in intimate contact with the case or leads of the semiconductor to achieve maximum uniform contact and maximum heat transfer. The matching surfaces are often lubricated with a substance having good thermal conductivity to reduce oxides or galvanic products from forming on the surfaces (*Dow-Corning Silicone Grease #200* is often used.)

Care must be exercised in the contact between dissimilar metals when mounting semiconductor devices, otherwise electrolytic action may take place at the joint, with subsequent corrosion of one or more surfaces. Many rectifiers come with plated finishes to provide a nonactive material to be placed in contact with the heat sink.

When it is necessary to electrically insulate the case of the semiconductor from the heat sink, a thin mica washer may be placed between the device and the heat sink after lubricating the surfaces with a thermal lubricant.

Diode Rectifiers Semiconductor power rectifiers are the most-used solid-state devices in the electronics industry. Copper-oxide disc rectifiers have been used for decades, as have selenium disc rectifiers. The germanium junction rectifier, too, has been used extensively in electronics; the representative type 1N91 is still available.

Almost all new rectifier system design today uses the *silicon junction* rectifier (figure 12). This device offers the most promising range of applications; from extreme cold to high temperature, and from a few watts of output power to very high voltage and currents. Inherent characteristics of silicon allow junction temperatures in the order of 200°C before the material exhibits intrinsic properties. This extends the operating range

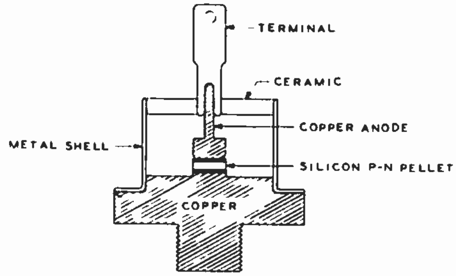


Figure 12

SILICON RECTIFIER

Silicon pellet is soldered to copper stud to provide low thermal resistance path between PN junction and heat sink. Copper anode is soldered to top of junction. Temperature of junction must be held to less than 200°C, as a result of increasing temperature on reverse current flow through junction.

of silicon devices beyond that of any other efficient semiconductor and the excellent thermal range coupled with very small size per watt of output power make silicon rectifiers applicable where other rectifiers were previously considered impractical.

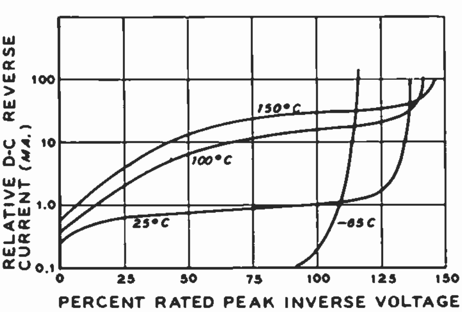
Silicon Current Density The current density of a silicon rectifier is very high, and on present designs ranges from 600 to 900 amperes per square inch of effective barrier layer. The usable current density depends on the general construction of the unit and the ability of the heat sink to conduct heat from the crystal. The small size of the crystal is illustrated by the fact that a rectifier rated at 15 d-c amperes, and 150 amperes peak surge current has a total cell volume of only .00023 inch. Peak currents are extremely critical because the small mass of the cell will heat instantaneously and could reach failure temperatures within a time lapse of microseconds.

Operating Characteristics The *reverse direction* of a silicon rectifier is characterized by extremely high resistance, up to 10^9 ohms below a critical voltage point. This point of *avalanche voltage* is the region of a sharp break in the resistance curve, followed by rapidly decreasing resistance (figure 13A). In practice, the peak inverse working voltage is usually set at least 20% below the avalanche point to provide a safety factor.

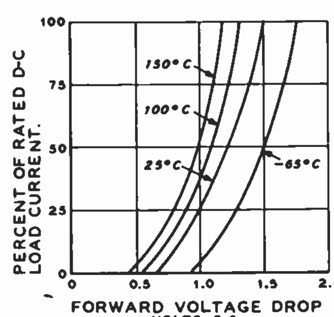
A limited reverse current, usually of the order of 0.5 ma or less flows through the silicon diode during the inverse-voltage cycle. The reverse current is relatively constant to the avalanche point, increasing rapidly as this reverse-voltage limit is passed. The maximum reverse current increases as diode temperature rises and, at the same time, the avalanche point drops, leading to a "runaway" reverse-current condition at high temperatures which can destroy the diode.

The forward characteristic, or resistance to the flow of forward current, determines the majority of power lost within the diode at operating temperatures. Figure 13B shows the static forward current characteristic relative to the forward voltage drop for a typical silicon diode. A small forward bias (a function of junction temperature) is required for conduction. The power loss of a typical diode rated at 0.5 ampere average forward current and operating at 100°C, for example, is about 0.6 watt during the conducting portion of the cycle. The forward voltage drop of silicon power rectifiers is carefully controlled to limit the heat dissipation in the junction.

Diode Ratings and Terms Silicon diodes are rated in terms similar to those used for vacuum-tube rectifiers. Some of the more important terms and their



A



B

Figure 13

SILICON RECTIFIER CHARACTERISTICS

A—Reverse direction of silicon rectifier is characterized by extremely high resistance up to point of avalanche voltage.

B—Threshold voltage of silicon cell is about 0.6 volt. Once device starts conducting the current increases exponentially with small increments of voltage, then nearly linearly on a very steep slope.

definitions follow: *Peak Inverse Voltage* (PIV). The maximum reverse voltage that may be applied to a specific diode type before the avalanche breakdown point is reached.

Maximum RMS Input Voltage—The maximum rms voltage that may be applied to a specific diode type for a resistive or inductive load. The PIV across the diode may be greater than the applied rms voltage in the case of a capacitive load and the maximum rms input voltage rating must be reduced accordingly.

Maximum Average Forward Current—The maximum value of average current allowed to flow in the forward direction for a specified junction temperature. This value is specified for a resistive load.

Peak Recurrent Forward Current—The maximum repetitive instantaneous forward current permitted to flow under stated conditions. This value is usually specified for 60 Hz and a specific junction temperature.

Maximum Single-Cycle Surge Current—The maximum one-cycle surge current of a 60-Hz sine wave at a specific junction temperature. Surge currents generally occur when the diode-equipped power supply is first turned on, or when unusual voltage transients are introduced in the supply line.

Derated Forward Current—The value of direct current that may be passed through a diode for a given ambient temperature. For higher temperatures, less current is allowed through the diode.

Maximum Reverse Current—The maximum leakage current that flows when the diode is biased to the peak-inverse voltage.

Silicon diodes may be mounted on a conducting surface termed a *heat sink* that, because of its large area and heat dissipating ability, can readily dispose of heat generated in the diode junction, thereby safeguarding the diode against damage by excessive temperature.

Improved Rectifier Types A recent silicon rectifier design has been developed having most of the advtantages of silicon, but also low forward voltage drop. This device is the Schottky-barrier or hot-carrier diode in a large format for power use. For two equal volume units, the Schottky-barrier type provides a higher current rating than does the equivalent silicon unit, brought about by the lower forward voltage drop.

The Schottky-barrier device is also a very fast rectifier; operation in high-frequency inverter circuits (up to several hundred kHz) is quite practical. So far the PIV of these diodes remains quite low (less than 50 volts).

A second semiconductor rectifier which combines most of the features of the Schottky-barrier and the common junction device is the *ion-implanted* diode. This diode has impurities implanted in the silicon by means of an "atom smasher." The impurity ions are fired from a particle accelerator into the silicon target wafer. The resultant silicon cystal lattice is modified in such a way as to cause the diodes made from this wafer to have a low forward drop and a fast recovery time (figure 14).

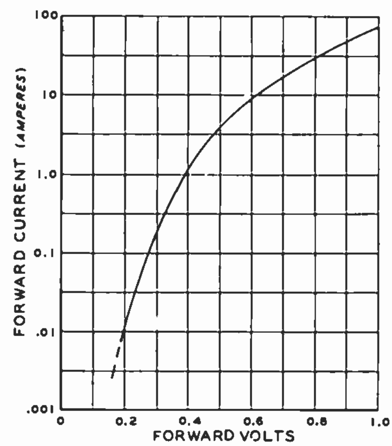


Figure 14

ION-IMPLANTED DIODE FEATURES LOW FORWARD DROP AND FAST RECOVERY TIME

SCR Devices The *thyristor* is a generic term for that family of multilayer semiconductors that comprise *silicon controlled rectifiers* (SCR's), *Triacs*, *Diacs*, *Four Layer Diodes* and similar devices. The SCR is perhaps the most important member of the family, at least economically, and is widely used in the control of large blocks of 60-Hz power.

The SCR is a three-terminal, three-junc-

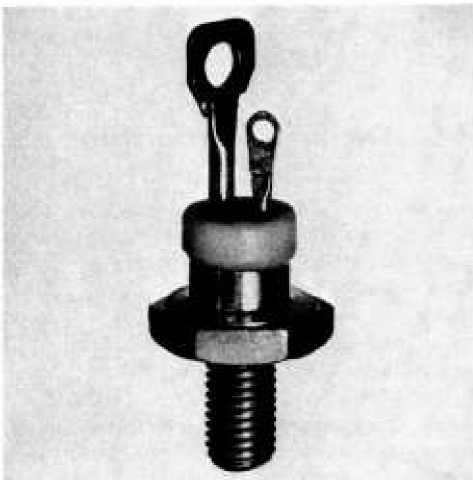


Figure 15

THE SILICON CONTROLLED RECTIFIER

This three-terminal semiconductor is an open switch until it is triggered in the forward direction by the gate element. Conduction will continue until anode current is reduced below a critical value.

tion semiconductor, which could be thought of as a solid-state thyratron. The SCR will conduct high current in the forward direction with low voltage drop, presenting a high impedance in the reverse direction. The three terminals (figure 15) of an SCR device are *anode, cathode,* and *gate.* Without gate current the SCR is an open switch in either direction. Sufficient gate current will close the switch in the forward direction only. Forward conduction will continue even with gate current removed until anode current is reduced below a critical value. At this point the SCR again blocks open. The SCR is therefore a high-speed unidirectional switch capable of being latched on in the forward direction.

The gate signal used to trigger an SCR may be an a-c wave, and the SCR may be used for dimming lights or speed control of small a-c universal series-wound motors, such as those commonly used in power tools. Several power-control circuits using SCR devices and *triacs* (bidirectional triode thyristors) are shown in figure 16.

The *triac* is similar to the SCR except that when its gate is triggered on, it will conduct either polarity of applied voltage. This makes full-wave control much easier to achieve than with an SCR. An example of the triac in a full-wave power control circuit is shown in figure 16C.

The *four layer diode* is essentially an SCR without a gate electrode. As the forward voltage is increased across it, no conduction occurs until the voltage rises to the holdoff value, above which the device conducts in much the same fashion an SCR does when its holdoff voltage has been exceeded.

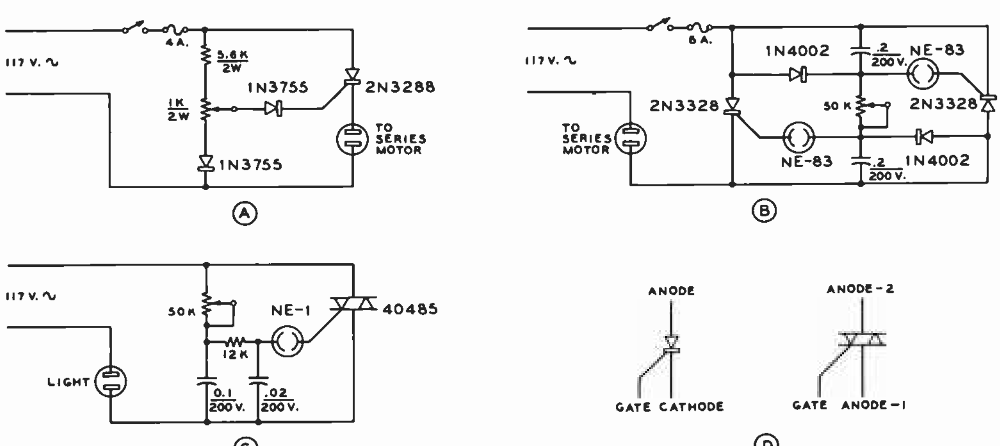


Figure 16

SCR CIRCUITS FOR MOTOR OR LIGHT CONTROL

A— Half-wave control circuit for series motor or light. B— Full-wave control circuit for series motor or light. C— Triac control light circuit. D— Symbols for SCR and Triac units.

The *diac* is analogous to the triac with no gate electrode. It acts like a four layer diode, except that it has similar holdoff in both directions. The diac is used principally to generate trigger pulses for triac gating circuits.

The *silicon unilateral switch* (SUS) is similar to the four layer diode and the *silicon bilateral switch* (SBS) is similar to the diac. There are also a number of other variously-named "trigger diodes" for use with thyristors, but they are all found to be functionally similar to the four layer diode or diac.

There exists one other thyristor of importance: it is the *silicon controlled switch* (SCS). This device has two electrodes: a *gate* to turn it on, and a second *port* to turn it off. The SCS has, so far, only been available in low-voltage low-current versions, as exemplified by the 3N81-3N85 series.

The Unijunction Transistor The *unijunction transistor* (UJT) was originally known as the double-base diode, and its terminal designations (emitter, base 1, base 2) still reflect that nomenclature. If a positive voltage is placed between B_2 and B_1, no conduction occurs until the emitter voltage rises to a fixed fraction of this voltage. The fixed fraction is termed η (the Greek letter *eta*) and is specified for each type of UJT. In the manner of the thyristor, when the emitter reaches η times the voltage between B_1 and B_2, the resistance between the base elements suddenly and markedly decreases. For this reason, the UJT makes a good relaxation oscillator. A simple relaxation oscillator and its transistor equivalent are shown in figure 17. Packaged equivalents are termed *programmed unijunction transistors* (PUT).

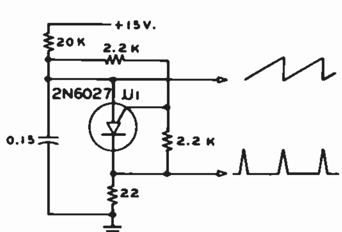


Figure 17

UNIJUNCTION TRANSISTOR SERVES AS RELAXATION OSCILLATOR

Sawtooth or spike waveforms are produced by this simple circuit using single 2N6027 UJT

4-5 The Bipolar Transistor

The decisive event in the creation of the modern semiconductor was the invention of the *transistor* in late 1947. In the last decade semiconductor devices have grown prodigiously in variety, complexity, power capability, and speed of operation. The transistor is a solid-state device having gain properties previously found only in vacuum tubes. The elements germanium and silicon are the principal materials exhibiting the proper semiconducting properties which permit their application in transistors. However, other semiconducting materials, including the compounds indium, antimony, and lead sulfide, have been used experimentally in the production of transistors.

Classes of Transistors Thousands of type numbers of transistors exist, belonging to numerous families of construction and use. The large classes of transistors, based on manufacturing processes are:

Point Contact Transistor—The original transistor was of this class and consisted of *emitter* and *collector* electrodes touching a small block of germanium called the *base.* The base could be either N-type or P-type material and was about .05" square. Because of the difficulty in controlling the characteristics of this fragile device, it is now considered obsolete.

Grown Junction Transistor—Crystals made by this process are grown from molten

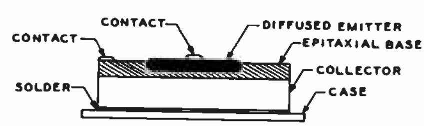


Figure 18

EPITAXIAL TRANSISTOR

Epitaxial, dual-epitaxial and overlay transistors are grown on semiconductor wafer in a lattice structure. After fabrication, individual transistors are separated from wafer and mounted on headers. Connector wires are bonded to metalized regions and unit is sealed in an inclosure.

germanium or silicon in such a way as to have the closely spaced junctions imbedded in the wafer. The impurity material is changed during the growth of the crystal to produce either PNP or NPN ingots, which are then sliced into individual wafers. Junction transistors may be subdivided into *grown junction*, *alloy junction*, or *drift field* types. The latter type transistor is an alloy junction device in which the impurity concentration is contained within a certain region of the base in order to enhance the high-frequency performance of the transistor.

Diffused Junction Transistor—This class of semiconductor has enhanced frequency capability and the manufacturing process has facilitated the use of silicon rather than germanium, which aids the power capability of the unit. Diffused junction transistors may be subdivided into *single diffused* (hometaxial), *double diffused*, *double diffused planar* and *triple diffused planar* types.

Epitaxial Transistors—These junction transistors are grown on a semiconductor wafer and photolithographic processes are used to define emitter and base region during growth. The units may be subdivided into *epitaxial-base*, *epitaxial-layer*, and *overlay* transistors. A representation of an epitaxial-layer transistor is shown in figure 18.

Field-Effect Transistors—Developed in the last decade from experiments conducted over forty years ago, the *field-effect (FET) transistor* may be expected to replace many more common transistor types. This majority carrier device is discussed in a later section of this Handbook.

Manufacturing techniques, transistor end-use, and patent restrictions result in a multitude of transistors, most of which fall into the broad groups discussed previously. Transistors, moreover, may be grouped in families wherein each member of the family is a unique type, but subtile differences exist between members in the matter of end-use, gain, capacitance, mounting, case, leads, breakdown-voltage characteristics, etc. The differences are important enough to warrant individual type identification of each member. In addition, the state of the art permits transistor parameters to be economically designed to fit the various equipment, rather than designing the equipment around available transistor types. This situation results in a great many transistor types having nearly identical general characteristics. Finally, improved manufacturing techniques may "obsolete" a whole family of transistors with a newer, less-expensive family. It is recommended, therefore, that the reader refer to one of the various transistor substitution manuals for up-to-date guidance in transistor classification and substitution.

Transistor Nomenclature Semiconductors are generally divided into product groups classified as "entertainment", "industrial," and "military." The latter classifications often call for multiple testing, tighter tolerances, and quality documentation; and transistors from the same production line having less rigorous specifications often fall into the first, and least-expensive, category. Semiconductors are type numbered by several systems. The oldest standard is the *JEDEC* system. The first number of the identifier establishes the number of junctions (1 = diode, 2 = triode, 3 = tetrode): the letter *N* stands for a semiconductor, followed by a sequential number under which the device was registered.

European manufacturers employ an identifier consisting of a type number composed of two or three letters followed by two or three numbers, the letters indicating the type of transistor and use and the numbers indicating the sequential number in the particular classification. Japanese transistors are usually identified by the code 2S, followed by an identifying letter and sequential number. In addition to these generally recognized codes, numerous codes adapted by individual manufacturers are also in use.

The Junction Transistor The junction transistor is fabricated in many forms, with the planar silicon type providing the majority of units. A pictorial equivalent of a silicon planar power transistor is shown in figure 19. In this type of transistor the emitter and base junctions are often formed by a photolithographic process in selected areas of the silicon dice. Many variations of this technique and design are in use.

The transistor has three essential actions which collectively are called *transistor action*. These are: minority carrier injection,

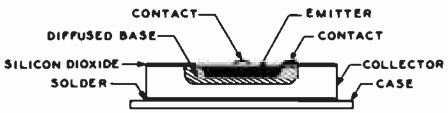


Figure 19

DIFFUSED JUNCTION TRANSISTOR

Emitter and base junctions are diffused into same side of semiconductor wafer which serves as collector. Junction heat is dissipated through solder joint between collector and package.

transport, and collection. Fig. 20 shows a simplified drawing of a PNP junction-type transistor, which can illustrate this collective action. The PNP transistor consists of a piece of N-type silicon on opposite sides of which a layer of P-type material has been grown by the fusion process. Terminals are connected to the two P-sections and to the N-type base. The transistor may be considered as two PN junction rectifiers placed in close juxtaposition with a semiconduction crystal coupling the two rectifiers together. The left-hand terminal is biased in the forward (or conducting) direction and is called the *emitter*. The right-hand terminal is biased in the back (or

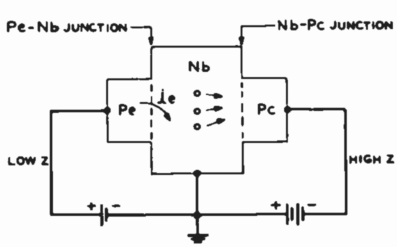


Figure 20

PICTORIAL EQUIVALENT OF PNP JUNCTION TRANSISTOR

reverse) direction and is called the *collector*. The operating potentials are chosen with respect to the *base terminal*, which may or may not be grounded. If an NPN transistor is used in place of the PNP, the operating potentials are reversed.

The P_e—N_b junction on the left is biased in the forward direction and holes from the $P_e$region are injected into the N_b region, producing therein a concentration of holes substantially greater than normally present in the material. These holes travel across the base region toward the collector, attracting neighboring electrons, finally increasing the available supply of conducting electrons in the collector loop. As a result, the collector loop possesses lower resistance whenever the emitter circuit is in operation. In junction transistors this *charge transport* is by means of diffusion wherein the charges move from a region of high concentration to a region of lower concentration at the collector. The collector, biased in the opposite direction, acts as a *sink* for these holes, and is said to collect them.

Alpha It is known that any rectifier biased in the forward direction has a very low internal impedance, whereas one biased in the back direction has a very high internal impedance. Thus, current flows into the transistor in a low-impedance circuit, and appears at the output as current flowing in a high-impedance circuit. The ratio of a change in d-c collector current to a change in emitter current is called the *current amplification*, or *alpha:*

$$\alpha = \frac{i_c}{i_e}$$

where,

α equals current amplification,
i_c equals change in collector current,
i_e equals change in emitter current.

Values of alpha up to 3 or so may be obtained in commercially available point-contact transistors, and values of alpha up to about 0.999 are obtainable in junction transistors.

Beta The ratio of change in d-c collector current to a change in base current(i_b) is a measure of amplification, or *beta:*

$$\beta = \frac{\alpha}{1 - \alpha} = \frac{i_c}{i_b}$$

Values of beta run to 100 or so in inexpensive junction transistors. The static d-c forward current gain of a transistor in the common-emitter mode is termed the *d-c beta* and may be designated β_F or h_{FE}.

Cutoff Frequencies *The alpha cutoff frequency* (f_{hfb}) of a transistor is that frequency at which the grounded-base current gain has decreased to 0.7 of the gain obtainable at 1 kHz. For audio transistors the alpha cutoff frequency is about 1 MHz. For r-f and switching transistors the alpha cutoff frequency may be 50 MHz or higher. The upper frequency limit of operation of the transistor is determined by the small but finite time it takes the majority carriers to move from one electrode to the other.

The *beta cutoff frequency* (f_{hfe}) is that frequency at which the grounded-emitter current gain has decreased to 0.7 of the gain obtainable at 1 kHz. *Transconductance cutoff frequency* (f_{gm}) is that frequency at which the transconductance falls to 0.7 of that value obtainable at 1 kHz. The *maximum frequency of oscillation* (f_{max}) is that frequency at which the maximum power gain of the transistor drops to unity.

Various internal time constants and transit times limit the high-frequency response of the transistor and these limitations are summarized in the *gain-bandwidth product* (f_t), which is identified by the frequency at which the beta current gain drops to unity. These various cutoff frequencies and the gain-bandwidth products are shown in figure 21.

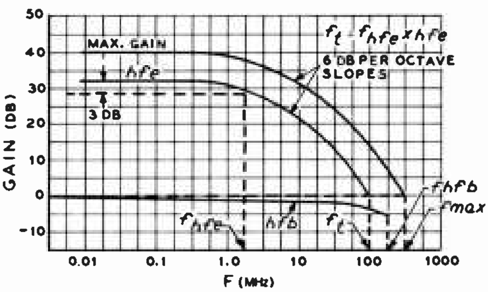


Figure 21

GAIN-BANDWIDTH CHART FOR TYPICAL H-F TRANSISTOR

The Transition Region A useful rule common to both PNP and NPN transistors is: *moving the base potential toward the collector voltage point turns the transistor on, while moving the base potential away from the collector voltage point turns the transistor off*. When fully on, the transistor is said to be *saturated*. When fully off, the transistor is said to be *cut off*. The region between these two extremes is termed the *transition region*. A transistor may be used as a switch by simply biasing the base-emitter circuit on and off. Adjusting the base-emitter bias to some point in the transition region will permit the transistor to act as a signal amplifier. For such operation, base-emitter d-c bias will be about 0.3 volt for many common germanium transistors, and about 0.6 volt for silicon transistors.

Handling Transistors Used in the proper circuit under correct operating potentials the life of a transistor is practically unlimited. Unnecessary transistor failure often occurs because the user does not know how to handle the unit or understand the limitations imposed on the user by virtue of the minute size of the transistor chip. Microwave transistors, in particular, are subject to damage due to improper handling. The following simple rules will help the user avoid unnecessary transistor failures:

Know how to handle the transistor. Static discharges may damage microwave transistors or certain types of field-effect transistors because of small emitter areas in the former and the thin active layer between the channel and the gate in the latter. The transistor should always be picked up by the case and not by the leads. The FET, moreover, should be protected against static electricity by wrapping the leads with tinfoil when it is not in use, or otherwise interconnecting the leads when the unit is moved about or stored. Finally, no transistor should be inserted into or removed from a socket when power is applied to the socket pins. *Never use an ohmmeter for continuity checks.* An ohmmeter may be used at some risk to determine if certain types of transistors are open or shorted. On the low ranges, however, an ohmmeter can supply over 250 milliamperes into a low-resistance load. Many small transistors are rated at a maximum emitter current of 20 to 50 milliamperes and should be tested only in a transistor test set wherein currents and voltages are adjustable and limited. *Don't solder transistor leads unless you can do it fast.* Always use a low-wattage (20 watts or so) pencil iron and a heat sink when soldering transistors into or removing them from the circuit. Long-nose pliers grasping the lead between iron and transistor body will help to prevent transistor chip temperature from becoming excessive. Make the joint fast so that time does not permit the chip to overheat.

In-circuit precautions should also be observed. Certain transistors may be damaged by applying operating potential of reversed polarity, applying an excessive surge of transient voltage, or subjecting the equipment to excessive heat. Dissipation of heat from intermediate-size and power transistors is vital and such units should never be run without an adequate heat-sink apparatus. Finally, a danger exists when operating a transistor close to a high-powered transmitter. The input circuit of the transistorized equipment may be protected by shunting it with two small diodes back to back to limit input voltage excursions.

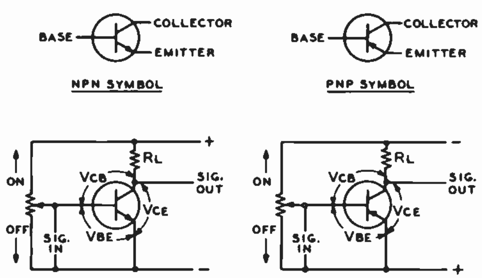


Figure 22

TRANSISTOR SYMBOLS AND BIAS

Moving the base potential toward the collector turns the transistor on. Moving the base potential away from the collector turns the transistor off. Voltage notations are: Collector-to-base voltage, V_{CB}; base-to-emitter voltage, V_{BE}; collector-to-emitter voltage, V_{CE}.

Transistor Symbols The electrical symbols for common three-terminal transistors are shown in figure 22. The left drawing is of a PNP transistor. The symbol for an NPN transistor is similar except that the direction of the arrow of the emitter points away from the base. This suggests that *the arrow points toward the negative terminal of the power source*, and the source potentials are reversed when going from NPN to PNP transistors, or vice-versa. As stated earlier, a useful rule-of-thumb common to both NPN and PNP transistors concerns the base-emitter bias: Moving the base toward the collector voltage turns the transistor *on*, and moving the base away from the collector voltage turns the transistor *off*. As shown in the illustration, capital letters are used for d-c voltages. The important d-c voltages existing in transistor circuitry are: *base-emitter voltage* (V_{BE}), *collector-emitter voltage* (V_{CE}), and *collector-base voltage* (V_{CB}). Signal and alternating voltages and currents are expressed by lower-case letters.

4-6 Transistor Characteristics

The transistor produces results that may be comparable to a vacuum tube, but there is a basic difference between the two devices. The vacuum tube is a voltage-controlled device whereas the transistor is a current-controlled device. A vacuum tube normally operates with its grid biased in the negative, or high-resistance, direction, and its plate biased in the positive, or low-resistance, direction. The tube conducts only by means of electrons, and has its conducting counterpart in the form of the NPN transistor, whose majority carriers are also electrons. There is no vacuum-tube equivalent of the PNP transistor, whose majority carriers are holes.

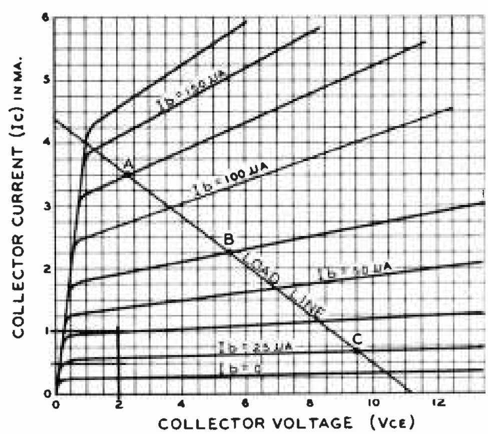


Figure 23

CHARACTERISTIC PLOT OF JUNCTION TRANSISTOR

Characteristics of junction transistor biased in active region may be expressed in terms of plot of collector voltage versus collector current. Load line and limits of operation (points A, C) are plotted, as well as operating point (B) in the manner shown in Chapter Six for vacuum-tube plots.

As discussed earlier, the transistor may be turned off and on by varying the bias on the base electrode in relation to the emitter potential. Adjusting the bias to some point approximately midway between cutoff and saturation will place the transistor in the *active* region of operation. When operated in this region the transistor is capable of amplification. The characteristics of a transistor biased in the active region may be expressed in terms of electrode voltages and currents as is done for vacuum tubes in Chapter Five. The plot of V_{CE} versus I_C (collector-emitter voltage versus collector current) shown in figure 23, for example, should be compared with figure 17, Chapter Five, the plot of I_b versus E_b (plate current versus plate voltage) for a pentode tube. Typical transistor graphs are discussed in this chapter, and the use of similar vacuum-tube plots is discussed in Chapter Six.

Transistor Analysis Transistor behavior may be analyzed in terms of mathematical equations which express the relationships among currents, voltages, resistances, and reactances. These relationships are termed *hybrid parameters* and define instantaneous voltage and current values existing in the circuit under examination. The parameters permit the prediction of the behavior of the particular circuit without actually constructing the circuit.

Equivalent circuits constructed from parameter data allow formulas to be derived

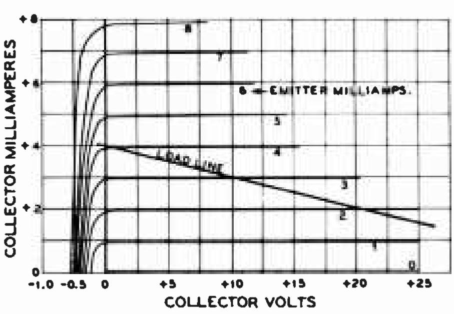


Figure 24

PLOT OF JUNCTION TRANSISTOR

Plot resembles that of a pentode tube except that emitter current, not grid voltage, defines each member of the curve family. Collector current is practically independent of collector voltage.

for current gain, voltage gain, power gain, and other important information necessary to establish proper transistor operation. A complete discussion of hybrid parameters and transistor circuitry may be obtained in the book *Basic Theory and Application of Transistors,* technical manual *TM-11-690,* available from the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402.

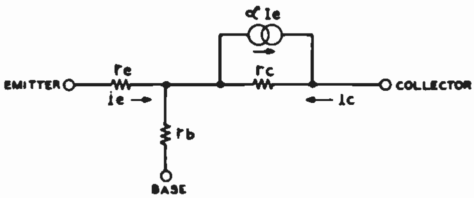
Some of the more useful parameters for transistor application are listed below:

The *resistance gain* of a transistor is expressed as the ratio of output resistance to input resistance. The input resistance of a typical transistor is low, in the neighborhood of 500 ohms, while the output resistance is relatively high, usually over 20,000 ohms. For a junction transistor, the resistance gain is usually over 50.

The *voltage gain* of a transistor is the product of *alpha* times *the resistance gain.* A junction transistor which has a value of alpha less than unity nevertheless has a resistance gain of the order of 2000 because of its extremely high output resistance, and the resulting voltage gain is about 1800 or so. For this type of transistor the *power gain* is the product of *alpha squared* times *the resistance gain* and is of the order of 400 to 500.

The ouput characteristics of the junction transistor are of great interest. A typical example is shown in figure 24. It is seen that the junction transistor has the characteristics of an ideal pentode vacuum tube. The collector current is practically independent of the collector voltage. The range of linear operation extends from a minimum voltage of about 0.2 volts up to the maximum rated collector voltage. A typical load line is shown, which illustrates the very high load impedance that would be required for maximum power transfer. A common-emitter circuit is usually used, since the output impedance is not as high as when a common-base circuit is used.

Equivalent Circuit of a Transistor As is known from network theory, the small-signal performance of any device in any network can be represented by means of an equivalent circuit. The most convenient equivalent circuit for the low-frequency small-signal performance of junc-



VALUES OF THE EQUIVALENT CIRCUIT

PARAMETER	JUNCTION TRANSISTOR (Ie = 1 MA., Vc = 3 V)
re - EMITTER RESISTANCE	(26/Ie)
rb - BASE RESISTANCE	300 Ω
rc - COLLECTOR RESISTANCE	1 MEGOHM
α - CURRENT AMPLIFICATION	0.97

Figure 25

LOW-FREQUENCY EQUIVALENT (COMMON-BASE) CIRCUIT FOR JUNCTION TRANSISTOR

Parameter r_e is equivalent to 52/Ie for silicon and 26/Ie for germanium

tion transistors is shown in figure 25. r_e, r_b, and r_c are dynamic resistances which can be associated with the emitter, base, and collector regions of the transistor. The current generator aI_e, represents the transport of charge from emitter to collector.

Transistor Configurations There are three basic transistor configurations; grounded-base connection, grounded-emitter connection, and grounded-collector connection. These correspond roughly to grounded-grid, grounded-cathode, and grounded-plate circuits in vacuum-tube terminology (figure 26).

The grounded-base circuit has a low input impedance and high output impedance, and no phase reversal of signal occurs from input to output circuit. The grounded-emitter circuit has a higher input impedance and a lower output impedance than the grounded-base circuit, and a reversal of phase between the input and output signal occurs. This usually provides maximum voltage gain from a transistor. The grounded-collector circuit has relatively high input impedance, low output impedance, and no phase reversal of signal from input to output circuit. Power and voltage gain are both low.

Bias Stabilization To establish the correct operating parameters of the transistor, a bias voltage must be established between the emitter and the base. Since transistors are temperature-sensitive devices, and since some variation in characteristics usually exists between transistors of a given type, attention must be given to the bias system to overcome these difficulties. The simple *self-bias* system is shown in figure 27A. The base is simply connected to the power supply through a large resistance which supplies a fixed value of base current to the transistor. This bias system is extremely sensitive to the current-transfer ratio of the transistor, and must be adjusted for optimum results with each transistor.

When the supply voltage is fairly high and wide variations in ambient temperature do not occur, the bias system of figure 27B may be used, with the bias resistor connected from base to collector. When the collector voltage is high, the base current is increased, moving the operating point of the transistor down the load line. If the collector voltage is low, the operating point moves upward along the load line, thus providing auto-

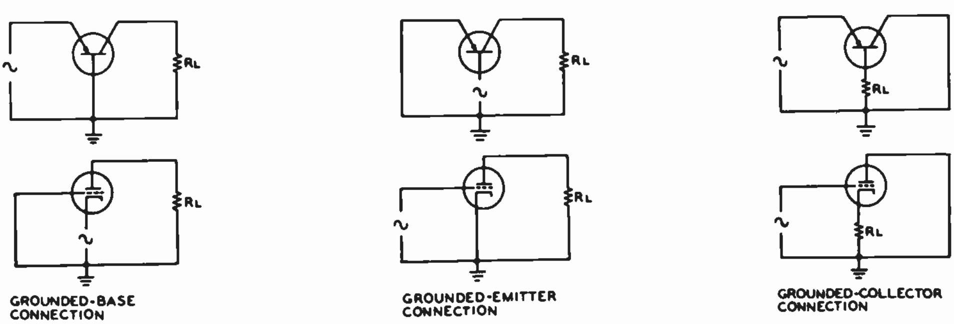


Figure 26

COMPARISON OF BASIC VACUUM-TUBE AND TRANSISTOR CONFIGURATIONS

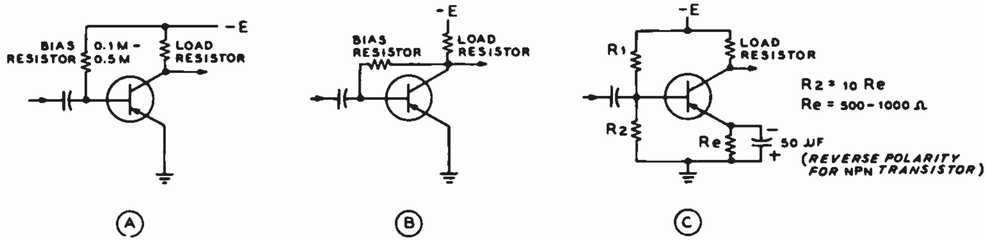


Figure 27

BIAS CONFIGURATIONS FOR TRANSISTORS

The voltage divider system of C is recommended for general transistor use. Ratio of R_1/R_2 establishes base bias, and emitter bias is provided by voltage drop across R_e. Battery polarity is reversed for NPN transistors.

matic control of the base bias voltage. This circuit is sensitive to changes in ambient temperature, and may permit transistor failure when the transistor is operated near maximum dissipation ratings.

These circuits are often used in small imported transistor radios and are not recommended for general use unless the bias resistor is selected for the value of current gain of the particular transistor in use. A better bias system is shown in figure 27C, where the base bias is obtained from a voltage divider, (R_1, R_2), and the emitter is forward-biased. To prevent signal degeneration, the emitter bias resistor is bypassed with a large capacitance. A high degree of circuit stability is provided by this form of bias, providing the emitter capacitance is of the order of 50 μfd for audio-frequency applications.

Bias Circuitry Calculation The voltage-divider bias technique illustrated in figure 27C is redrawn in generalized form in figure 28. This configuration divides the emitter resistor into two units (R_4 and R_5), one of which is bypassed. This introduction of a slight degree of feedback allows the designer more freedom to determine a-c gain, while maintaining good d-c stability. The assumption is made that a modern junction transistor is used having a h_{fe} of at least 40 and a low value of I_{CBO} (collector-cutoff current, emitter open). The procedure to determine bias circuitry is given in the following steps:

1. Collector current (I_c) is chosen from the data sheet.
2. Collector load resistor (R_3) is calculated so that the collector voltage is a little more than one-half the supply voltage.
3. A-c gain value (A) is chosen and emitter resistor R_4 calculated, letting $R_4 = R_3/A$.
4. Emitter resistor R_5 is calculated to raise emitter voltage (E_e) to about 10% to 15% of supply voltage:

$$R_5 = (E_e/I_e) - R_4$$

5. Total base voltage (E_b) is sum of E_e plus base-to-emitter voltage drop (about 0.7 volt for small-signal silicon devices).
6. The sum of base bias resistors R_1 and R_2 is such that one-tenth the value of the d-c collector current flows through the bias circuit.
7. Values of resistors R_1 and R_2 are calculated, knowing current and value of base voltage at mid-point of R_1 and R_2.
8. The a-c input impedance is approximately equal to the parallel combination of R_1, R_2, and $h_{fe} \times R_4$.

To illustrate the design method, an example based on the 2N3565 is chosen. It is assumed that 1 ma of collector-emitter current flows. Collector load resistor R_3 is estimated to be 6.2K, so that the voltage drop across it is 6.2 volts, placing the collector at a potential of 15 − 6.2 − 8.8 volts.

The data sheet of the 2N3565 shows that the range value of h_{fe} at 1 ma of collector current is 150 to 600. An a-c gain value (A) of 62 may be chosen, which is well below the ultimate current gain of the device. Emitter resistor R_4 is now calculated, being equal to R_3/A = 6200/62 = 100 ohms. Emitter resistor R_5 is now calculated to be 1.8K, which raises the emitter voltage to 1.9 volts.

The base-emitter drop is between 0.6 to 0.7 volt for small-signal silicon devices, so this places the base at approximately 2.6 volts. Assuming no base current, the values of resistors R_1 and R_2 can now be determined as they are a simple voltage divider. The series current through R_1 and R_2 is to be one-tenth of the collector current, or 100 μa. Resistor R_2 = 2.6v/.0001 ma = 26,000 ohms and R_1 = 15 − 2.6v/.0001 ma = 124,000 ohms. These are nonstandard values of resistance so 27K and 130K are used.

Once these calculations have been completed, the approximate value of the a-c input impedance may be determined. This is the parallel combination of R_1, R_2, and h_{fe} × R_4. Thus, R_1 and R_2 in parallel are

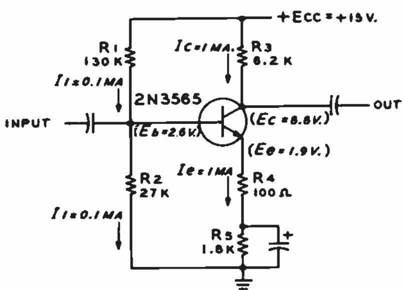


Figure 28

BIAS CIRCUITRY CALCULATION

Generalized form of voltage-divider bias technique.

22.3K and h_{fe} ×R_4 is 15K. Finally, 22.3K and 15K in parallel are 9K.

Actually, the a-c input impedance will be higher than 9K because a minimum value of h_{fe} was used. Also, it is worth noting that the d-c collector voltage is 8.8 volts. This is about half-way between +15v and +2.6v, permitting the collector to swing ± 6 volts in response to the a-c input voltage without clipping the peaks of the waveform.

This method of determining circuit parameters is quite simple and effective for RC amplifier design. With practice, the designer can juggle resistance values as calculations are made to avoid doing the design over at the end of the process.

Output Characteristic Curves Calculation of the current, voltage and power gain of a common-emitter amplifier may be accomplished by using the common-emitter output static

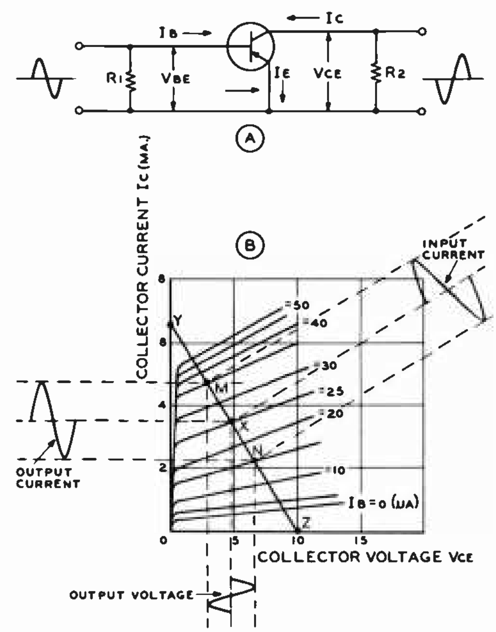


Figure 29

CHARACTERISTIC CURVES AND LOAD LINE FOR COMMON-EMITTER CIRCUIT

Calculation of current, voltage and power gain of a common-emitter transistor amplifier can be accomplished by using output characteristic curves as discussed in the text.

characteristic curves (figure 29) which plot collector current against collector voltage with the base current as a fixed value. In this example, the collector voltage supply is 10 volts, the load resistance is 1500 ohms, the emitter resistance is 500 ohms, the peak-to-peak input current is 20 microamperes and the *operating point* (*X*) is chosen at 25 microamperes of base current and 4.8 volts on the collector.

The first step is to establish a *load line* on the characteristic curves representing the voltage drop across the load resistor (R_2). When the collector current is zero, the total collector supply voltage (10 volts) equals the collector voltage, V_{CE}. Point Z (one point of the load line) then is at the 10-volt mark on the collector voltage axis (x-axis). When the collector current is zero, the total collector supply voltage (10 volts) is dropped across load resistor R_2. The total current (I_c) then is:

$$I_c = \frac{10}{1500} = 0.0066 \text{ amp} = 6.6 \text{ ma}$$

Point *Y* (a second point of the load line) then is at the 6.6-ma mark on the collector-current axis (y-axis). Connect points *Y* and *Z* to establish the load line. The operating point is located at point *X* on the load-line. Since the peak-to-peak input current is 20 microamperes, the deviation is 10 microamperes above the operating point (point *M*) and 10 microamperes below the operating point (point *N*).

The input current, output current, and output voltage waveforms may now be established by extending lines from the operating point perpendicular to the load line and to the y and x axes respectively and plotting the waveforms from each deviation point along the load-line excursions between points *M* and *N*.

Current gain (beta) in this configuration is the ratio of the change in collector current to the change in base current:

$$A_i = \frac{\Delta I_C}{\Delta I_B} = \frac{I_{C(max)} - I_{C(min)}}{I_{B(max)} - I_{B(min)}}$$

where,

A_i is current gain,
I_C is collector current,
I_B is base current,
Δ equals a small increment.

Substituting known values in the formula:

$$\text{Current Gain } (A_i) = \frac{4.7 - 2.1}{35 - 15} = \frac{2.6 \text{ ma}}{20\ \mu\text{a}} = 130$$

Voltage gain in this configuration is the ratio of the change in collector voltage to the change in base voltage:

$$A_v = \frac{\Delta V_{CE}}{\Delta V_{BE}} = \frac{V_{CE(max)} - V_{CE(min)}}{V_{BE(max)} - V_{BE(min)}}$$

where,

A_v is voltage gain,
V_{CE} is collector to emitter voltage,
V_{BE} is base to emitter voltage.

(Note: The change in input voltage is the change in input current multiplied by the input impedance. In this case the input voltage is: 20 microamperes times 500 ohms, or 0.01 volt).

Therefore:

$$\text{Voltage Gain } (A_v) = \frac{6.7 - 2.7}{0.01} = 400$$

Power gain is voltage gain times current gain:

$$\text{Power gain} = 130 \times 400 = 52{,}000$$

Power gain in decibels is:

$$\text{Gain} = 10 \log 52{,}000 = 10 \times 4.7 = 47 \text{ decibels}$$

Constant-Power-Dissipation Line Each transistor has a maximum collector power that it can safely dissipate without damage to the transistor. To ensure that the maximum collector dissipation rat-

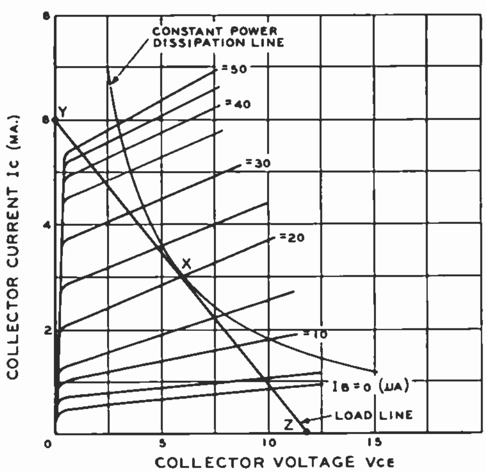


Figure 30

CONSTANT POWER-DISSIPATION LINE

Constant power-dissipation line is placed on output characteristic curves, with collector load line positioned so it falls within area bounded by vertical and horizontal axes and constant power-dissipation line. Load line tangent at (X) permits maximum power gain within maximum collector dissipation rating.

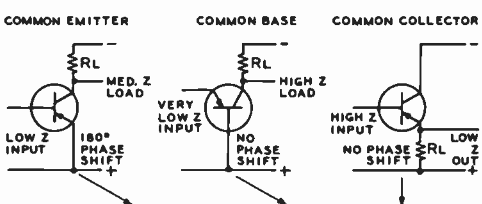
ing is not exceeded, a *constant-power-dissipation line* (figure 30) is drawn on the characteristic curves, and the collector load resistor is selected so that its load line falls in the area bounded by the vertical and horizontal axes and the constant-power-dissipation line. The dissipation line is determined by selecting points of collector voltage and current, the products of which are equal to the maximum collector power rating of the transistor. Any load line selected so that it is tangent to the constant-power-dissipation line will ensure maximum permissible power gain of the transistor while operating within the maximum collector power-dissipation rating. This is important in the design and use of power amplifiers.

4-7 Transistor Audio Circuitry

The transistor can be connected as either a common-base, common-collector, or common-emitter stage, as discussed previously. Similar to the case for vacuum tubes, choice of transistor circuit configuration depends on the desired operating characteristics of the stage. The over-all characteristics of these three circuits are summarized in figure 31. Common-emitter circuits are widely used for high gain amplification, and common-base circuits are useful for oscillator circuits and high-frequency operation, and common-collector circuits are used for various impedance transformation applications. Examples of these circuits will be given in this section.

Audio Circuitry As in the case of electron-tube amplifiers, transistor amplifiers can be operated Class A, class AB, class B, or class C. The first three classes are used in audio circuitry. The class-A transistor amplifier is biased so that collctor current flows continuously during the complete electrical cycle, even when no drive signal is present. The class-B transistor amplifier can be biased either for collector current cutoff or for zero collector voltage. The former configuration is most often used, since collector current flows only during that half-cycle of the input signal voltage that aids the forward bias. This bias technique is used because it results in the best power efficiency. Class-B transistor amplifiers must be operated in push-pull to avoid severe signal distortion. Class-AB transistor amplifiers can be biased so that either collector current or voltage is zero for less than half a cycle of the input signal, and the above statements for class-B service also apply for the class-AB mode.

A simple small-signal voltage amplifier is shown in figure 32A. Direct-current stabilization is employed in the emitter circuit. Operating parameters for the amplifier are given in the drawing. In this case, the input impedance of the amplifier is quite low. When used with a high-impedance driving source such as a crystal microphone,



INPUT Z	500-1.5K	30-150 Ω	20K-500K
OUTPUT Z	30K-50K	300K-500K	50Ω-1K
VOLTAGE GAIN	300-1000	50-1500	LESS THAN 1
CURRENT GAIN	25-50	LESS THAN 1	25-50

Figure 31

THREE BASIC TRANSISTOR CIRCUITS

Common-emitter circuits are used for high-gain amplification, common-base circuits are useful for oscillator circuits and common-collector circuits are used for various impedance transformations.

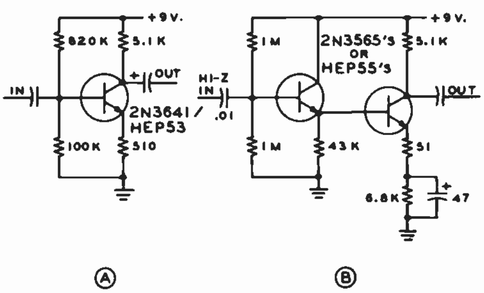


Figure 32

SMALL-SIGNAL VOLTAGE AMPLIFIERS

A—Low impedance, d-c stabilized amplifier
B—Two stage amplifier features high input impedance

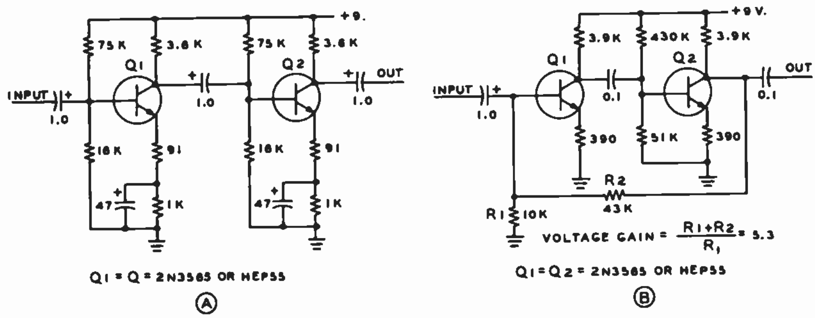


Figure 33

TWO STAGE RC AMPLIFIERS

A—Input impedance of amplifier is about 1600 ohms.
B—Feedback amplifier with feedback loop from collector of Q_2 to base of Q_1.

an emitter-follower input should be employed as shown in figure 32B.

The circuit of a two-stage resistance-coupled amplifier is shown in figure 33A. The input impedance is approximately 1600 ohms. Feedback may be placed around such

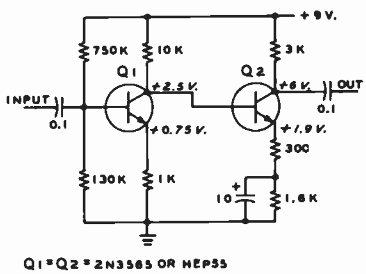


Figure 34

DIRECT-COUPLED TWO-STAGE AMPLIFIER

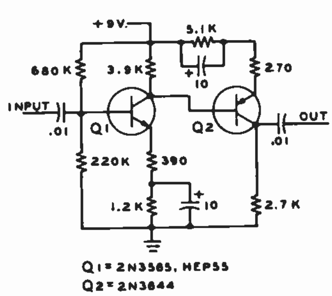


Figure 35

COMPLEMENTARY AMPLIFIER USING NPN AND PNP DEVICES

an amplifier from the collector of the second stage to the base of the first stage, as shown in figure 33B. A direct-coupled version of the resistance-coupled amplifier is shown in figure 34.

It is possible to employ NPN and PNP transistors in a common *complementary circuit* as shown in figure 35. There is no equivalent of this configuration in vacuum-tube technology. A variation of this interesting concept is the *complementary-symmetry* circuit of figure 36 which provides all the advantages of conventional push-pull operation plus direct coupling.

The Emitter Follower The *emitter-follower* configuration can be thought of as being very much like the vacuum-tube cathode follower, since both have a high input impedance and a relatively low output impedance. The basic emitter fol-

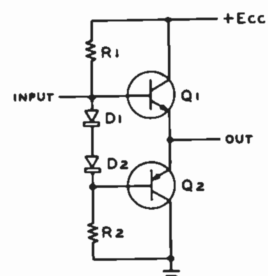


Figure 36

COMPLEMENTARY-SYMMETRY AMPLIFIER

Crossover distortion is reduced by use of diodes D_1 and D_2. Forward voltage drop in diodes is equal to the emitter-base forward voltage drop of transistors Q_1 and Q_2.

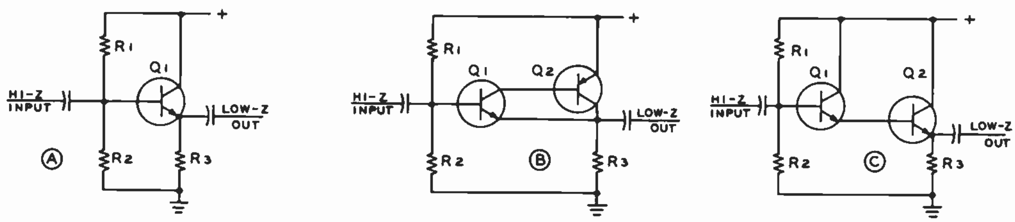


Figure 37

EMITTER-FOLLOWER CIRCUITS

A—Output voltage of emitter-follower is about 0.7 volt below input voltage
B—Complementary emitter follower
C—Darlington pair emitter follower. Q_1 and Q_2 are often on one chip

lower is shown in figure 37A. The output voltage is always 0.6 to 0.7 volt below the input (for silicon small-signal devices) and input and output impedances are approximately related by h_{fe}, the current gain of the transistor. Thus, a simple emitter follower with an emitter resistance of 500 ohms using a transistor having an h_{fe} of 150 can have an input impedance of over 75,000 ohms. A complementary emitter follower is shown in figure 37B.

A variation of the emitter-follower design is the *Darlington pair* (figure 37C). This arrangement cascades two emitter-follower stages with d-c coupling between the devices. Darlington-pair-wired dual transistors in monolithic form (for near-perfect temperature tracking) are available in both NPN and PNP pairs, even for power applications. A disadvantage of the Darlington pair emitter follower is that there are two emitter-base diode voltage drops between input and output. The high equivalent h_{fe} of the Darlington pair, however, allows for very large impedance ratios from input to output.

For power output stages another type of emitter follower is often used. A *push-pull* complementary emitter follower is shown in figure 38A. This circuit exhibits an inherent distortion in the form of a "dead zone" which exists when the input voltage is too low to turn on transistor Q_1 and too high to turn on transistor Q_2. Thus, a sine wave would be distorted so as to appear as shown in figure 38B. The circuit of figure 36 corrects this problem by making the forward voltage drop in diodes D_1 and D_2 equal to the emitter-base forward voltage drop of transistors Q_1 and Q_2.

Power-Amplifier Circuits The transistor may also be used as a class-A power amplifier as shown in figure 39.

Commercial transistors are available that will provide five or six watts of audio power when operating from a 22½-volt supply. The smaller units provide power levels of a few milliwatts. The correct operating point is chosen so that the output signal can swing equally in the positive and negative directions, as shown in the collector curves of figure 39B.

The proper primary impedance of the output transformer depends on the amount of power to be delivered to the load:

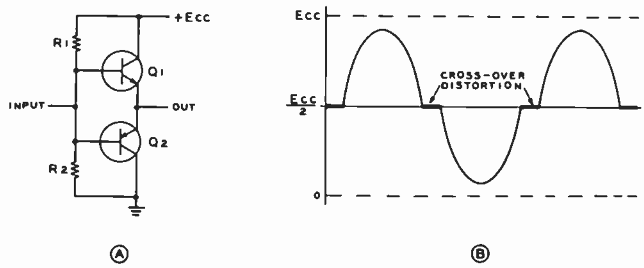


Figure 38

PUSH-PULL EMITTER-FOLLOWER OUTPUT STAGE

A—Crossover distortion exists when input voltage is too low to turn on Q_1 and too high to turn on Q_2.
B—Waveform distortion. Circuit of figure 36 corrects this problem.

$$R_p = \frac{E_c^2}{2P_o}$$

The collector current bias is:

$$I_c = \frac{2P_o}{E_c}$$

In a class-A output stage, the maximum a-c power output obtainable is limited to 0.5 the allowable dissipation of the transistor. The product I_cE_c determines the maximum collector dissipation, and a plot of these values is shown in figure 39B. The load line should always lie under the dissipation curve, and should encompass the maximum possible area between the axes of the graph for maximum output condition. In general, the load line is tangent to the dissipation curve and passes through the supply-voltage point at zero collector current. The d-c operating point is thus approximately one-half the supply voltage.

The circuit of a typical push-pull class-B transistor amplifier is shown in figure 40A. Push-pull operation is desirable for transistor operation, since the even-order harmonics are largely eliminated. This permits transistors to be driven into high collector-current regions without distortion normally caused by nonlinearity of the collector. Crossover distortion is reduced to a minimum by providing a slight forward base bias in addition to the normal emitter bias. The base bias is usually less than 0.5 volt in most cases. Excessive base bias will boost the quiescent collector current and thereby lower the over-all efficiency of the stage.

The operating point of the class-B amplifier is set on the $I_C = 0$ axis at the point where the collector voltage equals the supply voltage. The collector-to-collector impedance of the output transformer is:

$$R_{c\text{-}c} = \frac{2E_c^2}{P_o}$$

In the class-B circuit, the maximum a-c power input is approximately equal to three times the allowable collector dissipation of each transistor. Power transistors, such as the 2N514 have collector dissipation ratings of 80 watts and operate with class-B efficiency of about 67 percent. To achieve this level of operation the heavy-duty transistor relies on efficient heat transfer from the transistor case to the chassis, using the large thermal capacity of the chassis as a *heat sink*. An infinite heat sink may be approximated by mounting the transistor in the center of a 6″ × 6″ copper or aluminum sheet. This area may be part of a larger chassis.

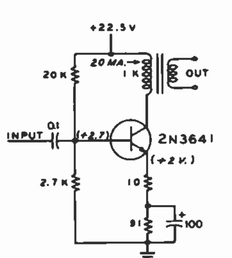
The collector of most power transistors is electrically connected to the case. For applications where the collector is not grounded a thin sheet of mica may be used between the case of the transistor and the chassis.

The "Bootstrap" Circuit The bipolar transistor in common-emitter configuration presents a low input impedance unsuitable for use with high-impedance driving sources such as a crystal microphone or a diode voltmeter probe. The *bootstrap* circuit of figure 41 provides a very high input impedance for these special circuits. The low-impedance base-bias network is isolated from the input circuit by the 100K resistor. The signal is fed to the base of the transistor and the output signal, taken across the emitter resistor, is also coupled to the bottom of the 100K isolating

Figure 39

TYPICAL CLASS-A AUDIO AMPLIFIER

Operating point is chosen so that output signal can swing equally in a positive or negative direction without exceeding maximum collector dissipation.



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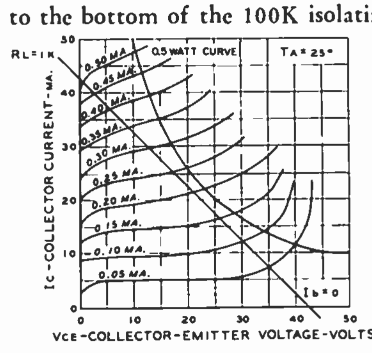

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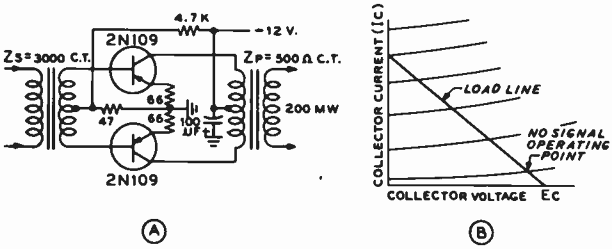


Figure 40

CLASS-B AUDIO AMPLIFIER CIRCUITRY

resistor via a capacitor. When a signal appears at the base, it also appears at the emitter in the same phase and almost the same amplitude. Thus, nearly identical signal voltages appear at the ends of the isolating resistor and little or no signal current flows through it. The resistor then resembles an infinitely high impedance to the signal current, thus effectively isolating the base-bias resistors. Since the isolating resistor has no effect on the bias level, the base bias remains unchanged. In practice, the signal voltage at the emitter is slightly less than at the base, thus limiting the over-all effectiveness of the circuit. For example, if the emitter-follower voltage gain is 0.99, and the value of the isolating resistor is 100K, the effective resistance to the a-c input signal is 100K raised to 10 megohms, an increase in value by a factor of 100 times.

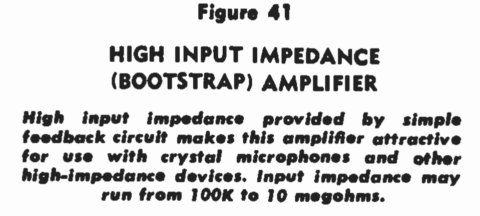


Figure 41

HIGH INPUT IMPEDANCE (BOOTSTRAP) AMPLIFIER

High input impedance provided by simple feedback circuit makes this amplifier attractive for use with crystal microphones and other high-impedance devices. Input impedance may run from 100K to 10 megohms.

4-8 R-F Circuitry

The bipolar transistor, almost from its commercial inception, proved to be operable up into the h-f range. The device has been refined and improved to the point where, now, operation into the gigahertz region is feasible. External feedback circuits are often used to counteract the effects of internal transistor feedback and to provide more stable performance at high gain figures. It should be noted, however, the bipolar transistor is not like a vacuum tube or FET device and must have its base-emitter junction forward-biased to display gain. The result of this requirement is that the driving stage is driving a nonlinear diode into forward conduction by the r-f signal intended to be amplified. This indicates the bipolar device is a nonlinear amplifier, to a greater or lesser degree. If the bipolar transistor is only required to amplify one frequency at a time, and that frequency is of constant amplitude, the bipolar transistor makes a satisfactory amplifier. When an ensemble of signals of different frequencies and/or amplitudes is present, the typical bipolar device will demonstrate the effect of its inherent nonlinearity in a high level of cross-modulation distortion. The fact the bipolar transistor exhibits such nonlinearity makes it useful as a frequency multiplier and mixer.

The severity of the nonlinearity of a bipolar device depends to a degree upon how it is used in a given circuit. The current gain

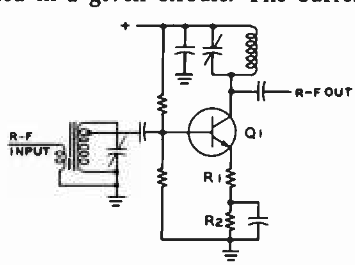


Figure 42

COMMON EMITTER R-F AMPLIFIER

Linearity is improved by leaving a portion of the emitter resistor unbypassed. Stage gain and cross-modulation are both reduced.

(h_{fe}) of a transistor drops rapidly with increasing frequency (figure 21) and the tendency is to use the transmitter in a common-emitter configuration to optimize gain. This circuit configuration also unfortunately optimizes nonlinearity. The common emitter circuit may be improved by leaving a portion of the emitter resistor unbypassed as shown in figure 42. This reduces stage gain, but also reduces nonlinearity and resultant crossmodulation problems to a greater degree. The unbypassed emitter resistor also boosts the input impedance at the base of the amplifier.

R-F Amplifiers A representative common-base r-f amplifier is shown in figure 43. This configuration generally has lower gain than the common-emitter circuit and is less likely to require neutralization. The linearity is better than that of the common-emitter circuit because of matching considerations. The input impedance of a common-base amplifier is in the region of 50 ohms, so no voltage step up is involved in matching the transistor to the common 50-ohm antenna circuit. In the common-emitter stage the input impedance of a small h-f transistor is about 500 ohms and a step-up impedance network must be used, causing the base voltage to be higher and aggravating the crossmodulation problem.

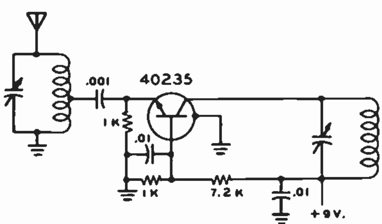


Figure 43

COMMON-BASE R-F AMPLIFIER

Linearity of this circuit is better than that of common-emitter configuration.

The relatively low gain of the common-base circuit may not be a detriment for h-f operation because good receiver design calls for only enough gain to overcome mixer noise at the frequency of operation.

Mixers and Converters As mentioned previously, the bipolar transistor is an inherently nonlinear device and, as such, can be used as an effective mixer or converter. Figure 44 shows two widely used transistor mixer circuits. The local oscillator signal can be injected into the base circuit in parallel with the r-f signal, or injected separately from a low-impedance source into the emitter circuit. The mixer products appear in the collector circuit and the desired one is taken from a selective output circuit.

A single transistor may be used in an *autodyne converter* circuit, as shown in figure 45. This is a common-emitter mixer with a tuned feedback circuit between emitter and collector and is often used in inexpensive transistorized broadcast receivers. The circuit has only economy to recommend it and often requires selection of transistors to make it oscillate.

Transistor Oscillators The bipolar transistor may be used in the oscillator circuits discussed in Chapter 11 (*Generation of Radio Frequency Energy*). Because of the base-emitter diode, the oscillator is of the self-limiting type, which produces a waveform with high harmonic content. A representative NPN transistor oscillator circuit is shown in figure 46. Sufficient coupling be-

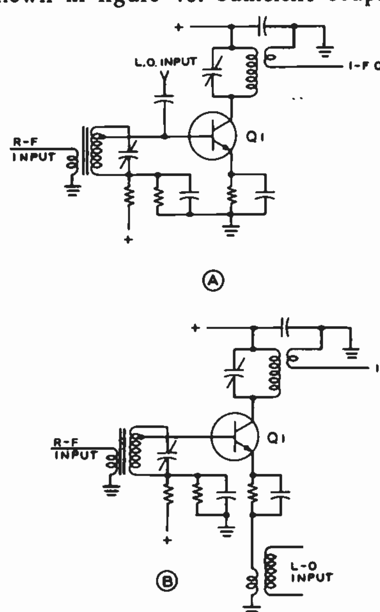


Figure 44

REPRESENTATIVE MIXER CIRCUITS

A—Base circuit injection of local oscillator.
B—Emitter injection from low-impedance source.

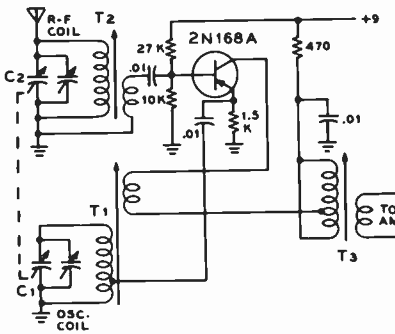


Figure 45

THE AUTODYNE CONVERTER CIRCUIT USING A 2N168A AS A MIXER

tween input and output circuits of the transistor via collector-base capacitance or via external circuitry will permit oscillation up to or slightly above the alpha-cutoff frequency.

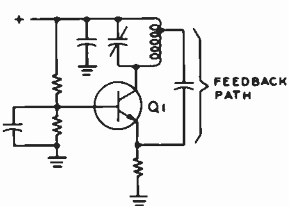


Figure 46

NPN OSCILLATOR CIRCUIT

External feedback path permits oscillation up to approximately the alpha-cutoff frequency of device.

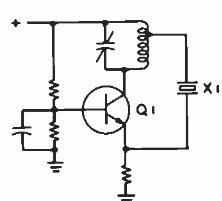


Figure 47

SERIES-MODE TRANSISTOR OSCILLATOR

Crystal is placed in feedback path and oscillates in series mode.

Because of the relatively low impedance associated with bipolar transistors, they are best used with crystals operating in the series mode, as shown in figure 47. If a standard parallel-mode type crystal is used in one of these series circuits, it will oscillate at its series-resonant frequency which is slightly lower than that frequency marked on the holder.

Transistor Detectors The bipolar device can be used as an amplitude detector, very much as a diode is used since the emitter-base junction is, after all, a diode. The transistor detector offers gain, however, since current passed by the base-emitter diode is multiplied by the factor h_{fe}. The detected signal is recovered at the collector. Since germanium transistors have a lower forward conduction voltage than silicon types, they are often used in this circuit. This allows the detector to operate on a few tenths of a volt (peak) as opposed to about 0.6 volt (peak) required for a silicon transistor. The bipolar transistor can also be used as a product detector for SSB and c-w, such as shown in figure 48.

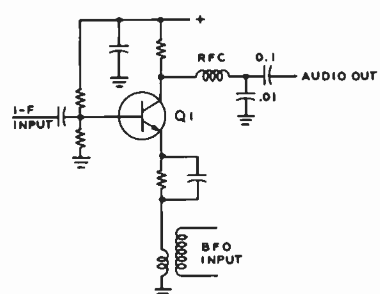


Figure 48

PRODUCT DETECTOR

Bfo is injected into the emitter circuit from a low-impedance source. Audio is recovered in the collector circuit.

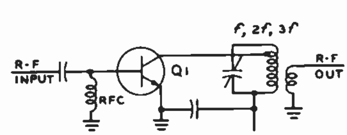


Figure 49

CLASS-C AMPLIFIER OR DOUBLER

Automatic Gain Control The gain of a transistor amplifier stage will decrease as the emitter current is decreased. This property can be used to control the gain of an r-f or i-f amplifier strip so that weak and strong signals will produce the

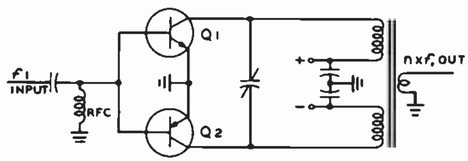


Figure 50

COMPLEMENTARY BASE-DRIVEN MULTIPLIER

Circuit may be considered to be either push pull or push push depending on phasing of the collector windings. Only one winding need be reversed to change mode of operation.

same audio output level. Automatic gain control voltage may be derived as described in Chapter 10 (*Radio Receiver Fundamentals*). If NPN transistors are used in the gain controlled stages, a negative agc voltage is required which reduces the fixed value of forward bias on the stage, decreasing the

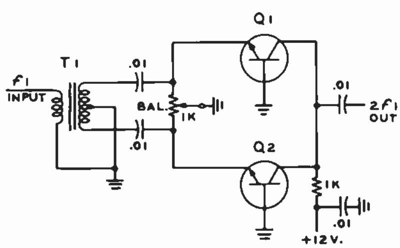


Figure 51

BROADBAND PUSH-PUSH DOUBLER

Balancing potentiometer permits attenuation of fundamental and third harmonic levels when circuit is used as a frequency doubler.

emitter current. If PNP transistors are used, a positive agc voltage is required.

Class-C Amplifier/ Multipliers As in vacuum-tube circuitry, class-C operation of a transistor implies conduction for less than 90 degrees of the operating cycle. This generally means zero-bias operation since it is necessary to place several tenths of a volt between emitter and base of a transistor (in forward conduction direction) before collector current flows. A typical class-C r-f stage is shown in figure 49; it can be either an amplifier or a frequency multiplier depending on the frequency to which the output circuit is adjusted.

The input and output impedances of such a class-C stage are generally quite low. Except for low power stages (under 100 mW or so) coupling networks other than the pi or tuned transformer are used, as discussed in the next section. This is not to say that the pi and tuned transformer will not work, but only that their element values often become unwieldy at the impedance levels and transformations encountered in solid-state circuitry.

Although single-transistor frequency multipliers are most common, it is possible to use the push-pull multiplier for high order odd multiples and the push-push multiplier for high order even multiples of the fundamental frequency (see Chapter 11, *Generation of R-F Energy*).

It is possible to build multipliers using bipolar transistors that are impossible to realize with tubes, because both NPN and PNP types of active devices are available.

Figure 50 shows a complementary base-driven frequency multiplier. It may be considered to be either a push-pull or a push-push configuration depending upon the phasing of the collector windings. Only one winding need be reversed to change from one design to the other since it is the balance of the circuit, in addition to the selectivity of the output tank, that attenuates adjacent harmonics in the output. A broadband h-f push-push doubler is shown in figure 51. In this configuration, the amplitude of the

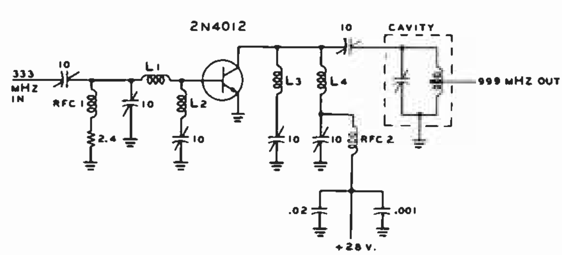


Figure 52

PARAMETRIC FREQUENCY MULTIPLIER

Bipolar transistor makes use of base-collector depletion capacitance to work as frequency multiplier. Idler circuits are used to reflect undesired harmonics back to collector-base capacitance.

fundamental and third-harmonic signals are respectively −28 db and 32 db below the level of the second harmonic output signal.

A second mechanism that may be used for frequency multiplication makes use of the base-collector depletion capacitance and is called *parametric multiplication* (figure 52). A number of idler circuits are used to reflect undesired harmonics back to the collector-base capacitance.

4-9 Silicon Power Transistors

Most high-frequency power transistors are silicon, planar, diffused NPN structures having a high ratio of active to physical area. Upwards of 200 watts average power at frequencies in the neighborhood of 450 MHz may be handled by modern silicon power transistors of advanced design. In the coming decade the efficiency, power gain, and temperature stability of these devices will lead to their use in many r-f amplifier applications heretofore solely reserved for electron tubes.

Circuit Considerations The power output capability of a transistor is determined by current and voltage limitations at the frequency of operation. The maximum current capacity is limited by emitter area and layer resistivity, and the voltage-handling capacity is limited by maximum breakdown limits imposed by layer resistivity and by the penetration of the junction. The *high-frequency current gain* figure of merit (f_T) defines the frequency at which the current gain is unity, and a high value of f_T at high emitter or collector current levels characterize a good r-f transistor.

In many cases, components and construction techniques used for vacuum tubes are not appropriate for transistor circuits. This variance in circuit considerations results mainly because of the lower circuit impedances encountered in transistor circuits. The most troublesome areas are power dissipation and parasitic oscillation. In the case of power dissipation, the levels reached under a given r-f power input are considerably higher than equivalent levels achieved under d-c operating conditions, since the junction temperature is a complex function of device dissipation, which includes r-f losses introduced in the pellet mounting structure. The package, then, is an integral part of the r-f power transistor having thermal, capacitive, and inductive properties. The most critical parasitic features of the package are the emitter and base lead inductances. These undesired parameters can lead to parasitic oscillations, most of which occur at frequencies *below* the frequency of operation because of the increased gain of the transistor at lower frequencies. Because transistor parameters change with power level, instabilities can be found in both common-emitter and common-base circuits. Some of the more common difficulties are listed below:

Parametric Oscillation—Parametric instability results because the transistor collector-base capacitance is nonlinear and can cause low-frequency modulation of the output frequency. This effect can be suppressed by careful selection of the bypass capacitors, and by the addition of a low-frequency bypass capacitor in addition to the high-frequency bypass capacitor (figure 53).

Low Frequency Oscillation—With transistor gain decreasing at about 6 decibels per octave, any parasitic low-frequency circuit can cause oscillation. Inadequate bypassing plus the use of high-Q, resonant r-f chokes can lead to this difficulty. This effect can be eliminated by placing small resistances in series with the r-f choke, or by the use of low-Q chokes of the ferrite-bead variety.

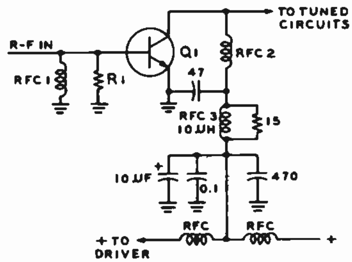


Figure 53

WIDEBAND DECOUPLING CIRCUIT FOR POWER TRANSISTOR

To suppress parametric oscillation collector bypass circuit must be effective at very low frequencies. Multiple bypass capacitors and series r-f chokes provide an adequate filter when used in conjunction with regular h-f and vhf filtering techniques.

Hysteresis—Hysteresis refers to discontinuous mode jumps in output power that occur when the input power or operating frequency is increased or decreased. This is caused by dynamic detuning resulting from nonlinear junction capacitance variation with change in r-f voltage. The tuned circuit, in other words, will have a different resonant frequency for a strong drive signal than for a weak one. Usually, these difficulties can be eliminated or minimized by careful choice of base bias, by proper choice of ground connections, and by the use of transistors having minimum values of parasitic capacitance and inductance. Circuit wiring should be short and direct as possible and all grounds should be concentrated in a small area to prevent chassis inductance from causing common-impedance gain degeneration in the emitter circuit. In common-emitter circuits, stage gain is dependent on series emitter impedance and small amounts of degeneration can cause reduced circuit gain at the higher frequencies and permit unwanted feedback between output and input circuits.

Thermal Considerations All semiconductor devices are temperature sensitive to a greater or lesser degree and the operating temperature and power dissipation of a given unit must be held below the maximum specified rating either by limiting the input power or by providing some external means of removing the excess heat generated during normal operation. Low power devices have sufficient mass and heat dissipation area to conduct away the heat energy formed at the junctions, but higher power devices must use a *heat sink* to drain away the excess heat.

Transistors of the 200-watt class, for example, have a chip size up to ¼-inch on a side and the excess heat must be removed from this very small area. For silicon devices, the maximum junction temperature is usually in the range of 135°C to 200°C. The heat generated in the chip is passed directly to the case through the collector-case bond.

The heat sink is a device which takes the heat from the transistor case and couples it into the surrounding air. Discrete heat sinks are available in various sizes, shapes, colors and materials. It is also common practice to use the chassis of the equipment as a heat sink. The heat dissipation capability of the heat sink is based on its *thermal resistance*, expressed in degrees per watt, where the watt is the rate of heat flow. Low power semiconductor devices commonly employ a clip-on heat sink while higher power units require a massive cast-aluminum, finned, radiator-style sink.

The interface between transistor case and sink is extremely important because of the problem of maintaining a low level of thermal resistance at the surfaces. If it is required to electrically insulate the device from the sink a mica washer may be used as an insulator and the mounting bolts are isolated with *nylon* or *teflon* washers. Some case designs may have a case mounting stud insulated from the collector so that it can be connected directly to the heat sink.

If the transistor is to be soldered into the circuit, the lead temperature during the soldering process is usually limited to about 250°C for not more than 10 seconds and the connections should not be made less than 1/32 inch away from the case.

The use of a thermal conductive compound such as a zinc-oxide, silicone compound (*Corning PC-4*), for example is recommended to fill the air insulating voids between the transistor case and the sink to achieve maximum heat transfer across the interface.

Figure 54 is a nomograph for obtaining the physical dimensions of a heat sink as a function of its thermal resistance. The data pertain to a convection- and radiation-cooled sink that is unpainted.

Input Circuits Once the dynamic input impedance has been determined from published data or from measurements, the input circuit may be designed. In practice, the input circuit must provide a match between a source impedance that is high compared to the input impedance of the transistor, which may be of the order of a few tenths of an ohm. Lumped LC circuits are used in the high-frequency region and air-line or strip-line circuits are used in the vhf region, as shown in figure 55.

The reactive portion of the input circuit is a function of the transistor package inductance and the chip capacitance; at the

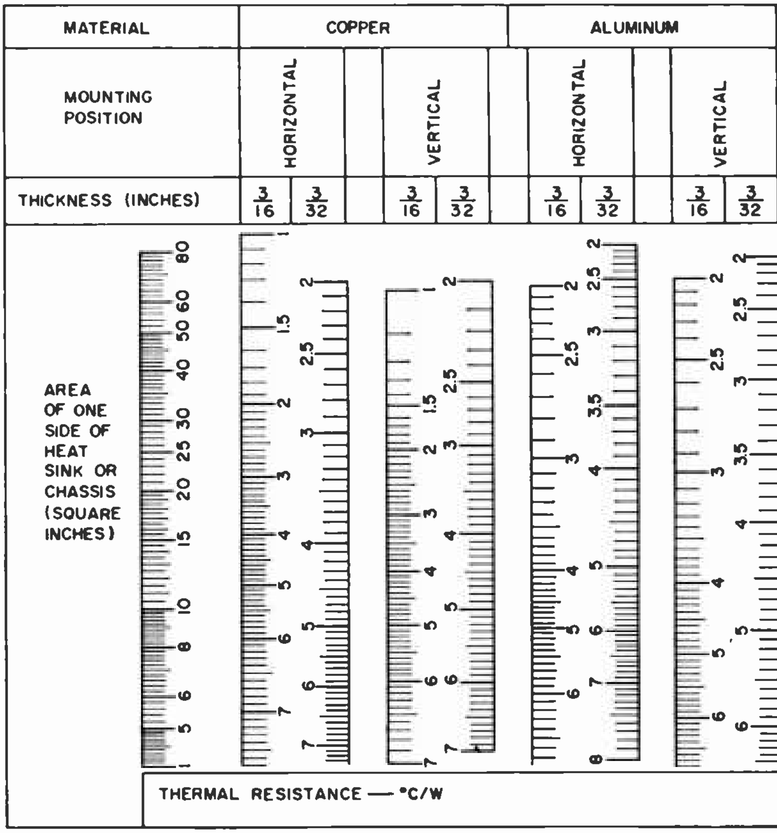


Figure 54

DIMENSIONS OF HEAT SINK AS FUNCTION OF THERMAL RESISTANCE

lower frequencies the input impedance is capacitive, and at the higher frequencies it becomes inductive; at some discrete intermediate frequency, it is entirely resistive. The inductive reactance present at the higher frequencies may be tuned out by means of a line section presenting capacitive reactance to the transistor. This advantageously results in an appreciable increase in over-all line length, as compared to the more common quarter-wave matching transformer (figure 55D).

For the low and medium-high frequencies where the input impedance of a power transistor is capacitive, the interstage networks shown in figure 56 are commonly used. The interstage network must tune out the capacitive portion of the input impedance (C_i) of the driven stage and provide a resistive load (R_1) to the collector of the driver stage. The collector of the driving transistor in each case is shunt-fed by a high impedance r-f choke.

At the very high frequencies, the input impedance of a power transistor is commonly inductive and the interstage network of figure 57 is often used. A representative 20-watt, 150-MHz silicon device may have a series input impedance of about $1 + j2$ ohms. Because of the low input impedance, network design and assembly is critical and care should be taken to observe the high circulating currents flowing in the final network loop, particularly through the shunt capacitance (C_3). Current values in the amperes range may flow through this capacitor at drive powers of well less than 5 watts or so. Special ceramic microwave capacitors

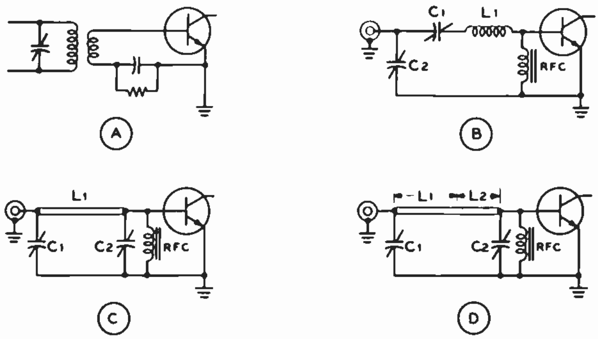


Figure 55

COMMON-EMITTER INPUT CIRCUITRY

Gain of common-emitter circuit is very dependent on emitter series impedance which should be low. Base input impedance is usually less than one ohm and a matching circuit must be provided from a source impedance that is high compared to input impedance. A low-impedance inductive circuit (A) may be used, or various tuned networks that combine impedance transformation with rejection of harmonic frequencies (B). A linear pi network is shown at C. If the input circuit is inductive, the reactance may be tuned out by means of a line section (L_2) that presents a capacitive reactance to the transistor (D).

having an extremely high value of Q and low lead inductance are available for configurations of this type. The low-loss porcelain units are expensive, but their cost is still small compared to the expensive transistors needed to produce appreciable power at the very high frequencies.

Output Circuits In most transistor power amplifiers, the lead impedance (R_L) presented to the collector is dictated by the required power output and the allowable peak d-c collector voltage, and thus is not made equal to the output resistance of the transistor. The peak a-c voltage is always less than the supply voltage and the collector load resistance may be expressed as:

$$R_L = \frac{(V_{CC})^2}{2 \times P_o}$$

$X_{L1} > X_{C1}$ AND $R1 > R2 = r_{bb'}$

1) $X_{L1} = Q_L R2 = Q_L r_{bb'}$

2) $X_{C1} = X_{Co}\left[\sqrt{\frac{(Q_L^2+1)\, r_{bb'}}{R1}} - 1\right]$

3) $X_{C2} = \frac{r_{bb'}(Q_L^2+1)}{Q_L} \cdot \frac{1}{1-\sqrt{\frac{R1 \cdot r_{bb'}(Q_L^2+1)}{X_{Co}^2 \cdot Q_L^2}}}$

$R1 > R2 = r_{bb'}$

1) $X_{L1} = \frac{R1}{Q_L}$

2) $X_{L2} = \frac{R2}{Q_L} \cdot \frac{\sqrt{\frac{R1}{R2}} - 1}{1 - \frac{R1}{Q_L \cdot X_{Co}}}$

3) $X_{C1} = \frac{R1}{Q_L} \cdot \frac{1 - \sqrt{\frac{R2}{R1}}}{1 - \frac{R1}{Q_L \cdot X_{Co}}}$

4) $X_{C2} = \frac{R1}{Q_L} \cdot \frac{\sqrt{\frac{R2}{R1}}}{1 - \frac{R1}{Q_L \cdot X_{Co}}}$

(A) **Figure 56** (B)

TRANSISTOR INPUT CIRCUIT COUPLING NETWORKS

A—Driver transistor shunt fed by high-impedance r-f choke. B—Driver transistor is parallel-tuned and coil L_1 serves in place of r-f choke.

where,

V_{CC} equals supply voltage,
P_o equals peak power output.

The nonlinear transfer characteristic of the transistor and the large dynamic voltage and current swings result in high-level harmonic currents being generated in the collector circuit. These currents must be suppressed by proper design of the output coupling network, which offers a relatively high impedance to the harmonic currents and a low impedance to the fundamental current (figure 58). Parallel-tuned, or pi-network circuitry may be used, with the reactive component of the output admittance tuned out by proper design of the series choke (RFC_1). At the lower frequencies, the collector of the transistor may be tapped down the tank coil as shown in the illustration. Capacitor C_1 provides tuning, and capacitor C_2 provides load matching. If the value of the inductor is properly chosen, harmonic suppression may be adequate. A form of this circuit is shown in figure 59, which provides better harmonic suppression with proper collector-circuit loading.

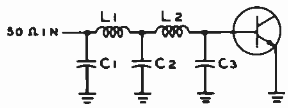


Figure 57

TRIPLE L-NETWORK INPUT CIRCUIT

Network steps down 50-ohm termination to low input impedance of base circuit. In the vhf region, the input impedance is commonly inductive, making up the missing series inductance of the third L network.

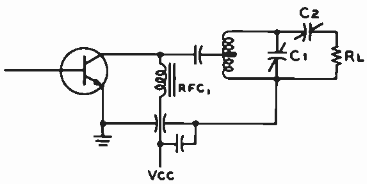


Figure 58

TRANSISTOR OUTPUT MATCHING CIRCUITRY

The reactive component of the output circuit of the transistor stage may be tuned out by proper design of the collector r-f choke (RFC_1). Tuning is accomplished by capacitor C_1 and load matching by capacitor C_2.

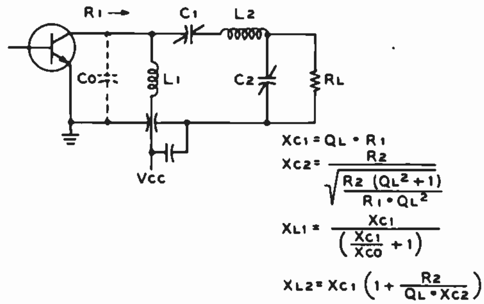
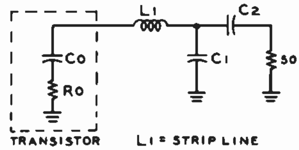


Figure 59

TRANSISTOR OUTPUT COUPLING NETWORK

This circuit provides proper collector loading and suppresses collector harmonic currents. The formulas for determination of constants are given in the illustration.

A second output network, especially suited to the vhf strip-line circuitry is shown in figure 60. Typically, the measured output impedance of a 25-watt, vhf power transistor is of the order of $3 + j2$ ohms. This compares favorably with the formula derived value given earlier in this section. The values of resistive and capacitive reactance may be used in figure 60 to determine the component parameters of the network. A circuit Q of 2 to 5 is usually chosen to provide adequate harmonic attenuation and practical component values. Component values and a schematic of a 40-watt, 175-MHz three stage



IF PARAMETERS ARE GIVEN IN PARALLEL FORM (R_P *AND* X_P):

$$R_O = \frac{R_P}{1+\left(\frac{R_P}{X_P}\right)^2} \quad \text{AND} \quad X_{CO} = R_O\left(\frac{R_P}{X_P}\right)$$

1) *LET* $B = R_O(1+Q_L^2)$

$$A = \sqrt{\frac{B}{R_L} - 1}$$

2) $X_{C2} = A\,R_L$

3) $X_{C1} = \frac{B}{Q_L - A}$

4) $X_{L1} = (Q_L \cdot R_O) + X_{C1}$

Figure 60

TRANSISTOR OUTPUT CIRCUIT COUPLING NETWORK

Network steps up low output impedance of collector to 50-ohm termination.

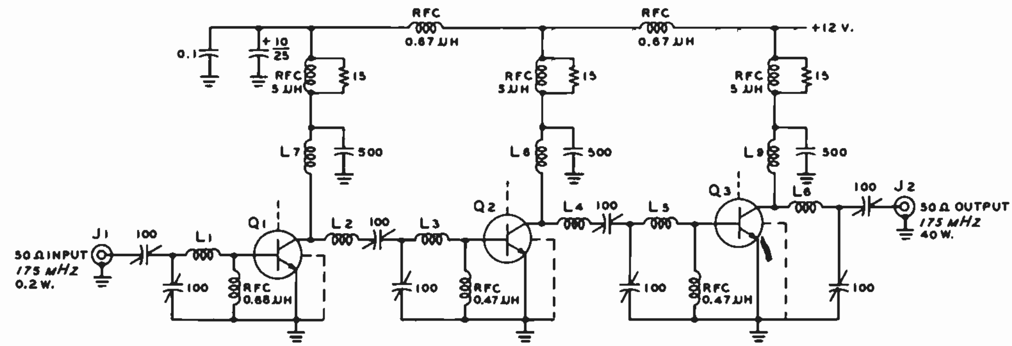


Figure 61

40-WATT, 175-MHz THREE STAGE AMPLIFIER

L_1—#16 wire, ½-inch long
L_2, L_4—#16 wire about ¾-inch long formed into "U"
L_3, L_5—¼" x ⅛" strap, .005" thick about ½" long
L_6—½" x ¼" strap, .005" thick about ½" long
L_7—8 turns #16 e., ¼" diam.
L_8—6 turns, as L_7
L_9—5 turns, as L_7
Q_1—CTC type B3-12
Q_2—CTC type B12-12
Q_3—CTC type B40-12
Note: 100-pF capacitors are mica compression type). (All transistors by Communications Transistor Corp.)

amplifier using interstage circuits of this type are shown in figure 61.

Mode of Operation From the stability standpoint, the common-emitter configuration provides a more stable circuit at the higher frequencies than does the common-base circuit. Collector efficiency in either case is about the same. Generally speaking, breakdown voltages under r-f conditions are considerably lower than the normal d-c breakdown voltages, and the capability of the r-f power transistor to work into loads having a high value of SWR is limited. A well-designed circuit operated at low supply voltage where power gain is not excessive is found to be less prone to SWR mismatch. High values of SWR mismatch lead to excessive r-f peak voltages, poor efficiency, and instability.

Single-sideband, linear operation calls for class-AB transistor operation. Most high-frequency power transistors are designed for on-off (class-C) operation and the forward bias necessary to place them in a class-AB mode leaves them susceptible to *second breakdown*, a destructive phenomenon characterized by localized heating within the transistor pellet, which leads to a regenerative layer damage.

Second breakdown may be controlled by the addition of emitter resistance of low value. A compromise amount is usually chosen as excessive emitter resistance can limit power gain and output. Developmental transistors designed for linear amplifier service have emitter resistance in the chip, in amounts of a fraction of an ohm. Other transistor types may incorporate a zener diode on the chip to provide controlled, positive base voltage.

The forward bias must, in any event, be maintained over a wide temperature range to prevent an increase in idling current accompanied by a rise in chip temperature, which leads to a destructive runaway condition under maximum output conditions when transistor temperature is highest.

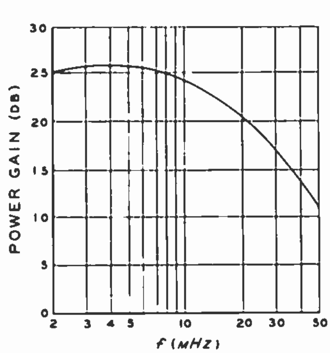
H-F Linear Power-Amplifier Design The operating parameters for linear service present severe circuit problems for the solid-state device, among which is the wide variation in the base input impedance, which may vary widely with frequency and tuning, because of the low value of impedance and the relatively large value of collector-base capacitance. A representative 50-watt transistor designed for linear service may have a series input impedance ranging from $4-j2$ ohms at 3.5 MHz to $0.5-j\,0.5$ ohms at 30 MHz.

The transistor for linear service should be chosen on the basis of good current-gain linearity at high values of collector current. A transistor having rapid h_{fe} falloff at high collector currents will generally have poor intermodulation distortion characteristics. In addition to good linearity, the device should have the ability to survive a mismatched load and maintain a low junction temperature at full power output. Transistors are available which combine these attributes, at power levels up to 100 watts PEP output, having intermodulation distortion levels of −30 db for the ratio of one distortion product to one of two test tones. Power gain and linearity are shown in figure 62 for the 2N5492 *Motorola* silicon transistor, specifically designed for linear amplifier service up to 30 MHz.

Operation of a solid-state linear amplifier at reduced collector voltage drastically reduces the maximum power output for a given degree of linearity since the device must deliver correspondingly higher collector peak currents for a given power output, thus placing a greater demand upon the h_{fe} linearity at high values of collector current.

Bias Considerations A typical class-C solid-state device is operated with both the base and emitter grounded and the transistor is cut off when no driving signal is present. The linearity of a solid-state device requires operation with forward bias, as stated previously. This implies a finite no-signal value of collector current. Optimum values of no-signal (quiescent) collector current range from 5 to 50 ma for devices in the 10- to 100-watt PEP range. Such values fall under the definition of class-B operation. Class-B operation is complicated by thermal runaway problems and large variations in the transistor base current as the r-f drive level is varied. For best linearity, the d-c base bias should remain constant as the r-f drive level is varied. This is in conflict with the conditions required to prevent thermal runaway. A representative bias circuit that meets these critical requirements is shown in figure 63. This circuit supplies an almost constant base bias by virtue of the zener diode (D_1) which is also used to temperature-compensate the transistor. The diode is thermally coupled to the transistor by mounting it on the same heat sink, thus providing temperature compensation due to its decrease in forward voltage drop with increasing temperature. Using this particular transistor, base current rises from the no-signal value of 3 ma to about 200 ma at 80 watts output with a two-tone test signal. The current through the diode at the no-signal condition is about 260 ma and when r-f drive is applied, the transistor receives its additional base current from the diode, since the voltage drop across the diode is always slightly greater than the base-emitter voltage of the transistor due to the voltage drop in choke RFC_1.

Resistor R_1 has a dual function in that it causes current to flow through RFC_1 in the no-signal condition and it also reduces the



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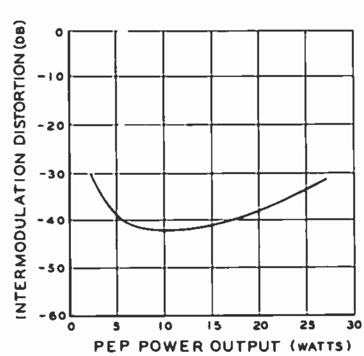

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Figure 62

POWER GAIN AND LINEARITY OF 2N5492

Motorola 2N5492 power transistor is designed for linear amplifier service up to 30 MHz and has intermodulation distortion level better than −30 db.

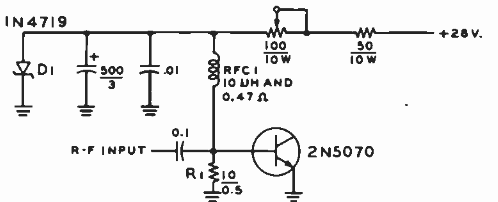


Figure 63

BASE BIAS CIRCUIT FOR 2N5070 IN LINEAR AMPLIFIER SERVICE

Zener diode D_1 is also used to temperature-compensate the transistor by mounting it on common heat sink.

impedance from base to ground, helping to improve the stability of the amplifier.

Wideband Circuitry The use of transmission-line type broadband transformers permits the construction of a wideband amplifier whose power gain versus frequency performance is shown in figure 62. The special transformers consisted of low-impedance, twisted wire transmission lines wound about a ferrite toroid. Impedance ratios of 4:1 or 9:1 may be achieved with the proper winding connections. Two series-connected transformers may be used to achieve greater step-down ratios, if required. The design of effective transformers for this class of service is covered in "Broadband 60-W HF Linear Amplifier," by Pitzalis, Horn, and Baranello in *IEEE Journal of Solid State Circuits*, Volume SC-6, No. 3; June, 1971. A representative broadband amplifier schematic is shown in figure 64.

4-10 VHF Circuitry

Power transistors are available that provide up to 150 watts power output to over 200 MHz and up to 100 watts power output to 500 MHz for class-C service. Experimental transistors can provide upward of 50 watts in class-C operation at frequencies in excess of 1000 MHz. These devices make practical, low cost solid-state power amplifiers for amateur f-m service up through 432 MHz.

Vhf power transistors are tailored for operation over certain popular frequency ranges (25-80 MHz, 100-200 MHz, or 200-600 MHz, for example) and the power capability and reliability require that the user operate the device within the intended range, since the ruggedness of the vhf power transistor is a function of *both* voltage and frequency. A transistor rated for operation near 175 MHz, will be less rugged at 100 MHz and may be too delicate for use at 30 MHz. In addition, the device must be operated well within the manufacturer's rating and due attention paid to the standing-wave ratio appearing on the transistor output load network.

For f-m service, the vhf transistor is operated in the zero bias, class-C mode and stripline circuitry is commonly employed.

Circuit Techniques Transistor input and output impedances are extremely low and stray circuit inductance and ground current return paths play a large role in circuit design. Impedance levels

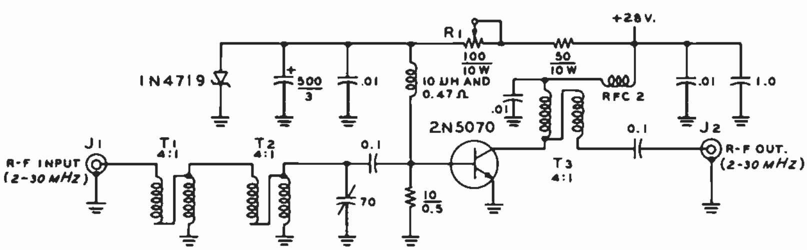


Figure 64

BROADBAND 2- TO 30-MHz LINEAR AMPLIFIER USING 2N5070

Nominal 50-ohm input is stepped down to the base impedance by series-connected 4:1 balun transformers. Single 4:1 balun transformer steps up collector impedance to 50-ohm level.

of one ohm, or less, are common and lead length in r-f circuitry of 0.1 inch or so become quite critical. Special vhf ceramic capacitors having ribbon leads may be used in impedance matching circuits and uncased mica/porcelain chip capacitors used for high r-f current paths. The technique of grounding the r-f components becomes a very critical aspect of the circuit design as a result of the very low impedance characteristics of the transistor.

The common-base or common-emitter lead should be grounded *at the body* of the transistor for proper performance. With the strip-line package, the device may be mounted to a ground plane (such as a

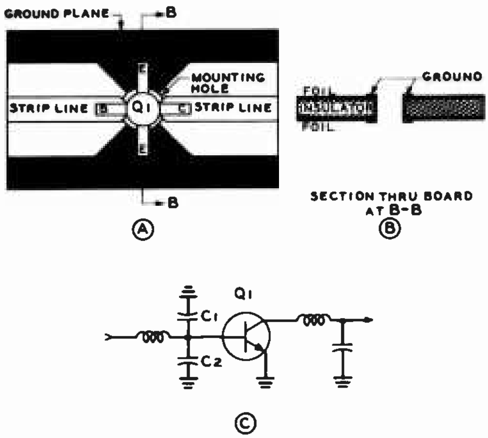


Figure 65

VHF TRANSISTOR MOUNTED IN STRIP-LINE CONFIGURATION

(A) Two emitter leads of transistor are connected to ground plane. Base and collector leads are soldered to resonant strip lines. Dual-surface board is used with top and bottom ground planes connected together with straps under each emitter lead (B). Small ceramic chip capacitors are often placed in parallel at base terminal to form portion of input matching network (C). Extremely low impedance to ground is required at this point because current flowing in capacitors is heavy.

printed-circuit board) as shown in figure 65. Dual-surface board is used, with the top and bottom ground planes connected together using straps under each emitter lead. Capacitors in the input matching network require a good ground and extremely low inductive impedance. Two small chip

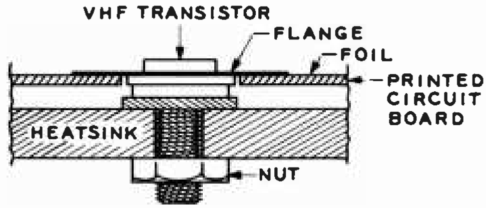


Figure 66

STUD-MOUNTED TRANSISTOR IS BOLTED TO HEAT SINK

Flange connections of transistor should not be twisted or bent. Printed-circuit board is elevated above the heat sink so that flange leads are not stressed and provide shortest possible connection to the strip line. Silicon grease is used on the stud to lower thermal resistance between transistor and heat sink.

capacitors are often used in parallel at this point, as shown in the illustration.

The stud-mounted transistor should be mounted on a flat surface (figure 66) for proper heat transfer. The flange connections should not be twisted or bent, and should not be stressed when the transistor is torqued to the heat sink. Silicone grease should always be used on the stud to lower the thermal resistance between transistor and sink.

The transistor user should remember that the vhf power transistor will not tolerate overload as the thermal time constant of the small chip is very fast, thus, the allowable dissipation rating of the transistor must be capable of handling momentary overloads. Generally speaking, for class-C operation, the r-f output level of the vhf power transistor should be held to about 50 percent of the power dissipation rating.

VHF Circuit Design Vhf transistor circuitry involves impedance matching networks and d-c feed systems. It is common practice to make networks up of simple, cascaded L-sections which provide low-pass filter characteristics and ample impedance transfer (figures 59 and 60). If the Q of each step of the network is held to a low figure (2 or 3), the bandwidth of the amplifier will be wide enough to cover any of the vhf amateur bands. Representative two-section networks for input and output terminations are shown in figure 67.

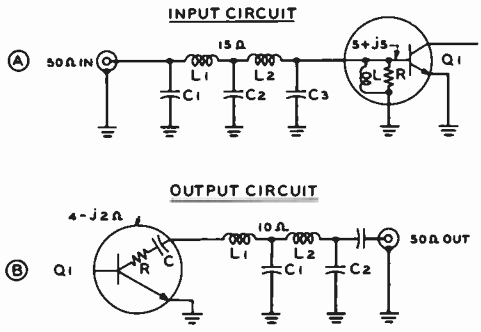


Figure 67

INPUT AND OUTPUT MATCHING NETWORKS

(A) Input impedance of vhf transistor, typically, is inductive. Two-section network with center impedance of 15 ohms matches 50-ohm input to the base circuit of the transistor. (B) Output impedance presents a low value of series reactance. Two-section network with center impedance of 10 ohms provides proper match to 50-ohm termination. Circuit Q of networks is held to 2 or 3 for optimum bandwidth.

The transistor input impedance in the vhf range is usually inductive and a shunt capacitor (circuit A, capacitor C_3) is used to cancel the reactive portion of the impedance. Two series-connected L-sections are used, the first matching the 50-ohm input impedance down to 15 ohms and the second matching down from 15 ohms to the 5-ohm impedance level of the transistor. The intermediate impedance point is often chosen as the mean value between the output and input impedance levels. If a strip-line configuration is used, line impedance may be taken as the mean value to simplify calculations.

The vhf transistor generally has a capacitive reactance and the proper load impedance is usually given by the manufacturer. A series inductance (circuit B, inductor L_1) equalizes the series capacitance of the device and two series-connected L-sections step the transistor impedance level up to 50 ohms.

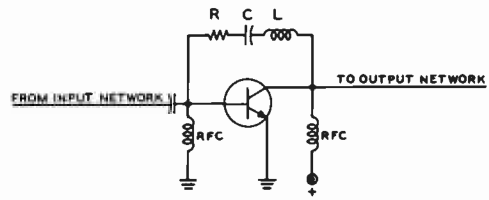


Figure 68

NEGATIVE COLLECTOR FEEDBACK DECREASES LOW-FREQUENCY STAGE GAIN

D-C Feed Systems Design The d-c feed network permits the operating voltages to be applied to the transistor without interfering with the r-f circuitry. Voltages may be fed to the transistor via r-f chokes, which must be carefully designed in order to prevent low-frequency parasitic oscillations. Transistor gain increases rapidly with decreasing signal frequency and a figure of 40 decibels is not uncommon for low-frequency gain. The d-c feed network therefore must present a load impedance which will not sustain low-frequency oscillation. This may be done by using as small r-f chokes as possible consistent with the operating frequency and impedance level and large bypass capacitors (figure 53).

In addition negative collector feedback can be used to decrease the stage gain below the design frequency (figure 68).

Part II—Field-Effect Devices, Integrated Circuits and Numeric Displays

4-11 Field-Effect Devices

The *junction field-effect transistor* (JFET), or unipolar transistors, was explored in 1928 but it was not until 1958 that the first practical field-effect transistor was developed. This device may be most easily visualized as a bar, or channel, of semiconductor material of either N-type or P-type silicon. An ohmic contact is made to each end of the bar as shown in figure 1A, which represents an N-type field-effect transistor in its simplest form. If two P-regions are diffused into a bar of N-material (from

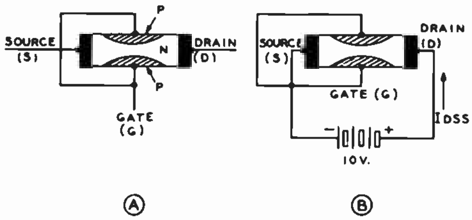


Figure 1
JUNCTION FIELD-EFFECT TRANSISTOR

A—Basic JFET is channel of N- or P-type material with contact at each end. Two P or N regions are diffused into the bar. B—If a positive voltage is applied across contacts a current flows through the gate region. Control of gate bias changes current flow from source contact to drain contact. Drain current is thus controlled by gate voltage.

opposite ends of the N-channel) and externally connected together electrically, a *gate* is produced. One contact is called the *source* and the other the *drain;* it matters not which if the gate diffusion is in the center of the device. If a positive voltage is applied between drain and source (figure 1B) and the gate is connected to the source, a current will flow. This is the most important definitive current in a field-effect device and is termed the *zero bias drain current* (I_{DSS}). This current represents the maximum current flow with the gate-source diode at zero bias. As the gate is made more negative relative to the source, the P-region expands cutting down the size of the N-channel through which current can flow. Finally, at a negative gate potential termed the *pinch-off voltage,* conduction in the channel ceases. The region of control for negative gate voltages lies between zero and the gate-to-source cutoff voltage ($V_{GS\text{-}off}$). These voltages cause the gate-source junction to be *back-biased,* a condition analogous to the vacuum tube, since drain current is controlled by gate voltage. In the vacuum tube a potential on the grid affects the plate current, however the charge carrying the signal does not flow in the region between cathode and plate to any significant extent.

It is possible to build a P-channel JFET device that requires a negative drain voltage and is biased with positive gate voltage. Combining both N-channel and P-channel JFET's makes it possible to design complementary circuits as in the manner previously

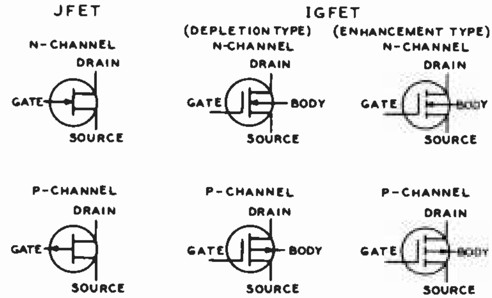


Figure 2

SYMBOLS AND NOMENCLATURE FOR FIELD-EFFECT TRANSISTORS

described for NPN and PNP bipolar transisters. The symbols used to depict N-channel and P-channel JFET's are shown in figure 2.

The *Insulated Gate Field-Effect Transistor* (IGFET) differs from the JFET in a number of ways. The gate element is insulated from the rest of the device and control is by means of capacitance variation. The IGFET may be visualized as in figure 3, again an N-channel device. The basic form of the device is P-type material, into which have been diffused two N-type regions to form the source and drain. The gate is a layer of metalization laid down directly over the P-type region between source and drain, but separated from the region by a thin layer of insulating silicon dioxide (silicon nitride is also used in some types). If a positive voltage is applied to the drain, relative

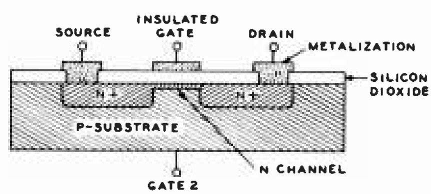


Figure 3

INSULATED-GATE FIELD-EFFECT TRANSISTOR

IGFET has insulated gate element and current control is by means of capacitance variation. Enhancement mode (positive gate control) and depletion mode (negative gate control) IGFETs are available. Gate voltage limitation is point of breakdown of oxide dielectric in the gate. Diode-protected IGFET has zener diodes on the chip to limit potential between gate and body of device.

to the source, and there is no potential difference between gate and substrate, no current will flow because the path appears as two back-to-back diodes (NP-PN). If a positive voltage is applied to the gate relative to the substrate, it will *induce* an N-region between source and drain and conduction will occur. This type of IGFET is termed an *enhancement mode* type; that is, application of forward bias to the gate enhances current flow from source to drain. (It is not possible

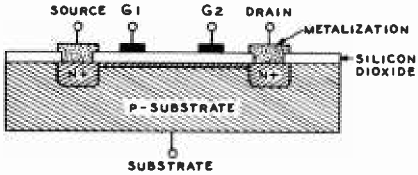


Figure 4

DUAL-GATE IGFET

Depletion type, dual-gate IGFET is intended for r-f use through the vhf range. One port is for input signal and the other for agc control.

to build an enhancement mode JFET because the gate is a diode which will conduct if forward-biased).

A *depletion mode* IGFET is built by diffusing a small N-region between the source

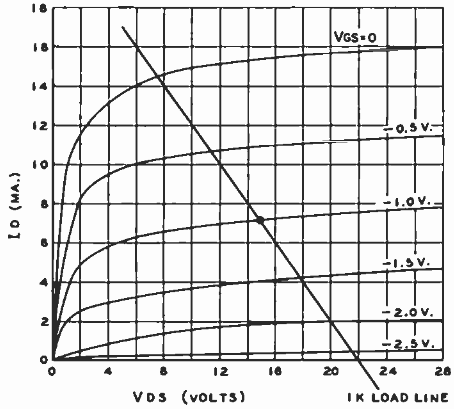


Figure 5

DRAIN CHARACTERISTICS OF E300 FET

Drain characteristic curves of FET resemble the characteristic curves of pentode vacuum tube as the current plots are nearly horizontal in slope above V_{DS} of about 6 volts. Load line is drawn on plot for gate bias of −1 volt, drain voltage of +15 volts, and drain current of 7 milliamperes.

and drain to cause conduction even if there is no voltage applied between gate and substrate. Similar to the JFET, this depletion mode IGFET must have its gate reverse-biased to reduce source-to-drain current. The depletion mode IGFET is used in the same manner as the JFET except that the gate may also be driven forward and the drain current can be increased to values even greater than the zero-bias drain current, I_{DSS}.

Gate voltage of the JFET is limited in the reverse direction by the avalanche breakdown potential of the gate-source and gate-drain circuits. In the IGFET, on the other hand, the gate voltage limitation is the point of destructive breakdown of the oxide dielectric under the gate. This breakdown must be avoided to prevent permanent damage to the oxide.

Static electricity represents the greatest threat to the gate insulation in IGFET devices. This type of charge accumulation can be avoided by wrapping the leads in tinfoil, or by otherwise connecting the leads when the devices are being transported and installed. The user of the device, moreover, may accumulate a static potential that will damage the IGFET when it is handled or installed and a grounding strap around the electrodes is recommended. Gate protection is often included within the device in the form of zener diodes on the chip between the gate and the body, forming a *diode-protected IGFET*.

FET Terminal Leads Note in figures 1 and 3 there are really four terminations associated with any FET device. In the JFET they are source, drain, and the two connections to the two P-diffusions made in the channel. In the IGFET they are source, drain, gate, and substrate. In some JFETs all four leads are brought out of the package and in others only three leads are available. In a three-lead configuration, it is considered that the two P-diffusion gate connections are tied together inside the package. In the case of the IGFET, all four leads are generally available for use; but more often than not, the substrate is externally connected to the source in the actual circuit. The advantage of the four-lead package is the ability to allow separate control ports, much like a multigrid vacuum tube.

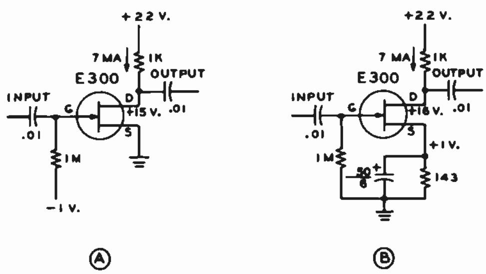


Figure 6

COMMON-SOURCE AMPLIFIERS USING E300 FET

Common-source amplifiers operating under conditions shown in figure 6. A—Separate gate bias. B—Source self-bias.

An improved *dual-gate IGFET* of the depletion type has recently become available, intended for r-f use through the vhf range. The 3N140, 3N141, and 40673 of RCA, and the Motorola MFE-3006 and MFE-3007 are representative types. Their construction is shown in figure 4. These devices serve where dual ports are required, such as in mixers, product detectors, and agc-controlled stages, with one gate used as the signal port and the other the control port.

4-12 FET Circuitry

JFET and depletion-mode IGFET devices are used in linear circuitry in very much the same way as are vacuum tubes, but at lower voltages. As an example, the drain characteristics of an inexpensive and popular FET (*Siliconix* E300) are shown in figure 5. The line that is labeled $V_{gs} = 0$ is the one that represents the zero-bias drain current state, or I_{dss}. At a drain to source potential of 10 volts, I_d is 15 milliamperes and, according to the data sheet, I_d could be any value between 6 ma and 30 ma at this potential. This *spread* of I_d is fairly typical of the lower cost FETs and the curve shown is also typical, as is the value of I_d read from it.

The E300 drain characteristics look very similar in shape to the characteristics of a pentode vacuum tube; that is, at V_{ds} (drain to source potential) greater than about 6 volts, the drain current curves are nearly horizontal in slope. The FET, then, like the pentode, is generally used in circuits in this so-called *constant current* region of the characteristics.

A 1000-ohm load line is drawn on the characteristic plot in the same manner as one is drawn on a vacuum-tube plate characteristic curve (see figure 17, Chapter 5 and figure 4, Chapter 7). The load line is marked for a gate-bias voltage of -1 volt, a drain voltage of $+15$ volts, and a resting drain current of 7 milliamperes. The circuit of a *common-source* amplifier operating under these conditions is shown in figure 6.

It can be observed from the load line that, at the bias point of -1 volt, as the input signal swings plus and minus 1 volt, the drain voltage will swing between $+8$ to $+20$ volts. The gate bias may be supplied either from a separate supply or from a source resistor (equivalent to a cathode resistor in vacuum-tube technology). Typical input impedance of the common-source small-signal audio amplifier is quite high, with 10 megohms a not-uncommon value for low leakage JFETs and values higher than this for IGFETs.

The *common-gate* configuration shown in figure 7 may be compared in performance to

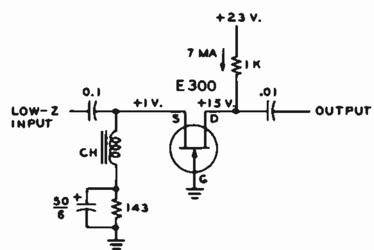


Figure 7

COMMON-GATE AMPLIFIER USING E300 FET

Input impedance of common-gate circuit is about 150 ohms. Stage gain is lower than common-source circuit.

the cathode-driven vacuum-tube amplifier, having a rather low value of input impedance. A typical value of input impedance is approximately $1/_{fs}$ where g_{fs} is the transconductance (similar to g_m in the vacuum tubes). The g_{fs} is approximately $1/g_{fs}$ for the E300 device is about 6600 microhmos; so the circuit of figure 7 will have an input impedance around 150 ohms. The common-gate configuration will have somewhat lower gain than the common-source circuit, but has other advantages discussed later.

The FET analogy to the cathode follower is shown in figure 8. This *source follower,* shown with self-bias, has a very high input impedance and very low output impedance $(1/g_{fs})$.

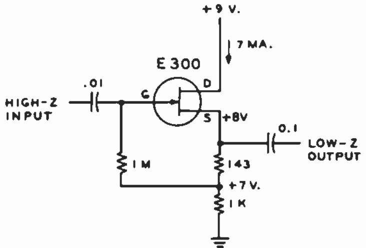


Figure 8

SOURCE-FOLLOWER AMPLIFIER USING E300 FET

Source-follower circuit has very high input impedance and low output impedance.

The FET in Specialized Circuits The FET makes a very good r-f device because of some of its unique characteristics. In particular, the FET has a transfer characteristic that is remarkably free of third-order curvature, which ensures that intermodulation distortion and crossmodulation will be at a minimum in a properly designed circuit. A typical IGFET (depletion mode) vhf r-f amplifier is shown in figure 9.

FET devices have second order curvature in their transfer functions and operate as good mixers having little intermodulation distortion. The use of FET devices in receivers is discussed in Chapter 10.

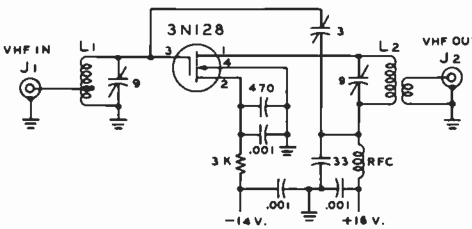


Figure 9

IGFET DEPLETION-MODE VHF AMPLIFIER

3N128 is neutralized for best circuitry stability and optimum noise figure.

Aside from common usage discussed elsewhere in this handbook, the characteristics of the FET permit it to do a good job in specialized circuits. A phase-shift audio oscillator using the HEP 801 is shown in figure 10. This configuration employs the *tapered RC network* wherein each RC pair has the same time constant but successively higher impedance. The bridge-T and Wien bridge circuits also adapt themselves easily to the FET as shown in figures 11 and 12.

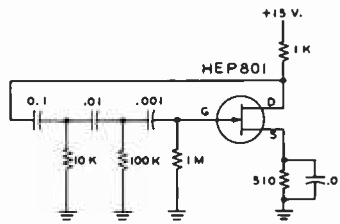


Figure 10

PHASE-SHIFT AUDIO OSCILLATOR WITH HEP 801

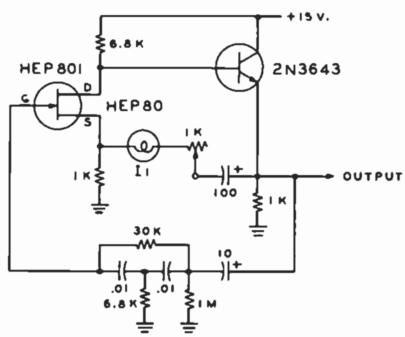


Figure 11

BRIDGE-T AUDIO OSCILLATOR USING HEP 801 AND 2N3643

I_1—Sylvania 120 MB lamp.

Since the FET is commonly operated in the constant-current region, it is often used as a constant-current generator with the gate and source connected together to form a two-terminal device. A linear ramp generator using a FET in place of a transistor to charge a capacitor is shown in figure 13. A

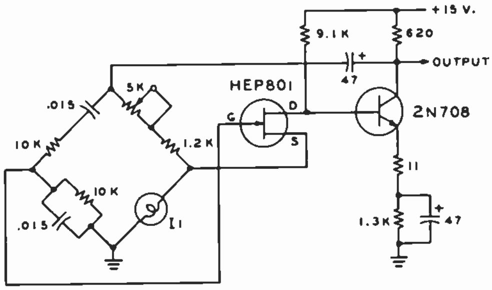


Figure 12

WIEN BRIDGE AUDIO OSCILLATOR USING HEP 801 AND 2N708

I_1—Sylvania 120 MB lamp.

unijunction transistor is used to discharge the capacitor.

A combination FET and zener diode circuit (figure 14A) provides improved regulation since the current flow through the zener is constant. Special JFETs that serve as *constant-current diodes* are available, but

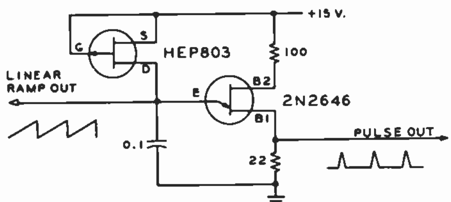


Figure 13

LINEAR RAMP GENERATOR

HEP 803 FET used as constant current source to generate linear ramp waveforms.

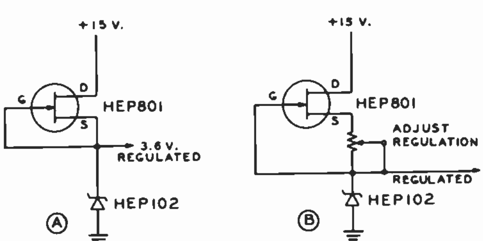


Figure 14

FET AND ZENER DIODE PROVIDE IMPROVED REGULATION

A—Constant current source. B—Variable current source.

the experimenter can use nearly any small JFET in a similar manner by connecting the gate to the source. If the FET is used with a variable resistance in the source lead, as shown in figure 14B, an adjustable but constant-current source is available.

The enhancement-mode IGFET (P-channel) is almost exclusively used as a switch for computing or for logic circuits and the basic building block upon which one form of logic integrated circuit is based, as discussed in a later chapter. Discrete en-

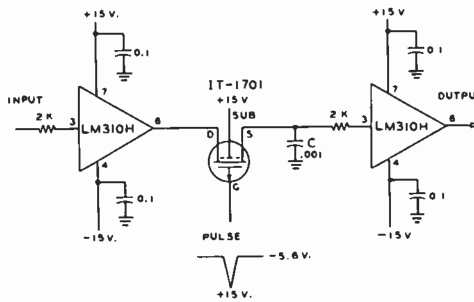


Figure 15

SAMPLE AND HOLD CIRCUIT WITH ENHANCEMENT MODE IGFET

Input waveform is sampled only when negative sample pulse applied between substrate and gate of IT-1701 IGFET is present. Capacitor C is then charged to value of input voltage and drives sensing amplifier through operational amplifier LM310H, at right. Capacitor holds charge because IGFET represents open circuit after pulse passes.

hancement-mode IGFETs are used in *sample and hold* circuits, such as shown in figure 15. The waveform at the input is sampled only when the negative sample pulse, applied between substrate and gate, is present. The capacitor (C) is then charged to whatever value the input received during the sample pulse, and holds this value because the IGFET represents an open circuit at all other times. The voltage on the capacitor may be used to drive another FET (depletion mode) so that the input impedance of the sensing amplifier does not discharge the capacitor to any degree during sampling times. The enhancement-mode IGFET also serves as a fast switch in chopper service or as a series switch in certain types of noise suppression devices.

As the technology of FET construction develops, JFETSs and IGFETs continue to invade new circuit areas. JFETs for 1-GHz operation are available and so are 10-watt stud-mounted types for lower-frequency power application. IGFETs are being designed for 1-GHz operation to satisfy the demands of UHF-TV reception. Some experimental FETs have been built to operate at 10 GHz. Other experimental JFETs available for low-frequency work can withstand 100 volts between source and drain.

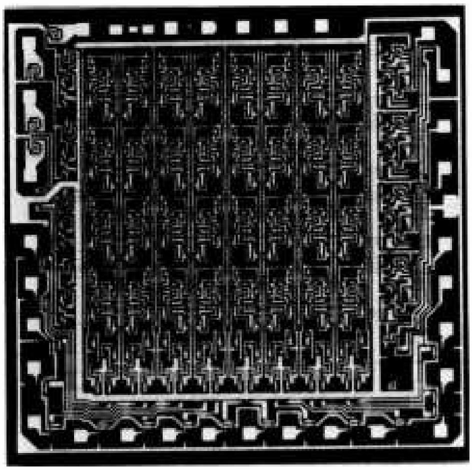


Figure 16

INTEGRATED CIRCUIT ASSEMBLY

This 36-lead integrated circuit complex is smaller than a postage stamp and includes 285 gates fabricated on a single chip. It is used for access to computer memory circuits. (Fairchild TTμL 9035).

It appears that virtually every circuit that can be realized with receiving type vacuum tubes can also be eventually duplicated with some sort of FET package and interesting variations of this efficient and inexpensive solid-state device that will apply to high-frequency communication are on the horizon.

The Fetron A JFET called a *Fetron* has been developed that replaces a vacuum tube in a circuit directly, without requiring major modifications in the circuit. High-voltage FETS are used and the Fetron can either be a single JFET or two cascode connected JFETs in a hybrid integrated circuit. The Fetron is packaged in an oversize metal can that has the same pin configuration as the tube it replaces. The JFET characteristics can be chosen to simulate the dynamic performance of a tube. Two JFETs are required to simulate the performance of a pentode. Fetrons feature long life, low aging, and reduced power consumption as compared to an equivalent vacuum tube.

Microwave FETs *Gallium Arsenide (GaAs) FETS* have been developed that promise superior low-noise performance for microwave applications. Typical noise figures for these devices are about 3 db at 4 GHz, 4 db at 8 GHz, and 5 db at 12.5 GHz. Developmental GaAs FETs with a Schottky-barrier gate exhibit a noise figure of 3.3 db at 10 GHz and a power gain of 9 db. Many of these new experimental FETs have an f_{max} in excess of 30 GHz. Enhanced noise figures have been produced by cooling the FET device with liquid nitrogen to 77° K.

4-13 Integrated Circuits

The *integrated circuit* (IC) comprises a family in the field of microelectronics in which small, conventional components are combined in an orderly fashion in compact, high-density assemblies (*micromodules*) as shown in figure 16. Integrated circuits may be composed of passive elements (resistors, capacitors, and interconnections), and active elements such as diodes and transistors. The IC family may be divided into *monolithic* and multichip, or *hybrid*, circuits. The former category consists of an entire circuit function constructed in a single semiconductor block. The latter consists of two or more semiconductor blocks, each containing active or passive elements interconnected to form a complete circuit and assembled in a single package.

Integrated circuits offer relief in complex systems by permitting a reduction in the number of pieces and interconnections making up the system, a reduction in overall system size, better transistor matching and potentially lower system cost.

Using very small monolithic IC's makes it possible to make thousands of circuits simultaneously. For example, several hundred *dice* (plural of die) may be produced side by side from a single silicon slice in the simultaneous processing of about a hundred slices. Each die contains a complete circuit made up of ten to one hundred or more active and inactive components.

The silicon slice is prepared by an *epitaxial* process, which is defined as "the placement of materials on a surface." Epitaxy is used to grow thin layers of silicon on the slice, the layer resistivity controlled by the addition of N-type or P-type impurities (*diffusion*) to the silicon atoms being deposited. When localized regions are diffused into the base material (*substrate*), isolated circuits are achieved. Diffusion of additional P-type or N-type regions forms transistors.

Once the die is prepared by successive diffusions, a photomasking and etching process cuts accurately sized-and-located windows in the oxide surface, setting the circuit element dimensions simultaneously on every circuit in the slice. The wafer is then coated with an insulating oxide layer which can be opened in areas to permit metalization and interconnection.

The metalization process follows next, connecting circuit elements in the substrate. Electrical *isolation barriers* (insulators) may be provided in the form of reverse-biased PN junctions, or the resistance of the substrate may be used. Dielectric insulation, making use of a formed layer around a sensitive region is also employed. Successive diffusion processes produce transistors and circuit elements of microscopic size, ready to have external leads bonded to them, and suitable for encapsulation.

Typical IC dice range in size from less than 0.02″ square up to 0.08″ × 0.2″. Many package configurations are used, the most popular being the *multipin* TO-5 package, the *dual in-line package*, the *flat package*, and the inexpensive *epoxy package*.

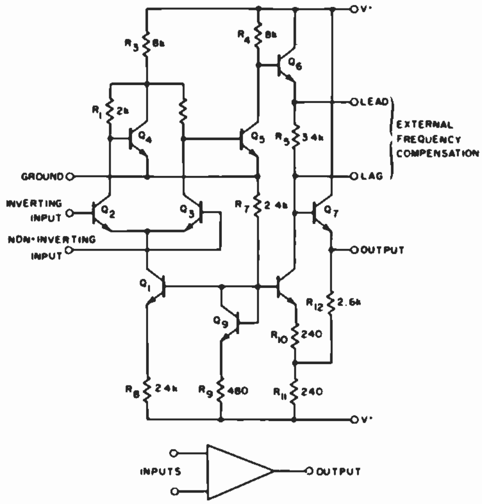


Figure 18

OPERATIONAL INTEGRATED-CIRCUIT AMPLIFIER

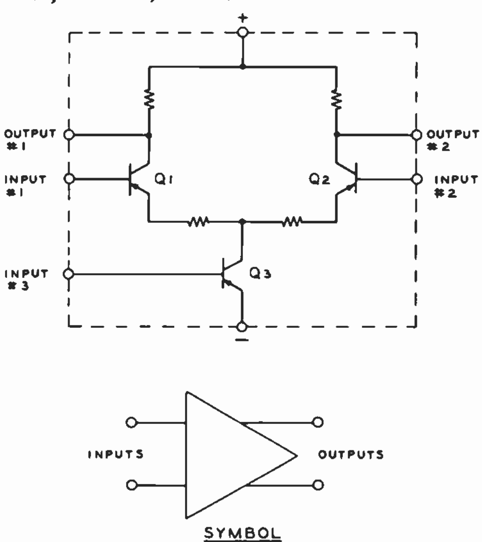


Figure 17

DIFFERENTIAL INTEGRATED-CIRCUIT AMPLIFIER

Digital and Linear IC's Integrated circuits may be classified in terms of their functional end-use into two families:

Digital—A family of circuits that operate effectively as "on-of" switches. These circuits are most frequently used in computers to count in accord with the absence or presence of a signal.

Linear (Analog)—A family of circuits that operate on an electrical signal to change its shape, increase its amplitude, or modify it for a specific use.

The *differential amplifier* is a basic circuit configuration for ICs used in a wide variety of linear applications (figure 17). The circuit is basically a balanced amplifier

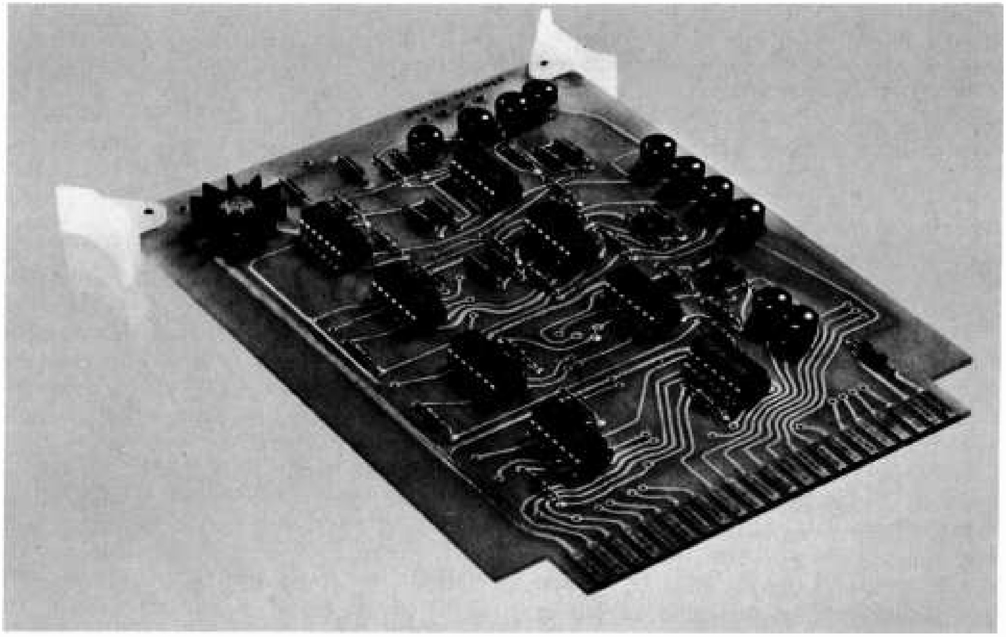


Figure 19

I-C CIRCUIT BOARD PERFORMS AS VOLTAGE REGULATOR

Complicated circuitry is reduced to printed-circuit board, eight "in-line" IC's and ten TO-5 style IC's. Transistor version would occupy many times this volume and have hundreds of discrete components. Final voltage regulator IC is at left with heat sink.

in which the currents to the emitter-coupled differential pair of transistors are supplied from a constant-current source, such as a transistor. An *operational amplifier* is a high-gain direct-coupled amplifier utilizing frequency compensation (feedback) for control of response characteristics (figure 18). The circuit symbol for these amplifiers is a triangle, with the apex pointing in the direction of operation.

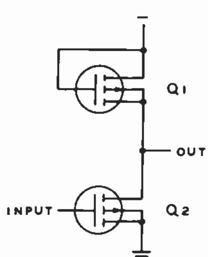


Figure 20

BASIC MOS INTEGRATED CIRCUIT

Device Q_2 serves as active device and Q_1 serves as drain resistor.

The MOSFET IC The basic monolithic bipolar IC requires a seven-mask process; that is, seven different photographic masks (negatives) must be used in diffusion, etching, and oxidizing cycles. The necessity for all of these masks to *exactly* overlay (or register) is one very critical factor in getting the yield of an IC fabrication process up to a reasonable percentage of functional chips.

Another monolithic IC, that is more simple to fabricate, is the MOSFET type. The *MOSFET IC* is principally used in logic type functional blocks. Unlike the bipolar monolithic IC, no separate diffusion is necessary to make resistors—FETs are used as resistors as well as active devices. Since MOSFET's have capacitors inherent in them (gate to channel capacitance), the small capacitors needed are already present. So, with every device on the chip a MOSFET, only several maskings must be made. The smaller number of mask processes has the effect of increasing yields, or alternately allowing more separate elements to be put on the chip.

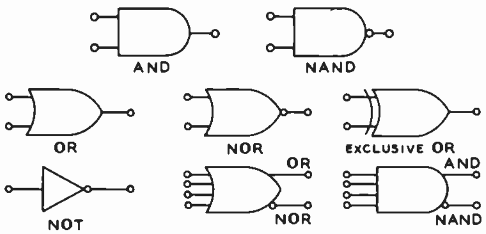


Figure 21

EXAMPLES OF SYMBOLIC LOGIC CIRCUITRY

A simple MOS-IC circuit is shown in figure 20. This is a digital inverter, Q_1 serving as the active device and Q_2 functioning as a drain resistor. A typical MOS-IC chip has literally hundreds or thousands of circuits such as this on it, interconnected as a relatively complex circuit system block, such as a shift register.

4-14 Digital-Logic ICs

An electronic system that deals with discrete events based on digits functions on an "on-off" principle wherein the active devices in the system are either operating in one of two modes: *cutoff* or *saturation* (on). Operation is based on *binary* mathematics using only the digits *zero* and *one*. In general, zero is indicated by a low signal voltage and one by a higher signal voltage. In a negative logic system the reverse is true, one being indicated by the most negative voltage.

In either case, the circuits that perform digital logic exercises may be made up of hundreds or thousands of discrete components, both active and inactive. Logic diagrams show symbols based on the specific functions performed and not on the component configuration which may consist of many microscopic particles on a semiconductor chip. Typical examples of symbolic circuitry are shown in figure 21.

RTL Logic The earliest practical IC logic form was *resistor-transistor logic* (RTL). A basic building block of RTL is the inverter or *NOT gate* (figure 22A), whose output is the opposite or complement of the input level. The output and input levels, thus, are *not* the same. The *NOR* gate is shown in figure 22B. These gates, plus the *NAND* gate permit the designer to build up *OR* and *AND* gates, plus multivibrators and even more complicated logic functions.

The NOR gate (not OR) makes use of two or more bipolar devices. If both NOR inputs are at ground (state "0"), then the output level is at +3.6 volt in this example (state "1"). However, if either input A or input B is at a positive level, then the output level drops to a voltage near ground. The logic statement expressed in binary mathematics by the NOR gate is (in Boolean algebra): $A+B=\bar{C}$, or if A or B is one, then C is *zero*. Simply, the statement says input at gate 1 or gate 2 yields a *zero* (NOR) at the output.

By adding a NOT circuit after the NOR, an *OR* circuit is formed (figure 22C); now if either A or B are *one*, then C is *one*. In Boolean notation: $A \cdot B=C$.

If *one* is termed *true* and *zero* termed *false*, these terms relate the circuits to logic in the common sense of the word. An *AND* gate is shown in figure 22D.

These simple AND, OR, and NOT circuits can be used to solve complex problems, and systems may be activated by the desired combination of true and false input statements. In addition to use in logic functions, NAND, NOR, and NOT gates can be wired as astable (free-running) multivibrators, monostable (one-shot) multivibrators, and Schmitt triggers. Representative examples of such functions are shown in figure 23.

DTL Logic Some logic ICs are *diode transistor logic* (DTL) as shown in figure 24. Illustration A shows one-quarter of a quadruple-two-input NAND gate. The DTL configuration behaves differently than the RTL devices. If the two inputs of figure 24A are open ("high," or *one*), the output is "low," (or *zero*). If any input is grounded (*zero*), the output remains high. Current has to flow out of the diode inputs to place the output level at zero. This action is termed *current sinking*.

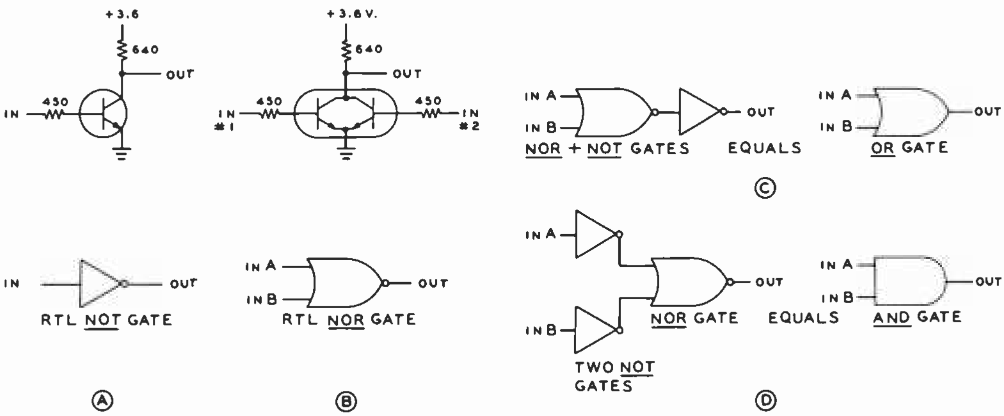


Figure 22

RTL LOGIC

A—Inverter, or NOT gate. B—Noninverting NOR gate. C—NOR plus NOT gates form OR gate. D—Two NOT gates plus NOR gate form AND gate.

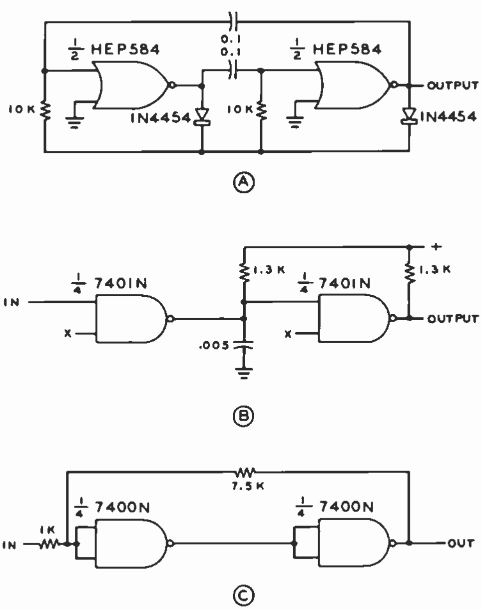


Figure 23

RTL GATES USED AS MULTIVIBRATORS AND TRIGGERS

A—Free-running multivibrator using RTL dual gate. B—Monostable multivibrator (one-shot) made from half of a TTL quad-gate. C—Schmitt trigger made from half of a TTL quad-gate.

The portion of the two-input NAND gate shown in figure 24B is a member of the TTL family, all of which can be interfaced electrically with each other and with DTL as far as signal levels are concerned. It is possible to use logic ICs in linear circuits and figure 25 shows two crystal oscillators built around RTL and TTL integrated circuits.

RTL and DTL devices are inexpensive and easily used in system designs. The RTL devices require a +3.6 volt supply and the DTL devices require a +5.0 volt supply. Both these families suffer the disadvantage of low immunity to transient noise and are sensitive to r-f pickup.

Flip Flops and Counters A *flip flop* is a device which provides two outputs which can be driven to *zero*- and *one*-level combinations. Usually when one output is *zero*, the other is *one*. Flip-flop devices may be interconnected to provide a *decade counter* (a divide-by-ten operation with ten input pulses required to provide one output pulse). A programmed counter can be used to divide frequencies by 2^n, 10, or any programmed number for service in frequency counters and synthesizers. A decade divider made up of four flip flops is shown in figure 26. These flip flops are

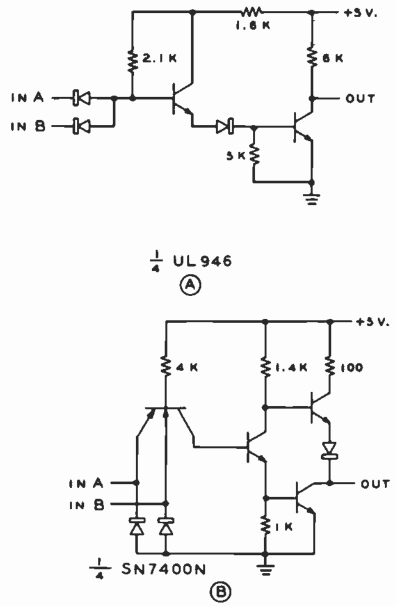


Figure 24

DTL LOGIC GATES

A—DTL two input NAND gate using ¼ of μL946.
B—TTL two input NAND gate using ¼ of SN7400N.

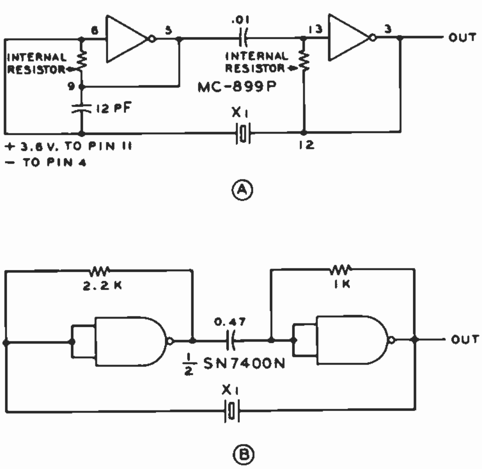


Figure 25

CRYSTAL OSCILLATORS USING RTL AND TTL INTEGRATED CIRCUITS

A—7 MHz oscillator using RTL dual buffer.
B—1 MHz oscillator using TTL gates.

toggled or clocked devices which change state as a result of an input change.

Flip-flop devices to divide by a common integer are available on a single chip, a divide-by-ten counter such as shown being representative.

HTL Logic Another form of DTL type logic device is designed to operate at a higher signal level for noise and transient immunity. *High Threshold Logic* (HTL) and *High Noise Immunity Logic* (HNIL) are devices often used in circuits that have relays and control power, such as those found in industrial systems. These families of ICs are generally operated from +12 to + 15 volts and special HTL/HNIL devices are available to interface with the less expensive RTL, DTL, and TTL families.

ECL Logic *Emitter coupled logic* (ECL) is a very high speed system capable of operation up to 350 MHz with certain devices. A typical ECL configuration is shown in figure 27. ECL operates on the principle of nonsaturation of the internal transistors. Logic swings are reduced in amplitude and the fact that the stored charge of a saturated transistor does not have to be discharged results in the speed increase. ECL is, by convention, operated from a −5.2 volt source and the swing from *zero* to *one* in logic levels is comparatively small; *zero* being −1.55 volt and *one* being −0.75 volt. This is still considered to be "positive" logic because the most negative voltage level is defined as *zero*.

Representative nonlogic IC usage as a crystal-controlled oscillator and an astable multivibrator is shown in figure 28. Interface ICs are available to or from ECL and RTL, DTL, and TTL.

4-15 MOS Logic

Digital MOS devices have been recently developed that handle logic problems whose solution is impractical in other logic families, such as problems requiring very high capacity memories. *Complementary* MOS (CMOS) will interface directly with RTL, DTL, TTL, or HTL if operated on a common power buss. Because of the low power

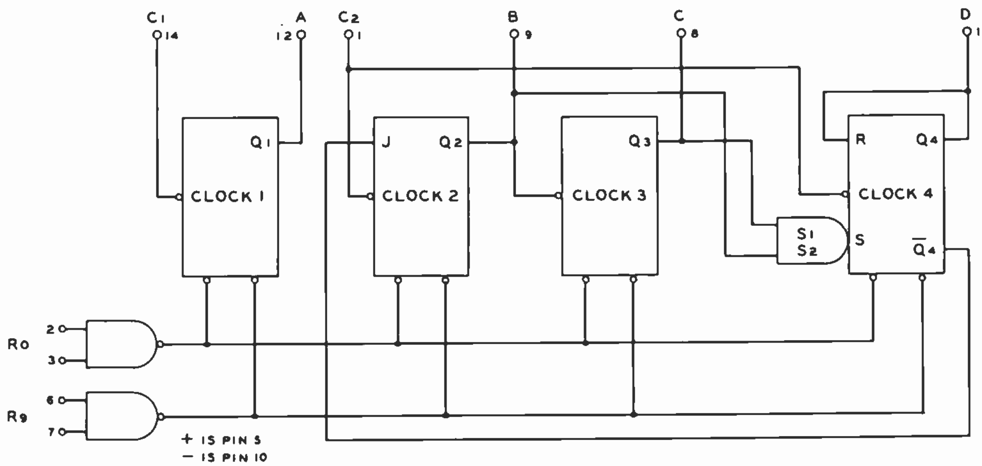


Figure 26

SN 7490N USED AS DECADE DIVIDER

Decade divider is made up of four flip-flop devices which provide zero and one level combinations. If R_0 and R_9 terminals are grounded and terminals 1 and 12 jumpered, input frequency applied to terminal 14 will be divided by 10 and appear at terminal 11. Output waveform has 20% on-cycle.

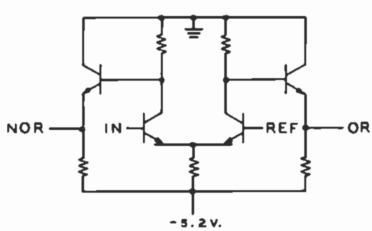


Figure 27

HIGH SPEED ECL LOGIC CIRCUIT

ECL device operates up to 350 MHz with nonsaturation of internal transistors.

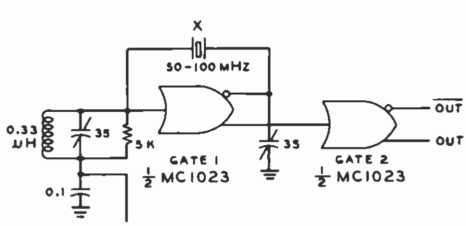


Figure 28

ECL CRYSTAL-CONTROLLED OSCILLATOR

Frequency range is 50 MHz to 100 MHz dependent on crystal and resonant circuit tuning.

consumption of CMOS, it is widely used for the frequency-divider IC in quartz-crystal-controlled watches.

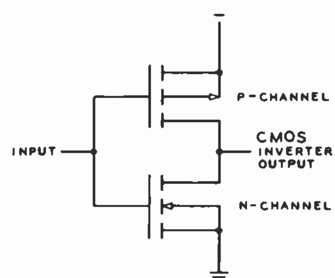


Figure 29

CMOS INVERTER

CMOS device makes use of P-channel, N-channel, enhancement-mode devices and provides low current consumption which is proportional to switching speed.

A typical CMOS inverter is shown in figure 29. It makes use of a P-channel, N-channel pair (both enhancement-mode types). If the gates are high (*one*), then the N-channel MOSFET is on and the P-channel is off, so the output is low (*zero*). If the gates are low (*zero*), then the P-channel MOSFET is on and the N-channel is off, so the output is high (*one*). Note that in

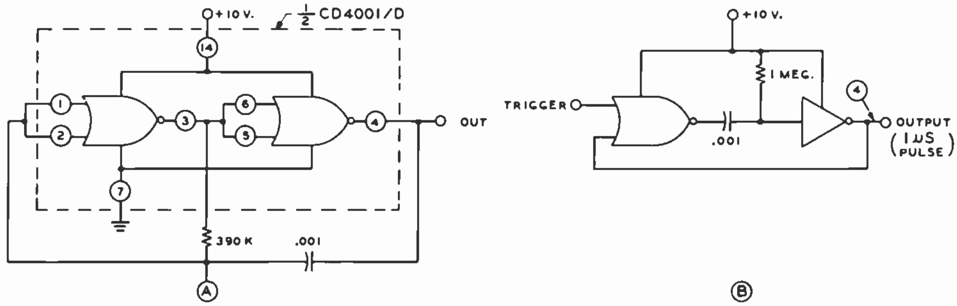


Figure 30

CMOS GATES USED AS MULTIVIBRATORS

A—Astable multivibrator using CD4001/D dual gates. B—One-shot multivibrator using dual CMOS gates.

either state one device or the other is off and the inverter pair draws only a very small leakage current, with appreciable current being drawn only during the transition from *one* to *zero* and vice versa. The more transitions per second, the higher is the average current drawn, thus the power consumption of CMOS is directly proportional to the frequency at which it is switched.

As a result of the low power consumption and the simplifications of MOS-type fabrication CMOS is moving rapidly through medium scale integration (MSI), with hundreds of FETs per chip, into large scale integration (LSI), with thousands of FETs per chip—all in one package and at a relatively low cost.

The CMOS devices now available allow for quite a large variety of circuitry, and like the types previously discussed, they may be used in nonlogic ways. Figure 30 shows how CMOS gates may be used as an astable multivibrator and a one-shot multivibrator.

P-MOS (Memory) Logic Conventional *P*-MOS (P-channel, enhancement mode) logic provides low cost, high capacity *shift registers* and *memories*. The shift register is a unique form of memory device which has one input and one output, plus a *clock* (timing) input. One commonly used P-MOS shift register has 256 bits of storage in it. The shift register may be compared to a piece of pipe just long enough to hold 256 marbles which are randomly colored white and black. The black marbles indicate a *one* value and the

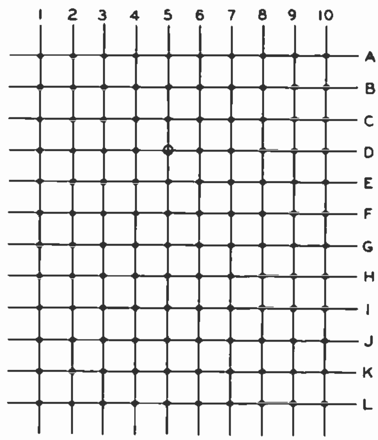


Figure 31

FERRITE-CORE MEMORY

Representation of core memory showing cores and sensing wires. Address of representative sample core is D-5. This configuration is termed a matrix.

white marbles indicate a *zero* value. The sum of marbles makes up a 256-bit binary *word*. The pipe is assumed to be opaque so the sequence of marbles cannot be seen. In order to determine the binary word, it is necessary to push 256 marbles in at the input end of the pipe and observe each marble exiting from the output, noting the binary sequence of the marbles. Each marble pushed in the pipe is the equivalent of a clock pulse. In a real shift register the output is wired back to the input, 256 clock pulses are triggered, and the content of the register is read and the binary word is loaded back into the register.

The shift register form of memory represents a valid way of storing binary information but it is slow because interrogating the register takes as many clock pulses as the register is long. To speed up access to the content of a memory, it is possible to array the bits of storage in better ways.

A more efficient organization of a large memory bank is the use of a *ferrite-core memory,* such as shown in figure 31. A bit of information can be permanently stored in a core by having it magnetized or not magnetized. If the memory has a 30 × 30 *matrix,* there are 900 cores and 900 bits of storage. Any X-line and Y-line combination locates one particular core; this location is referred to as the core *address.*

If, instead of ferrite memory cores, a large number of MOS two-state circuits are arranged in a similar matrix, an *IC memory* is produced. Most small ICs, however, are pin-limited by their packaging and to bring out 60 leads from one package is a mechanical problem. The common package has 10 leads brought out for addressing purposes; five leads for the X-line, and five for the Y-line. By using all the lines in X and Y to define a location, $2^5 = 32$ X and Y coordinates are available, thus the total bit storage is thus $2^5 \times 2^5 = 1024$ bits of information.

The Random-Access Memory A *random-access* memory device (RAM) is organized in the above fashion and 32×32 is a common bit size. These memories can be written-into and read-out of, and are used for purposes where the stored information is of a changing nature, such as in signal processing systems. For this reason a RAM is often referred to as a *scratch-pad memory.*

There is a feature about MOS devices which is unique and which allows the manufacture of shift registers and RAMs that are unlike any other semiconductor memory. Sinre the gate of a MOSFET is a capacitor it will store a charge, making a complete two-state flip flop to store *ones* and *zeros* unnecessary if the data rate is high enough. Such a *dynamic register* will only hold data for about one millisecond. Each cell of the dynamic shift register is simpler than a cell of a static shift register so the dynamic type permits more bits on a chip and is cheaper per bit to manufacture.

The Read-Only Memory The *read-only memory* (ROM) can only be programmed once and is read in sequence. Certain ROMs, however, are made in reprogrammable versions, where the stored information can be changed. The ROM is used in a type of Morse code automatic keyer which employs a 256-bit device custom-programmed to send a short message, such as: CQ CQ DE W6SAI K. This type of program is permanently placed in the chip matrix in the manufacturing process by a photomask process. However, at least one semiconductor manufacturer makes a *programmable* ROM (pROM) that may be programmed in the field. The way in which

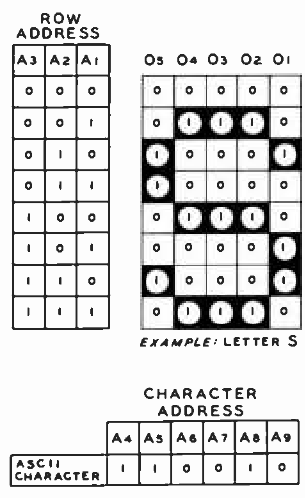


Figure 32

TELETYPE-TO-CODE CONVERTER

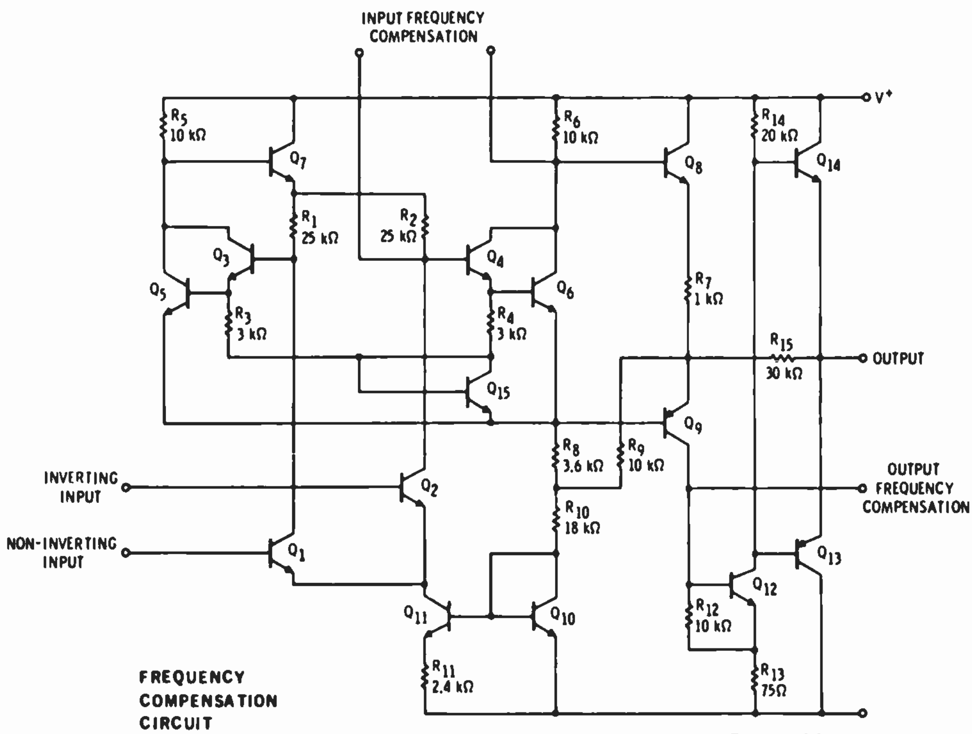
Signetics 2513 ROM device produces letters and figures on screen of a cathode-ray tube from an ASCII teletype code input. ROM illustrates letter "S" readout.

a pROM is programmed is by subjecting the bits desired to be *zeros* to a pulse of current which burns out a fusible link of nichrome on the chip. Some manufacturers will program a pROM for the buyer to his specification for a nominal charge.

Another type of pROM has been developed that is not only programmable, but which may be erased and reprogrammed. The *avalanche-induced charge-migration* pROM is initially all *zeros*. By pulsing high current into each location where a *one* is desired, the device is programmed. This charge is apparently permanent, until a flash of ultraviolet light is directed through the quartz window atop the chip. Following the ultraviolet erasure, the pROM can be programmed again. Some pROMs are available in up to 2048 bits, with 4096-bit capacity expected shortly.

Other ROM Devices There are several standard ROMS available that have factory mask programs of potential interest to the radio amateur. The *character generator* is useful for presenting letters and numerals on a cathode-ray tube such as is done in various electronic RTTY (radio teletype) terminal units. An example of such an ROM is the *Signetics 2513* which creates readable characters from an ASCII 8-level teletype code used in most time-shared computer terminals (figure 32).

Radio amateurs use the older 5-level *Baudot code* in their RTTY systems, but another ROM device can make the translation from Baudot to ASCII code. Still another ROM is now available to generate "The quick brown fox jumps over the lazy dog 1 2 3 4 5 6 7 8 9 0."



FREQUENCY COMPENSATION CIRCUIT

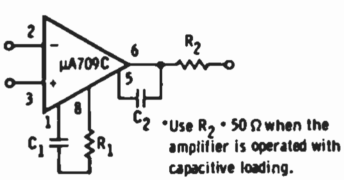


Figure 33

FAIRCHILD μA709 INTERNAL SCHEMATIC

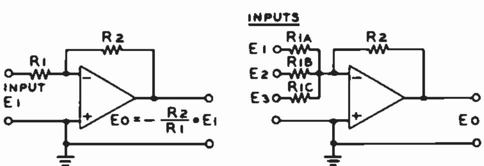
Integrated circuits are designed to replace discrete components and perform functions heretofore unavailable.

4-16 Linear ICs

The *linear integrated circuit* is a device whose output signal is a replica of the input signal. Some linear ICs are designed to replace nearly all the discrete components used in earlier composite equipment. Others perform unique functions heretofore unavailable.

Operational amplifiers, differential amplifiers and *diode-transistor arrays* are important members of the linear IC family.

The *Fairchild* μA700 series of linear monolith IC devices and particularly the μA709, are the most widely used linear IC types and more recent IC *operational amplifiers* (op-amps) are compatible in their pin configuration to this basic family of devices. The basic μA709 schematic is shown in figure 33, along with the equivalent op-amp symbol. Compensating networks may be required for stable operation and some of the newer op-amps have the necessary compensation built inside the package.



$$E_o = -\left[E_1 \cdot \frac{R_2}{R_{1A}} + E_2 \cdot \frac{R_2}{R_{1B}} + E_3 \cdot \frac{R_2}{R_{1C}}\right]$$

Ⓐ Ⓑ

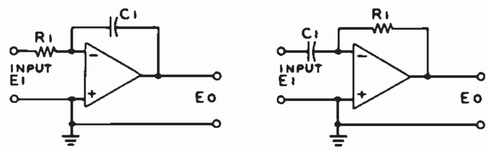
Figure 34

OPERATIONAL AMPLIFIER (OP-AMP) SYMBOL

A—Differential amplifier in inverting mode. B—Summing amplifier. If input is applied to positive gate, output is subtractive.

The Operational Amplifier The perfect operational amplifier is a high-gain d-c coupled amplifier having two differential inputs of infinite impedance, infinite gain, zero output impedance, and no phase shift. (Phase shift is 180° between the output and inverting input and 0° between the output and non-inverting input).

Two voltages may be *added* in a differential amplifier as shown in figure 34. In illustration A, the noninverting (plus) in-



$$E_o = \frac{-1}{R_1 C_1} \int E_1 dt \qquad E_o = -R_1 C_1 \frac{dE_1}{dt}$$

Ⓐ Ⓑ

Figure 35

INTEGRATING AND DIFFERENTIATING AMPLIFIERS

A—Inverting integrating circuit. B—Inverting differentating circuit.

put is grounded and the amplifier is in the *inverting* mode. The stage gain is the ratio R_2/R_1 and the input impedance is R_1. The circuit may be modified so that input to the positive gate is *subtractive*. Other inputs may be connected (illustration B) and the op-amp is now considered to be a *summing amplifier*.

The op-amp can be connected to perform the *integral* or *differential* of the input voltage as shown in figure 35. By combining these operations in a number of coordinated op-amps an *analog computer* may be con-

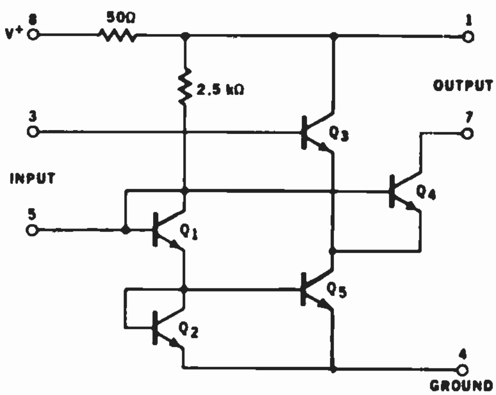


Figure 36

DIFFERENTIAL AMPLIFIER

The differential op-amp is a dual input d-c coupled amplifier comparable to a push-pull stage fed from a constant-current source.

structed. This type of machine represents the use of an electrical system as a model for a second system that is usually more difficult

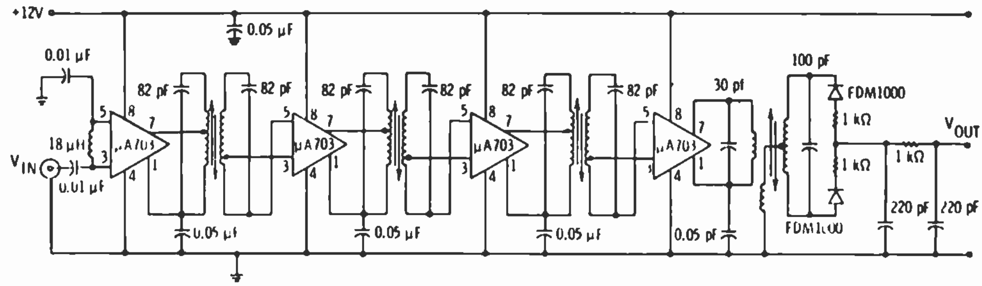


Figure 37

DIFFERENTIAL AMPLIFIERS IN R-F SERVICE

FAIRCHILD µA703 op-amps used in f-m i-f amplifier and limiter.

or more expensive to construct or measure, and that obeys the equations of the same form. The term *analog* implies similarity of relations or properties between the systems.

The Differential Amplifier The *differential amplifier* is a d-c coupled amplifier having similar input circuits. The amplifier responds to the difference between two input voltages or currents (figure 36). The differential amplifier may

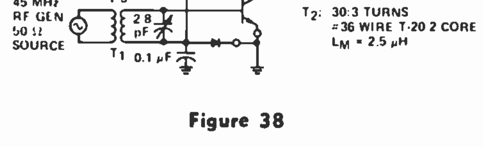


Figure 38

HEP-50 OP-AMP USED AS A-M MODULATOR

be compared to a push-pull stage fed from a constant current source.

Differential amplifiers are useful linear devices over the range from d-c to the vhf spectrum and are useful as product detectors, mixers, limiters, frequency multipliers and r-f amplifiers. Various versions of the differential amplifier are discussed in the following sections.

A widely used differential amplifier is the r-f/i-f amplifier device used as an i-f amplifier at 10.7 MHz in f-m tuners. The Fairchild µA703, Motorola HEP-590 and the Signetics NE-510 are typical examples of this device. A representative amplifier-limiter is shown in figure 37. These ICs can be used for a variety of other purposes and an a-m modulator using the HEP-50 is shown in figure 38.

The National Semiconductor LM-373 IC may be used for the detection of a-m, f-m, c-w, or SSB signals, as shown in figure 39. Note that the gain of the LM-373 has been divided into two blocks, with provisions for insertion of an i-f bandpass filter between the blocks.

Various ICs have been developed for use as i-f/f-m detectors in TV receivers. One unit comprises a complete 4.5-MHz TV sound system using the quadrature method of f-m detection similar to that employed with the 6BN6 tube. A second unit has a quadrature f-m detector, 10.7-MHz i-f, and limiter in one package (figure 40).

An IC package that is useful in signal processing applications—especially SSB—is shown in figure 41. The circuit is a balanced modulator for SSB detection.

The PLL IC A recent development is the *phase-locked loop integrated circuit* which performs a remarkable range of functions: selective amplifier, f-m detector, frequency multiplier, touchtone decoder, a-m detector, frequency synthesizer, and many more. The Signetics NE-560B shown

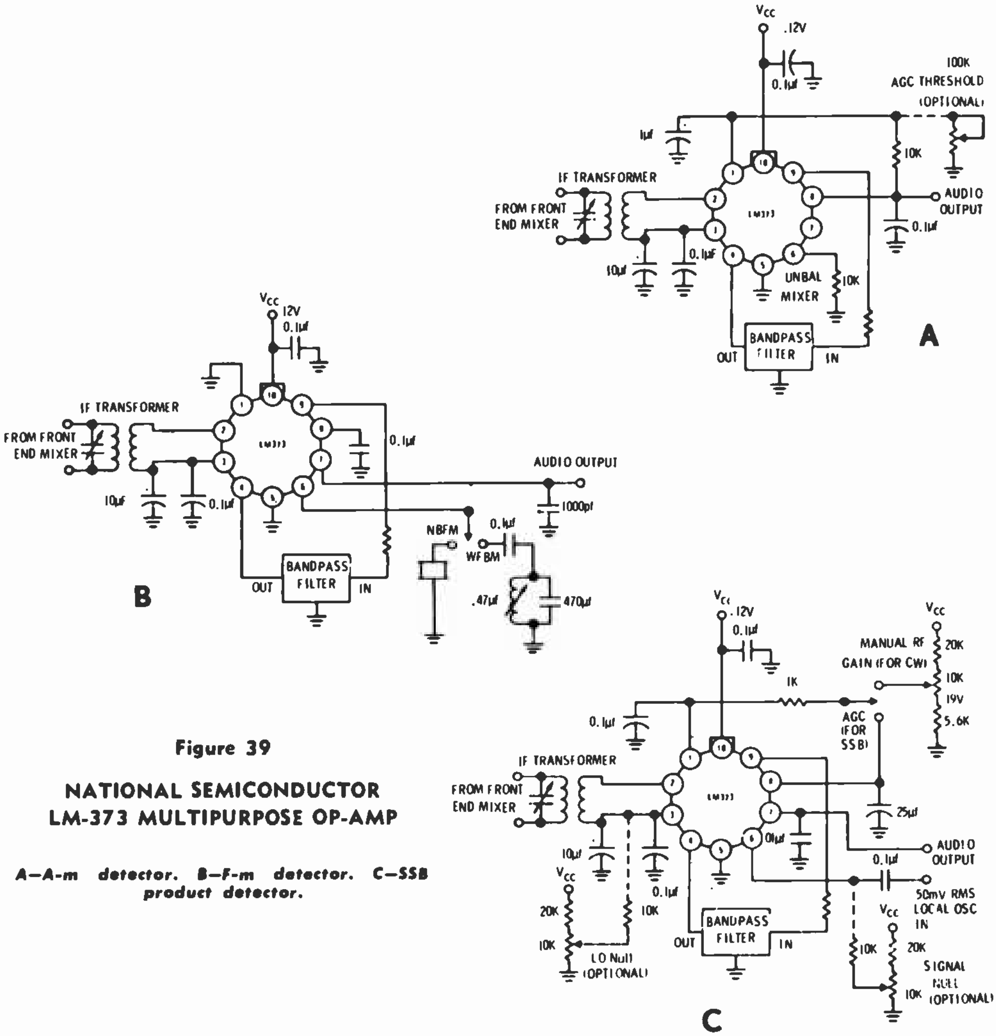


Figure 39

NATIONAL SEMICONDUCTOR LM-373 MULTIPURPOSE OP-AMP

A—A-m detector. B—F-m detector. C—SSB product detector.

in figure 42 is configured as an f-m detector. In this circuit the voltage-controlled oscillator (VCO) in the PLL locks itself into a 90° phase relationship with the incoming carrier signal. Variations of this circuit are useful in solid-state color-TV receivers.

Diode-Transistor Arrays A category of linear ICs that is of great use comprises the *diode-transistor array* family, or *array* for short. The various types of arrays available contain a number of bipolar transistors inside the package which are more or less uncommitted to any particular configuration. Because of pin limitations there are necessarily some interconnections inside the package but there is still great flexibility to interconnect the transistors for a specific purpose. Examples of these array devices are the CA 3018, CA 3036, etc. of RCA. A voltage regulator built around the CA 3018 is shown in figure 43. Note that one of the internal transistor base-emitter junctions of the IC has been used as a breakdown diode for a voltage reference. This is only one of many circuits possible using an IC array.

Many other types of linear ICs exist: video amplifiers, logarithmic amplifiers, TV

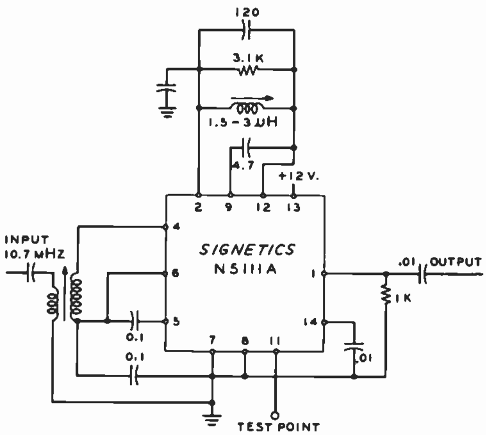


Figure 40

SIGNETICS N5111A AS QUADRATURE F-M DETECTOR AT 10.7 MHz

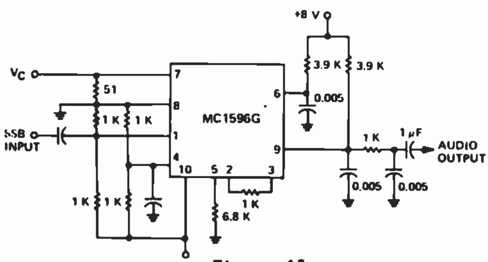


Figure 41

MC-1596G AS BALANCED MODULATOR FOR SSB GENERATION OR DETECTION

chroma demodulators, stereo-multiplex demodulators, squelch amplifiers, and so on. These represent special interest areas and it would be impossible to treat each category here. Looking at the large market areas wherein linear electronics is used, the experimenter will find ICs available or being designed for TV receivers, auto ignition systems, CATV distribution, a-m/f-m radios, stereo gear, and camera equipment. Doubtless many of the ICs developed for these markets will be readily usable in the radio communications field.

4-17 Solid-State Light Sources and Numeric Displays

A recent development is the *light-emitting diode* (LED) which promises to replace the incandescent lamp as a light source in displays—especially those subject to heavy vibration.

The first LEDs were a deep red in color and made of Gallium-Arsenide-Phosphide and produced about 30 to 100 microwatts of light power output. More recently, the green LED of Gallium-Phosphide and the amber LED of Gallium-Arsenide-Phosphide have been made available.

Small LEDs have a forward voltage drop of about 1.5 to 2.0 volts and they can be driven up to about 40 ma. The LED does not have a sudden end-of-life as does an in-

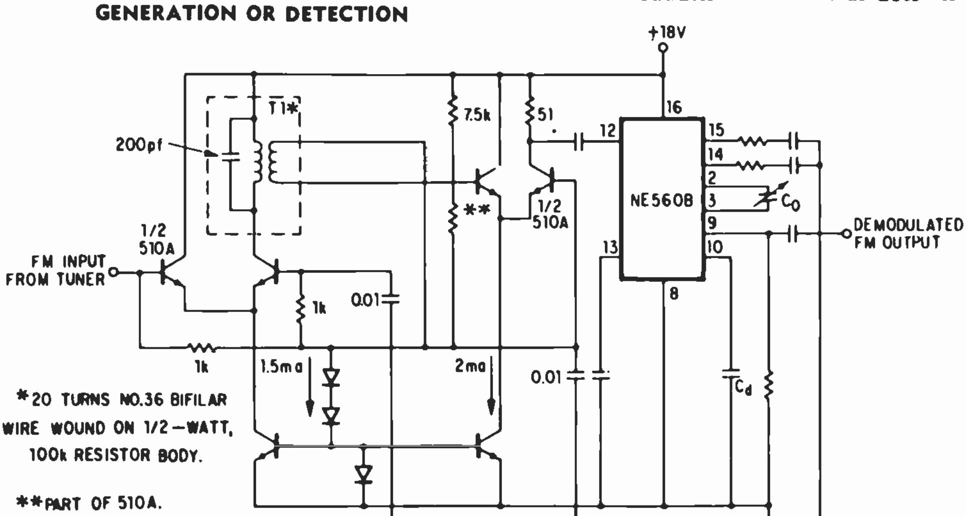


Figure 42

SIGNETICS NE-506B PHASE-LOCKED LOOP AS VOLTAGE-CONTROLLED F-M DETECTOR

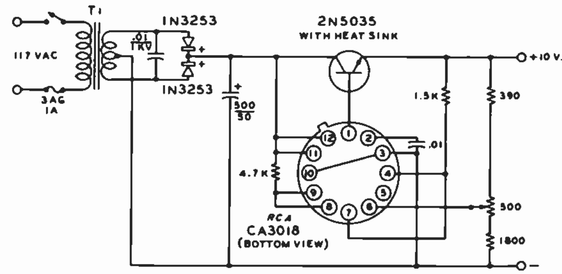


Figure 43

RCA CA-3018 AS VOLTAGE REGULATOR

T_1—10/20 40 volts center tap. Triad F-91X. Use red and yellow leads.

candescent lamp, instead the LED loses brilliance with age. Predicted life (to half brilliance) of a typical LED is 10^6 hours.

Another type of LED is the *infrared diode* which has maximum radiation at about 9000 Angstrom units (10^{-10} meters) wavelength

Figure 44

HEWLETT-PACKARD SOLID-STATE NUMERIC INDICATORS

in the near-infrared region. Because it radiates just outside the visible spectrum, the infrared produced by this Gallium-Arsenide diode is treated in the same manner as visible light, using conventional optics. The IR output of these diodes is very close to the optimum sensitivity of most LEDs, light-sensitive transistors, and FETs. The IR diode can be modulated (even at megahertz rates) and serves as a transmitter in voice and data links or as an intruder alarm. A Gallium Arsenide emitter and Silicon detector may be combined in an *optically coupled isolator* (opto-electronic switch) which combines the pair in an opaque, plastic package. Light then couples the input circuit of the emitter to the output circuit of the detector, with no electrical coupling between the ports. This isolator is the equivalent of a relay, with none of the mechanically fragile components.

An array of LEDs can be configured as a seven-segment display for numeric indication and integrated circuits are available that will convert the binary-coded decimal system to the seven-segment coding required for this display.

A solid-state numeric indicator is shown in figure 44. This small unit is a hybrid microcircuit consisting of a decoder-driver and an array of light-emitting diodes. The numeric indicator is enabled by a pulse and the display will follow changes on the logic inputs as long as the enable port is held at *zero* (low). In this mode the device is operated as a real-time display. When the enable line rises (high), the latches retain the current inputs and the display is no longer affected by changes on the logic input ports. The decimal point voltage low corresponds to point illumination.

CHAPTER FIVE

Vacuum-Tube Principles

In the previous chapters we have seen the manner in which an electric current flows through a metallic conductor as a result of an electron drift. This drift, which takes place when there is a difference in potential between the ends of the metallic conductor, is in addition to the normal random electron motion between the molecules of the conductor.

The electron may be considered as a minute negatively charged particle, having a mass of 9×10^{-28} gram, and a charge of 1.59×10^{-19} coulomb. Electrons are always identical, regardless of the source from which they are obtained.

An electric current can be caused to flow through other media than a metallic conductor. One such medium is an ionized solution, such as the sulfuric acid electrolyte in a storage battery. This type of current flow is called *electrolytic conduction.* Further, it was shown at about the turn of the century that an electric current can be carried by a stream of free electrons in an evacuated chamber. The flow of a current in such a manner is said to take place by *electronic conduction.* The study of electron tubes (also called vacuum tubes, or valves) is actually the study of the control and use of electronic currents within an evacuated or partially evacuated chamber.

Since the current flow in an electron tube takes place in an evacuated chamber, there must be located within the enclosure both a source of electrons and a collector for the electrons which have been emitted. The electron source is called the *cathode,* and the electron collector is usually called the *anode.* Some external source of energy must be applied to the cathode in order to impart sufficient velocity to the electrons within the cathode material to enable them to overcome the surface forces and thus escape into the surrounding medium. In the usual types of electron tubes the cathode energy is applied in the form of heat; electron emission from a heated cathode is called *thermionic emission.* In another common type of electron tube, the photoelectric cell, energy in the form of light is applied to the cathode to cause *photoelectric emission.*

5-1 Thermionic Emission

Electron Emission Emission of electrons from the cathode of a thermionic electron tube takes place when the cathode of the tube is heated to a temperature sufficiently high that the free electrons in the emitter have sufficient velocity to overcome the restraining forces at the surface of the material. These surface forces vary greatly with different materials. Hence different types of cathodes must be raised to different temperatures to obtain adequate quantities of electron emission. The several types of emitters found in common types of transmitting and receiving tubes will be described in the following paragraphs.

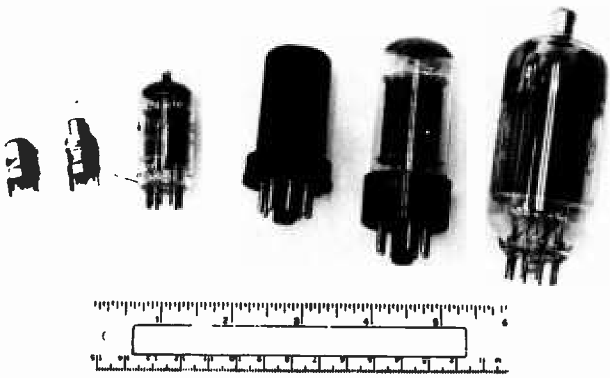


Figure 1

ELECTRON-TUBE TYPES

At the left are two Nuvistor types intended for vhf operation in TV tuners and receivers up to 450 MHz or so. Next is a typical miniature glass receiving tube alongside an old-style metal, octal-based tube. At the right are an octal-based glass audio tube and (at far right) a nine-pin based sweep tube intended for use in color television receivers (type 6LQ6). This type of tube is often used as a linear amplifier in amateur-type single-side-band transmitting equipment.

Cathode Types The emitters or cathodes as used in present-day thermionic electron tubes may be classified into two groups; the directly heated or *filament type* and the indirectly heated or *heater-cathode type*. Directly heated emitters may be further subdivided into three important groups, all of which are commonly used in modern vacuum tubes. These classifications are: the pure-tungsten filament, the thoriated-tungsten filament, and the oxide-coated filament.

The Pure-Tungsten Filament Pure-tungsten wire was used as the filament in nearly all the earlier transmitting and receiving tubes. However, the thermionic efficiency of tungsten wire as an emitter (the number of milliamperes emission per watt of filament-heating power) is quite low; the filaments become fragile after use; their life is rather short, and they are susceptible to burnout at any time. Pure-tungsten filaments must be run at bright white heat (about 2500° Kelvin). For these reasons, tungsten filaments have been replaced in all applications where another type of filament could be used. They are, however, occasionally employed in large water-cooled tubes and in certain large, high-power air-cooled triodes where another filament type would be unsuitable. Tungsten filaments are the most satisfactory for high-power, high-voltage tubes where the emitter is subjected to positive ion bombardment caused by the residual gas content of the tubes. Tungsten is not adversely affected by such bombardment.

The Thoriated-Tungsten Filament In the course of experiments made upon tungsten emitters, it was found that filaments made from tungsten having a small amount of thoria (thorium oxide) as an impurity had much greater emission than those made from the pure metal. Subsequent development has resulted in the highly efficient carburized thoriated-tungsten filament as used in many medium-power transmitting tubes today.

Thoriated-tungsten emitters consist of a tungsten wire containing from 1% to 2% thoria. The activation process varies between different manufacturers of vacuum tubes, but it is essentially as follows: (1) the tube is evacuated; (2) the filament is burned for a short period at about 2800° Kelvin to clean the surface and reduce some of the thoria within the filament to metallic thorium; (3) the filament is burned for a longer period at about 2100° Kelvin to form a layer of thorium on the surface of the tungsten; (4) the temperature is reduced to about 1600° Kelvin and some pure hydrocarbon gas is admitted to form a layer of tungsten carbide on the surface of the tungsten. This layer of tungsten carbide reduces the rate of thorium evaporation from the surface at the normal operating temperature of the filament and thus increases the operating life of the vacuum tube. Thorium evaporation from the surface is a natu-

Figure 2

VHF and UHF TUBE TYPES

At the left is an 8058 nuvistor tetrode, representative of the family of small vhf types useful in receivers and low power transmitters. The second type is an 6816 planar tetrode rated at 180 watts input to 1215 MHz. The third tube from the left is a 3CX100A5 planar triode, an improved and ruggedized version of the 2C39A, and rated at 100 watts input to 2900 MHz. The fourth tube from the left is the X-843 (Eimac) planar triode designed to deliver over 100 watts at 2100 MHz. The tube is used in a grounded-grid cavity configuration. The tube to the right is a 7213 planar tetrode, rated at 2500 watts input to 1215 MHz. All of these vhf/uhf negative-grid tubes make use of ceramic insulation for lowest envelope loss at the higher frequencies and the larger ones have coaxial bases for use in resonant cavities.

ral consequence of the operation of the thoriated-tungsten filament. The carburized layer on the tungsten wire plays another role in acting as a reducing agent to produce new thorium from the thoria to replace that lost by evaporation. This new thorium continually diffuses to the surface during the normal operation of the filament.

The last process, (5), in the activation of a thoriated-tungsten filament consists of re-evacuating the envelope and then burning or aging the new filament for a considerable period of time at the normal operating temperature of approximately 1900° K.

One thing to remember about any type of filament, particularly the thoriated type, is that the emitter deteriorates practically as fast when "standing by" (no plate current) as it does with any normal amount of emission load. Also, a thoriated filament may be either temporarily or permanently damaged by a heavy overload which may strip the surface layer of thorium from the filament.

Reactivating Thoriated-Tungsten Filaments Thoriated-tungsten filaments (and *only* thoriated-tungsten filaments) which have lost emission as a result of insufficient filament voltage, a severe temporary overload, a less severe extended overload, or even normal operation may quite frequently be reactivated to their original characteristics by a process similar to that of the original activation. However, only filaments which have not approached too close to the end of their useful life may be successfully reactivated.

The actual process of reactivation is relatively simple. The tube which has gone "flat" is placed in a socket to which only the two filament wires have been connected. The filament is then "flashed" for about 20 to 40 seconds at about 1½ times normal rated voltage. The filament will become extremely bright during this time and, if there is still some thoria left in the tungsten and if the tube did not originally fail as a result of an air leak, some of this thoria will be reduced to metallic thorium. The filament is then burned at 15 to 25 percent overvoltage for from 30 minutes to 3 to 4 hours to bring this new thorium to the surface.

The tube should then be tested to see if it shows signs of renewed life. If it does, but is still weak, the burning process should be continued at about 10 to 15 percent overvoltage for a few more hours. This should bring it back almost to normal. If the tube checks still very low after the first attempt at reactivation, the complete process can be repeated as a last effort.

The Oxide-Coated Filament The most efficient of all modern filaments is the oxide-coated type which consists of a mixture of barium and strontium oxides coated on a nickel alloy wire or strip. This type of filament operates at a dull-red to orange-red temperature (1050° to 1170° K) at which temperature it will emit large quantities of electrons. The oxide-coated filament is somewhat more efficient than the thoriated-tungsten type in small sizes and it is considerably less expensive to manufacture. For this reason all receiving tubes and quite a number of the low-powered transmitting tubes use the oxide-coated filament. Another advantage of the oxide-coated emitter is its extremely long life — the average tube can be expected to run from 3000 to 5000 hours, and when loaded very lightly, tubes of this type have been known to give 50,000 hours of life before their characteristics changed to any great extent.

Oxide filaments are unsatisfactory for use at very high plate voltage because: (1) their activity is seriously impaired by the high temperature necessary to de-gas the high-voltage tubes and, (2) the positive ion bombardment which takes place even in the best evacuated high-voltage tube causes destruction of the oxide layer on the surface of the filament.

Oxide-coated emitters have been found capable of emitting an enormously large current pulse with a high applied voltage for a very short period of time without damage. This characteristic has proved to be of great value in radar work. For example, the relatively small cathode in a microwave magnetron may be called on to deliver 25 to 50 amperes at an applied voltage of perhaps 25,000 volts for a period in the order of one microsecond. After this large current pulse has been passed, plate voltage normally will be removed for 1000 microseconds or more so that the cathode surface may recover in time for the next pulse of current. If the cathode were to be subjected to a continuous current drain of this magnitude, it would be destroyed in an exceedingly short period of time.

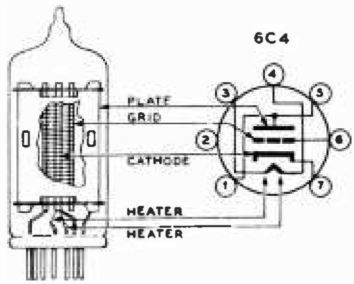


Figure 3

CUTAWAY DRAWING OF A 6C4 TRIODE

The activation of oxide-coated filaments also varies with tube manufacturers but consists essentially in heating the wire which has been coated with a mixture of barium and strontium carbonates to a temperature of about 1500° Kelvin for a time and then applying a potential of 100 to 200 volts through a protective resistor to limit the emission current. This process thermally reduces the carbonates to oxides, cleans the filament surface of foreign materials, and activates the cathode surface.

Reactivation of oxide-coated filaments is not possible since there is always more than sufficient reduction of the oxides and diffusion of the metals to the surface of the filament to meet the emission needs of the cathode.

The Heater Cathode The heater-type cathode was developed as a result of the requirement for a type of emitter which could be operated from alternating current and yet would not introduce a-c ripple modulation even when used in low-level stages. It consists essentially of a small nickel-alloy cylinder with a coating of strontium and barium oxides on its surface similar to the coating used on the oxide-coated filament. Inside the cylinder is an insulated heater element consisting usually of a double spiral of tungsten wire. The heater may operate on any voltage from 2 to 117 volts, although 6.3 is the most common value. The heater is operated at quite a high temperature so that the cathode itself usually may be brought to operating temperature in a matter of 15 to 30 seconds. Heat-coupling between the heater and the

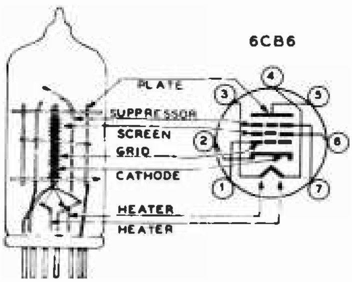


Figure 4

CUTAWAY DRAWING OF A 6CB6 PENTODE

cathode is mainly by radiation, although there is some thermal conduction through the insulating coating on the heater wire, since this coating is also in contact with the cathode thimble.

Indirectly heated cathodes are employed in all a-c operated tubes which are designed to operate at a low level either for r-f or a-f use. However, some receiver power tubes use heater cathodes (6L6, 6V6, 6F6, and 6K6-GT) as do some of the low-power transmitter tubes (802, 807, 815, 3E29, 2E26, 5763, 6146, etc.). Heater cathodes are employed almost exclusively when a number of tubes are to be operated in series as in an a-c/d-c receiver. A heater cathode is often called a *unipotential cathode* because there is no voltage drop along its length as there is in the directly heated or filament cathode.

The Emission Equation The emission of electrons from a heated cathode is quite similar to the evaporation of molecules from the surface of a liquid. The molecules which leave the surface are those having sufficient kinetic (heat) energy to overcome the forces at the surface of the liquid. As the temperature of the liquid is raised, the average velocity of the molecules is increased, and a greater number of molecules will acquire sufficient energy to be evaporated. The evaporation of electrons from the surface of a thermionic emitter is similarly a function of average electron velocity, and hence is a function of the temperature of the emitter.

Electron emission per unit area of emitting surface is a function of the temperature (T) in degrees Kelvin, the work function of emitting surface b (which is a measure of the surface forces of the material and hence of the energy required of the electron before it may escape), and of the constant (A) which also varies with the emitting surface. The relationship between emission current in amperes per square centimeter (I) and the above quantities can be expressed as:

$$I = AT^2 \varepsilon^{-b/T}$$

Secondary Emission The bombarding of most metals and a few insulators by electrons will result in the emission of other electrons by a process called *secondary emission.* The secondary electrons are literally knocked from the surface layers of the bombarded material by the primary electrons which strike the material. The number of secondary electrons emitted per primary electron varies from a very small percentage to as high as 5 to 10 secondary electrons per primary.

The phenomena of secondary emission is undesirable for most thermionic electron tubes. However, the process is used to advantage in certain types of electron tubes such as the *image orthicon* (TV camera tube) and the *electron-multiplier* type of photoelectric cell. In types of electron tubes which make use of secondary emission, such as the type 931 photocell, the secondary-electron emitting surfaces are specially treated to provide a high ratio of secondary to primary electrons. Thus a high degree of

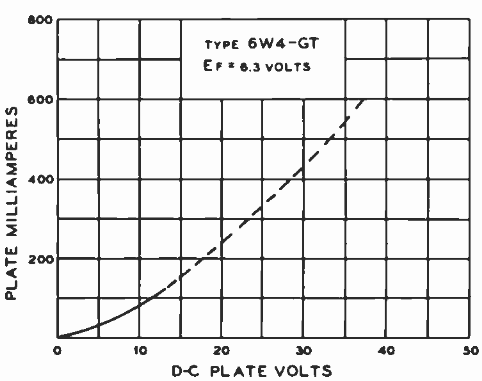


Figure 5

AVERAGE PLATE CHARACTERISTICS OF A POWER DIODE

current amplification in the electron-multiplier section of the tube is obtained.

The Space-Charge Effect As a cathode is heated so that it begins to emit, those electrons which have been discharged into the surrounding space form a negatively charged cloud in the immediate vicinity of the cathode. This cloud of electrons around the cathode is called the *space charge*. The electrons comprising the charge are continuously changing, since those electrons making up the original charge fall back into the cathode and are replaced by others emitted by it.

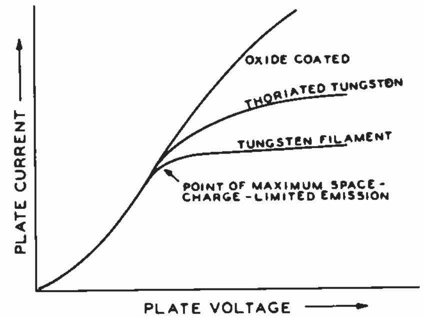


Figure 6

MAXIMUM SPACE-CHARGE-LIMITED EMISSION FOR DIFFERENT TYPES OF EMITTERS

5-2 The Diode

If a cathode capable of being heated either indirectly or directly is placed in an evacuated envelope along with a plate, such a two-element vacuum tube is called a *diode*. The diode is the simplest of all vacuum tubes and is the fundamental type from which all the others are derived.

Characteristics of the Diode When the cathode within a diode is heated, it will be found that a few of the electrons leaving the cathode will leave with sufficient velocity to reach the plate. If the plate is electrically connected back to the cathode, the electrons which have had sufficient velocity to arrive at the plate will flow back to the cathode through the external circuit. This small amount of initial plate current is an effect found in all two-element vacuum tubes.

If a battery or other source of d-c voltage is placed in the external circuit between the plate and cathode so that it places a positive potential on the plate, the flow of current from the cathode to plate will be increased. This is due to the strong attraction offered by the positively charged plate for any negatively charged particles (figure 5).

The Three-Halves Power Law At moderate values of plate voltage the current flow from cathode to anode is limited by the space charge of electrons around the cathode. Increased values of plate voltage will tend to neutralize a greater portion of the cathode space charge and hence will cause a greater current to flow.

Under these conditions, with plate current limited by the cathode space charge, the plate current is not linear with plate voltage. In fact it may be stated in general that the plate-current flow in diode tubes does not obey Ohm's Law. Rather, plate current increases as the three-halves power of the plate voltage. The relationship between plate voltage, (E) and cathode current (I) can be expressed as:

$$I = K E^{3/2}$$

where,

K is a constant determined by the geometry of the element structure within the diode tube.

Plate-Current Saturation As plate voltage is raised to the potential where the cathode space charge is neutralized, all the electrons that the cathode is capable of emitting are being attracted to the plate. The electron tube is said then to have reached *saturation* plate current. Further increase in plate voltage will cause only a relatively small increase in plate current. The initial point of plate-current saturation is sometimes called the point of *Maximum Space-Charge-Limited Emission.*

The degree of flattening in the plate-voltage plate-current curve after the limited-emission point will vary with different types of cathodes. This effect is shown in figure 6. The flattening is quite sharp with a pure tungsten emitter. With thoriated tungsten the flattening is smoothed somewhat, while with an oxide-coated cathode the flattening is quite gradual. The gradual saturation in emission with an oxide-coated emitter is generally considered to result from a lowering of the surface work function by the field at the cathode resulting from the plate potential.

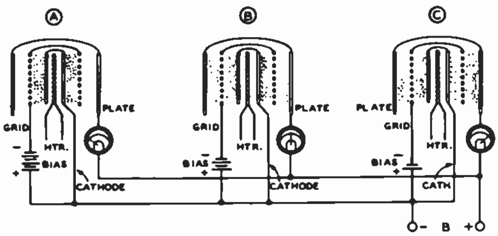


Figure 7

ACTION OF THE GRID IN A TRIODE

(A) shows the triode tube with cutoff bias on the grid. Note that all the electrons emitted by the cathode remain inside the grid mesh. (B) shows the same tube with an intermediate value of bias on the grid. Note the medium value of plate current and the fact that there is a reserve of electrons remaining within the grid mesh. (C) shows the operation with a relatively small amount of bias which with certain tube types will allow substantially all the electrons emitted by the cathode to reach the plate. Emission is said to be saturated in this case. In a majority of tube types a high value of positive grid voltage is required before plate-current saturation takes place.

Electron Energy Dissipation The current flowing in the plate-cathode space of a conducting electron tube represents the energy required to accelerate electrons from the zero potential of the cathode space charge to the potential of the anode. Then, when these accelerated electrons strike the anode, the energy associated with their velocity is immediately released to the anode structure. In normal electron tubes this energy release appears as heating of the plate or anode structure.

5-3 The Triode

If an element consisting of a mesh or spiral of wire is inserted concentric with the plate and between the plate and the cathode, such an element will be able to control by electrostatic action the cathode-to-plate current of the tube. The new element is called a *grid*, and a vacuum tube containing a cathode, grid, and plate is commonly called a *triode*.

Action of the Grid If this new element through which the electrons must pass in their course from cathode to plate is made negative with respect to the cathode, the negative charge on this grid will effectively repel the negatively charged electrons (like charges repel; unlike charges attract) back into the space charge surrounding the cathode. Hence, the number of electrons which are able to pass through the grid mesh and reach the plate will be reduced, and the plate current will be reduced accordingly. If the charge on the grid is made sufficiently negative, all the electrons leaving the cathode will be repelled back to it and the plate current will be reduced to zero. Any d-c voltage placed on a grid is called a *bias* (especially so when speaking of a control grid). The smallest negative voltage which will cause cutoff of plate current at a particular plate voltage is called the value of *cutoff bias* (figure 7).

Amplification Factor The amount of plate current in a triode is a result of the net field at the cathode from interaction between the field caused by the grid bias and that caused by the plate voltage. Hence, both grid bias and plate voltage affect the plate current. In all normal tubes a small change in grid bias has a considerably greater effect than a similar change in plate voltage. The ratio between the change in grid bias and the change in plate current which will cause the same small change in plate current is called the ***amplification factor*** or μ of the electron tube. Expressed as an equation:

$$\mu = -\frac{\Delta E_b}{\Delta E_c}$$

with I_b constant (Δ represents a small increment).

The μ can be determined experimentally by making a small change in grid bias, thus slightly changing the plate current. The plate current is then returned to the original value by making a change in the plate voltage. The ratio of the change in plate voltage to the change in grid voltage is the μ of the tube under the operating conditions chosen for the test. The μ of modern triodes ranges from 5 to 200.

Current Flow in a Triode In a diode it was shown that the electrostatic field at the cathode was proportional to the plate potential (E_b) and that the total cathode current was proportional to the three-halves power of the plate voltage. Similarly, in a triode it can be shown that the field at the cathode space charge is proportional to the equivalent voltage ($E_c + E_b/\mu$), where the amplification factor (μ) actually represents the relative effectiveness of grid potential and plate potential in producing a field at the cathode.

It would then be expected that the cathode current in a triode would be proportional to the three-halves power of ($E_c + E_b/\mu$). The cathode current of a triode can be represented with fair accuracy by the expression:

$$\text{cathode current} = K\left(E_c + \frac{E_b}{\mu}\right)^{3/2}$$

where,

K is a constant determined by element geometry within the triode.

Plate Resistance The *dynamic plate resistance* of a vacuum tube is the ratio of a change in plate voltage to the change in plate current which the change in plate voltage produces. To be accurate, the changes should be very small with respect to the operating values. Expressed as an equation:

$$r_p = \frac{\Delta E_b}{\Delta I_c}$$

The dynamic plate resistance can also be determined by the experiment mentioned above. By noting the change in plate current as it occurs when the plate voltage is changed (grid voltage held constant), and by dividing the latter by the former, the plate resistance can be determined. Plate resistance is expressed in ohms.

Transconductance The *mutual conductance*, also referred to as *transconductance*, is the ratio of a change in the plate current to the change in grid voltage which brought about the plate-current change, the plate voltage being held constant. Expressed as an equation:

$$G_m = \frac{\Delta I_b}{\Delta E_c}$$

where,

E_b is held constant.

The transconductance is also numerically equal to the amplification factor divided by the plate resistance. $G_m = \mu/r_p$.

Transconductance is most commonly expressed in microreciprocal-ohms or *micromhos*. However, since transconductance expresses change in plate current as a function of a change in grid voltage, a tube is often said to have a transconductance of so many milliamperes per volt. If the transconductance in milliamperes per volt is multiplied by 1000 it will then be expressed in micromhos. Thus the transconductance of a 6A3 could be called either 5.25 ma/volt or 5250 micromhos.

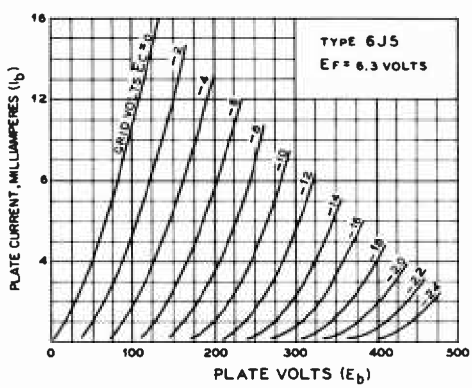


Figure 8

NEGATIVE-GRID CHARACTERISTICS (I_b VS. E_b CURVES) OF A TYPICAL TRIODE

Average plate characteristics of this form are most commonly used in determining the class-A operating characteristics of a triode amplifier stage.

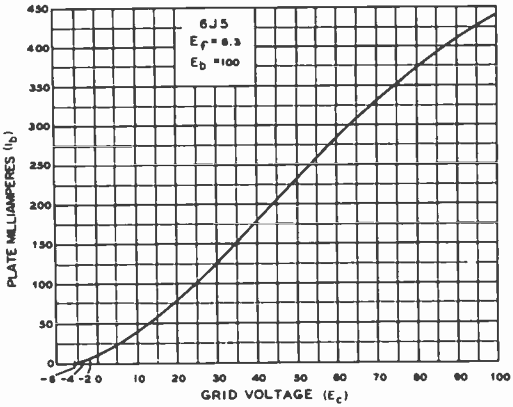


Figure 9

POSITIVE-GRID CHARACTERISTICS (I_b vs. E_c) OF A TYPICAL TRIODE

Plate characteristics of this type are most commonly used in determining the pulse-signal operating characteristics of a triode amplifier stage. Note the large emission capability of the oxide-coated heater cathode in tubes of the general type of the 6J5.

Characteristic Curves of a Triode Tube The operating characteristics of a triode tube may be summarized in three sets of curves. The I_b vs. E_b curve (figure 8), the I_b vs. E_c curve (figure 9) and the E_b vs. E_c curve (figure 10). The *plate resistance* (r_p) of the tube may be observed from the I_b vs. E_b curve, the *transconductance* (G_m) may be observed from the I_b vs. E_c curve and the *amplification factor* (μ) may be determined from the E_b vs. E_c curve.

The Load Line A *load line* is a graphical representation of the voltage on the plate of a vacuum tube and the current passing through the plate circuit of the tube for various values of plate load resistance and plate supply voltage. Figure 11 illustrates a triode tube with a resistive plate load, and a supply voltage of 300 volts. The voltage at the plate of the tube (e_b) may be expressed as:

$$e_b = E_b - \left(i_b \times R_L\right)$$

where,

E_b is the plate supply voltage,

i_b is the plate current,

R_L is the load resistance in ohms.

Assuming various values of i_b flowing in the circuit, controlled by the internal resistance of the tube (a function of the grid

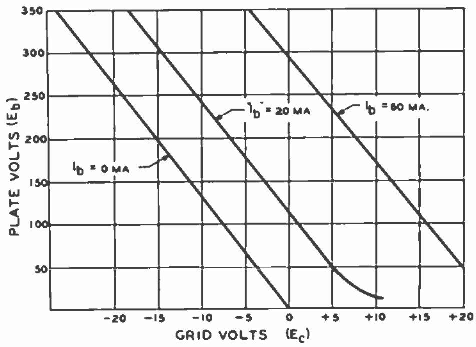
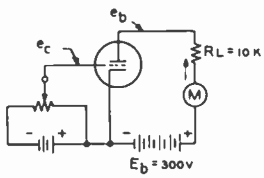


Figure 10

CONSTANT CURRENT (E_b vs. E_c) CHARACTERISTICS OF A TYPICAL TRIODE TUBE

This type of graphical representation is used for class-C amplifier calculations since the operating characteristic of a class-C amplifier is a straight line when drawn on a constant-current graph.

bias), values of plate voltage may be plotted as shown for each value of plate current (i_b). The line connecting these points is called the *load line* for the particular value of plate load resistance used. The *slope* of



I_b (MA)	e_b
0	300
5	250
10	200
15	150
20	100
25	50
30	0

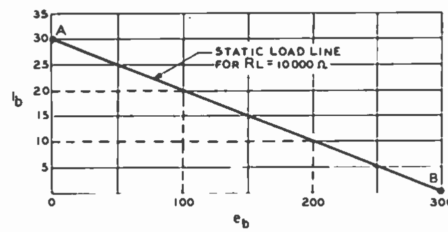


Figure 11

The static load line for a typical triode tube with a plate load resistance of 10,000 ohms.

the load line is equal to the ratio of the lengths of the vertical and horizontal projections of any segment of the load line. For this example it is:

$$\text{slope} = -\left(\frac{.01 - .02}{100 - 200}\right) = -.0001 = -\frac{1}{10{,}000}$$

The slope of the load line is equal to $-1/R_L$. At point A on the load line, the voltage across the tube is zero. This would be true for a perfect tube with zero internal voltage drop, or if the tube is short-circuited from cathode to plate. Point B on the load line corresponds to the cutoff point of the tube, where no plate current is flowing. The operating range of the tube lies between these two extremes. For additional information regarding *dynamic* load lines, the reader is referred to the *Radiotron Designer's Handbook* distributed by Radio Corporation of America.

Application of Tube Characteristics As an example of the application of tube characteristics, the constants of the triode amplifier circuit shown in figure 12 may be considered. The plate supply is 300 volts, and the plate load is 8000 ohms. If the tube is considered to be an open circuit no plate current will flow, and there is no voltage drop across the plate load resistor (R_L). The plate voltage on the tube is therefore 300 volts. If, on the other hand, the tube is considered to be a short circuit, maximum possible plate current flows and the full 300 volt drop appears across R_L. The plate voltage is zero, and the plate current is 300/1000, or 37.5 milliamperes. These two extreme conditions define the ends of the load line on the I_b vs. E_b characteristic curve figure 13.

For this application the grid of the tube is returned to a steady biasing voltage of -4 volts. The steady or quiescent operation of the tube is determined by the intersection of the load line with the -4-volt curve at point Q. By projection from point Q

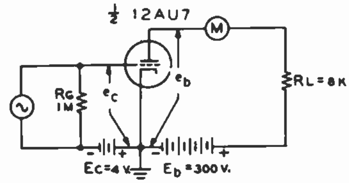


Figure 12

TRIODE TUBE CONNECTED FOR DETERMINATION OF PLATE-CIRCUIT LOAD LINE AND OPERATING PARAMETERS OF THE CIRCUIT

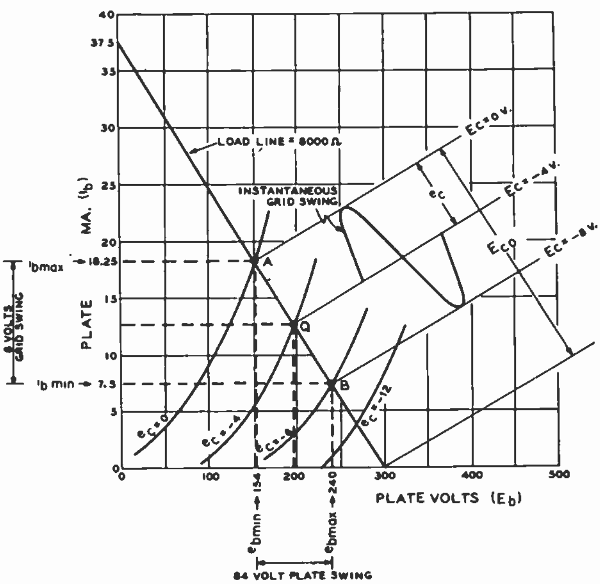


Figure 13

APPLICATION OF I_b vs. E_b CHARACTERISTICS OF A VACUUM TUBE

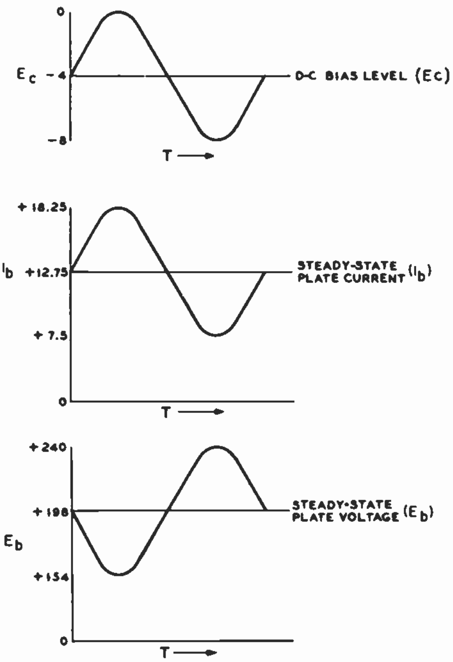


Figure 14

POLARITY REVERSAL BETWEEN GRID AND PLATE VOLTAGES

through the plate-current axis it is found that the value of plate current with no signal applied to the grid is 12.75 milliamperes. By projection from point Q through the plate-voltage axis it is found that the quiescent plate voltage is 198 volts. This leaves a drop of 102 volts across R_L which is borne out by the relation 0.01275 × 8000 = 102 volts.

An alternating voltage of 4 volts maximum swing about the normal bias value of −4 volts is applied now to the grid of the triode amplifier. This signal swings the grid in a positive direction to 0 volts, and in a negative direction to −8 volts, and establishes the *operating region* of the tube along the load line between points A and B. Thus the maxima and minima of the plate voltage and plate current are established. By projection from points A and B through the plate-current axis the maximum instantaneous plate current is found to be 18.25 milliamperes and the minimum is 7.5 milliamperes. By projections from points A and B through the plate-voltage axis the minimum instantaneous plate-voltage swing is found to be 154 volts and the maximum is 240 volts.

By this graphical application of the I_b vs. E_b characteristic of the 6SN7 triode the operation of the circuit illustrated in figure 12 becomes apparent. A voltage variation of 8 volts (peak to peak) on the grid produces a variation of 84 volts at the plate.

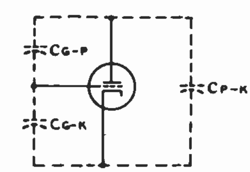


Figure 15

SCHEMATIC REPRESENTATION OF INTERELECTRODE CAPACITANCE

Polarity Inversion When the signal voltage applied to the grid has its maximum positive instantaneous value the plate current is also maximum. Reference to figure 12 shows that this maximum plate current flows through plate-load resistor R_L, producing a maximum voltage drop across it. The lower end of R_L is connected to the plate supply, and is therefore held at a constant potential of 300 volts. With maximum voltage drop across the load resistor, the upper end of R_L is at a minimum instantaneous voltage. The plate of the tube is connected to this end of R_L and is therefore at the same minimum instantaneous potential.

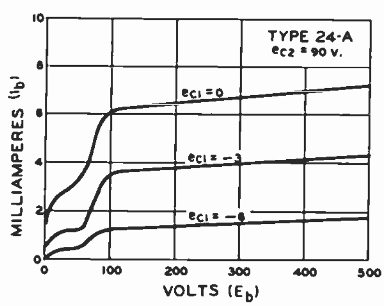


Figure 16

TYPICAL I_b vs. E_b TETRODE CHARACTERISTIC CURVES

This polarity reversal between instantaneous grid and plate voltages is further clarified by a consideration of Kirchhoff's law as it applies to series resistance. The sum of the *IR* drops around the plate circuit must at all times equal the supply voltage of 300 volts. Thus when the instantaneous voltage drop across R_L is maximum, the voltage drop across the tube is minimum, and their sum must equal 300 volts. The variations of grid voltage, plate current and plate voltage about their steady-state values is illustrated in figure 14.

Interelectrode Capacitance Capacitance always exists between any two pieces of metal separated by a dielectric. The exact amount of capacitance depends on the size of the metal pieces, the dielectric between them, and the type of dielectric. The electrodes of a vacuum tube have a similar characteristic known as *interelectrode capacitance*, illustrated in figure 15. These direct capacitances in a triode are: grid-to-cathode capacitance, grid-to-plate capacitance, and plate-to-cathode capacitance. The interelectrode capacitance, though very small, has a coupling effect, and often can cause unbalance in a particular circuit. At very-high frequencies (vhf), interelectrode capacitances become very objectionable and prevent the use of conventional tubes at these frequencies. Special vhf tubes must be used which are characterized by very small electrodes and close internal spacing of the elements of the tube.

5-4 Tetrode and Screen-Grid Tubes

Many desirable characteristics can be obtained in a vacuum tube by the use of more than one grid. The most common multielement tube is the tetrode (four electrodes). Other tubes containing as many as eight electrodes are available for special applications.

The Tetrode The quest for a simple and easily usable method of eliminating the effects of the grid-to-plate capacitance of the triode led to the development of the *screen-grid* tube, or *tetrode*. When another grid is added between the grid and plate of a vacuum tube the tube is called a tetrode, and because the new grid

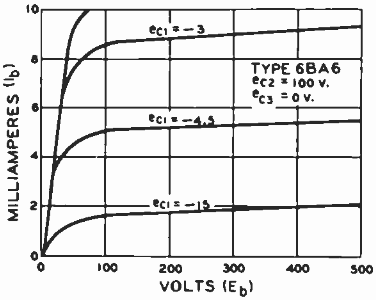


Figure 17

TYPICAL I_b vs. E_b PENTODE CHARACTERISTIC CURVES

is called a *screen*, as a result of its screening or shielding action, the tube is often called a screen-grid tube. The interposed screen grid acts as an electrostatic shield between the grid and plate, with the consequence that the grid-to-plate capacitance is reduced. Although the screen grid is maintained at a positive voltage with respect to the cathode of the tube, it is maintained at ground potential with respect to r.f. by means of a bypass capacitor of very low reactance at the frequency of operation.

In addition to the shielding effect, the screen grid serves another very useful purpose. Since the screen is maintained at a positive potential, it serves to increase or accelerate the flow of electrons to the plate. There being large openings in the screen mesh, most of the electrons pass through it and on to the plate. Due also to the screen, the plate current is largely independent of plate voltage, thus making for high amplification. When the screen voltage is held at a constant value, it is possible to make large changes in plate voltage without appreciably affecting the plate current, (figure 16).

When the electrons from the cathode approach the plate with sufficient velocity, they dislodge electrons on striking the plate. This effect of *bombarding* the plate with high-velocity electrons, with the consequent dislodgement of other electrons from the plate, gives rise to the condition of secondary emission which has been discussed in a previous paragraph. This effect can cause no particular difficulty in a triode because the secondary electrons so emitted are eventually attracted back to the plate. In the screen-grid tube, however, the screen is close to the plate and is maintained at a positive potential. Thus, the screen will attract these electrons which have been knocked from the plate, particularly when the plate voltage falls to a lower value than the screen voltage, with the result that the plate current is lowered and the amplification is decreased.

In the application of tetrodes, it is necessary to operate the plate at a high voltage in relation to the screen in order to overcome these effects of *secondary emission.*

The Pentode The undesirable effects of secondary emission from the plate can be greatly reduced if yet another element is added between the screen and plate. This additional element is called a *suppressor*, and tubes in which it is used are called *pentodes*. The suppressor grid is sometimes connected to the cathode within the tube; sometimes it is brought out to a connecting pin on the tube base, but in any case it is established negative with respect to the minimum plate voltage. The secondary electrons that would travel to the screen if there were no suppressor are diverted back to the plate. The plate current is, therefore, not reduced and the amplification possibilities are increased (figure 17).

Pentodes for audio applications are designed so that the suppressor increases the limits to which the plate voltage may swing; therefore the consequent power output and gain can be very great. Pentodes for radio-frequency service function in such a manner that the suppressor allows high voltage gain, at the same time permitting fairly high gain at low plate voltage. This holds true even if the plate voltage is the same or slightly lower than the screen voltage.

Remote-Cutoff Tubes *Remote-cutoff* tubes (variable-μ) are screen-grid tubes in which the control grid structure has been physically modified so as to cause the plate current of the tube to drop off gradually, rather than to have a well-defined cutoff point (figure 18). A non-uniform control-grid structure is used, so that the amplification factor is different for different parts of the control grid.

Remote-cutoff tubes are used in circuits where it is desired to control the amplification by varying the control-grid bias. The characteristic curve of an ordinary screen-grid tube has considerable curvature near the plate-current cutoff point, while the curve of a remote-cutoff tube is much more linear (figure 19). The remote-cutoff tube minimizes cross-talk interference that would otherwise be produced. Examples of remote cutoff tubes are: 6BD6, 6BA6, 6SG7 and 6SK7.

Beam-Power Tubes A *beam-power* tube makes use of another method of suppressing secondary emission. In this tube there are four electrodes: a cathode, a grid, a screen, and a plate, so spaced and placed that secondary emission from the plate is suppressed without actual power loss. Because of the manner in which the electrodes are spaced, the electrons which travel to the plate are slowed down when the plate voltage is low, almost to zero velocity in a certain region between screen and plate. For this reason the electrons form a stationary cloud, or *space charge*. The effect of this

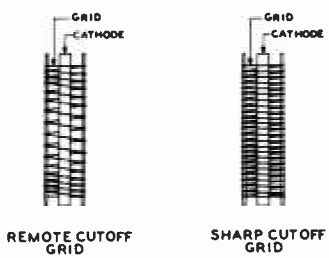


Figure 18

REMOTE-CUTOFF GRID STRUCTURE

space charge is to repel secondary electrons emitted from the plate and thus cause them to return to the plate. In this way, secondary emission is suppressed.

Another feature of the beam-power tube is the low current drawn by the screen. The screen and the grid are spiral wires wound so that each turn in the screen is shaded from the cathode by a grid turn. This alignment of the screen and the grid causes the electrons to travel in sheets between the turns of the screen so that very few of them strike the screen itself. This formation of the electron stream into sheets or beams increases the charge density in the screen-plate region and assists in the creation of the space charge in this region.

Because of the effective suppressor action provided by the space charge, and because of the low current drawn by the screen, the beam-power tube has the advantages of high power output, high power sensitivity, and high efficiency. The 6AQ5 is such a beam-power tube, designed for use in the power-amplifier stages of receivers and speech amplifiers or modulators. Larger tubes employing the beam-power principle are being made by various manufacturers for use in the radio-frequency stages of transmitters. These tubes feature extremely high power sensitivity (a very small amount of driving power is required for a large output), good plate efficiency, and low grid-to-plate capacitance. Examples of these tubes are 813, 4-250A, 4CX250B, etc.

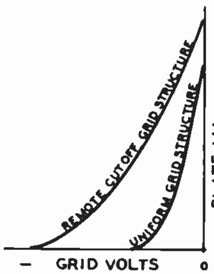


Figure 19
ACTION OF A REMOTE-CUTOFF GRID STRUCTURE

Grid-Screen Mu Factor The *grid-screen μ factor* (μ_s) is analogous to the amplification factor in a triode, except that the screen of a pentode or tetrode is substituted for the plate of a triode. μ_s denotes the ratio of a change in grid voltage to a change in screen voltage, each of which will produce the same change in screen current. Expressed as an equation:

$$\mu_s = \frac{\Delta E_{c2}}{\Delta E_{c1}}$$

where

I_{c2} is held constant.

The grid-screen μ factor is important in determining the operating bias of a tetrode or pentode tube. The relationship between control-grid potential and screen potential determines the plate current of the tube as well as the screen current since the plate current is essentially independent of the plate voltage in tubes of this type. In other words, when the tube is operated at cutoff bias as determined by the screen voltage and the grid-screen μ_s factor (determined in the same way as with a triode, by dividing the operating voltage by the μ factor), the plate current will be substantially at cutoff, as will be the screen current. The grid-screen μ_s factor is numerically equal to the amplification factor of the same tetrode or pentode tube when it is triode connected.

Current Flow in Tetrodes and Pentodes The following equation is the expression for total cathode current in a triode tube. The expression for the total cathode current of a tetrode and a pentode tube is the same, except that the screen-grid voltage and the grid-screen μ_s factor are used in place of the plate voltage and μ of the triode.

$$\text{Cathode current} = K\left(E_{c1} + \frac{E_{c2}}{\mu_s} + \frac{E_b}{\mu}\right)^{3/2}$$

Cathode current, of course, is the sum of the screen and plate currents plus control-grid current in the event that the control grid is positive with respect to the cathode. *It will be noted that total cathode current is independent of plate voltage in a tetrode or pentode.* Also, in the usual tetrode or pentode the *plate* current is substantially independent of plate voltage over the usual operating range—which means simply that the effective plate resistance of such tubes is relatively high. However, when the plate voltage falls below the normal operating range, the plate current falls sharply, while the screen current rises to such a value that the total cathode current remains substantially constant. Hence, the screen grid in a tetrode or pentode will almost invariably be damaged by excessive dissipation if the plate voltage is removed while the screen voltage is still being applied from a low-impedance source.

The Effect of Grid Current The current equations show how the total cathode current in triodes, tetrodes, and pentodes is a function of the potentials applied to the various electrodes. If only one electrode is positive with respect to the cathode (such as would be the case in a triode acting as a class-A amplifier) all the cathode current goes to the plate. But when both screen and plate are positive in a tetrode or pentode, the cathode current divides between the two elements. Hence the screen current is taken from the total cathode current, while the balance goes to the plate. Further, if the control grid in a tetrode or pentode is operated at a positive potential the total cathode current is divided between all three elements which have a positive potential. In a tube which is receiving a large excitation voltage, it may be said that the control grid robs electrons from the output electrode during the period that the grid is positive, making it always necessary to limit the peak-positive excursion of the control grid.

Coefficients of Tetrodes and Pentodes In general it may be stated that the amplification factor of tetrode and pentode tubes is a coefficient which is not of much use to the designer. In fact the amplification factor is seldom given on the design-data sheets of such tubes. Its value

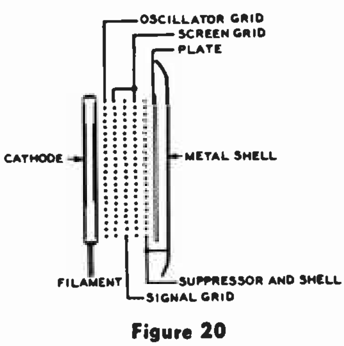


Figure 20

GRID STRUCTURE OF 6SA7 CONVERTER TUBE

is usually very high, due to the relatively high plate resistance of such tubes, but bears little relationship to the stage gain which actually will be obtained with such tubes.

On the other hand, the *grid-plate transconductance* is the most important coefficient of pentode and tetrode tubes. Gain per stage can be computed directly when the G_m is known. The grid-plate transconductance of a tetrode or pentode tube can be calculated through use of the expression:

$$G_m = \frac{\Delta I_b}{\Delta E_c}$$

with E_{c2} and E_b constant.

The plate resistance of such tubes is of less importance than in the case of triodes, though it is often of value in determining the amount of damping a tube will exert on the impedance in its plate circuit. Plate resistance is calculated from:

$$r_p = \frac{\Delta E_b}{\Delta I_b}$$

with E_{c1} and E_{c2} constant.

5-5 Mixer and Converter Tubes

The superheterodyne receiver always includes at least one stage for changing the frequency of the incoming signal to the fixed frequency of the main intermediate-frequency amplifier in the receiver. This frequency-changing process is accomplished by selecting the beat-note difference frequency between a locally generated oscillation and the incoming signal frequency. If the oscillator signal is supplied by a separate tube, the frequency changing tube is called a *mixer*. Alternatively, the oscillation may be generated by additional elements within the frequency-changer tube. In this case the frequency changer is commonly called a *converter* tube.

Conversion Conductance The *conversion conductance* (G_c) is a coefficient of interest in the case of mixer or converter tubes, or of conventional triodes, tetrodes, or pentodes operating as frequency changers. The conversion conductance is the ratio of a change in the signal-grid voltage at the input frequency to a change in the output current at the converted frequency. Hence G_c in a mixer is essentially the same as transconductance in an amplifier, with the exception that the input signal and the output current are on different frequencies. The value of G_c in conventional mixer tubes is from 300 to 3000 micromhos.

The value of G_c in an amplifier tube operated as a mixer is approximately 0.3 the G_m of the tube operated as an amplifier. The voltage gain of a mixer stage is equal to $G_c \times R_L$, where R_L is the impedance of the plate load into which the mixer tube operates.

The Diode Mixer The simplest mixer tube is the diode. The noise figure, or figure of merit, for a mixer of this type is not as good as that obtained with other more complex mixers; however, the diode is useful as a mixer in uhf and vhf equipment where low interelectrode capacities are vital to circuit operation. Since the diode impedance is low, the local oscillator must furnish considerable power to the diode mixer. A good diode mixer has an over-all gain of about 0.5.

The Triode Mixer A triode mixer has better gain and a better noise figure than the diode mixer. At low frequencies, the gain and noise figure of a triode mixer closely approaches those figures obtained when the tube is used as an amplifier. In the uhf and vhf range, the efficiency of the triode mixer deteriorates rapidly. The optimum local-oscillator voltage for a triode mixer is about 0.7 as large as the cutoff bias of the triode. Very little local-oscillator power is required by a triode mixer.

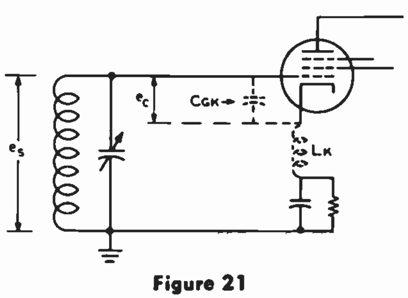


Figure 21

SHOWING THE EFFECT OF CATHODE LEAD INDUCTANCE

The degenerative action of cathode-lead inductance tends to reduce the effective grid-to-cathode voltage with respect to the voltage available across the input tuned circuit. Cathode-lead inductance also introduces undesirable coupling between the input and the output circuits.

Pentode Mixers and Converter Tubes A common multigrid converter tube for broadcast or shortwave use is the *pentagrid converter*, typified by the 6BE6, 6BA7, and 6SA7 tubes (figure 20). Operation of these converter tubes and pentode mixers will be covered in the Receiver Fundamentals Chapter.

5-6 Electron Tubes at Very-High Frequencies

As the frequency of operation of the usual type of electron tube is increased above about 20 MHz, certain assumptions which are valid for operation at lower frequencies must be re-examined. First, we find that lead inductances from the socket connections to the actual elements within the envelope no longer are negligible. Second, we find that electron transit time no longer may be ignored; an appreciable fraction of a cycle of input signal may be required for an electron to leave the cathode space charge, pass through the grid wires, and travel through the space between grid and plate.

Effects of Lead Inductance The effect of lead inductance is twofold. First, as shown in figure 21, the combination of grid-lead inductance, grid-cathode capacitance, and cathode-lead inductance tends to reduce the effective grid-cathode signal voltage for a constant voltage at the tube terminals as the frequency is increased. Second, cathode-lead inductance tends to introduce undesired coupling between the various elements within the tube.

Tubes especially designed for vhf and uhf use have had their lead inductances minimized. The usual procedures for reducing lead inductance are: (1) using heavy lead conductors or several leads in parallel (examples are the 6J4 and 6AK5), (2) scaling down the tube in all dimensions to reduce both lead inductances and interelectrode capacitances (examples are the 6CW4, 6F4, and other nuvistor and miniature tubes), and (3) the use of very low-inductance extensions of the elements themselves as external connections (examples are lighthouse tubes such as the 2C40, planar tubes such as the 2C29, and many types of vhf transmitting tubes).

Effect of Transit Time When an electron tube is operated at a frequency high enough that electron transit time between cathode and plate is an appreciable fraction of a cycle at the input frequency, several undesirable effects take place. First, the grid takes power from the input signal even though the grid is negative at all times. This comes about since the grid will have changed its potential during the time required for an electron to pass from cathode to plate. Due to interaction, and a resulting phase difference between the field associated with the grid and that associated with a moving electron, the grid presents a resistance to an input signal in addition to its normal "cold" capacitance. Further, as a result of this action, plate current no longer is in phase with grid voltage.

An amplifier stage operating at a frequency high enough that *transit time* is appreciable:

(a) Is difficult to excite as a result of grid loss from the equivalent input grid resistance,

(b) Is capable of less output since transconductance is reduced and plate current is not in phase with grid voltage.

The effects of transit time increase with the square of the operating frequency, and they increase rapidly as frequency is increased above the value where they become just appreciable. These effects may be reduced by scaling down tube dimensions; a procedure which also reduces lead inductance. Further, transit-time effects may be reduced by the obvious procedure of increasing electrode potentials so that electron velocity will be increased. However, due to the law of electron motion in an electric field, transit time is increased only as the square root of the ratio of operating potential increase; therefore this expedient is of limited value due to other limitations on operating voltages of small electron tubes.

5-7 Special Microwave Electron Tubes

Due primarily to the limitation imposed by transit time, conventional negative-grid electron tubes are capable of affording worthwhile amplification and power output only up to a definite upper frequency. This upper frequency limit varies from perhaps 100 MHz for conventional tube types to about 4000 MHz for specialized types such as the lighthouse tube. Above the limiting frequency, the conventional negative-grid tube no longer is practicable and recourse must be taken to totally different types of electron tubes in which electron transit time is not a limitation to operation. Three of the most important of such microwave tube types are the *klystron*, the *magnetron*, and the *traveling-wave tube*.

The Power Klystron The klystron is a type of electron tube in which electron transit time is used to advantage. Such tubes comprise, as shown in figure 22, a cathode, a focusing electrode, a resonator connected to a pair of grids which afford *velocity modulation* of the electron beam (called the "buncher"), a *drift space*, and another resonator connected to a pair of grids (called the "catcher"). A *collector* for the expended electrons may be included at the end of the tube, or the catcher may also perform the function of electron collection.

The tube operates in the following manner: The cathode emits a stream of electrons which is focused into a beam by the focusing electrode. The stream passes through the buncher where it is acted upon by any field existing between the two grids of the buncher cavity. When the potential between the two grids is zero, the stream passes through without change in velocity. But

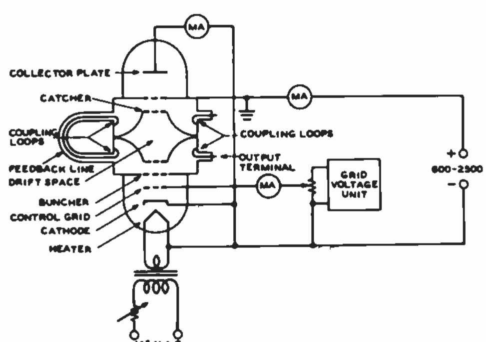


Figure 22

TWO-CAVITY KLYSTRON OSCILLATOR

A conventional two-cavity klystron is shown with a feedback loop connected between the two cavities so that the tube may be used as an oscillator.

when the potential between the two grids of the buncher is increasingly positive in the direction of electron motion, the velocity of the electrons in the beam is increased. Conversely, when the field becomes increasingly negative in the direction of the beam (corresponding to the other half-cycle of the exciting voltage from that which produced electron acceleration) the velocity of the electrons in the beam is decreased.

When the velocity-modulated electron beam reaches the drift space where there is no field, those electrons which have been sped up on one half-cycle overtake those immediately ahead which were slowed down on the other half-cycle. In this way, the beam electrons become bunched together. As the bunched groups pass through the two grids of the catcher cavity, they impart pulses of energy to these grids. The catcher-grid space is charged to different voltage levels by the passing electron bunches, and a corresponding oscillating field is set up in the catcher cavity. The catcher is designed to resonate at the frequency of the velocity-modulated beam, or at a harmonic of this frequency.

In the klystron amplifier, energy delivered by the buncher to the catcher grids is greater than that applied to the buncher cavity by the input signal. In the klystron oscillator a feedback loop connects the two cavities. Coupling to either buncher or catcher is provided by small loops which enter the cavities by way of concentric lines.

The klystron is an electron-coupled device. When used as an oscillator, its output voltage is rich in harmonics. Klystron oscillators of various types afford power outputs ranging from less than 1 watt to many thousand watts. Operating efficiency varies between 5 and 50 percent. Frequency may be shifted to some extent by varying the beam voltage. Tuning is carried on mechanically in some klystrons by altering (by means of knob settings) the shape of the resonant cavity.

The Reflex Klystron The multicavity klystron as described in the preceding paragraphs is primarily used as a transmitting device since quite reasonable amounts of power are made available in its output circuit. However, for applications where a much smaller amount of power is

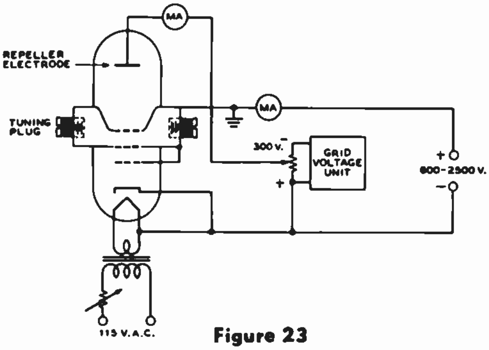


Figure 23

REFLEX KLYSTRON OSCILLATOR

A conventional reflex klystron oscillator of the type commonly used as a local oscillator in superheterodyne receivers operating above about 2000 MHz is shown above. Frequency modulation of the output frequency of the oscillator, or afc operation in a receiver, may be obtained by varying the negative voltage on the repeller electrode.

required — power levels in the milliwatt range — for low-power transmitters, receiver local oscillators, etc., another type of klystron having only a single cavity is more frequently used.

The theory of operation of the single-cavity klystron is essentially the same as the multicavity type with the exception that the velocity-modulated electron beam, after having left the buncher cavity is reflected back into the area of the buncher again by a repeller electrode as illustrated in figure 23. The potentials on the various electrodes are adjusted to a value such that proper bunching of the electron beam will take place just as a particular portion of the velocity-modulated beam re-enters the area of the resonant cavity. Since this type of klystron has only one circuit it can be used only as an oscillator and not as an amplifier. Effective modulation of the frequency of a single-cavity klystron for f-m work can be obtained by modulating the repeller electrode voltage.

The Magnetron The *magnetron* is a *uhf* oscillator tube normally employed where very-high values of peak power or moderate amounts of average power are required in the range from perhaps 700 MHz to 30,000 MHz. Special magnetrons were developed for wartime use in radar equipment which had peak power capabilities of several million watts (megawatts) output at frequencies in the vicinity of 3000 MHz. The normal duty cycle of oper-

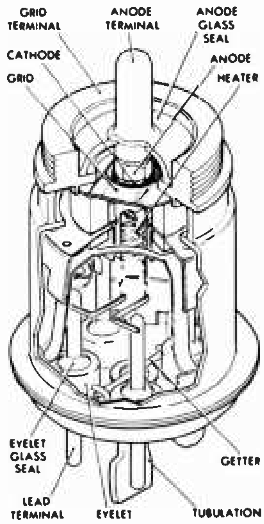


Figure 24

CUTAWAY VIEW OF WESTERN ELECTRIC 416-B/6280 VHF PLANAR TRIODE TUBE

The 416-B, designed by the Bell Telephone Laboratories is intended for amplifier or frequency multiplier service in the 4000 MHz region. Employing grid wires having a diameter equal to fifteen wavelengths of light, the 416-B has a transconductance of 50,000. Spacing between grid and cathode is .0005", to reduce transit-time effects. Entire tube is gold plated.

ation of these radar units was approximately 1/10 of one percent (the tube operated about 1/1000 of the time and rested for the balance of the operating period) so that the average power output of these magnetrons was in the vicinity of 1000 watts.

In its simplest form the magnetron tube is a filament-type diode with two half-cylindrical plates or anodes situated coaxially with respect to the filament. The construction is illustrated in figure 25A. The anodes of the magnetron are connected to a resonant circuit as illustrated in figure 25B. The tube is surrounded by an electromagnet coil which, in turn, is connected to a low-voltage d-c energizing source through a rheostat (R) for controlling the strength of the magnetic field. The field coil is oriented so that the lines of magnetic force it sets up are parallel to the axis of the electrodes.

Under the influence of the strong magnetic field, electrons leaving the filament are deflected from their normal paths and move in circular orbits within the anode cylinder. This effect results in a negative

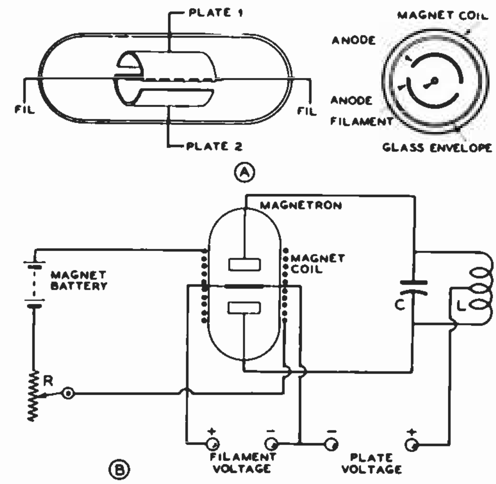


Figure 25

SIMPLE MAGNETRON OSCILLATOR

An external tank circuit is used with this type of magnetron oscillator for operation in the lower uhf range.

resistance which sustains oscillations. The oscillation frequency is very nearly the value determined by L and C. In other magnetron circuits, the frequency may be governed by the electron rotation, no external tuned circuits being employed. Wavelengths of less than 1 centimeter have been produced with such circuits.

More complex magnetron tubes employ no external tuned circuit, but utilize instead one or more resonant cavities which are integral with the anode structure. Figure 26 shows a magnetron of this type having a multicellular anode of eight cavities. It will be noted, also, that alternate cavities (which would operate at the same polarity when the tube is oscillating) are strapped together. Strapping was found to improve the efficiency and stability of high-power radar magnetrons. In most radar applications of magnetron oscillators, a powerful permanent magnet of controlled characteristics is employed to supply the magnetic field, rather than the use of an electromagnet.

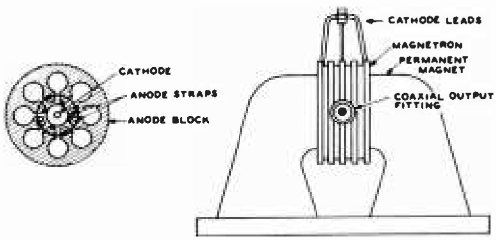


Figure 26

MODERN MULTICAVITY MAGNETRON

Illustrated is an external-anode strapped magnetron of the type commonly used in radar equipment for the 10-cm. range. An integral permanent magnet is shown in the righthand portion of the drawing, with the magnetron in place between the pole pieces of the magnet.

The Traveling-Wave Tube The *Traveling-Wave Tube* (figure 27) consists of a helix located within an evacuated envelope. Input and output terminations are affixed to each end of the helix. An electron beam passes through the helix and interacts with a wave traveling along the helix to produce broadband amplification at microwave frequencies.

When the input signal is applied to the gun end of the helix, it travels along the helix wire at approximately the speed of light. However, the signal velocity measured along the axis of the helix is considerably lower. The electrons emitted by the cathode gun pass axially through the helix to the collector, located at the output end of the helix. The average velocity of the electrons depends on the potential of the collector with respect to the cathode. When the average velocity of the electrons is greater than the velocity of the helix wave, the eletrons become crowded together in the various regions of retarded field, where they impart energy to the helix wave. A power gain of 100 or more may be produced by this tube.

5-8 The Cathode-Ray Tube

The *cathode-ray tube* is a special type of electron tube which permits the visual observation of electrical signals. It may be incorporated into an oscilloscope for use as a test instrument or it may be the display device for radar equipment or television.

Operation of the CRT A cathode-ray tube always includes an *electron gun* for producing a stream of electrons, a *grid* for controlling the intensity of the electron beam, and a *luminescent screen* for converting the impinging electron beam into visible light. Such a tube always operates in conjunction with either a built-in or an external means for focusing the electron stream into a narrow beam, and a means for deflecting the electron beam in accordance with an electrical signal.

The main electrical difference between types of cathode-ray tubes lies in the means

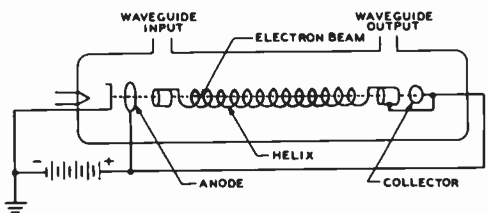


Figure 27

THE TRAVELING-WAVE TUBE

Operation of this tube is the result of interaction between the electron beam and wave traveling along the helix.

employed for focusing and deflecting the electron beam. The beam may be focused and/or deflected either electrostatically or magnetically, since a stream of electrons can be acted on by either an electrostatic or a magnetic field. In an electrostatic field the electron beam tends to be deflected toward the positive termination of the field (figure 28). In a magnetic field the stream tends to be deflected at right angles to the field. Further, an electron beam tends to be deflected so that it is normal (perpendicular) to the equipotential lines of an electrostatic field—and it tends to be deflected so that it is parallel to the lines of force in a magnetic field.

Large cathode-ray tubes used as *kinescopes* in television receivers usually are both focused and deflected magnetically. On the other hand, the medium-size CR tubes used in oscilloscopes and small television receivers usually are both focused and deflected electrostatically. Cathode-ray tubes for special applications may be focused magnetically and deflected electrostatically or vice versa.

There are advantages and disadvantages to both types of focusing and deflection. However, it may be stated that electrostatic deflection is much better than magnetic deflection when high-frequency waves are to be displayed on the screen; hence the almost universal use of this type of deflection for oscillographic work. When a tube is operated at a high value of accelerating potential so as to obtain a bright display on the face of the tube as for television or radar work, the use of magnetic deflection becomes desirable since it is relatively easier to deflect a high-velocity electron beam magnetically than electrostatically. An *ion trap* is required with magnetic deflection since the heavy negative ions emitted by the cathode are not materiallv deflected

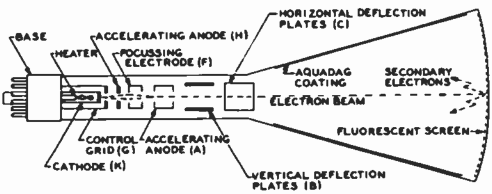


Figure 28

TYPICAL ELECTROSTATIC CATHODE-RAY TUBE

by the magnetic field and would burn an *ion spot* in the center of the luminescent screen. With electrostatic deflection the heavy ions are deflected equally as well as the electrons in the beam so that an ion spot is not formed.

Construction of Electrostatic CRT The construction of a typical electrostatic-focus, electrostatic - deflection cathode-ray tube is illustrated in the pictorial diagram of figure 28. The *indirectly heated cathode* (K) releases free electrons when heated by the enclosed filament. The cathode is surrounded by a cylinder (G) which has a small hole in its front for the passage of the electron steam. Although this element is not a wire mesh as is the usual grid, it is known by the same name because its action is similar: it controls the electron stream when its negative potential is varied.

Next in order, is found the first *accelerating anode* (H) which resembles another disk or cylinder with a small hole in its center. This electrode is run at a high or moderately high positive voltage, to accelerate the electrons toward the far end of the tube.

The *focusing electrode* (F) is a sleeve which usually contains two small disks, each with a small hole.

After leaving the focusing electrode, the electrons pass through another *accelerating anode* (A) which is operated at a high positive potential. In some tubes this electrode is operated at a higher potential than the first accelerating electrode (H) while in other tubes both accelerating electrodes are operated at the same potential.

The electrodes which have been described up to this point constitute the *electron gun*, which produces the free electrons and focuses them into a slender, concentrated, rapidly traveling stream for projecting onto the viewing screen.

Electrostatic Deflection To make the tube useful, means must be provided for deflecting the electron beam along two axes at right angles to each other. The more common tubes employ *electrostatic deflection plates*, one pair to exert a force on the beam in the vertical plane and one pair to exert a force in the horizontal plane. These plates are designated as B and C in figure 28.

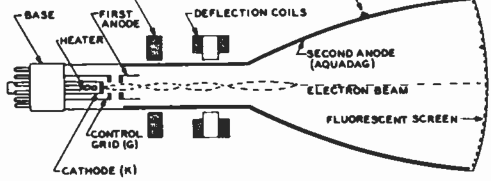


Figure 29

TYPICAL ELECTROMAGNETIC CATHODE-RAY TUBE

Standard oscilloscope practice with small cathode-ray tubes calls for connecting one of the B plates and one of the C plates together and to the high-voltage accelerating anode. With the newer three-inch tubes and with five-inch tubes and larger, all four deflection plates are commonly used for deflection. The *positive* high voltage is grounded, instead of the negative as is common practice in amplifiers, etc., in order to permit operation of the deflecting plates at a d-c potential at or near ground.

An *Aquadag* coating is applied to the inside of the envelope to attract any secondary electrons emitted by the fluorescent screen.

In the average electrostatic-deflection CR tube the spot will be fairly well centered if all four deflection plates are returned to the potential of the second anode (ground). However, for accurate centering and to permit moving the entire trace either horizontally or vertically to permit display of a particular waveform, horizontal- and vertical-centering controls usually are provided on the front of the oscilloscope.

After the spot is once centered, it is necessary only to apply a positive or negative voltage (with respect to ground) to one of the ungrounded or "free" deflector plates in order to move the spot. If the voltage is positive with respect to ground, the beam will be attracted toward that deflector plate. If it is negative, the beam and spot will be repulsed. The amount of deflection is directly proportional to the voltage (with respect to ground) that is applied to the free electrode.

With the larger-screen higher-voltage tubes it becomes necessary to place deflecting voltage on both horizontal and both vertical plates. This is done for two reasons: First, the amount of deflection voltage required by the high-voltage tubes is so great that a transmitting tube operating from a high-voltage supply would be required to attain this voltage without distortion. By using push-pull deflection with two tubes feeding the deflection plates, the necessary plate-supply voltage for the deflection amplifier is halved. Second, a certain amount of defocusing of the electron stream is always present on the extreme excursions in deflection voltage when this voltage is applied only to one deflecting plate. When the deflecting voltage is fed in push-pull to both deflecting plates in each plane, there is no defocusing because the *average* voltage acting on the electron stream is zero, even though the *net* voltage (which causes the deflection) acting on the stream is twice that on either plate.

The fact that the beam is deflected by a magnetic field is important even in an oscilloscope which employs a tube using electrostatic deflection, because it means that precautions must be taken to protect the tube from the transformer fields and sometimes even the earth's magnetic field. This normally is done by incorporating a magnetic shield around the tube and by placing any transformers as far from the tube as possible, oriented to the position which produces minimum effect on the electron stream.

Construction of Electromagnetic CRT The electromagnetic cathode-ray tube allows greater definition than does the electrostatic tube. Also, electromagnetic definition has a number of advantages when a rotating radial sweep is required to give polar indications.

The production of the electron beam in an electromagnetic tube is essentially the same as in the electrostatic tube. The grid structure is similar, and controls the electron beam in an identical manner. The elements of a typical electromagnetic tube are shown in figure 29. The *focus coil* is wound on an iron core which may be moved along the neck of the tube to focus the electron beam. For final adjustment, the current flowing in the coil may be varied. A second pair of coils, the *deflection coils*, are mounted at right angles to each other around the neck of the tube. In some cases, these coils can rotate around the axis of the tube.

Two *anodes* are used for accelerating the electrons from the cathode to the screen. The second anode is a graphite coating (*Aquadag*) on the inside of the glass envelope. The function of this coating is to attract any secondary electrons emitted by the fluorescent screen, and also to shield the electron beam.

In some types of electromagnetic tubes, a first, or *accelerating anode* is also used in addition to the *Aquadag*.

Electromagnetic Deflection A magnetic field will deflect an electron beam in a direction which is at right angles to both the direction of the field and the direction of motion of the beam.

In the general case, two pairs of deflection coils are used (figure 30). One pair is for horizontal deflection, and the other pair is for vertical deflection. The two coils in a pair are connected in series and are wound in such directions that the magnetic field flows from one coil, through the electron beam to the other coil. The force exerted on the beam by the field moves it to any point on the screen by application of the proper currents to these coils.

The Trace The human eye retains an image for about one-sixteenth second after viewing. In a CRT, the spot can be moved so quickly that a series of adjacent spots can be made to appear as a line, if the beam is swept over the path fast enough. As long as the electron beam strikes in a given place at least sixteen times a second, the spot will appear to the human eye as a source of continuous light with very little flicker.

Screen Materials—"Phosphors" At least five types of luminescent screen materials are commonly available on the various types of CR tubes commercially available. These screen materials are called *phosphors;* each of the five phosphors is best suited to a particular type of application. The P-1 phosphor, which has a green fluorescence with medium persistence, is almost invariably used for oscilloscope tubes for visual observation. The P-4 phosphor, with white fluorescence and medium persistence, is used on television viewing tubes (*Kinescopes*). The P-5 and P-11 phosphors, with blue fluorescence and very short persistence, are used primarily in oscilloscopes where photographic recording of the trace is to be obtained. The P-7 phosphor, which has a blue flash and a long-persistence greenish-yellow persistence, is used primarily for radar displays where retention of the image for several seconds after the initial signal display is required.

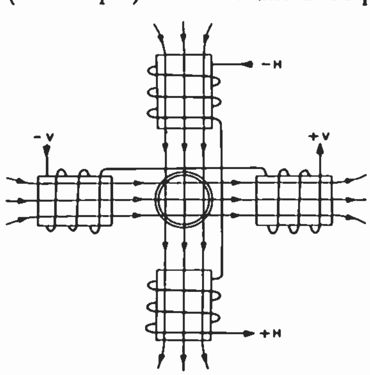


Figure 30

Two pairs of coils arranged for electromagnetic deflection in two directions.

5-9 Gas Tubes

The space charge of electrons in the vicinity of the cathode in a diode causes the plate-to-cathode voltage drop to be a function of the current being carried between the cathode and the plate. This voltage drop can be rather high when large currents are being passed, causing a considerable amount of energy loss which shows up as plate dissipation.

Action of Positive Ions The negative space charge can be neutralized by the presence of the proper density of positive ions in the space between the cathode and anode. The positive ions may be obtained by the introduction of the proper amount of gas or a small amount of mercury into the envelope of the tube. When the voltage drop across the tube reaches the ionization potential of the gas or mercury vapor, the gas molecules will become ionized to form positive ions. The positive ions then tend to neutralize the space charge in the vicinity of the cathode. The voltage drop across the tube then remains constant at the ionization potential of the gas, up to a current drain equal to the maximum emission capability of the cathode. The voltage drop varies between 10 and 20 volts, depending on the particular gas employed, up to the maximum current rating of the tube.

Mercury-Vapor Tubes Mercury-vapor tubes, although very widely used, have the disadvantage that they must be operated within a specific temperature range (25° to 70° C) in order that the mercury-vapor pressure within the tube shall be within the proper range. If the temperature is too low, the drop across the tube becomes too high causing immediate overheating and possible damage to the elements. If the temperature is too high, the vapor pressure is too high, and the voltage at which the tube will "flash back" is lowered to the point where destruction of the tube may take place. Since the ambient temperature range specified above is within the normal room temperature range, no trouble will be encountered under normal operating conditions. However, by the substitution of xenon gas for mercury it is possible to produce a rectifier with characteristics comparable to those of the mercury-vapor tube except that the tube is capable of operating over the range from approximately −70° to +90° C. The 3B25 rectifier is an example of this type of tube.

Thyratron Tubes If a grid is inserted between the cathode and plate of a mercury-vapor gaseous-conduction rectifier, a negative potential placed on the added element will increase the plate-to-cathode voltage drop required before the tube will ionize or "fire." The potential on the control grid will have no effect on the plate-to-cathode drop after the tube has ionized. However, the grid voltage may be adjusted to such a value that conduction will take place only over the desired portion of the cycle of the a-c voltage being impressed on the plate of the rectifier.

Voltage-Regulator Tubes In a glow-discharge gas tube the voltage drop across the electrodes remains constant over a wide range of current passing through the tube. This property exists because the degree of ionization of the gas in the tube varies with the amount of current passing through the tube. When a large current is passed, the gas is highly ionized and the internal impedance of the tube is low. When a small current is passed, the gas is lightly ionized and the internal impedance of the tube is high. Over the operating range of the tube, the product (*IR*) of the current through the tube and the internal impedance of the tube is very nearly constant. Examples of this type of tube are the OB2, OC2, and VR-150.

Vacuum-Tube Classification Vacuum tubes are grouped into three major classifications: commercial, ruggedized, and premium (or reliable). Any one of these three groups may also be further classified for military duty (MIL spec. or JAN classification). To qualify for MIL classification, sample lots of the particular tube must have passed special qualification tests at the factory. It should not be construed that a MIL-type tube is better than a commercial tube, since some commercial tests and specifications are more rigid than the corresponding MIL specifications. The MIL stamped tube has merely been accepted under a certain set of conditions for military service.

Ruggedized or Premium Tubes Radio tubes are being used in increasing numbers for industrial applications, such as computing and control machinery, and in aviation and marine equipment. When a tube fails in a home radio receiver, it is merely inconvenient, but a tube failure in industrial applications may bring about stoppage of some vital process, resulting in financial loss, or even danger to life.

To meet the demands of these industrial applications, a series of tubes was evolved incorporating many special features designed to ensure a long and predetermined operating life, and uniform characteristics among similar tubes. Such tubes are known as *ruggedized* or *premium* tubes. Early attempts to select reliable specimens of tubes from ordinary stock tubes proved that in the long run the selected tubes were no better than tubes picked at random. Long life and ruggedness had to be built into the tubes by means of proper choice and 100% inspection of all materials used in the tube, by critical processing inspection and assembling, and by conservative ratings of the tube.

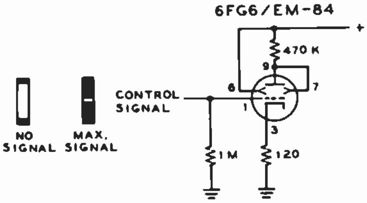


Figure 31

SCHEMATIC REPRESENTATION OF "MAGIC EYE" TUBE

Pure tungsten wire is used for heaters in preference to alloys of lower tensile strength. Nickel tubing is employed around the heater wires at the junction to the stem wires to reduce breakage at this point. Element structures are given extra supports and bracing. Finally, all tubes are given a 50-hour test run under full operating conditions to eliminate early failures. When operated within their ratings, ruggedized or premium tubes should provide a life well in excess of 10,000 hours.

Ruggedized tubes will withstand severe impact shocks for short periods, and will operate under conditions of vibration for many hours. The tubes may be identified in many cases by the fact that their nomenclature includes a "W" in the type number, as in 807W, 5U4W, etc. Some ruggedized tubes are included in the "5000" series nomenclature. The 5654 is a ruggedized version of the 6AK5, the 5692 is a ruggedized version of the 6SN7, etc.

5-10 Miscellaneous Tube Types

Electron-Ray Tubes The electron-ray tube or *magic eye* contains two sets of elements, one of which is a triode amplifier and the other a cathode-ray indicator. The plate of the triode section is connected to the ray-control electrodes (figure 31), so that as the plate voltage varies in accordance with the applied signal, the voltage on the ray-control electrode also varies. The electrons which strike the anode cause it to fluoresce, or give off light, so that the deflection caused by the ray-control electrodes, which prevents electrons from striking part of the anode, produces an electrical shadow on the fluorescent anode. The size of this shadow is determined by the voltage on the ray electrodes. When these electrodes are at the same potential as the fluorescent anode, the shadow disappears; if the ray electrode is less positive than the anode, a shadow appears, the length of which is proportional to the voltage on the ray electrodes.

Controlled Warmup Tubes Series heater strings are employed in a-c/d-c radio receivers and television sets to reduce the cost, size, and weight of the equipment. Voltage surges of great magnitude occur in series-operated filaments because of variations in the rate of warm-up of the various tubes. As the tubes warm up, the heater resistance changes. This change is not the same between tubes of various types, or even between tubes of the same type made by different manufacturers. Some 6-volt tubes show an initial surge as high as 9 volts during warm-up, while slow-heating tubes such as the 25BQ6 are underheated during the voltage surge on the 6-volt tubes.

Standardization of heater characteristics in a new group of tubes designed for series heater strings has eliminated this trouble. The new tubes have either 600 ma or 400 ma heaters, with a controlled warm-up time of approximately 11 seconds. The 5U8, 6CG7, and 12BH7-A are examples.

Digital Readout Tubes *Register tubes*, or *Nixies*, are glow tubes that provide the direct display of characters for data presentation. Nixies have stacked internal elements in the form of metallic numerals with a common anode. When negative voltage is applied to a selected character, it glows like the cathode of a gas-discharge tube. Usually only the selected numeral is visible in the viewing area because the visual glow discharge is larger than its metallic source. The *Nixie* tube requires careful control of cathode current for long life and reliability.

Register tubes are available with up to 10 characters and require a potential of about 200 volts for proper character formation. In addition to digits (0 to 9), some devices display letters of the alphabet or special characters.

CHAPTER SIX

Vacuum-Tube Amplifiers

6-1 Vacuum-Tube Parameters

The ability of the control grid of a vacuum tube to control large amounts of plate power with a small amount of grid energy allows the vacuum tube to be used as an amplifier. It is this ability of vacuum tubes to amplify an extremely small amount of energy up to almost any level, without change in anything except amplitude, which makes the vacuum tube such an extremely valuable adjunct to modern electronics and communication.

Symbols for Vacuum-Tube Parameters As an assistance in simplifying and shortening expressions involving vacuum-tube parameters, the symbols used throughout this book are shown in the Glossary at the front of this book.

Vacuum-Tube Constants The relationships between certain of the electrode potentials and currents within a vacuum tube are reasonably constant under specified conditions of operation. These relationships are called *vacuum-tube constants* and are listed in the data published by the manufacturers of vacuum tubes. The defining equations for the basic vacuum-tube constants are given in Chapter Five.

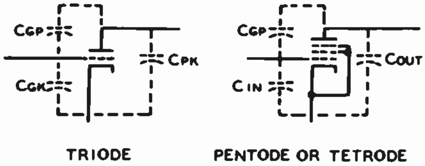


Figure 1

STATIC INTERELECTRODE CAPACITANCES WITHIN A TRIODE, PENTODE, OR TETRODE

Interelectrode Capacitances and Miller Effect The values of interelectrode capacitance published in vacuum-tube tables are the static values measured, in the case of triodes for example, as shown in figure 1. The static capacitances are simply as shown in the drawing, but when a tube is operating as amplifier there is another consideration known as *Miller Effect* which causes the dynamic input capacitance to be different from the static value. The output capacitance of an amplifier is essentially the same as the static value given in the published tube tables. The grid-to-plate capacitance is also the same as the published static value, but since C_{gp} acts as a small capacitance coupling energy back from the plate circuit to the grid circuit, the dynamic input capacitance is equal to the static value plus an amount (frequently much greater in the case of a triode) determined by the gain of the stage, the plate load impedance, and the C_{gp} feedback capacitance. The total value for an audio-amplifier stage can be expressed in the following equation:

$$C_{gk(dynamic)} = C_{gk(static)} + (A + 1)\, C_{gp}$$

This expression assumes that the vacuum tube is operating into a resistive load such as would be the case with an audio stage working into a resistance plate load in the middle audio range.

The more complete expression for the input admittance (vector sum of capacitance and resistance) of an amplifier operating into any type of plate load is as follows:

$$\text{input capacitance} = C_{gk} + (1 + A\cos\theta)\, C_{gp}$$

$$\text{input resistance} = \frac{-\left(\dfrac{1}{\omega C_{gp}}\right)}{A'\sin\theta}$$

where,

A' equals voltage amplification of the tube alone,

θ equals angle of the plate-load impedance, positive for inductive loads, negative for capacitive.

It can be seen from the above that if the plate-load impedance of the stage is capacitive or inductive, there will be a resistive component in the input admittance of the stage. The resistive component of the input admittance will be positive (tending to load the circuit feeding the grid) if the load impedance of the plate is capacitive, or it will be negative (tending to make the stage oscillate) if the load impedance of the plate is inductive.

Neutralization of Interelectrode Capacitance Neutralization of the effects of interelectrode capacitance is employed most frequently in the case of radio-frequency power amplifiers. Before the introduction of the tetrode and pentode tube, triodes were employed as neutralized class-A amplifiers in receivers. Except for vhf operation of low-noise triodes, this practice has been largely superseded through the use of tetrode and pentode tubes in which the C_{gp} or feedback capacitance has been reduced to such a low value that neutralization of its effects is not necessary to prevent oscillation and instability.

6-2 Classes and Types of Vacuum-Tube Amplifiers

Vacuum-tube amplifiers are grouped into various classes and subclasses according to the type of work they are intended to perform. The difference between the various classes is determined primarily by the angle of plate-current flow, the value of average grid bias employed, and the maximum value of the exciting signal to be impressed on the grid.

Class-A Amplifier A *class-A amplifier* is an amplifier biased and supplied with excitation of such amplitude that plate current flows continuously (360° of the exciting voltage waveshape) and grid current does not flow at any time. Such an amplifier is normally operated in the center of the grid-voltage plate-current transfer characteristic and gives an output waveshape which is a substantial replica of the input waveshape.

Class-A operation is employed in most small-signal applications such as in receivers and exciters. This mode of operation is characterized by high gain, low distortion, and low efficiency. Class-A mode may be further subdivided into A_1 and A_2 operation signifying the degree of grid drive on the stage, with the A_2 mode signifying grid drive approaching the class-AB_1 mode.

Class-AB_1 Amplifier *Class-AB_1* signifies an amplifier operated under such conditions of grid bias and exciting voltage that plate current flows for more than one-half the input voltage cycle but for less than the complete cycle. In other words the operating angle of plate current flow is appreciably greater than 180° but less than 360°. The suffix $_1$ indicates that grid current does not flow over any portion of the input cycle.

Class-AB_1 operation is utilized in most high quality, medium-power audio amplifiers and linear r-f amplifiers. Gain is lower and distortion higher than for class-A amplifiers.

Class-AB_2 Amplifier A *Class-AB_2 amplifier* is operated under essentially the same conditions of grid bias as the class-AB_1 amplifier mentioned above, but the exciting voltage is of such amplitude that grid current flows over an appreciable portion of the input wave cycle.

Class-B Amplifier A *class-B amplifier* is biased substantially to cutoff of plate current (without exciting voltage) so that plate current flows essentially over one-half the input voltage cycle. The operating angle of plate-current flow is 180°. The class-B amplifier is usually excited to the extent that grid current flows.

Class-C Amplifier A *class-C amplifier* is biased to a value greater than the value required for plate-current cutoff and is excited with a signal of such amplitude that grid current flows over an appreciable period of the input-voltage waveshape. The angle of plate-current flow in a class-C amplifier is appreciably less than 180°, or in other words, plate current flows less than one-half the time. Class-C amplifiers are not capable of linear amplification as their output waveform is not a replica of the input voltage for all signal amplitudes.

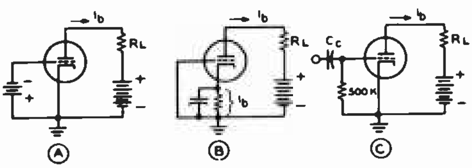


Figure 2

TYPES OF BIAS SYSTEMS

A - Grid bias
B - Cathode bias
C - Grid resistor bias

Types of Amplifiers There are three general types of amplifier circuits in use. These types are classified on the basis of the *return* for the input and output circuits. Conventional amplifiers are called *grid-driven* amplifiers, with the cathode acting as the common return for both the input and output circuits. The second type is known as a plate-return amplifier or *cathode follower* since the plate circuit is effectively at ground for the input and output signal voltages and the output voltage or power is taken between cathode and plate. The third type is called a *cathode-driven* or *grounded-grid* amplifier since the grid is effectively at ground potential for input and output signals and output is taken between grid and plate.

6-3 Biasing Methods

The difference in average potential between grid and cathode is called the *grid bias* of a vacuum tube. There are three general methods of providing this bias voltage. In each of these methods the purpose is to establish the grid at a potential with respect to the cathode which will place the tube in the desired operating condition as determined by its characteristics.

Grid bias may be obtained from a source of voltage specially provided for this purpose, such as a battery or other d-c power supply. This method is illustrated in figure 2A, and is known as *fixed bias*.

A second biasing method is illustrated in figure 2B which utilizes a cathode resistor across which an *IR* drop is developed as a result of plate current flowing through it. The cathode of the tube is held at a positive potential with respect to ground by the amount of the *IR* drop because the grid is at ground potential. Since the biasing voltage depends on the flow of plate current the tube cannot be held in a cutoff condition by means of the *cathode bias* voltage developed across the cathode resistor. The value of this resistor is determined by the bias required and the plate current which flows at this value of bias, as found from the tube characteristic curves. A capacitor is shunted across the bias resistor to provide a low-impedance path to ground for the a-c component of the plate current which results from an a-c input signal on the grid.

The third method of providing a biasing voltage is shown in figure 2C, and is called *grid-resistor* bias. During the portion of the input cycle which causes the grid to be positive with respect to the cathode, grid current flows from cathode to grid, charging capacitor C_C. When the grid draws current, the grid-to-cathode resistance of the tube drops from an infinite value to a very low value (on the order of 1000 ohms or so) making the charging time constant of the capacitor very short. This enables C_C to charge up to essentially the full value of the positive input voltage and results in the grid (which is connected to the low-potential plate of the capacitor) being held essentially at ground potential. During the negative swing of the input signal no grid current flows and the discharge path of C_G is through the grid resistance which has a value of 500,000 ohms or so. The discharge time constant for C_G is, therefore, very long in comparison to the period of the input signal and only a small part of the charge on C_G is lost. Thus, the bias voltage developed by the discharge of C_G is substantially constant and the grid is not permitted to follow the positive portion of the input signal.

6-4 Distortion in Amplifiers

There are three main types of distortion that may occur in amplifiers: *frequency* distortion, *phase* distortion and *amplitude* distortion.

Frequency Distortion *Frequency distortion* may occur when some frequency components of a signal are amplified more than others. Frequency distortion occurs at low frequencies if coupling capacitors between stages are too small, or it may occur at high frequencies as a result of the shunting effects of the distributed capacities in the circuit.

Phase Distortion In figure 3 an input signal consisting of a fundamental and a third harmonic is passed through a two-stage amplifier. Although the amplitudes of both components are amplified by identical ratios, the output waveshape is considerably different from the input signal because the phase of the third-harmonic signal has been shifted with respect to the fundamental signal. This phase shift is known as *phase distortion,* and is caused principally by the coupling circuits between the stages of the amplifier. Most coupling circuits shift the phase of a sine wave, but this has no effect on the shape of the output wave. However, when a complex wave is passed through the same coupling circuit each component frequency of the wave shape may be shifted in phase by a different amount so that the output wave is not a faithful reproduction of the input waveshape.

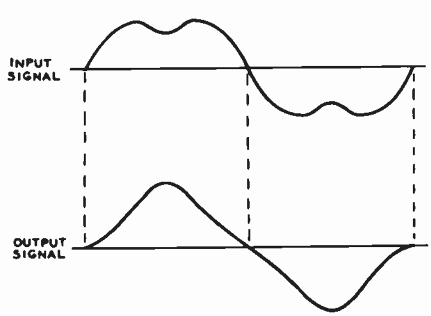


Figure 3

Illustration of the effect of phase distortion on input wave containing a third-harmonic signal

Amplitude Distortion If a signal is passed through a vacuum tube that is operating on any nonlinear part of its characteristic, *amplitude distortion* will occur. In such a region, a change in grid voltage does not result in a change in plate current which is directly proportional to the change in grid voltage. For example, if an amplifier is excited with a signal that overdrives the tubes, the resultant signal is distorted in amplitude, since the tubes are then operating over a nonlinear portion of their characteristic.

6-5 Resistance-Capacitance Coupled Audio-Frequency Amplifiers

Present practice in the design of audio-frequency voltage amplifiers is almost exclusively to use resistance-capacitance coupling between the low-level stages. Both triodes and pentodes are used; triode amplifier stages will be discussed first.

RC-Coupled Triode Stages Figure 4 illustrates the standard circuit for a resistance-capacitance coupled amplifier stage utilizing a triode tube with cathode

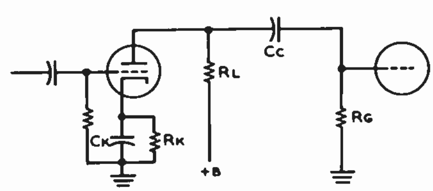
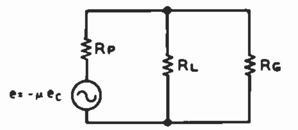


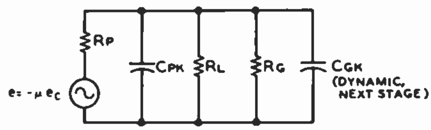
Figure 4

STANDARD CIRCUIT FOR RESISTANCE-CAPACITANCE COUPLED TRIODE AMPLIFIER STAGE



$$A = \frac{\mu R_L R_G}{R_P (R_L + R_G) + R_L R_G}$$

MID-FREQUENCY RANGE

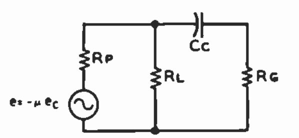


$$\frac{A_{\text{HIGH FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1 + (R_{EQ}/X_S)^2}}$$

$$R_{EQ} = \frac{R_L}{1 + \frac{R_L}{R_G} + \frac{R_L}{R_P}}$$

HIGH-FREQUENCY RANGE

$$X_S = \frac{1}{2\pi f\,(C_{PK} + C_{GK}\ (\text{DYNAMIC}))}$$



$$\frac{A_{\text{LOW FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1 + (X_C/R)^2}}$$

$$X_C = \frac{1}{2\pi f C_C}$$

LOW-FREQUENCY RANGE

$$R = R_G + \frac{R_L R_P}{R_L + R_P}$$

Figure 5

Equivalent circuits and gain equations for a triode RC-coupled amplifier stage. In using these equations, be sure the values of μ and R_p are proper for the static current and voltages with which the tube will operate. These values may be obtained from curves published in the RCA Receiving Tube Manual (series RC).

bias. In conventional audio-frequency amplifier design such stages are used at medium voltage levels (from 0.01 to 5 volts peak on the grid of the tube) and use medium-μ triodes such as the 6C4 or high-μ triodes such as the 6AB4 or 12AT7. Normal voltage gain for a single stage of this type is from 10 to 70, depending on the tube chosen and its operating conditions. Triode tubes are normally used in the last voltage-amplifier stage of an RC amplifier since their harmonic distortion with large output voltage (25 to 75 volts) is less than with a pentode tube.

Voltage Gain per Stage The voltage gain per stage of a resistance-capacitance coupled triode amplifier can be calculated with the aid of the equivalent circuits and expressions for the mid-frequency, high-frequency, and low-frequency ranges given in figure 5.

A triode RC-coupled amplifier stage is normally operated with values of cathode resistor and plate-load resistor such that the actual voltage on the tube is approximately one-half the d-c plate-supply voltage. To assist the designer of such stages, data on operating conditions for commonly used tubes is published in the *RCA Receiving Tube Manual*. It is assumed, in the case of the gain equations of figure 5, that the cathode bypass capacitor (C_k) has a reactance that is low with respect to the cathode resistor at the lowest frequency to be passed by the amplifier stage.

RC Coupled Pentode Stages Figure 6 illustrates the standard circuit for a resistance-capacitance coupled pentode amplifier stage. Cathode bias is used and the screen voltage is supplied through a dropping resistor from the plate-voltage supply. In conventional audio-frequency amplifier design such stages are normally used at low voltage levels (from 0.00001 to 0.1 volts

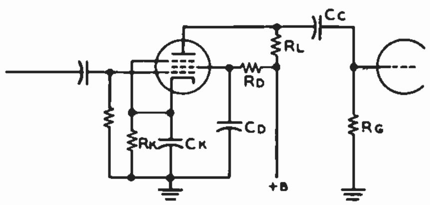


Figure 6

STANDARD CIRCUIT FOR RESISTANCE-CAPACITANCE COUPLED PENTODE AMPLIFIER STAGE

peak on the grid of the tube) and use moderate-G_m pentodes such as the 6AU6. Normal voltage gain for a stage of this type is from 60 to 250, depending on the tube chosen and its operating conditions. Pentode tubes are ordinarily used in the first stage of an RC amplifier, where the high gain which they afford is of greatest advantage, and where only a small voltage output is required from the stage.

The voltage gain per stage of a resistance-capacitance coupled pentode amplifier can be calculated with the aid of the equivalent circuits and expressions for the mid-frequency, high-frequency, and low-frequency ranges given in figure 7.

To assist the designer of such stages, data on operating conditions for commonly used types of tubes is published in the RCA Receiving Tube Manual, RC-series. It is assumed, in the case of the gain equations of figure 7, that cathode bypass capacitor C_k has a reactance that is low with respect to the cathode resistor at the lowest frequency to be passed by the stage. It is additionally assumed that the reactance of screen bypass capacitor C_d is low with respect to screen dropping resistor R_d at the lowest frequency to be passed by the amplifier stage.

Cascaded Voltage-Amplifier Stages When voltage-amplifier stages are operated in such a manner that the output voltage of the first is fed to the grid of the second, and so forth, such stages are said to be *cascaded.* The total voltage gain of cascaded amplifier stages is obtained by taking the product of the voltage gains of each of the successive stages.

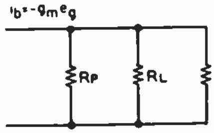
Sometimes the voltage gain of an amplifier stage is rated in decibels. Voltage gain is converted into decibel gain through the use of the following expression: db $= 20 \log_{10} A$, where A is the voltage gain of the stage. The total gain of cascaded voltage-amplifier stages can be obtained by *adding* the number of db gain in each of the cascaded stages.

RC Amplifier Response A typical frequency-response curve for an RC-coupled audio amplifier is shown in figure 8. It is seen that the amplification is poor for the extreme high and low frequencies. The reduced gain at the low frequencies is caused by the loss of voltage across the coupling capacitor. In some cases, a low-value coupling capacitor is deliberately chosen to reduce the response of the stage to hum, or to attenuate the lower voice frequencies for communication purposes. For high-fidelity work the product of the grid resistor in ohms times the coupling capacitor in microfarads should equal 25,000 (i.e.: 500,000 ohms $\times$ 0.05 μfd $=$ 25,000).

The amplification of high frequencies falls off because of the Miller effect of the subsequent stage, and the shunting effect of residual circuit capacities. Both of these effects may be minimized by the use of a low-value plate-load resistor.

Grid-resistor Bias for High-Mu Triodes The correct operating bias for a high-mu triode such as the 12AT7, is fairly critical, and will be found to be highly variable from tube to tube because of minute variations in contact potential within the tube itself. A satisfactory bias method is to use grid-resistor bias, with a resistor of one to ten megohms connected directly between grid and cathode of the tube with the cathode grounded. Grid current flows at all times, and the effective input resistance is about one-half the resistance value of the grid resistor. This circuit is particularly well suited as a high-gain amplifier following low-output devices, such as crystal, or dynamic microphones.

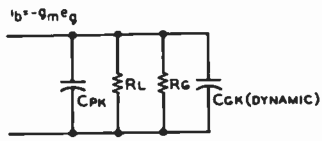
RC Amplifier General Characteristics A resistance - capacitance coupled amplifier can be designed to provide a good frequency response for almost any desired range. For instance, such an amplifier can be built to provide a fairly



MID-FREQUENCY RANGE

$$A = G_M R_{EQ}$$

$$R_{EQ} = \frac{R_L}{1 + \frac{R_L}{R_G} + \frac{R_L}{R_P}}$$

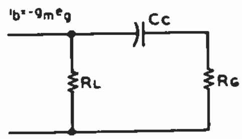


HIGH-FREQUENCY RANGE

$$\frac{A_{\text{HIGH FREQ}}}{A_{\text{MID FREQ}}} = \frac{1}{\sqrt{1 + (R_{EQ}/X_S)^2}}$$

$$R_{EQ} = \frac{R_L}{1 + \frac{R_L}{R_G} + \frac{R_L}{R_P}}$$

$$X_S = \frac{1}{2\pi f\,(C_{PK} + C_{GK\,(\text{DYNAMIC})})}$$



LOW-FREQUENCY RANGE

$$\frac{A_{\text{LOW FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1 + (X_C/R)^2}}$$

$$X_C = \frac{1}{2\pi f\, C_C}$$

$$R = R_G + \frac{R_L R_P}{R_L + R_P}$$

Figure 7

Equivalent circuits and gain equations for a pentode RC-coupled amplifier stage. In using these equations be sure to select the values of g_m and R_p which are proper for the static currents and voltages with which the tube will operate. These values may be obtained from curves published in the RCA Receiving Tube Manual Series RC.

uniform amplification for frequencies in the audio range of about 100 to 20,000 Hz. Changes in the values of coupling capacitors and load resistors can extend this frequency range to cover the very wide range required for video service. However, extension of the range can only be obtained at the cost of reduced over-all amplification. Thus RC coupling allows good frequency response with minimum distortion, but low amplification. Phase distortion is less with RC coupling than with other types, except direct coupling. The RC amplifier may exhibit tendencies to *motorboat* or oscillate if it is used with a high-impedance plate supply.

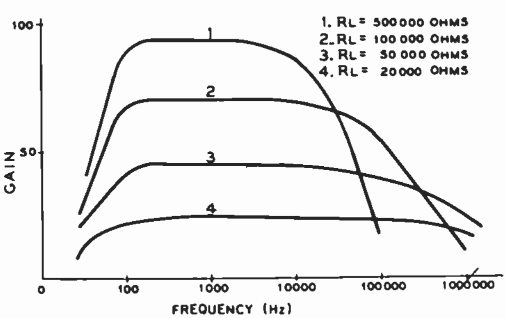


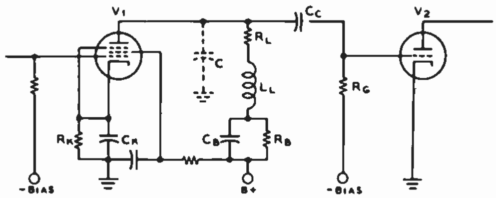
Figure 8

The variation of stage gain with frequency in an RC-coupled pentode amplifier for various values of plate load resistance.

6-6 Video-Frequency Amplifiers

A *video-frequency amplifier* is one which has been designed to pass frequencies from the lower audio range (lower limit perhaps 50 Hz) to the middle r-f range (upper limit perhaps 4 to 6 MHz). Such amplifiers, in addition to passing such an extremely wide frequency range, must be capable of amplifying this range with a minimum of amplitude, phase, and frequency distortion. Video amplifiers are commonly used in television, pulse communication, and radar work.

Tubes used in video amplifiers must have a high ratio of G_m to capacitance if a usable gain per stage is to be obtained. Commonly available tubes which have been designed for or are suitable for use in video amplifiers are: 6AU6, 6AG5, 6AK5, 6CB6, 6BC5, 6DE6, and 6AH6. Since, at the upper frequency limits of a video amplifier the input and output shunting capacitances of the amplifier tubes have rather low values of reactance, low values of coupling resistance, along with peaking coils or other special interstage coupling impedances, are usually used to flatten out the gain/frequency and hence the phase/frequency characteristic of the amplifier. Recommended operating conditions along with expressions for calcula-



MID-FREQUENCY GAIN = $G_M v_1 R_L$

HIGH-FREQUENCY GAIN = $G_M v_1 Z_{\text{COUPLING NETWORK}}$

$C = C_{\text{OUT}}\,v_1 + C_{\text{IN}}\,v_2 + C_{\text{DISTRIBUTED}}$

FOR COMPROMISE HIGH-FREQUENCY EQUALIZATION:

$X_{L_L} = 0.5\,X_C$ AT f_C

$R_L = X_C$ AT f_C

WHERE f_C = CUTOFF-FREQUENCY OF AMPLIFIER

L_L = PEAKING INDUCTOR

FOR COMPROMISE LOW-FREQUENCY EQUALIZATION

$R_B = R_K\,(G_M v_1 R_L)$

$R_B C_B = R_K C_K$

C_K = 25 TO 50 µFD IN PARALLEL WITH .001 MICA

C_B = CAPACITANCE FROM ABOVE WITH .001 MICA IN PARALLEL

Figure 9

SIMPLE COMPENSATED VIDEO AMPLIFIER CIRCUIT

Resistor R_L in conjunction with coil L_L serves to flatten the high-frequency response of the stage, while C_B and R_B serve to equalize the low-frequency response of this simple video amplifier stage.

tion of gain and circuit values are given in figure 9. Only a simple two-terminal interstage coupling network is shown in this figure.

The performance and gain per stage of a video amplifier can be improved by the use of increasingly complex two-terminal interstage coupling networks or through the use of four-terminal coupling networks or filters between successive stages. The reader is referred to Terman's *"Radio Engineer's Handbook"* for design data on such interstage coupling networks.

6-7 Other Interstage Coupling Methods

Figure 10 illustrates, in addition to resistance-capacitance interstage coupling, seven additional methods in which coupling between two successive stages of an audio-frequency amplifier may be accomplished. Although RC coupling is most commonly used, there are certain circuit conditions wherein coupling methods other than RC are more effective.

Transformer Coupling *Transformer coupling,* as illustrated in figure 10B, is seldom used at the present time between two successive single-ended stages of an audio amplifier. There are several reasons why resistance coupling is favored over transformer coupling between two successive single-ended stages. These are: (1) a transformer having frequency characteristics comparable with a properly designed RC stage is very expensive; (2) transformers, unless they are very well shielded, will pick up inductive hum from nearby power and filament transformers; (3) the phase characteristics of step-up interstage transformers are poor, making very difficult the inclusion of a transformer of this type within a feedback loop; and (4) transformers are heavy.

However, there is one circuit application where a step-up interstage transformer is of considerable assistance to the designer; this is the case where it is desired to obtain a large amount of voltage to excite the grid of a cathode follower or of a high-power class-A amplifier from a tube operating at a moderate plate voltage. Under these conditions it is possible to obtain a peak voltage on the secondary of the transformer of a value somewhat greater than the d-c plate-supply voltage of the tube supplying the primary of the transformer.

Push-Pull Transformer Interstage Coupling *Push-pull transformer coupling* between two stages is illustrated in figure 10C. This interstage coupling arrangement is fairly commonly used. The system is particularly effective when it is desired, as in the system just described, to obtain a rather high voltage to excite the grids of a high-power audio stage. The arrangement is also very good when it is desired to apply feedback to the grids of the push-pull stage by applying the feedback voltage to the low-potential sides of the two push-pull secondaries.

Impedance Coupling *Impedance coupling* between two stages is shown in figure 10D. This circuit arrangement is seldom used, but it offers one strong advantage over RC interstage coupling. This advantage is the fact that the operating voltage on the tube with the impedance in the plate circuit is equal to the plate-supply voltage, and it is possible to obtain approximately twice the peak voltage output that is possible to obtain with RC coupling. This is because, as has been mentioned before, the d-c plate voltage on an RC stage is approximately one-half the plate supply voltage.

Impedance-Transformer and Resistance-Transformer Coupling These two circuit arrangements, illustrated in figures 10E and 10F, are employed when it is desired to use transformer coupling for the reasons cited above, but where it is desired that the d-c plate current of the amplifier stage be isolated from the primary of the coupling transformer. With most types of high-permeability wide response transformers it is necessary that there be no d-c flow through the windings of the transformer. The impedance-transformer arrangement of figure 10E will give a higher voltage output from the stage but is not often used since the plate coupling impedance (choke) must have very high inductance and very low distributed capacitance in order not to restrict the range of the transformer which it and its associated tube feed. The resistance-transformer arrangement of figure 10F is ordinarily satisfactory where it is desired to feed a transformer from a voltage-amplifier stage with no direct current in the transformer primary.

Cathode Coupling The *cathode-coupling* arrangement of figure 10G has been widely used only comparatively recently. One outstanding characteristic of such a circuit is that there is no phase reversal between the grid and the plate circuit. All other common types of interstage coupling are accompanied by a 180° phase reversal between the grid circuit and the plate circuit of the tube.

Figure 11 gives the expressions for determining the appropriate factors for an equivalent triode obtained through the use of a pair of similar triodes connected in the cathode-coupled circuit shown. With these equivalent triode factors it is possible to use the expressions shown in figure 5 to determine the gain of the stage at different frequencies. The input capacitance of such a stage is less than that of one of the triodes, the effective grid-to-plate capacitance is very much less (it is so much less that such a stage may be used as an r-f amplifier without neutralization), and the output capacitance is approximately equal to the grid-to-plate capacitance of one of the triode sections. This circuit is particularly effective with tubes such as the 6J6, 12AU7, and 12AT7, which have two similar triodes in one envelope. An appropriate value of cathode resistor to use for such a stage is the value which would be used for the cathode resistor of a conventional amplifier using *one* of the same type tubes with the values of plate voltage and load resistance to be used for the cathode-coupled stage.

Inspection of the equations in figure 11 shows that as the cathode resistor is made smaller to approach zero, G_m approaches zero, the plate resistance approaches the R_p of one tube, and the μ approaches zero. Since the cathode resistor is made very large the G_m approaches one-half that of a single tube of the same type, the plate resistance approaches twice that of one tube, and the μ approaches the same value as one tube. But since the G_m of each tube decreases as the cathode resistor is made larger (the plate current will decrease on each tube) the optimum value of cathode resistor will be found to be in the vicinity of the value mentioned in the previous paragraph.

Direct Coupling *Direct coupling* between successive amplifier stages (plate of first stage connected directly to the grid of the succeeding stage) is complicated by the fact that the grid of an amplifier stage must be operated at an average negative potential with respect to the cathode of that stage. However, if the cathode of the second amplifier stage can be operated at a potential more positive than the plate of the preceding stage by the amount of the grid bias on the second amplifier stage, this direct connection between the plate of one stage and the grid of the succeeding stage can be used. Figure 10H illustrates an application of this principle in the coupling of a pentode amplifier stage to the grid of a *hot-cathode* phase inverter. In this arrangement the values of cathode, screen, and plate re-

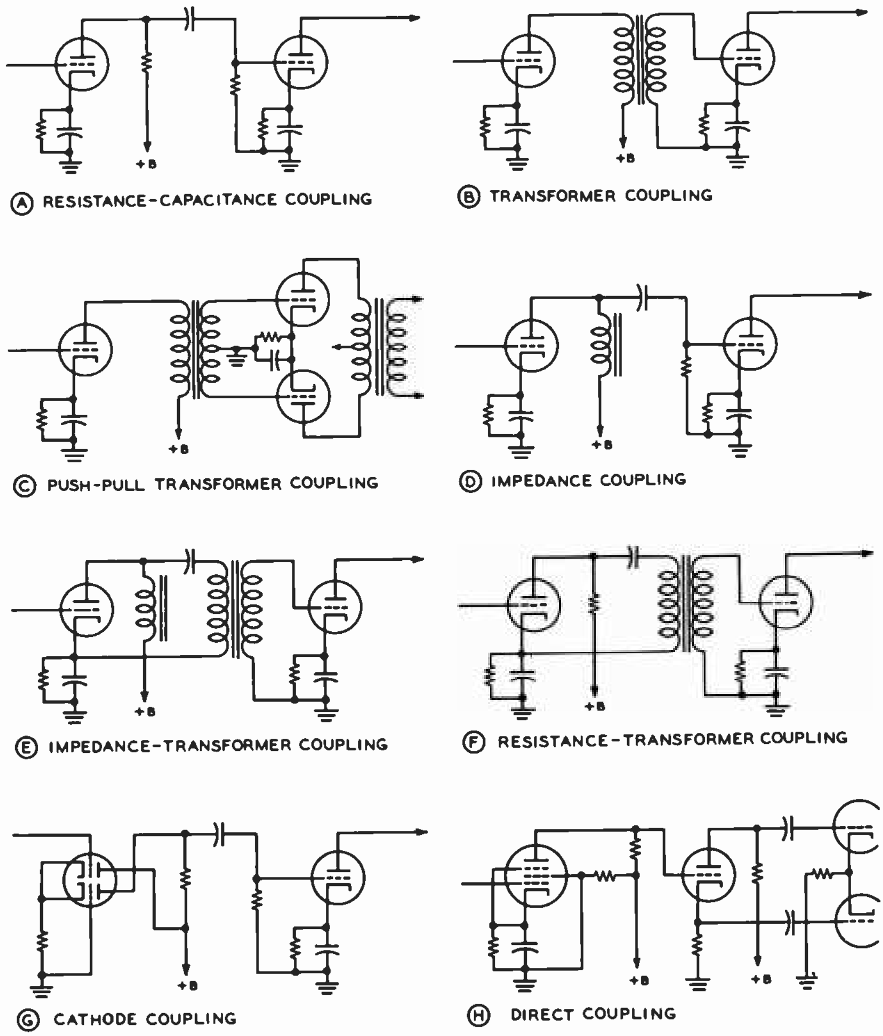


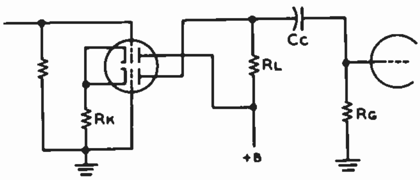
Figure 10

INTERSTAGE COUPLING METHODS FOR AUDIO-FREQUENCY VOLTAGE AMPLIFIERS

sistors in the pentode stage are chosen so that the plate of the pentode is at approximately one-third of the plate supply potential. The succeeding phase-inverter stage then operates with conventional values of cathode and plate resistor (same value of resistance) in its normal manner. This type of phase inverter is described in more detail in the section to follow.

6-8 Phase Inverters

In order to excite the grids of a push-pull stage it is necessary that voltages equal in amplitude and opposite in polarity be applied to the two grids. These voltages may be obtained through the use of a push-pull input transformer such as is shown in figure 10C. It is possible also, without the attendant bulk and expense of a push-pull input transformer, to obtain voltages of the proper polarity and phase through the use of a so-called *phase-inverter* stage. There are a large number of phase-inversion circuits which have been developed and applied but the three shown in figure 12 have been found



$$G_M' = -G_M \frac{G}{2G+1} \qquad G = R_K G_M \left(1+\frac{1}{\mu}\right)$$

R_K = CATHODE RESISTOR

$$R_P' = R_P \frac{2G+1}{G+1} \qquad G_M = G_M \text{ OF EACH TUBE}$$

$\mu = \mu$ OF EACH TUBE

$$\mu' = -\mu \frac{G}{G+1} \qquad R_P = R_P \text{ OF EACH TUBE}$$

EQUIVALENT FACTORS INDICATED ABOVE BY (') ARE THOSE OBTAINED BY USING AN AMPLIFIER WITH A PAIR OF SIMILAR TUBE TYPES IN CIRCUIT SHOWN ABOVE.

Figure 11

Equivalent factors for a pair of similar triodes operating as a cathode-coupled audio-frequency voltage amplifier.

over a period of time to be the most satisfactory from the point of view of the number of components required and from the standpoint of the accuracy with which the two out-of-phase voltages are held to the same amplitude with variations in supply voltage and changes in tubes.

All of these vacuum-tube phase inverters are based on the fact that a 180° phase shift occurs within a vacuum tube between the grid input voltage and the plate output voltage. In certain circuits, the fact that the grid input voltage and the voltage appearing across the cathode bias resistor are in phase, is used for phase-inversion purposes.

"Hot-Cathode" Phase Inverter Figure 12A illustrates the *hot-cathode* type of phase inverter. This phase inverter is the simplest of the three types since it requires only one tube and a minimum of circuit components. It is particularly simple when directly coupled from the plate of a pentode amplifier stage as shown in figure 10H. The circuit does, however, possess the following two disadvantages: (1) the cathode of the tube must run at a potential of approximately one-third the plate supply voltage above the heater when a grounded common heater winding is used for this tube as well as the other heater-cathode tubes in a receiver or amplifier; (2) the circuit actually has a loss in voltage from its input to either of the output grids —about 0.9 times the input voltage will be applied to each of these grids. This does represent a voltage gain of about 1.8 in *total* voltage output with respect to input (grid-to-grid output voltage) but it is still small with respect to the other two phase-inverter circuits shown.

Recommended component values for use with a 6C4 tube in this circuit are shown in figure 12A. If it is desired to use another tube in this circuit, appropriate values for the operation of that tube as a conventional amplifier can be obtained from manufacturer's tube data. The designated value of R_L should be divided by two, and this new value of resistance placed in the circuit as R_L. The value of R_k from tube-manual tables should then be used as R_{k1} in this circuit, and the total of R_{k1} and R_{k2} should be equal to R_L.

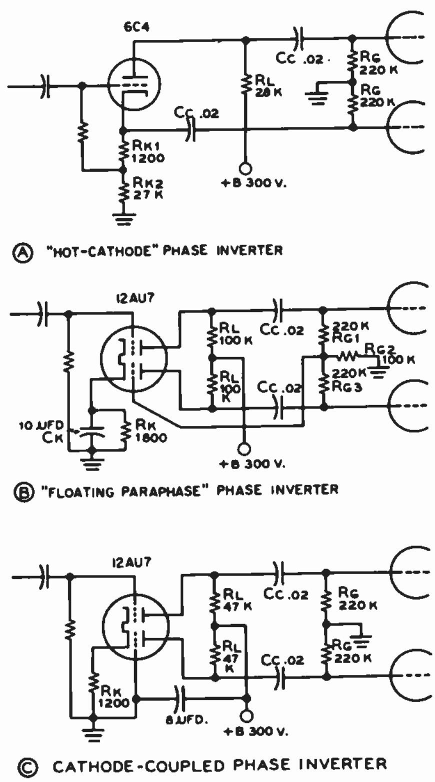


Figure 12

THREE TYPICAL PHASE-INVERTER CIRCUITS WITH RECOMMENDED VALUES FOR CIRCUIT COMPONENTS

"Floating Paraphase" Phase Inverter An alternate type of phase inverter sometimes called the *floating paraphase* is illustrated in figure 12B. This circuit is quite often used with a 12AU7 tube, and appropriate values for this tube in a typical inverter circuit are shown. Using the component values given will provide a voltage gain of approximately 12 from the input grid to each of the grids of the succeeding stage. It is capable of approximately 70 volts peak output to each grid.

The circuit inherently has a small unbalance in output voltage. This unbalance can be eliminated, if it is required for some special application, by making the resistor R_{g1} a few percent lower in resistance value than R_{g3}.

Cathode-Coupled Phase Inverter The circuit shown in figure 12C gives approximately one half the voltage gain from the input grid to either of the grids of the succeeding stage that would be obtained from a single tube of the same type operating as a conventional RC amplifier stage. Thus, with a 12AU7 tube as shown (two 6C4's in one envelope) the voltage gain from the input grid to either of the output grids will be approximately 7—the gain is, of course, 14 from the input to both output grids. The phase characteristics are such that the circuit is commonly used in deriving push-pull deflection voltage for a cathode-ray tube from a single-ended input signal.

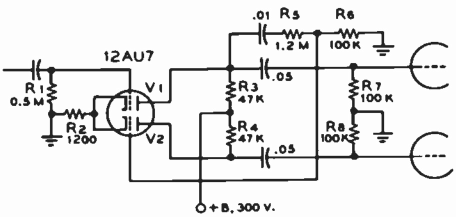


Figure 13

VOLTAGE-DIVIDER PHASE INVERTER

The first section of the 12AU7 is used as an amplifier to increase the amplitude of the applied signal to the desired level. The second section of the 12AU7 is used as an inverter and amplifier to produce a signal of the same amplitude but of opposite polarity. Since the common cathode resistor (R_k) is not bypassed the voltage across it is the algebraic sum of the two plate currents and has the same shape and polarity as the voltage applied to the input grid of the first half of the 12AU7. When a signal (e) is applied to the input circuit, the effective grid-cathode voltage of the first section is $Ae/2$, when A is the gain of the first section. Since the grid of the second section of the 12AU7 is grounded, the effect of the signal voltage across R_k (equal to $e/2$ if R_k is the proper value) is the same as though a signal of the same amplitude but of opposite polarity were applied to the grid. The output of the second section is equal to $-Ae/2$ if the plate load resistors are the same for both tube sections.

Voltage-Divider Phase Inverter A commonly used phase inverter is shown in figure 13. The input section (V_1) is connected as a conventional amplifier. The output voltage from V_1 is impressed on the voltage divider R_5-R_6. The values of R_5 and R_6 are in such a ratio that the voltage impressed on the grid of V_2 is $1/A$ times the output voltage of V_1, where A is the amplification factor of V_1. The output of V_2 is then of the same amplitude as the output of V_1, but of opposite phase.

6-9 D-C Amplifiers

Direct-current amplifiers are special types used where amplification of very slow variations in voltage, or of d-c voltages is desired. A simple d-c amplifier consists of a single tube with a grid resistor across the input terminals, and the load in the plate circuit.

Basic D-C Amplifier Circuit A simple d-c amplifier circuit is shown in figure 14, wherein the grid of one tube is connected directly to the plate of the preceding tube in such a manner that voltage changes on the grid of the first tube will be amplified by the system. The voltage drop across the plate coupling resistor is impressed directly on the grid of the second tube, which is provided with enough negative grid bias to balance out the excessive voltage drop across the coupling resistor. The grid of the

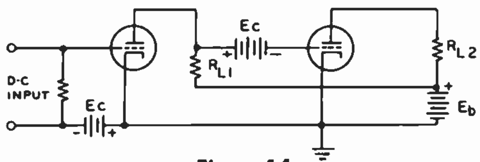


Figure 14

DIRECT-COUPLED D-C AMPLIFIER

second tube is thus maintained in a slightly negative position.

The d-c amplifier will provide good low-frequency response, with negligible phase distortion. High-frequency response is limited by the shunting effect of the tube capacitances, as in the normal resistance-coupled amplifier.

A common fault with d-c amplifiers of all types is static instability. Small changes in the filament, plate, or grid voltages cannot be distinguished from the exciting voltage. Regulated power supplies and special balancing circuits have been devised to reduce the effects of supply variations on these amplifiers. A successful system is to apply the plate potential in phase to two tubes, and to apply the exciting signal to a push-pull grid-circuit configuration. If the two tubes are identical, any change in electrode voltage is balanced out. The use of negative feedback can also greatly reduce drift problems.

The "Loftin-White" Circuit Two d-c amplifier stages may be arranged, so that their plate supplies are effectively in series, as illustrated in figure 15. This is known as a *Loftin-White* amplifier. All plate and grid voltages may be obtained from one master power supply instead of separate grid and plate supplies. A push-pull version of this amplifier (figure 16) can be used to balance out the effects of slow variations in the supply voltage.

6-10 Single-Ended Triode Amplifiers

Figure 17 illustrates five circuits for the operation of class-A triode amplifier stages. Since the cathode current of a triode class-A (no grid current) amplifier stage is constant with and without excitation, it is common practice to operate the tube with cathode bias. Recommended operating conditions in regard to plate voltage, grid bias, and load

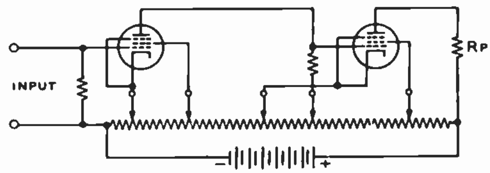


Figure 15

LOFTIN-WHITE D-C AMPLIFIER

impedance for conventional triode amplifier stages are given in the RCA Receiving Tube Manuals.

Extended Class-A Operation It is possible, under certain conditions, to operate single-ended triode amplifier stages (and pentode and tetrode stages as well) with grid excitation of sufficient amplitude that grid current is taken by the tube on peaks. This type of operation is called class-A_2 and is characterized by increased plate-circuit efficiency over straight class-A amplification without grid current. The normal class-A amplifier power stage will operate with a plate-circuit efficiency of from 20 percent to perhaps 35 percent.

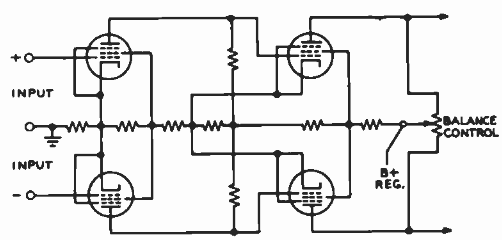


Figure 16

PUSH-PULL D-C AMPLIFIER WITH EITHER SINGLE-ENDED OR PUSH-PULL INPUT

Through the use of class-A_2 operation it is possible to increase this plate-circuit efficiency to approximately 38 to 45 percent. However, such operation requires careful choice of the value of plate load impedance, a grid-bias supply with good regulation (since the tube draws grid current on peaks although the plate current does not change with signal), and a driver tube with moderate power capability to excite the grid of the class A_2 tube.

Figures 17D and 17E illustrate two methods of connection for such stages. Tubes such as the 845, 450TL, and 304TL are suitable for these circuits. In each case the grid bias is approximately the same as would be used for a class-A amplifier using the same tube, and as mentioned before, fixed bias must be used along with an audio driver of good regulation—preferably a triode stage with a 1:1 or step-down driver transformer. In each case it will be found that the correct value of plate load impedance will be increased about 40 percent over the value

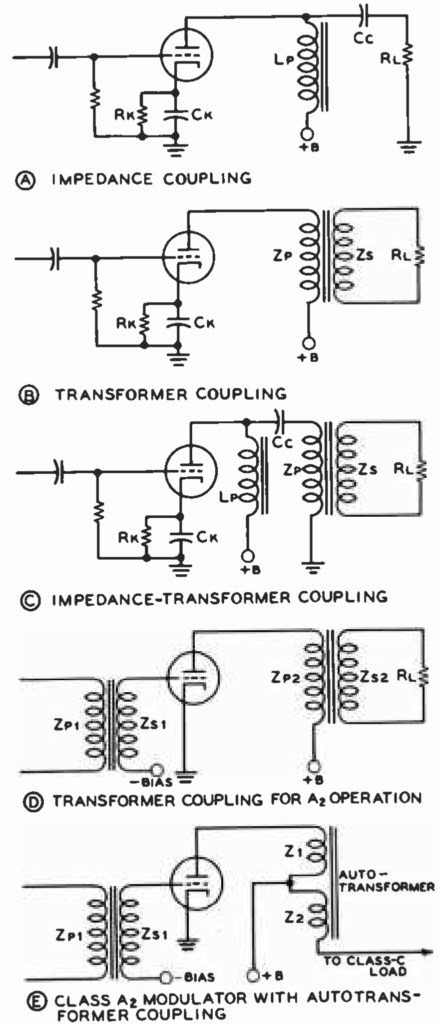


Figure 17

Output coupling arrangements for single-ended class-A triode audio-frequency power amplifiers.

recommended by the tube manufacturer for class-A operation of the tube.

Operation Characteristics of a Triode Power Amplifier A class-A power amplifier operates in such a wav as to amplify as faithfully as possible the waveform applied to the grid of the tube. Large power output is of more importance than high voltage amplification, consequently gain characteristics may be sacrificed in power-tube design to obtain more important power-handling capabilities. Class-A power tubes, such as the 12BY4A, 2A3, and 6AS7G, are characterized by a low amplification factor, high plate dissipation, and relatively high filament emission.

The operating characteristics of a class-A triode amplifier employing an output-transformer coupled load may be calculated from the plate family of curves for the particular tube in question by employing the following steps:

1. The load resistance should be approximately twice the plate resistance of the tube for maximum undistorted power output. Remember this fact for a quick check on calculations.
2. Calculate the zero-signal bias voltage (E_{c1}).

$$E_{c1} = \frac{-(0.68 \times E_b)}{\mu}$$

3. Locate the E_{c1} bias point on the I_b versus E_b graph where the E_c bias line crosses the plate-voltage line, as shown in figure 18. Call this point *P*.
4. Locate on the plate family of curves the value of zero-signal plate current, (I_b) corresponding to operating point *P*.
5. Locate 2 × I_b (twice the value of I_b) on the plate-current axis (*Y* axis). This point corresponds to the value of maximum-signal plate current ($i_{b\,max}$).
6. Locate point *x* on the d-c bias curve at zero volts ($E_c = 0$), corresponding to the value of $i_{b\,max}$.
7. Draw a straight line ($x - y$) through points *x* and *P*. This line is *the load-resistance line*. Its slope corresponds to the value of the load resistance.
8. Load resistance, (in ohms) equals:

$$R_L = \frac{e_{b\,max} - e_{b\,min}}{i_{b\,max} - i_{b\,min}}$$

9. Check: Multiply the zero-signal plate current (I_b) by the operating plate voltage, (E_b). If the plate dissipation rating of the tube is exceeded, it is necessary to increase the bias (E_c) on the tube so that the plate dissipation falls within the maximum rating of the tube. If this step is taken, operations 2 through 8 must be repeated with the new value of E_c.
10. For maximum power output, the peak a-c grid voltage on the tube should swing to $2E_c$ on the negative cycle, and to zero-bias on the positive cycle. At the peak of the negative swing, the plate voltage reaches $e_{b\,max}$ and the plate current drops to $i_{b\,min}$. On the positive swing of the grid signal, the plate voltage drops to $e_{b\,min}$ and the plate current reaches $i_{b\,max}$. The power output of the tube in watts is:

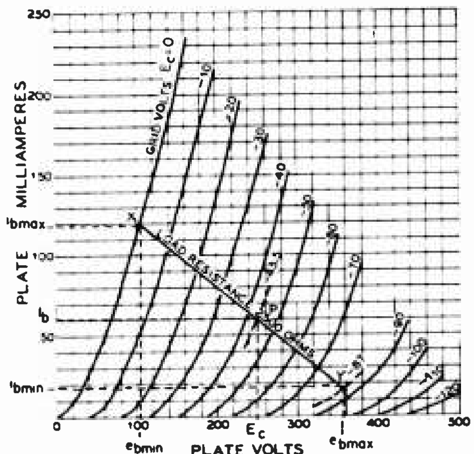
$$P_o = \frac{(i_{b\,max} - i_{b\,min}) \times (e_{b\,max} - e_{b\,min})}{8}$$

where,
i is in amperes,
e is in volts.

11. The second-harmonic distortion generated in a single-ended class-A triode amplifier, expressed as a percentage of the fundamental output signal is:
% 2nd harmonic =

$$\frac{\dfrac{(i_{b\,max} - i_{b\,min})}{2} - I_b}{i_{b\,max} - i_{b\,min}} \times 100$$

Figure 18 illustrates the above steps as applied to a single class-A 2A3 amplifier stage.



AVERAGE PLATE CHARACTERISTICS – 2A3

μ = 4.2 R_p = 800 OHMS
PLATE DISSIPATION = 15 WATTS

LOAD RESISTANCE

$$R_L = \frac{e_{bmax} - e_{bmin}}{i_{bmax} - i_{bmin}} \text{ OHMS}$$

POWER OUTPUT

$$P_o = \frac{(i_{bmax} - i_{bmin})\ (e_{bmax} - e_{bmin})}{8} \text{ WATTS}$$

SECOND-HARMONIC DISTORTION

$$D_2 = \frac{\dfrac{(i_{bmax} + i_{bmin})}{2} - i_b}{i_{bmax} - i_{bmin}} \times 100 \text{ PERCENT}$$

Figure 18

Formulas for determining the operating conditions of a class-A triode single-ended audio-frequency power output stage. A typical load line has been drawn on the average plate characteristics of a type 2A3 tube to illustrate the procedure.

6-11 Single-Ended Pentode Amplifiers

Figure 19 illustrates the conventional circuit for a single-ended tetrode or pentode amplifier stage. Tubes of this type have largely replaced triodes in the output stage of receivers and amplifiers due to the higher plate efficiency (30%—40%) at which they operate. Tetrode and pentode tubes do, however, introduce a considerably greater amount of harmonic distortion in their output circuit, particularly odd harmonics. In addition, their plate-circuit impedance (which acts in an amplifier to damp speaker overshoot and ringing, and acts in a driver stage to provide good regulation) is many times higher than that of an equivalent triode. The application of negative feedback acts both to reduce distortion and to reduce the effective plate-circuit impedance of these tubes.

Operating Characteristics of a Pentode Power Amplifier The operating characteristics of pentode power amplifiers may be obtained from the plate family of curves, much as in the manner applied to triode tubes. A typical family of pentode plate curves is shown in figure 20.

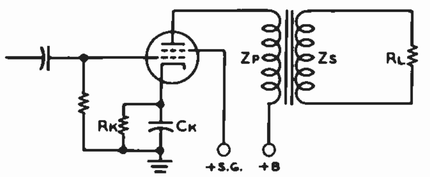


Figure 19

Conventional single-ended pentode or beam tetrode audio-frequency power-output stage.

The plate current of the pentode tube is relatively independent of the applied plate voltage, but is sensitive to screen voltage. In general, the correct pentode load resistance is about

$$\frac{0.9\,E_b}{I_b}$$

and the power output is somewhat less than

$$\frac{E_b \times I_b}{2}$$

These formulas may be used for a quick check on more precise calculations. To obtain the operating parameters for class-A pentode amplifiers, the following steps are taken:

1. The $i_{b\,max}$ point is chosen so as to fall on the zero-bias curve, just above the "knee" of the curve (point *A*, figure 20).
2. A preliminary operating point (*P*) is determined by the intersection of the plate-voltage line (E_b) and the line of $i_{b\,max}/2$. The grid-voltage curve that this point falls on should be one that is about ½ the value of E_c required to cut the plate current to a very low value (point *B*). Point *B* represents $i_{b\,min}$ on the plate-current axis (*y* axis). The line $i_{b\,max}/2$ should be located halfway between $i_{b\,max}$ and $i_{b\,min}$.
3. A trial load line is constructed about point *P* and point *A* in such a way that the lengths *AP* and *PB* are approximately equal.
4. When the most satisfactory load line has been determined, the load resistance may be calculated:

$$R_L = \frac{e_{b\,max} - e_{b\,min}}{i_{b\,max} - i_{b\,min}}$$

5. The operating bias (E_c) is the bias at point *P*.
6. The power output is:

$$\frac{(i_{b\,max} - i_{b\,min}) + 1.41\,(I_x - I_y)^2 \times R_L}{32}$$

where,
I_x is the plate current at the point on the load line where the grid voltage (e_c) is equal to: $E_c - 0.7\,E_c$,
I_y is the plate current at the point where, e_c is equal to: $E_c + 0.7\,E_c$.

7. The percentage harmonic distortion is:
% 2nd harmonic distortion =

$$\frac{i_{b\,max} - i_{b\,min} - 2I_{bo}}{i_{b\,max} - i_{b\,min} + 1.41\,(I_x - I_y)} \times 100$$

where,
I_{bo} is the static plate current of the tube.
% 3rd harmonic distortion =

$$\frac{i_{b\,max} - i_{b\,min} - 1.41\,(I_x - I_y)}{i_{b\,max} - i_{b\,min} + 1.41\,(I_x - I_y)} \times 100$$

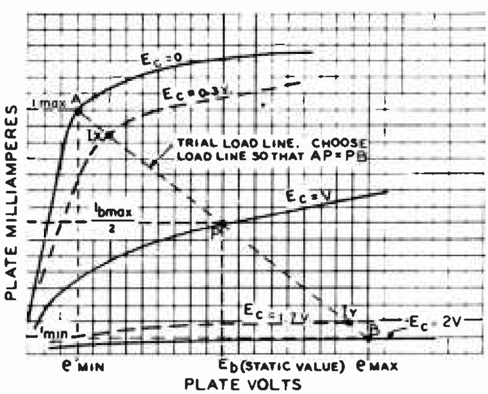
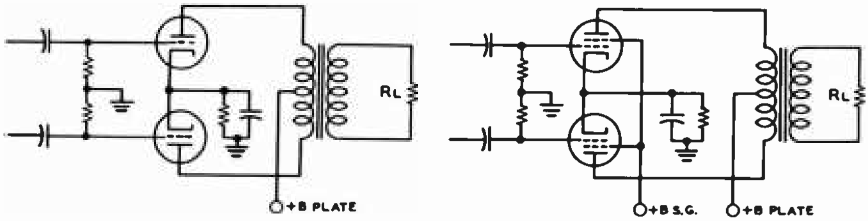


Figure 20

GRAPHIC DETERMINATION OF OPERATING CHARACTERISTICS OF A PENTODE POWER AMPLIFIER

"V" is the negative control grid voltage at the operating point P.



PUSH-PULL TRIODE AND TETRODE

Figure 21

CONVENTIONAL PUSH-PULL CIRCUITS

6-12 Push-Pull Audio Amplifiers

A number of advantages are obtained through the use of the push-pull connection of two or four tubes in an audio-frequency power amplifier. Two conventional circuits for the use of triode and tetrode tubes in the push-pull connection are shown in figure 21. The two main advantages of the push-pull circuit arrangement are: (1) the magnetizing effect of the plate currents of the output tubes is cancelled in the windings of the output transformer; (2) even harmonics of the input signal (second and fourth harmonics primarily) generated in the push-pull stage are cancelled when the tubes are balanced.

The cancellation of even harmonics generated in the stage allows the tubes to be operated class AB—in other words the tubes may be operated with bias and input signals of such amplitude that the plate current of alternate tubes may be cut off during a portion of the input voltage cycle. If a tube were operated in such a manner in a single-ended amplifier the second-harmonic amplitude generated would be prohibitively high.

Push-pull class-AB operation allows a plate circuit efficiency of from 45 to 60 percent to be obtained in an amplifier stage depending on whether or not the exciting voltage is of such amplitude that grid current is drawn by the tubes. If grid current is taken on input voltage peaks the amplifier is said to be operating class-AB_2 and the plate-circuit efficiency can be as high as the upper value just mentioned. If grid current is not taken by the stage it is said to be operating class-AB_1 and the plate-circuit efficiency will be toward the lower end of the range just quoted. In all class-AB amplifiers the plate current will increase from 40 to 150 percent over the no-signal value when full excitation voltage is applied.

Operating Characteristics of Push-Pull Class-A Triode Power Amplifier The operating characteristics of push-pull class-A amplifiers may also be determined from the plate family of curves for a particular triode tube by the following steps:

1. Erect a vertical line from the plate-voltage axis (x-axis) at 0.6 E_b (figure 22), which intersects the $E_c = 0$ curve. This point of intersection (P), interpolated to the plate current axis (y-axis), may be taken as $i_{b\,max}$. It is assumed for simplification that $i_{b\,max}$ occurs at the point of the zero-bias curve corresponding to 0.6 E_b.
2. The power output obtainable from the two tubes is:

$$(P_o) = \frac{i_{b\,max} \times E_b}{5}$$

where,

P_o is expressed in watts,
$i_{b\,max}$ is in amperes,
E_b is the applied plate voltage.

3. Draw a preliminary load line through point P to the E_b point located on the x-axis (the zero plate-current line). This load line represents ¼ of the actual plate-to-plate load of the class-A tubes. Therefore:

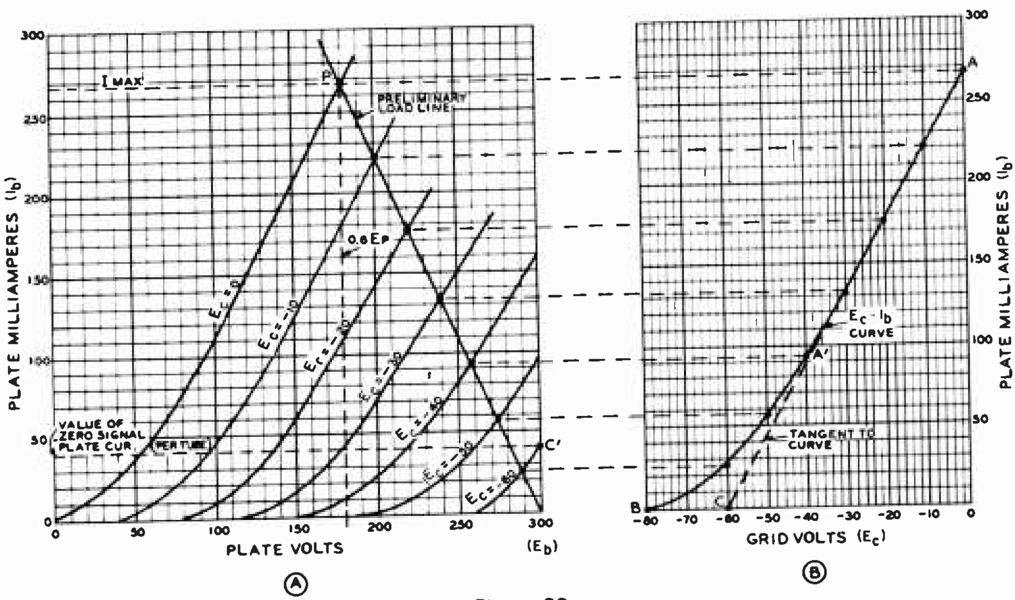


Figure 22

DETERMINATION OF OPERATING PARAMETERS FOR PUSH-PULL CLASS-A TRIODE TUBES

$$R_L \text{ (plate-to-plate)} = 4 \times \frac{E_b - 0.6\,E_b}{i_{b\,max}}$$

$$= \frac{1.6\,E_b}{i_{b\,max}}$$

Figure 22 illustrates the above steps applied to a push-pull class-A amplifier using two 2A3 tubes.

4. The average plate current is 0.636 $i_{b\,max}$, and multiplied by plate voltage E_b, will give the average watts input to the plates of the two tubes. The power output should be subtracted from this value to obtain the total operating plate dissipation of the two tubes. If the plate dissipation is excessive, a slightly higher value of R_L should be chosen to limit the plate dissipation.
5. The correct value of operating bias, and the static plate current for the push-pull tubes may be determined from the E_C versus I_b curves, which are a derivation of the E_b versus I_b curves for various values of E_c.
6. The E_c versus I_b curve may be constructed in this manner: Values of grid bias are read from the intersection of each grid-bias curve with the load line. These points are transferred to the E_c versus I_b graph to produce a curved line, *A-B*. If the grid bias curves of the E_b versus I_c graph were straight lines, the lines of the E_c versus I_b graph would also be straight. This is usually not the case. A tangent to this curve is therefore drawn, starting at point *A'*, and intersecting the grid-voltage abscissa (*x*-axis). This intersection (*C*) is the operating-bias point for fixed-bias operation.
7. This operating-bias point may now be plotted on the original E_c versus I_b family of curves (*C'*), and the zero-signal current produced by this bias is determined. This operating bias point (*C'*) does not fall on the operating load line, as in the case of a single-ended amplifier.
8. Under conditions of maximum power output, the exciting signal voltage swings from zero-bias voltage to zero-bias voltage for each of the tubes on each half of the signal cycle. Second-harmonic distortion is largely cancelled out.

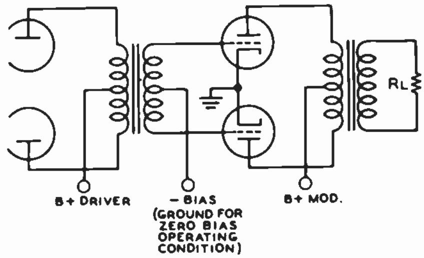


Figure 23

CLASS-B AUDIO-FREQUENCY POWER AMPLIFIER

6-13 Class-B Audio-Frequency Power Amplifiers

The *class-B audio-frequency power amplifier* (figure 23) operates at a higher plate-circuit efficiency than any of the previously described types of audio power amplifiers. Full-signal plate-circuit efficiencies of 60 to 70 percent are readily obtainable with the tube types presently available for this mode of operation. Since the plate-circuit efficiency is higher, smaller tubes of lower plate dissipation may be used in a class-B power amplifier of a given power output than can be used in any other conventional type of audio amplifier. An additional factor in favor of the class-B audio amplifier is the fact that the power input to the stage is relatively low under no-signal conditions. It is for these reasons that this type of amplifier has largely superseded other types for the generation of audio-frequency levels from perhaps 100 watts on up to levels of approximately 150,000 watts as required for large short-wave broadcast stations.

Disadvantages of Class-B Amplifier Operation There are attendant disadvantageous features to the operation of a power amplifier of this type; but all these disadvantages can be overcome by proper design of the circuits associated with the power-amplifier stage. These disadvantages are: (1) The class-B audio amplifier requires driving power in its grid circuit; this requirement can be overcome by the use of an oversize power stage preceding the class-B stage with a step-down transformer between the driver stage and the class-B grids. Degenerative feedback is sometimes employed to reduce the plate impedance of the driver stage and thus to improve the voltage regulation under the varying load presented by the class-B grids. (2) The class-B stage requires a constant value of average grid bias to be supplied in spite of the fact that the grid current of the stage is zero over most of the cycle but rises to value as high as one-third of the peak plate current at the peak of the exciting voltage cycle. Special regulated bias supplies have been used for this application, or B batteries can be used. However, a number of tubes especially designed for class-B audio amplifiers have been developed which require zero average grid bias for their operation. The 811A, 805, 3-400Z, and 3-1000Z are examples of this type of tube. All these so-called *zero-bias* tubes have rated operating conditions up to moderate plate voltages wherein they can be operated without grid bias. As the plate voltage is increased to the maximum ratings, however, a small amount of grid bias, such as could be obtained from a regulated bias supply, is required. (3), A class-B audio-frequency power amplifier or modulator requires a source of plate-supply voltage having reasonably good regulation. This requirement led to the development of the *swinging choke.* The swinging choke is essentially a conventional filter choke in which the core air gap has been reduced. This reduction in the air gap allows the choke to have a much greater value of inductance with low-current values such as are encountered with no signal or small signal being applied to the class-B stage.

With a higher value of current such as would be taken by a class-B stage with full signal applied, the inductance of the choke drops to a much lower value. With a swinging choke of this type, having adequate current rating, as the input inductor in the filter system for a rectifier power supply, the regulation will be improved to a point which is normally adequate for a power supply for a class-B amplifier or modulator stage.

Calculation of Operating Conditions of Class-B Power Amplifiers The following procedure can be used for the calculation of the operating conditions of class-B power amplifiers when they are to operate into a resistive load such as presented by a class-C power amplifier. This procedure will be found quite satisfactory for the application of vacuum tubes as class-B modulators when it is desired to operate the tubes under conditions which are not specified in the tube operating characteristics published by the tube manufacturer. The same procedure can be used with equal effectiveness for the calculation of the operating conditions of beam tetrodes as class-AB_2 amplifiers or modulators when the resting plate current of the tubes (no-signal condition) is less than 25 or 30 percent of the maximum-signal plate current.

1. With the average plate characteristics of the tube as published by the manufacturer before you, select a point on the $E_b = E_c$ (diode bend) line at about twice the plate current you expect the tubes to draw under modulation peaks. If beam tetrode tubes are concerned, select a point at about the same amount of plate current mentioned above, just to the right of the region where the I_b line takes a sharp curve downward. This will be the first trial point, and the plate voltage at the point chosen should be not more than about 20 percent of the d-c voltage applied to the tubes if good plate-circuit efficiency is desired.
2. Note down the value of $i_{b\,max}$ and $e_{b\,min}$ at this point.
3. Subtract the value of $e_{b\,min}$ from the d-c plate voltage on the tubes.
4. Substitute the values obtained in the following equations:

$$P_o \text{ (2 tubes)} = \frac{i_{b\,max}\,(E_b - e_{b\,min})}{2}$$

$$R_L \text{ (2 tubes)} = 4\,\frac{(E_b - e_{b\,min})}{i_{b\,max}}$$

$$\text{Full signal efficiency } (N_p) =$$

$$78.5\left(1 - \frac{(e_{b\,min})}{E_b}\right)$$

Effects of Speech Clipping All the above equations are true for sine-wave operating condition of the tubes concerned. However, if a speech clipper is being used in the speech amplifier, or if it is desired to calculate the operating conditions on the basis of the fact that the ratio of peak power to average power in a speech wave is approximately 4 to 1 as contrasted to the ratio of 2 to 1 in a sine wave — in other words, when nonsinusoidal waves such as plain speech or speech that has passed through a clipper are concerned, we are no longer concerned with average power output of the modulator as far as its capability of modulating a class-C amplifier is concerned; we are concerned with its *peak power output* capability.

Under these conditions we call on other, more general relationships. The first of these is: it requires a *peak* power output *equal* to the class-C stage input to modulate that input fully.

The second relationship is: the average power output required of the modulator is equal to the shape factor of the modulating wave multiplied by the input to the class-C stage. The shape factor of unclipped speech is approximately 0.25. The shape factor of a sine wave is 0.5. The shape factor of a speech wave that has been passed through a clipper-filter arrangement is somewhere between 0.25 and 0.9 depending on the amount of clipping that has taken place. With 15 or 20 db of clipping the shape factor may be as high as the figure of 0.9 mentioned above. This means that the audio power output of the modulator will be 90% of the input to the class-C stage. Thus with a kilowatt input we would be putting 900 watts of audio into the class-C stage for 100 percent modulation as contrasted to perhaps 250 watts for unclipped speech modulation of 100 percent.

Sample Calculation for 811A Tubes Figure 24 shows a set of plate characteristics for a type 811A tube with a load line for class-B operation. Figure 25 lists a sample calculation for determining the proper operating conditions for obtaining approximately 185 watts output from a pair of the tubes with 1000 volts d-c plate potential. Also shown in figure 25 is the method of determining the proper ratio for the modulation transformer to couple between the 811's or 811A's and the anticipated final amplifier which is to operate at 2000 plate volts and 175 ma plate current.

Modulation Transformer Calculation The method illustrated in figure 25 can be used in general for the determination of the proper transformer ratio to couple between the modulator tube and the amplifier to be modulated. The procedure can be stated as follows: (1) Determine the proper plate-to-plate load impedance for the modulator tubes either by the use of the type of calculation shown in figure 25, or by reference to the published characteristics on the tubes to be used. (2) Determine the load impedance which will be presented by the class-C amplifier stage to be modulated by dividing the operating plate voltage on that stage by the operating value of plate current in *amperes.* (3) Divide the class-C load impedance determined in (2) above by the plate-to-plate load impedance for the modulator tubes determined in (1) above. The ratio determined in this way is the secondary-to-primary *impedance* ratio. (4) Take the square root of this ratio to determine the secondary-to-primary *turns ratio.* If the turns ratio is greater than unity, the use of a step-up transformer is required. If the turns ratio as determined in this way is less than unity, a step-down transformer is called for.

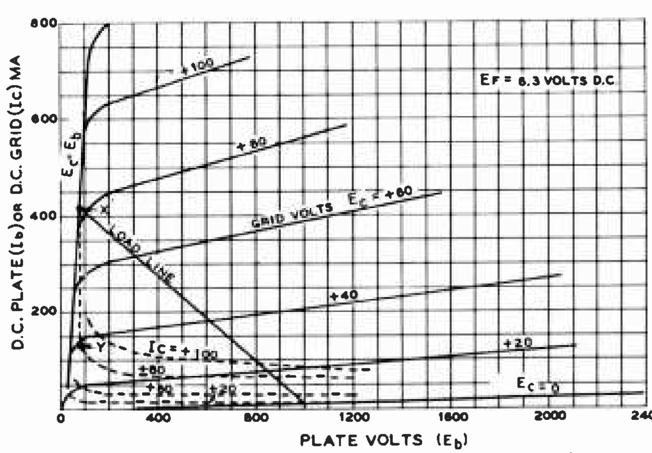
If the procedure shown in figure 25 has been used to calculate the operating conditions for the modulator tubes, the transformer ratio calculation can be checked in the following manner: Divide the plate voltage on the modulated amplifier by the total voltage swing on the modulator tubes ($2 \times [E_b - e_{b\,min}]$). This ratio should be quite close numerically to the transformer turns ratio as previously determined. The reason for this condition is that the ratio between the total primary voltage and the d-c plate-supply voltage on the modulated stage is equal to the turns ratio of the transformer, since a peak secondary voltage equal to the plate voltage on the modulated stage is required to modulate this stage 100 percent.

Use of Clipper Speech Amplifier with Tetrode Modulator Tubes When a clipper speech amplifier is used in conjunction with a class-B modulator stage, the plate current on that stage will rise to a higher value with modulation (due to the greater average power output and input) but the plate dissipation on the tubes will ordinarily be less than with sine-wave modulation. However, when tetrode tubes are used as modulators, the screen dissipation will be much greater than with sine-wave modulation. Care must be taken to ensure that the screen dissipation rating on the modulator tubes is not exceeded under full modulation conditions with a clipper speech amplifier. The screen dissipation is equal to screen voltage times screen current.

Practical Aspects of Class-B Modulators As stated previously, a class-B audio amplifier requires the driving

Figure 24

Typical class-B a-f amplifier load line. The load line has been drawn on the average characteristics of a type 811 tube.



AVERAGE PLATE CHARACTERISTICS TYPE 811 AND 811-A

SAMPLE CALCULATION

CONDITION: 2 TYPE 811 TUBES, $E_b = 1000$

INPUT TO FINAL STAGE, 350 W.

PEAK POWER OUTPUT NEEDED = 350 + 6% = 370 W.

FINAL AMPLIFIER E_b = 2000 V.

FINAL AMPLIFIER I_b = .175 A.

FINAL AMPLIFIER $Z_L = \frac{2000}{.175} = 11400\ \Omega$

EXAMPLE: CHOSE POINT ON 811 CHARACTERISTICS JUST TO RIGHT OF $E_b = E_c$. (POINT X, FIG. 24)

$I_{b\,max} = .410$ A. $\qquad E_{b\,min} = +100$

$I_{c\,max} = .100$ A. $\qquad E_{c\,min.} = +80$

PEAK $P_O = .410 \times (1000-100) = .410 \times 900 = 369$ W.

$R_L = 4 \times \frac{900}{.410} = 8800\ \Omega$

$N_P = 78.5\left(1 - \frac{100}{1000}\right) = 78.5\,(.9) = 70.5\ \%$

W_O (AVERAGE WITH SINE WAVE) $= \frac{P_{O(PEAK)}}{2} = 184.5$ W

$W_{IN} = \frac{184.5}{70.5} = 260$ W.

I_b (MAXIMUM WITH SINE WAVE) = 260 MA

W_G PEAK $= .100 \times 80 = 8$ W.

DRIVING POWER $= \frac{W_{G\,PK}}{2} = 4$ W.

TRANSFORMER:

$\frac{Z_S}{Z_P} = \frac{11400}{8800} = 1.29$

TURNS RATIO $= \sqrt{\frac{Z_S}{Z_P}} = \sqrt{1.29} = 1.14$ STEP UP

Figure 25

Typical calculation of operating conditions for a class-B a-f power amplifier using a pair of type 811 or 811A tubes. Plate characteristics and load line are shown in figure 24.

stage to supply well-regulated audio power to the grid circuit of the class-B stage. Since the performance of a class-B modulator may easily be impaired by an improperly designed driver stage, it is well to study the problems incurred in the design of the driver stage.

The grid circuit of a class-B modulator may be compared to a variable resistance which decreases in value as the exciting grid voltage is increased. This variable resistance appears across the secondary terminals of the driver transformer so that the driver stage is called on to deliver power to a varying load. For best operation of the class-B stage, the grid excitation voltage should not drop as the power taken by the grid circuit increases. These opposing conditions call for a high order of voltage regulation in the driver-stage plate circuit. In order to enhance the voltage regulation of this circuit, the driver tubes must have low plate resistance, the driver transformer must have as large a step-down ratio as possible, and the d-c resistance of both primary and secondary windings of the driver transformer should be low.

The driver transformer should reflect into the plate circuit of the driver tubes a load of such value that the required driving power is just developed with full excitation applied to the driver grid circuit. If this is done, the driver transformer will have as high a step-down ratio as is consistent with the maximum drive requirements of the class-B stage. If the step-down ratio of the driver transformer is too large, the driver plate load will be so high that the power required to drive the class-B stage to full output cannot be developed. If the step-down ratio is too small the regulation of the driver stage will be impaired.

Driver-Stage Calculations The parameters for the driver stage may be calculated from the plate characteristic curve, a sample of which is shown in figure 24. The required positive grid voltage ($e_{c\,max}$) for the 811A tubes used in the sample calculation is found at point *X*, the intersection of the load line and the peak plate current as found on the *y*-axis. This is +80 volts. If a vertical line is dropped from point *X* to intersect the dotted grid-current curves, it will be found that the grid current for a single 811A at this value of grid voltage is 100 milliamperes (point *Y*). The peak grid-driving power is therefore $80 \times 0.100 = 8$ watts. The approximate *average* driving power is 4 watts. This is an approximate figure because the grid impedance is not constant over the entire audio cycle.

A pair of 2A3 tubes will be used as drivers, operating class-A, with the maximum excitation to the drivers occurring just below the point of grid-current flow in the 2A3 tubes. The driver plate voltage is 300 volts, and the grid bias is −62 volts. The peak power (P_p) developed in the primary winding of the driver transformer is:

$$(P_p) = 2R_L\left(\frac{\mu e_{c\,max}}{R_p + R_L}\right)^2$$

where,

μ is the amplification factor of the driver tubes (4.2 for 2A3),

e_c is the peak grid swing of the driver stage (62 volts),

R_p is the plate resistance of one driver tube (800 ohms),

R_L is ½ the plate-to-plate load of the driver stage,

P_p (peak power in watts) is 8 watts.

Solving the above equation for R_L, we obtain a value of 14,500 ohms load, plate to plate for the 2A3 driver tubes.

The peak primary voltage (e_{pri}) is then found from the formula:

$$e_{pri} = 2R_L \times \frac{\mu e_{c\,max}}{R_p + R_L} = 493 \text{ volts}$$

and the turns ratio of the driver transformer (primary to ½ secondary) is:

$$\frac{e_{pri}}{e_{c\,max}} = \frac{493}{80} = 6.15{:}1$$

Plate Circuit Impedance Matching One of the most common causes of distortion in a class-B modulator is incorrect load impedance in the plate circuit. The purpose of the class-B modulation transformer is to take the power developed by the modulator (which has a certain operating impedance) and transform it to the operating impedance imposed by the modulated amplifier stage.

If the transformer in question has the same number of turns on the primary winding as it has on the secondary winding, the turns ratio is 1:1, and the impedance ratio is also 1:1. If a 10,000-ohm resistor is placed across the secondary terminals of the transformer, *a reflected load* of 10,000 ohms would appear across the primary terminals. If the resistor is changed to one of 2376 ohms, the reflected primary impedance would also be 2376 ohms.

If the transformer has twice as many turns on the secondary as on the primary, the turns ratio is 2:1. The impedance ratio is the square of the turns ratio, or 4:1. If a 10,000-ohm resistor is now placed across the secondary winding, a reflected load of 2500 ohms will appear across the primary winding.

Effects of Plate Circuit Mismatch It can be seen from the above paragraphs that the class-B modulator plate load is entirely dependent on the load placed on the secondary terminals of the class-B modulation transformer. If the secondary load is incorrect, certain changes will take place in the operation of the class-B modulator stage.

When the modulator load impedance is too low, the efficiency of the class-B stage is reduced and the plate dissipation of the tubes is increased. Peak plate current of the modulator stage is increased, and saturation of the modulation transformer core may result. "Talk-back" of the modulation transformer may result if the plate load impedance of the modulator stage is too low.

When the modulator load impedance is too high, the maximum power capability of the stage is reduced. An attempt to increase the output by increasing grid excitation to the stage will result in peak clipping of the audio wave. In addition, high peak voltages may be built up in the plate circuit that may damage the modulation transformer.

6-14 Cathode-Follower Power Amplifiers

The *cathode follower* is essentially a power output stage in which the exciting signal is applied between grid and ground. The plate is maintained at ground potential with respect to input and output signals, and the output signal is taken between cathode and ground.

Types of Cathode-Follower Amplifiers Figure 26 illustrates four types of cathode-follower power amplifiers in common usage and figure 27 shows the output impedance (R_L), and stage gain (A) of both triode and pentode (or tetrode) cathode-follower stages. It will be seen by inspection of the equations that the stage voltage gain is always less than unity, and that the output impedance of the stage is much less than the same stage operated as a conventional cathode-return amplifier. The output impedance for conventional tubes will be somewhere between 100 and 1000 ohms, depending primarily on the transconductance of the tube.

This reduction in gain and output impedance for the cathode follower comes about since the stage operates as though it has 100 percent degenerative feedback applied between its output and input circuit. Even though the voltage gain of the stage is reduced to a value less than unity by the action of the degenerative feedback, the power gain of the stage (if it is operating class-A) is not reduced. Although more voltage is required to excite a cathode-follower amplifier than appears across the load circuit (since the cathode "follows" along with the grid) the relative grid-to-cathode voltage is essentially the same as in a conventional amplifier.

Use of Cathode-Follower Amplifiers Although the cathode follower gives no voltage gain, it is an effective power amplifier where it is desired to feed a low-impedance load, or where it is desired to feed a load of varying impedance with a signal having good regulation. This latter capability makes the cathode follower particularly effective as a driver for the grids of a class-B modulator stage.

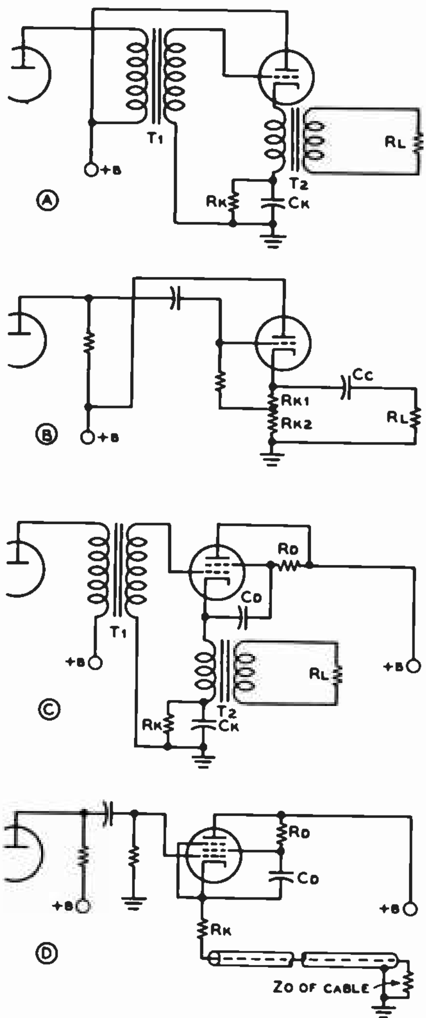


Figure 26

CATHODE-FOLLOWER OUTPUT CIRCUITS FOR AUDIO OR VIDEO AMPLIFIERS

TRIODE:

$$\mu_{CF} = \frac{\mu}{\mu+1} \qquad A = \frac{\mu R_L}{R_L(\mu+1)+R_P}$$

$$R_{O(CATHODE)} = \frac{R_P}{\mu+1} \qquad R_L = \frac{(R_{K1}+R_{K2})R_L'}{R_{K1}+R_{K2}+R_L'}$$

PENTODE:

$$R_{O(CATHODE)} = \frac{1}{G_M} \qquad R_{EQ} = \frac{R_L}{1+R_L G_M}$$

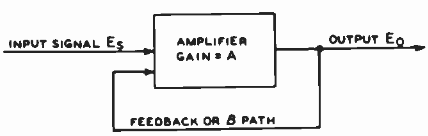
$$A = G_M R_{EQ}$$

Figure 27

Equivalent factors for pentode (or tetrode) cathode-follower power amplifiers

The circuit of figure 26A is the type of amplifier, either single-ended or push-pull, which may be used as a driver for a class-B modulator or which may be used for other applications such as feeding a speaker where unusually good damping of the speaker is desired. If the d-c resistance of the primary of the transformer (T_2) is approximately the correct value for the cathode bias resistor for the amplifier tube, the components R_k and C_k need not be used. Figure 26B shows an arrangement which may be used to feed directly a value of load impedance which is equal to or higher than the cathode impedance of the amplifier tube. The value of C_c must be quite high, somewhat higher than would be used in a conventional circuit, if the frequency response of the circuit when operating into a low-impedance load is to be preserved.

Figures 26C and 26D show cathode-follower circuits for use with tetrode or pentode tubes. Figure 26C is a circuit similar to that shown in 26A and essentially the same comments apply in regard to components R_k and C_k and the primary resistance of transformer T_2. Notice also that the screen of the tube is maintained at the same signal potential as the cathode by means of coupling capacitor C_d. This capacitance should be large enough so that at the lowest frequency it is desired to pass through the stage, its reactance will be low with respect to the dynamic screen-to-cathode resistance in parallel with R_d. T_2 in this stage as well as in the circuit of figure 26A should have the proper turns (or impedance) ratio to give the desired step-down or step-up from the cathode circuit to the load. Figure 26D is an arrangement frequently used in video systems for feeding a coaxial cable of relatively low impedance from a vacuum-tube amplifier. A pentode or tetrode tube with a cathode impedance as a cathode follower ($1/G_m$) of approximately the same impedance as the cable should be chosen. The 12BY7A and 6CL6 have cathode impedances of the same order as the surge impedances of certain types of low-capacitance coaxial cable. An arrangement such as 26D is also usable for feeding coaxial cable with audio or r-f energy where it is desired to transmit the output signal over moderate distances. The resistor R_k is added to the circuit as shown if the cathode impedance of the tube used is lower than the characteristic impedance of the cable. If the output impedance of the stage is higher than the cable impedance, a resistance of appropriate value is sometimes placed in parallel with the input end of the cable. The values of C_d and R_d should be chosen with the same considerations in mind as mentioned in the discussion of the circuit of figure 26C.



VOLTAGE AMPLIFICATION WITH FEEDBACK = $\frac{A}{1-A\beta}$

A = GAIN IN ABSENCE OF FEEDBACK

β = FRACTION OF OUTPUT VOLTAGE FED BACK

β IS NEGATIVE FOR NEGATIVE FEEDBACK

FEEDBACK IN DECIBELS = $20 \log (1-A\beta)$

$= 20 \log \frac{\text{MID-FREQ. GAIN WITHOUT FEEDBACK}}{\text{MID-FREQ. GAIN WITH FEEDBACK}}$

DISTORTION WITH FEEDBACK = $\frac{\text{DISTORTION WITHOUT FEEDBACK}}{(1-A\beta)}$

$$R_O = \frac{R_N}{1 - A\beta\left(1 + \frac{R_N}{R_L}\right)}$$

WHERE:

R_O = OUTPUT IMPEDANCE OF AMPLIFIER WITH FEEDBACK

R_N = OUTPUT IMPEDANCE OF AMPLIFIER WITHOUT FEEDBACK

R_L = LOAD IMPEDANCE INTO WHICH AMPLIFIER OPERATES

Figure 28

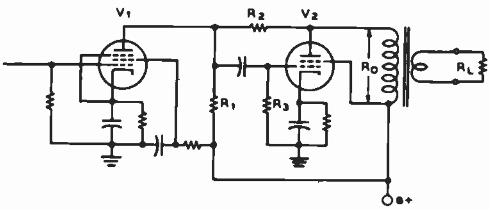
FEEDBACK AMPLIFIER RELATIONSHIPS

The Cathode Follower in R-F Stages The cathode follower may conveniently be used as a method of coupling r-f or i-f energy between two units separated a considerable distance. In such an application a coaxial cable should be used to carry the r-f or i-f energy. One such application would be for carrying the output of a vfo to a transmitter located a considerable distance from the operating position. Another application would be where it is desired to feed a single-sideband demodulator, an f-m adaptor, or another accessory with an intermediate-frequency signal from a communications receiver. A tube such as a 6CB6 connected in a manner such as is shown in figure 26D would be adequate for the i-f amplifier coupler, while a 6AQ5 or a 6CL6 could be used in the output stage of a vfo as a cathode follower to feed the coaxial line which carries the vfo signal from the control unit to the transmitter proper.

6-15 Feedback Amplifiers

It is possible to modify the characteristics of an amplifier by feeding back a portion of the output to the input. All components, circuits, and tubes included between the point where the feedback is taken off and the point where the feedback energy is inserted are said to be included within the feedback loop. An amplifier containing a feedback loop is said to be a *feedback amplifier*. One stage or any number of stages may be included within the feedback loop. However, the difficulty of obtaining proper operation of a feedback amplifier increases with the bandwidth of the amplifier, and with the number of stages and circuit elements included within the feedback loop.

Gain and Phase Shift in Feedback Amplifiers The gain and phase shift of any amplifier are functions of frequency. For any amplifier containing a feedback loop to be completely stable, the gain of such an amplifier, as measured from the input back to the point where the feedback circuit connects to the input, must be less than unity at the frequency where the feedback voltage is in phase with the input voltage of the amplifier. If the gain is equal to or more than unity at the frequency where the feedback voltage is in phase with the input, the amplifier will oscillate. This fact imposes a limitation on the amount of feedback which may be employed in an amplifier which is to remain stable. If the reader is desirous of designing amplifiers in which a large amount of feedback is to be employed he is referred to a book on the subject by H. W. Bode.*



$$\text{DB FEEDBACK} = 20 \log \left[\frac{R_2 + R_A (G_{M V_2} R_O)}{R_2} \right]$$

$$= 20 \log \left[\frac{R_2 + R_A [\text{VOLTAGE GAIN OF } V_2]}{R_2} \right]$$

$$\text{GAIN OF BOTH STAGES} = \left[G_{M V_1} \left(\frac{R_B \times R_A}{R_B + R_A} \right) \right] \times (G_{M V_2} R_O)$$

WHERE:

$$R_A = \frac{R_1 \times R_3}{R_1 + R_3}$$

$$R_B = \frac{R_2}{G_{M V_2} R_O}$$

R_O = REFLECTED LOAD IMPEDANCE ON V_2

R_2 = FEEDBACK RESISTOR (USUALLY ABOUT 500 K)

$$\text{OUTPUT IMPEDANCE} = \frac{R_N R_2}{\left[R_2 + R_A (G_{M V_2} R_O) \right] \times \left(1 + \frac{R_N}{R_O}\right)}$$

R_N = PLATE IMPEDANCE OF V_2

Figure 29

SHUNT FEEDBACK CIRCUIT FOR PENTODES OR TETRODES

This circuit requires only the addition of one resistor (R_2) to the normal circuit for such an application. The plate impedance and distortion introduced by the output stage are materially reduced.

Types of Feedback Feedback may be either negative or positive, and the feedback voltage may be proportional either to output voltage or output current. The most commonly used type of feedback with a-f or video amplifiers is *negative feedback* proportional to output voltage. Figure 28 gives the general operating conditions for feedback amplifiers. Note that the reduction in distortion is proportional to the reduction in gain of the amplifier, and also that the reduction in the output impedance of the amplifier is somewhat greater than the reduction in the gain by an amount which is a function of the ratio of the output impedance of the amplifier without feedback to the load impedance. The reduction in noise and hum in those stages included within the feedback loop is proportional to the reduction in gain. However, due to the reduction in gain of the output section of the amplifier somewhat increased gain is required of the stages preceding the stages included within the feedback loop. Therefore the noise and hum output of the entire amplifier may or may not be reduced dependent on the relative contributions of the first part and the latter part of the amplifier to hum and noise. If most of the noise and hum is coming from the stages included within the feedback loop the undesired signals will be reduced in the output from the complete amplifier. It is most frequently true in conventional amplifiers that the hum and distortion come from the latter stages, hence these will be reduced by feedback, but thermal agitation and microphonic noise come from the first stage and will not be reduced but may be increased by feedback unless the feedback loop includes the first stage of the amplifier.

Figure 29 illustrates a very simple and effective application of negative-voltage feedback to an output pentode or tetrode amplifier stage. The reduction in hum and distortion may amount to 15 to 20 db. The reduction in the effective plate impedance of the stage will be by a factor of 20 to 100 depending on the operating conditions. The circuit is commonly used in commercial equipment with tubes such as the 6AU6 for V_1 and the 6AQ5 for V_2.

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*H. W. Bode, *Network Analysis and Feedback Amplifier Design*. D. Van Nostrand Company, Inc. Princeton, New Jersey.

CHAPTER SEVEN

Radio-Frequency Power Amplifiers

All modern radio transmitters consist of a comparatively low-level source of radio-frequency energy which is amplified in strength and mixed or multiplied in frequency to achieve the desired power level and operating frequency. Microwave transmitters may be of the self-excited oscillator type, but when it is possible to use r-f amplifiers in uhf transmitters the flexibility of their application is increased.

Radio-frequency power amplifiers are generally classified according to frequency range (hf, vhf, uhf, etc.), power level, type of tube used, and type of service (a-m, f-m, c-w, SSB). In addition, the amplifier may be classified according to mode, or dynamic operating characteristic of the tube (Class AB_1, B, or C); and according to circuitry (grid driven or cathode driven). Each mode of operation and circuit configuration has its distinct advantages and disadvantages, and no one mode or circuit is superior in all respects to any other. As a result, modern transmitting equipments employ various modes of operation, intermixed with various tubes and circuit configurations. The following portion of this chapter will be devoted to the calculation of dynamic characteristics for some of the more practical modes of tuned power amplifier operation.

7-1 Class-C R-F Power Amplifiers

It is often desired to operate the r-f power amplifier in the class-B or class-C mode since such stages can be made to give high plate-circuit efficiency. Hence, the tube cost and cost of power to supply the stage is least for any given power output. Nevertheless, the class-C amplifier provides less *power gain* than either a class-A or class-B amplifier under similar conditions. The grid of the class-C amplifier must be driven highly positive over the small portion of the exciting signal when the instantaneous plate voltage on the tube is at its lower point, and is at a large negative potential over a major portion of the operating cycle. As a result, no plate current will flow except during the time plate voltage is very low. Comparatively large amounts of drive power are necessary to achieve this mode of operation. Class-C operational efficiency is high because no plate current flows except when the plate-to-cathode voltage drop across the tube is at its lowest value, but the price paid for stage efficiency is the large value of drive power required to achieve this mode of operation.

The gain of a class-B amplifier is higher than that of the class-C stage, and driving power is less in comparison. In addition, the class-B amplifier may be considered to be linear; that is, the output voltage is a replica of the input voltage at all signal levels up to overload. This is not true in the case of the class-C amplifier whose output waveform consist of short pulses of current, as discussed later in this chapter.

The gain of a class-A amplifier is higher than that of the class-B or class-C stage, but the efficiency is the lowest of the three modes of operation. As with the class-B stage, the class-A amplifier is considered to be linear with respect to input and output waveforms.

Relationships in Class-C Stage The class-C amplifier is analyzed as its operation provides an all-inclusive case of the study of class-B and class-AB_1 r-f amplifiers.

The class-C amplifier is characterized by the fact that the plate current flows in pulses which, by definition, are less than one-half of the *operating cycle.* The operating cycle is that portion of the electrical cycle in which the grid is driven in a positive direction with respect to the cathode. The operating cycle is considered in terms of the plate or grid *conduction angle* (θ). The conduction angle is an expresion of that fraction of time (expressed in degrees of the electrical cycle) that the tube conducts plate or grid current as compared to the operating cycle of the input voltage waveform.

The theoretical efficiency of any power amplifier depends on the magnitude of the conduction angle; a tuned class-A amplifier having a large conduction angle with a maximum theoretical efficiency of 50 percent; a class-B amplifier with an angle of 180 degrees, and efficiency of 78.5 percent; and a class-C amplifier with an angle of about 160 degrees and efficiency of about 85 percent.

Figure 1 illustrates a transfer curve representing the relationships between grid and plate voltages and currents during the operating cycle of a class-C amplifier. Symbols shown in figure 1 and given in the following discussion are defined and listed in the *Glossary of Terms* included at the front of this Handbook.

The plot is of the *transfer curve* of a typical triode tube, and represents the change in plate current, (i_b) for a given amount of grid voltage (e_c). The representation is of the form of the I_b versus E_c plot for a triode shown in figure 9, chapter 4.

The *operating point,* or grid-bias level (E_c), is chosen at several times cutoff bias (E_{co}), and superimposed on the operating point is one-half cycle of the grid exciting voltage, $e_{g\ max}$. A sample point of grid voltage, e_{cx}, is shown to produce a value of instantaneous plate current, i_{bx}. All other points on the grid-voltage curve relate to corresponding points on the plate-current curve.

As the grid is driven considerably positive, grid current flows, causing the plate current to be "starved" at the peak of each cycle, thus the plate-current waveform pulse is slightly indented at the top. As the waveform is poor and the distortion high, class-C operation is restricted to r-f amplification where high efficiency is desirable and when the identity of the output waveform to the input waveform is relatively unimportant.

The relation between grid and plate voltages and currents is more fully detailed in the graphs of figures 2 and 3, which illustrate in detail the various voltage and cur-

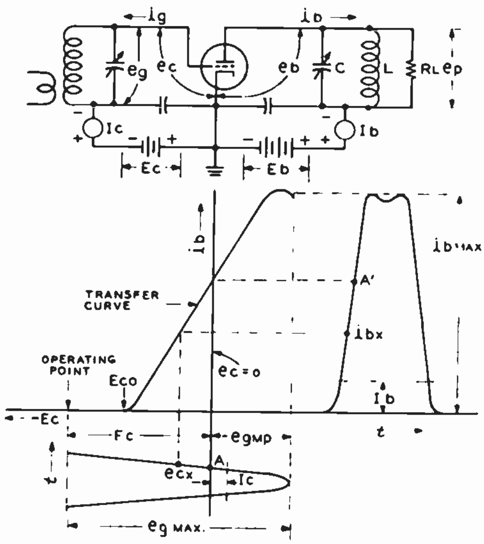


Figure 1

TRANSFER CURVE FOR OPERATING CYCLE OF CLASS-C AMPLIFIER

Typical class-C amplifier (less neutralizing circuits) is shown with various average and instantaneous voltages noted. A summary of symbols is given in the glossary of terms. The plot is of the transfer curve, representing the change in plate current for a given grid voltage. The grid signal ($e_{g\ max}$) is represented by a pulse of voltage along the y-axis, with the operating point determined by the amount of grid bias, E_c. As the waveform rises in amplitude, a corresponding pulse of plate current is developed across the plate load impedance, (R_L). A single point of grid voltage (A) represents a corresponding value of instantaneous plate current (A'). All other points on the grid-voltage curve relate to corresponding points on the plate-current curve.

rent variations during one electrical cycle of the exciting signal.

Voltage at the Grid The curves of figure 2 represent the grid voltage and current variations with respect to time. The x-axis for grid voltage is E_{c1} with a secondary axis (E_{co} = 0) above it, the vertical distance between axes representing the fixed grid-bias voltage (E_c). At the beginning of the operating cycle ($t = o$) the exciting voltage (e_g) is zero and increases in amplitude, until at *point A* it equals in magnitude the value of the bias voltage. At this point, the instantaneous voltage on the grid of the tube is zero with respect to the cathode, and plate current has already begun to flow (*point A* in figure 1), as the exciting signal is already greater in magnitude than the cut-off grid voltage (E_{co}). The relations are normally such that at the crest of the positive grid voltage cycle, e_{cmp} (or $e_{g\ max}$ positive), the grid is driven appreciably positive with respect to the cathode and consequently draws some grid current, i_g. The d-c component of grid current, I_c, may be read on the grid meter shown in figure 1.

The grid draws current only over that portion of the operating cycle when it is positive with respect to the cathode (that portion of the curve above the $E_c = 0$ axis in graph A). This portion of the exciting voltage is termed the *maximum positive grid voltage* (e_{cmp}).

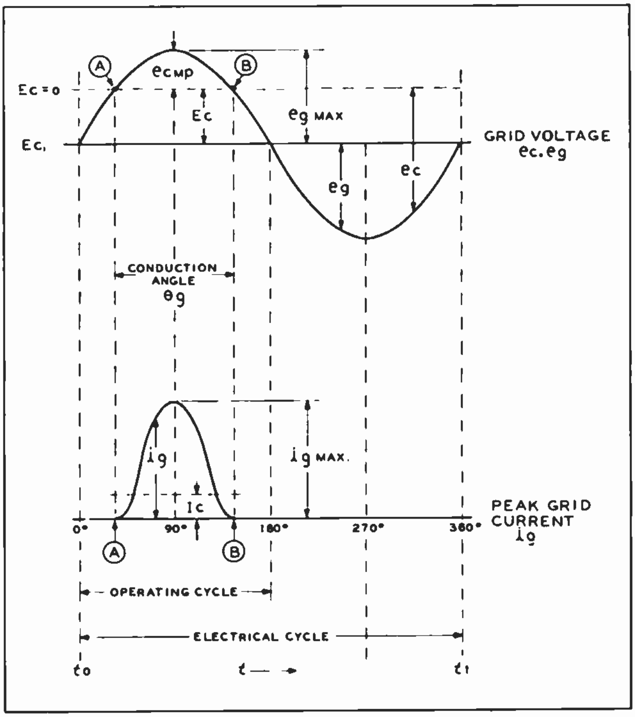
Voltage at the Plate The voltage at the plate of the tube responds to the changes in grid voltage as shown in figure 3. Instantaneous plate voltage (e_b), consists of the d-c plate voltage (E_b) less the a-c voltage drop across the plate load impedance (e_p). As the grid element becomes more positive, a greater flow of electrons reach the plate, instantaneous plate current increases, and the voltage drop across the plate load impedance (R_L) rises. The phase relations are such that the minimum instantaneous plate potential ($e_{b\ min}$) and the maximum instantaneous grid potential ($e_{g\ max}$) occur simultaneously. The corresponding instantaneous plate current (i_b) for this sequence is shown in the current plot of figure 3.

As plate current is conducted only be-

Figure 2

INSTANTANEOUS GRID VOLTAGE AND CURRENT OF A CLASS-C R-F POWER AMPLIFIER

Grid voltage and current variations with respect to time are shown. The grid is negatively biased by the amount E_c. As soon as the positive value of grid exciting voltage (e_g) exceeds E_c (point A) the grid starts to draw current, as it is positive with respect to the filament. Grid current flows from point A to point B of the grid voltage plot. This portion of the grid cycle is termed the conduction angle. Average value of grid current (I_c) may be read on a d-c meter in series with grid return line to bias supply. For typical class-C performance, grid current flows over a portion of the operating cycle, which is less than half the electrical cycle.



tween *points* A and B of the grid-voltage excursion, it can be seen that the plate-current pulse exists only over a portion (θ_b) of the complete plate *operating cycle*. (The operating cycle is taken to be that half-cycle of grid voltage having a positive excursion of the drive voltage.) The opposite

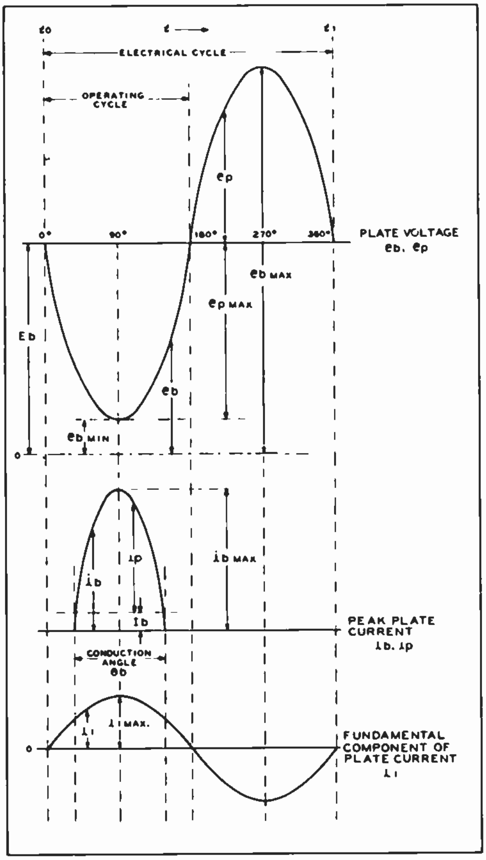


Figure 3

INSTANTANEOUS PLATE VOLTAGE AND CURRENT OF CLASS-C POWER AMPLIFIER

Instantaneous plate voltage and current responds to the changes in grid voltage shown in figure 2. As grid becomes more positive, the peak plate current rises, causing an increased voltage drop across the plate load impedance (R_L, figure 1). Maximum peak plate current flows at condition of minimum instantaneous plate voltage ($e_{b\,min}$) and maximum voltage drop across load impedance ($e_{p\,max}$). Plate-current pulse exists only over a portion of the operating cycle (conduction angle). Usable power is derived from the fundamental component of the plate current which is a sine wave developed across the resonant tank circuit. $e_{p\,max}$ equals $e_{p\,min}$.

half of the electrical cycle is of little interest, as the grid merely assumes a more negative condition and no flow of plate current is possible.

Peak plate current pulses, then, flow as pictured in figure 3 over the *conduction angle* of each operating cycle. *The fundamental component of plate current* (i_1) however, is a sine wave since it is developed across a resonant circuit (*LC*). The resonant circuit, in effect, acts as a "flywheel," holding r-f energy over the pulsed portion of the operating cycle, and releasing it during the quiescent portion of the electrical cycle.

The patterns of grid voltage and current shown in figure 2 are important in determining grid-circuit parameters, and the patterns of plate voltage and current shown in the illustrations can be used to determine plate-circuit parameters, as will be discussed later.

The various manufacturers of vacuum tubes publish data sheets listing in adequate detail various operating conditions for the tubes they manufacture. In addition, additional operating data for special conditions is often available for the asking. It is, nevertheless, often desirable to determine optimum operating conditions for a tube under a particular set of circumstances. To assist in such calculations the following paragraphs are devoted to a method of calculating various operating conditions which is moderately simple and yet sufficiently accurate for all practical purposes. It is based on wave-analysis techniques of the peak plate current of the operating cycle, adapted from Fourier analysis of a fundamental wave and its accompanying harmonics. Considerable ingenuity has been displayed in devising various graphical ways of evaluating the waveforms in r-f power amplifiers. One of these techniques, a *Tube Performance Calculator*, for class-AB_1 class-B, and Class-C service may be obtained at no cost by writing: Application Engineering Dept., Eimac Division of Varian, San Carlos, Calif. 94070.

7-2 Constant-Current Curves

Although class-C operating conditions can be determined with the aid of conventional grid-voltage versus plate-current operating curves (figure 9, chapter 4), the calculation is simplified if the alternative *constant current* graph of the tube in question is used figure 4). This representation is a graph of constant plate current on a grid-voltage versus plate-voltage plot, as previously shown in figure 10, chapter 4. The constant-current plot is helpful as the *operating line* of a tuned power amplifier is a straight line on a set of such curves and lends itself readily to graphic computations. Any point on the operating line, moreover, defines the instantaneous values of plate, screen and grid current which must flow when these particular values of plate, screen and grid voltages are applied to the tube. Thus, by taking off the values of the currents and plotting them against time, it is possible to generate a curve of instantaneous electrode currents, such as shown in figures 1 and 2. An analysis of the curve of instantaneous current values will derive the d-c components of the currents, which may be read on a d-c ammeter. In addition, if the plate current flows through a properly loaded resonant r-f circuit, the amount of power delivered to the circuit may be predicted, as well as drive power, and harmonic components of drive and output voltage.

A set of typical constant-current curves for the 304-TH medium-μ triode is shown in figure 5, with a corresponding set of

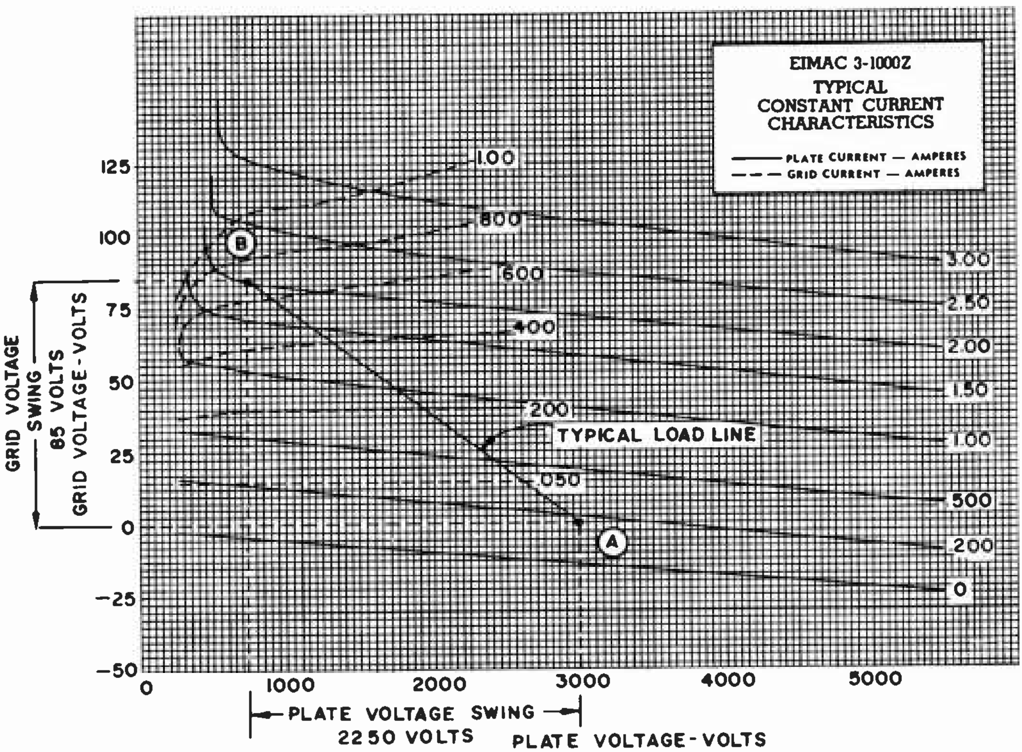


Figure 4

CONSTANT-CURRENT CHART FOR 3-1000Z HIGH-μ TRIODE

The constant-current chart is a plot of constant plate-current lines for various values of grid voltage and plate current. At the start of operation (quiescent point A) the tube rests at a plate voltage of 3000 and zero grid voltage. At a positive grid potential of 85 volts (point B), the plate current has increased to 2 amperes, and the plate voltage has dropped to 750, by virtue of the voltage drop across the plate load impedance. As the grid voltage rises from zero to maximum, the operating point passes from A to B along the load line. By examining representative samples of plate voltage and current along the load line, typical operating characteristics may be derived for the given set of conditions shown on the graph.

curves for the 304-TL low-μ triode shown in figure 6. The graphs illustrate how much more plate current can be obtained from the low-μ tube without driving the grid into the positive-grid region, as contrasted to the higher-μ tube. In addition, more bias voltage is required to cut off the plate current of the low-μ tube, as compared to the higher-μ tube for a given value of plate voltage. With the higher value of bias, a corresponding increase in grid-voltage swing is required to drive the tube up to the zero grid-voltage point on the curve. Low-μ tubes thus, by definition, have lower voltage gain, and this can be seen by comparing the curves of figures 5 and 6.

Low-μ (3-15) power triodes are chosen for class-A amplifiers and series-pass tubes in voltage regulators, as they operate well over a wide range of load current with low plate voltage drop. Medium-μ (15-50) triodes are generally used in r-f amplifiers and oscillators, as well as class-B audio modulators. High-μ (50-200) triodes have high power gain and are often used in cathode-driven ("grounded-grid") r-f amplifiers. If the amplification factor (μ) is sufficiently high, no external bias supply is required, and no protective circuits for loss of bias or drive are necessary. A set of constant-current curves for the 3-500Z high-μ triode is given in figure 7.

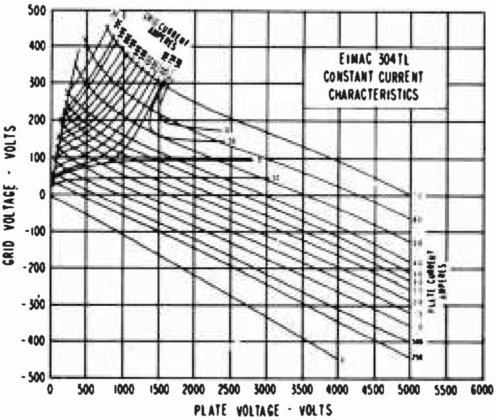


Figure 6

CONSTANT CURRENT CHART FOR LOW-μ TRIODE

Constant-current plot for a 304TL triode with a μ of 12. Note that more plate current at a given plate voltage can be obtained from the low-μ triode without driving the grid into the positive voltage region. In addition, more bias voltage is required to cut off the plate current at a given plate voltage. With this increased value of bias there is a corresponding increase in grid-voltage swing required to drive up to the zero grid-voltage point on the graph.

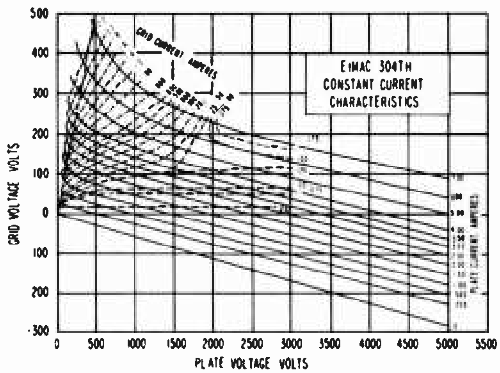


Figure 5

CONSTANT-CURRENT CHART FOR MEDIUM-μ TRIODE

Constant current plot for a 304TH triode with a μ of 20. Note that the lines of constant plate current have a greater slope than the corresponding lines of the high-μ triode (3-1000Z) and that for a given value of positive grid potential, and plate potential, the plate current of this tube is higher than that of the higher-μ tube.

The amplification factor of a triode is a function of the physical size and location of the grid structure. The upper limit of amplification factor is controlled by grid dissipation, as high-μ grid structures require many grid wires of small diameter having relatively poorer heat-conduction qualities as compared to a low-μ structure, made up of fewer wires of greater diameter and better heat conductivity. A set of constant-current curves for the 250TH power triode with a sample load line drawn thereon is shown in figure 8.

7-3 Class-C Amplifier Calculations

In calculating and predicting the operation of a vacuum tube as a class-C radio-frequency amplifier, the considerations which determine the operating conditions are plate efficiency, power output required, maximum

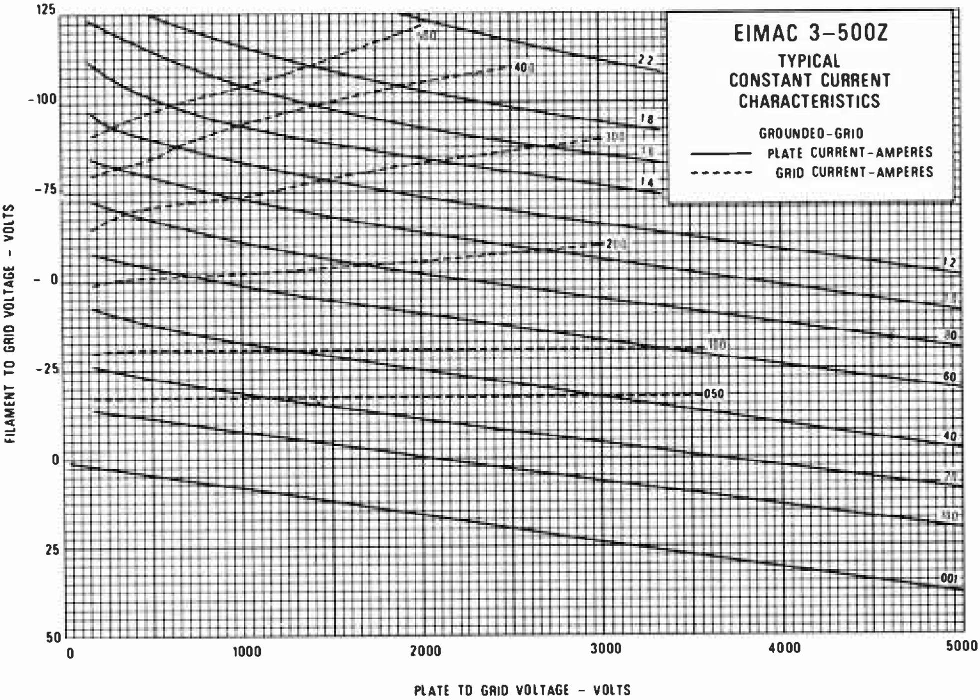


Figure 7

CONSTANT-CURRENT CHART FOR HIGH-μ TRIODE

Constant current plot for a 3-500Z triode with μ of 160. The 3-500Z is considered to be "zero bias" up to a plate potential of about 3000. Resting plate current at this value of plate voltage is approximately 160 milliamperes. This plot is for grounded-grid, cathode-driven use, and grid-voltage axis is defined in terms of filament to grid voltage (negative) instead of grid-to-filament voltage (positive). Grid and screen currents are usually logged on constant-current plots, along with plate current.

allowable plate and grid dissipation, maximum allowable plate voltage, and maximum allowable plate current. The values chosen for these factors will depend on the demands of a particular application of the tube.

The plate and grid currents of a class-C amplifier tube are periodic pulses, the durations of which are always less than 180 degrees. For this reason the average grid current, average plate current, power output, driving power, etc., cannot be directly calculated but must be determined by a Fourier analysis from points selected at proper intervals along the line of operation as plotted on the constant-current characteristics. This may be done either analytically or graphically. While the Fourier analysis has the advantage of accuracy, it also has the disadvantage of being tedious and involved.

The approximate analysis which follows has proved to be sufficiently accurate for most applications. This type of analysis also has the advantage of giving the desired information at the first trial. The system is direct in giving the desired information since the important factors, power output, plate efficiency, and plate voltage are arbitrarily selected at the beginning.

Method of Calculation The first step in the method to be described is to determine the power which must be delivered by the class-C amplifier. In making this determination it is well to remember that ordinarily from 5 to 10 percent of the power delivered by the amplifier tube or tubes will be lost in well-designed tank and coupling circuits at frequencies below 20 MHz. Above 20 MHz the tank and circuit losses are ordinarily somewhat above 10 percent.

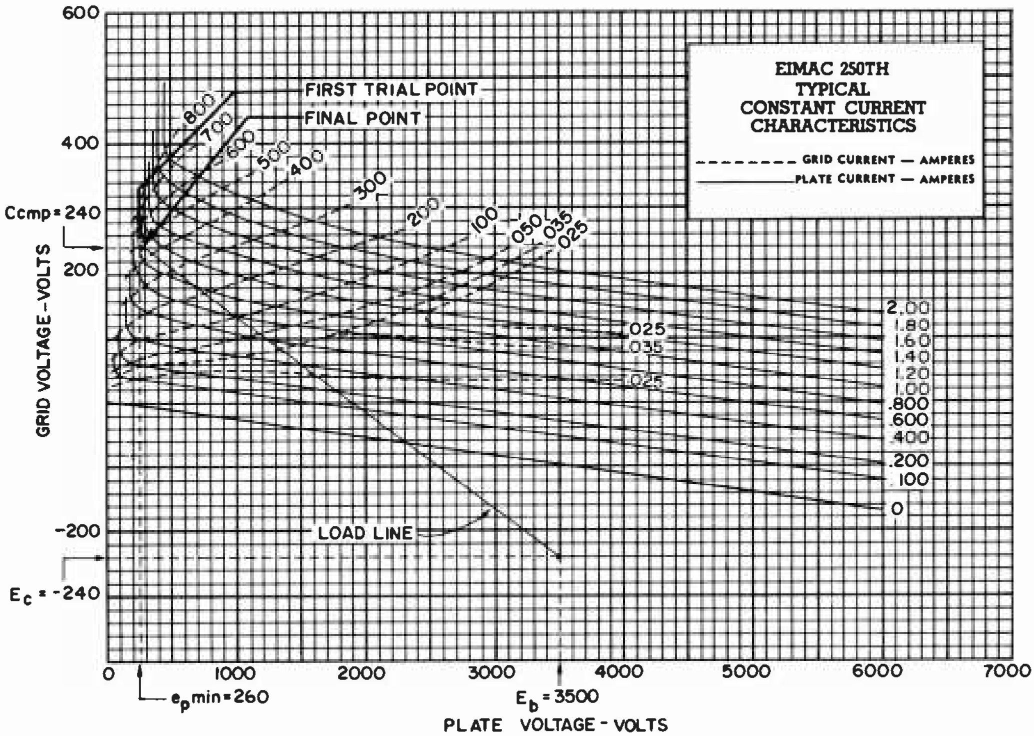


Figure 8

CONSTANT-CURRENT CHART FOR 250TH

Active portion of load line for an Eimac 250TH class-C r-f power amplifier, showing first trial point and final operating point for calculation of operating parameters at a power input of 1000 watts.

The plate power input necessary to produce the desired output is determined by the plate efficiency: $P_i = P_o/N_p$, assuming 100-percent tank circuit efficiency.

For most applications it is desirable to operate at the highest practicable efficiency. High-efficiency operation usually requires less-expensive tubes and power supplies, and the amount of external cooling required is frequently less than for low-efficiency operation. On the other hand, high-efficiency operation usually requires more driving power and involves the use of higher plate voltages and higher peak tube voltages. The better types of triodes will ordinarily operate at a plate efficiency of 75 to 85 percent at the highest rated plate voltage, and at a plate efficiency of 65 to 75 percent at intermediate values of plate voltage.

The first determining factor in selecting a tube or tubes for a particular application is the amount of plate dissipation which will be required of the stage. The total plate dissipation rating for the tube or tubes to be used in the stage must be equal to or greater than that calculated from: $P_p = P_i - P_o$.

After selecting a tube or tubes to meet the power output and plate dissipation requirements it becomes necessary to determine from the tube characteristics whether the tube selected is capable of the desired operation and, if so, to determine the driving power, grid bias, and grid dissipation.

The complete procedure necessary to determine a set of class-C amplifier operating conditions is given in the following steps:

1. Select the plate voltage, power output and efficiency.

2. Determine plate input from:
   $P_i = P_o/N_p$
3. Determine plate dissipation from:
   $P_p = (P_i - P_o)/1.1$
   (P_p must not exceed maximum rated plate dissipation for selected tube or tubes. Tank circuit efficiency assumed to be 90%).
4. Determine average plate current (I_b) from: $I_b = P_i/E_b$.
5. Determine approximate peak plate current ($i_{b\,max}$) from:

$$i_{b\,max} = 4.9\ I_b \text{ for } N_p = 0.85$$
$$i_{b\,max} = 4.5\ I_b \text{ for } N_p = 0.80$$
$$i_{b\,max} = 4.0\ I_b \text{ for } N_p = 0.75$$
$$i_{b\,max} = 3.5\ I_b \text{ for } N_p = 0.70$$
$$i_{b\,max} = 3.1\ I_b \text{ for } N_p = 0.65$$

   Note: A figure of $N_p = 0.75$ is often used for class-C service, and a figure of $N_p = 0.65$ is often used for class-B and class-AB service.
6. Locate the point on the constant-current chart where the constant-current plate line corresponding to the appropriate value of $i_{b\,max}$ determined in step 5 crosses the point of intersection of equal values of plate and grid voltage. (The locus of such points for all these combinations of grid and plate voltage is termed the *diode line*). Estimate the value of $e_{p\,min}$ at this point.
   In some cases, the lines of constant plate current will inflect sharply upward before reaching the diode line. If so, $e_{p\,min}$ should not be read at the diode line but at a point to the right where the plate-current line intersects a line drawn from the origin through these points of inflection.
7. Calculate $e_{b\,min}$ from:

$$e_{b\,min} = E_b - e_{p\,min}.$$

8. Calculate the ratio: $i_{1\,max} / I_b$ from:

$$\frac{i_{1\,max}}{I_b} = \frac{2\ N_p \times E_b}{e_{p\,min}}$$

   (where $i_{1\,max}$ = peak fundamental component of plate current).
9. From the ratio of $i_{1\,max} / I_b$ calculated in step 8 determine the ratio: $i_{b\,max}/I_b$ from the graph of figure 9.
10. Derive a new value for $i_{b\,max}$ from the ratio found in step 9:
    $i_{b\,max}$ = (ratio found in step 9) $\times\ I_b$
11. Read the values of maximum positive grid voltage, $e_{g\,max}$ and peak grid current ($i_{g\,max}$) from the chart for the values of $e_{p\,min}$ and $i_{b\,max}$ found in steps 6 and 10 respectively.
12. Calculate the cosine of one-half the angle of plate-current flow (one-half the operating cycle, $\theta_p/2$).

$$\cos\frac{\theta_p}{2} = 2.32\left(\frac{i_{1\,max}}{I_b} - 1.57\right)$$

13. Calculate the grid bias voltage (E_c) from:

$$E_c = \frac{1}{1 - \cos\frac{\theta_p}{2}} \times \left[\cos\frac{\theta_p}{2}\left(\frac{e_{b\,min}}{\mu} - e_{cmp}\right) - \frac{E_b}{\mu}\right]$$

    for triodes.

$$E_{c1} = \frac{1}{1 - \cos\frac{\theta_p}{2}} \times \left[- e_{cmp} \times \cos\frac{\theta_p}{2} - \frac{E_{c2}}{\mu_s}\right]$$

    for tetrodes, where μ_s is the grid-screen amplification factor.
14. Calculate the peak fundamental grid voltage, $e_{g\,max}$ from:
    $e_{g\,max} = e_{cmp} - (-E_c)$, using negative value of E_c.
15. Calculate the ratio $e_{g\,max}/E_c$ for the values of E_c and $e_{g\,max}$ found in steps 13 and 14.
16. Read the ratio $i_{g\,max}/I_c$ from figure 10 for the ratio $e_{g\,max} / E_c$ found in step 15.
17. Calculate the average grid current (I_c) from the ratio found in step 16 and the value of $i_{g\,max}$ found in step 11:

$$I_c = \frac{i_{g\,max}}{\text{(ratio found in step 16)}}$$

18. Calculate approximate grid driving power from:

$$P_d = 0.9\, e_{g\,max} \times I_c$$

19. Calculate grid dissipation from:

$$P_g = P_d - (-E_c \times I_c)$$

(P_g must not exceed the maximum rated grid dissipation for the tube or tubes selected).

Sample Calculation A typical example of class-C amplifier calculation is shown in the following example. Reference is made to figures 8, 9, and 10 in the calculation. The steps correspond to those in the previous section.

1. Desired power output—800 watts.
2. Desired plate voltage—3500 volts. Desired plate efficiency—80% ($N_p = 0.8$). $P_i = 800/0.8 = 1000$ watts.
3. $P_p = \dfrac{1000 - 800}{1.1} = 182$ watts.

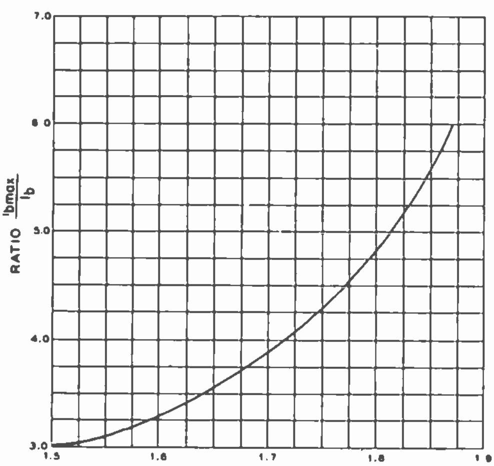


Figure 9

Relationship between the peak value of the fundamental component of the tube plate current, and average plate current; as compared to the ratio of the instantaneous peak value of tube plate current, and average plate current value.

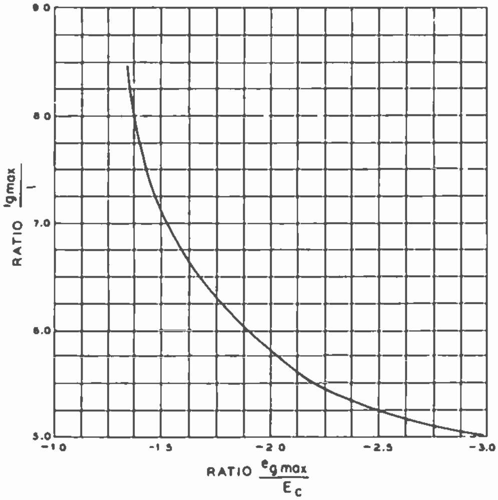


Figure 10

Relationship between the ratio of the peak value of the fundamental component of the grid excitation voltage, and the average grid bias; as compared to the ratio between instantaneous peak grid current and average grid current

(Use 250TH; max P_p = 250W; $\mu = 37$).

4. I_b = 1000/3500 = 0.285 ampere (285 ma). (Maximum rated I_b for 250TH = 350 ma).
5. Approximate $i_{b\,max}$: $0.285 \times 4.5 = 1.28$ amp
6. $e_{b\,min}$ = 260 volts (see figure 8, first trial point).
7. $e_{p\,min} = 3500 - 260 = 3240$ volts.
8. $i_{l\,max} / I_b = (2 \times 0.8 \times 3500) / 3240 = 1.73$.
9. $i_{b\,max} / I_b = 4.1$ (from figure 9).
10. $i_{b\,max} = 4.1 \times 0.285 = 1.17$.
11. e_{cmp} = 240 volts
    $i_{g\,max}$ = 0.43 amp
    (Both read from final point on figure 8).
12. $\cos\dfrac{\theta_b}{2} = 2.32\,(1.73 - 1.57) = 0.37$

    $\left(\dfrac{\theta_b}{2} = 68.3^\circ \text{ and } \theta_b = 136.6^\circ\right)$
13. $$E_c = \frac{1}{1 - 0.37} \times \left[0.37\left(\frac{3240}{37} - 240\right) - \frac{3500}{37}\right]$$

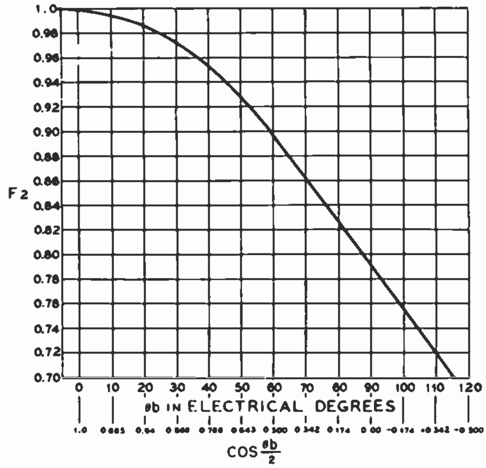


Figure 11

Relationship between factor F_2 and the half-angle of plate-current flow in an amplifier with sine-wave input and output voltage, operating at a grid-bias voltage greater than cutoff

$= -\ 240$ volts.

14. $e_{g\ max} = 240 - (-240) = 480$ volts.
15. $e_{g\ max}/E_c = 480/-240 = -2.$
16. $i_{g\ max}/I_c = 5.75$ (from figure 10).
17. $I_c = 0.43/5.75 = 0.075$ amp (75
18. $P_d = 0.9 \times 480 \times 0.075 = 32.5$ watts.
19. $P_g = 32.5 + (-240 \times 0.075) = 14.5$ watts (Maximum rated P_g for 250TH = 40 watts).
20. The power output of any type of r-f amplifier is equal to:

$$P_o = \frac{i_{1\ max} \times e_{p\ min}}{2}$$

($i_{1\ max}$ can be determined by multiplying the ratio determined in step 8 by I_b. Thus $= 1.73 \times 0.285 = 0.495$).
$P_o = (0.495 \times 3240)/2 = 800$ watts

21. The plate load impedance of any type of r-f amplifier is equal to:

$$R_L = \frac{e_{p\ min}}{i_{1\ max}}$$

$$R_L = \frac{3240}{0.495} = 6550 \text{ ohms}$$

An alternative equation for the approximate value of R_L is:

$$R_L \cong \frac{E_b}{1.8 \times I_b}$$

$$R_L \cong \frac{3500}{1.8 \times 0.285} = 6820 \text{ ohms}$$

Q of Amplifier Tank Circuit In order to obtain proper plate tank-circuit tuning and low radiation of harmonics from an amplifier it is necessary that the plate tank circuit have the correct Q. Charts giving compromise values of Q for class-C amplifiers are given in the chapter, *Generation of R-F Energy*. However, the amount of inductance required for a special tank-circuit Q under specified operating conditions can be calculated from the following expression:

$$\omega L = \frac{R_L}{Q}$$

where,

ω equals $2\pi \times$ operating frequency,
L equals tank inductance,
R_L equals required tube load impedance,
Q equals effective tank circuit Q.

A tank circuit Q of 12 to 20 is recommended for all normal conditions. However, if a balanced push-pull amplifier is employed the tank receives two impulses per cycle and the circuit Q may be lowered somewhat from the above values.

Quick Method of Calculating Amplifier Plate Efficiency The plate-circuit efficiency of a class-B or class-C r-f amplifier is approximately equal to the product of two factors: F_1, which is equal to the ratio of $e_{p\ max}$ to E_b ($F_1 = e_{p\ max}/E_b$) and F_2, which is proportional to the one-half angle of plate current flow $\theta_b/2$. A graph of F_2 versus both $\theta_b/2$ and $\cos \theta_b/2$ is given in figure 11. Either $\theta_b/2$ or $\cos \theta_b/2$ may be used to determine F_2. $\cos \theta_b/2$ may be determined either from the procedure previously given for making class-C amplifier computations or it may be determined from the following expression:

$$\cos \frac{\theta_b}{2} = -\frac{\mu E_c + E_b}{\mu \times e_{g\ max} - e_{p\ max}}$$

Example of Method It is desired to know the one-half angle of plate-current flow and the plate-circuit efficiency for an 812 tube operating under the following class-C conditions which have been assumed from inspection of the data and curves given in the RCA Transmitting Tube Handbook:

1. $E_b = 1100$ volts
   $E_c = -40$ volts
   $\mu = 29$
   $e_{g\,max} = 120$ volts
   $e_{p\,max} = 1000$ volts
2. $F_1 = \frac{e_{p\,max}}{E_b} = 0.91$
3. $\cos\frac{\theta_b}{2} = -\left[\frac{(29 \times 40) + 1100}{(29 \times 120) - 1000}\right] = \frac{60}{2480} = 0.025$
4. $F_2 = 0.79$ (by reference to figure 11)
5. $N_p = F_1 \times F_2 = 0.91 \times 0.79 = 0.72$ (72 percent efficiency)

F_1 could be called the *plate-voltage-swing efficiency factor*, and F_2 can be called the *operating-angle efficiency factor* or the maximum possible efficiency of any stage running with that value of half-angle of plate current flow.

N_p is, of course, only the ratio between power output and power input. If it is desired to determine the power input, exciting power, and grid current of the stage, these can be obtained through the use of steps 7, 8, 9, and 10 of the previously given method for determining power input and output; and knowing that $i_{g\,max}$ *is* 0.095 ampere, the grid-circuit conditions can be determined through the use of steps 15, 16, 17, 18, and 19.

7-4 Class-B Radio-Frequency Power Amplifiers

Radio-frequency power amplifiers operating under class-B conditions of grid bias and excitation voltage are used in various types of applications in transmitters. The first general application is as a buffer-amplifier stage where it is desired to obtain a high value of power amplification in a particular stage without regard to linearity. A particular tube type operated with a given plate voltage will be capable of somewhat greater output for a certain amount of excitation power when operated as a class-B amplifier than when operated as a class-C amplifier.

Calculation of Operating Characteristics Calculation of the operating conditions for this type of class-B r-f amplifier can be carried out in a manner similar to that described in the previous paragraphs, except that the grid-bias voltage is set on the tube before calculation at the value: $E_c = -E_b/\mu$. Since the grid bias is set at cutoff the one-half angle of plate-current flow is 90°; hence cos $\theta_b/2$ is fixed at 0.00. The plate-circuit efficiency for a class-B r-f amplifier operated in this manner can be determined in the following manner:

$$N_p = 78.5 \times \frac{e_{p\,max}}{E_b}$$

Note: In reference to figure 3, $e_{p\,max}$ is equal in magnitude to $e_{p\,min}$ and absolute value should be used.

The "Class-B Linear" The second type of class-B r-f amplifier is the so-called *class-B linear amplifier* which is often used in transmitters for the amplification of a single-sideband signal or a conventional amplitude-modulated wave. Calculation of operating conditions may be carried out in a manner similar to that previously described with the following exceptions: The first trial operating point is chosen on the basis of the 100-percent positive modulation peak (or PEP condition) of the exciting wave. The plate-circuit and grid-peak voltages and currents can then be determined and the power input and output calculated. Then (in the case for an a-m linear) with the exciting voltage reduced to one-half for the no-modulating condition of the exciting wave, and with the same value of load resistance reflected on the tube, the a-m plate input and plate efficiency will drop to approximately one-half the values at the 100-percent positive modulation peak and the power output of the stage will drop to one-fourth the peak-modulation value. On the negative modulation peak the input, efficiency and output all drop to zero.

In general, the proper plate voltage, bias voltage, load resistance, and power output listed in the tube tables for class-B audio work will also apply to class-B linear r-f application.

Calculation of Operating Parameters for a Class-B Linear Amplifier The class-B linear amplifier parameters may be calculated from constant-current curves, as suggested, or may be derived from the E_b vs I_b curves, as outlined in this section.

Figure 12 illustrates the characteristic curves for an 813 tube. Assume the plate supply to be 2000 volts, and the screen supply to be 400 volts. To determine the operating parameters of this tube as a class-B linear SSB r-f amplifier, the following steps should be taken:

1. The grid bias is chosen so that the resting plate current will produce approximately 1/3 of the maximum plate dissipation of the tube. The maximum dissipation of the 813 is 125 watts, so the bias is set to allow one-third of this value, or 42 watts of resting dissipation. At a plate potential of 2000 volts, a plate current of 21 milliamperes will produce this figure. Referring to figure 12, a grid bias of −45 volts is approximately correct.
2. A practical class-B linear r-f amplifier runs at an efficiency of about 66% at full output (the carrier efficiency dropping to about 33% with a *modulated* exciting signal). In the case of single-sideband suppressed-carrier excitation, the linear amplifier runs at the resting or quiescent input of 42 watts with no exciting signal. The peak allowable power input to the 813 is:

$$\text{PEP input power } (p_i) = \frac{\text{plate dissipation} \times 100}{(100 - \%\text{ plate efficiency})} =$$

$$\frac{125 \times 100}{33} = 378 \text{ watts PEP}$$

3. The maximum d-c signal plate current is:

$$I_{b\,max} = \frac{p_i}{E_b} = \frac{378}{2000} = 0.189 \text{ ampere}$$

(Single-tone drive signal condition)

4. The plate-current conduction angle (θ_b) of the class-B linear amplifier is *approximately* 180°, and the peak

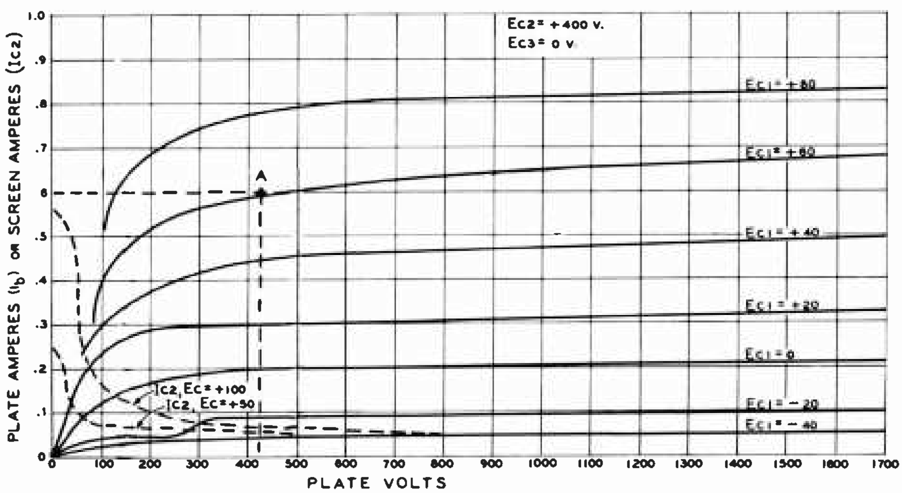


Figure 12

AVERAGE PLATE CHARACTERISTICS OF 813 TUBE

plate-current pulses have a maximum value of about 3.14 times $I_{b\,max}$:

$$i_{b\,max} = 3.14 \times 0.189 = 0.593 \text{ amp.}$$

5. Referring to figure 12, a current of about 0.6 ampere (*Point A*) will flow at a positive grid potential of 60 volts and a minimum plate potential of 420 volts. The grid is biased at −45 volts, so a peak r-f grid voltage of 60 + 45 volts, or 105 volts, swing is required.
6. The grid driving power required for the class-B linear stage may be found by the aid of figure 13. It is one-third the product of the peak grid current times the peak grid swing.

$$P_d = \frac{0.015 \times 105}{3} = 0.525 \text{ watt}$$

7. The single-tone (peak) power output of the 813 is:

$$P_o = .785\ (E_b - e_{b\,min}) \times I_{b\,max}$$

$$P_o = .785\ (2000 - 420) \times 0.189 = 235 \text{ watts PEP}$$

8. The plate load resistance is:

$$R_L \cong \frac{E_b}{1.8 \times I_b} = \frac{2000}{1.8 \times 0.188} = 5870 \text{ ohms}$$

9. If a loaded plate tank circuit Q of 12 is desired, the reactance of the plate tank capacitor of a parallel tuned circuit at resonance is:

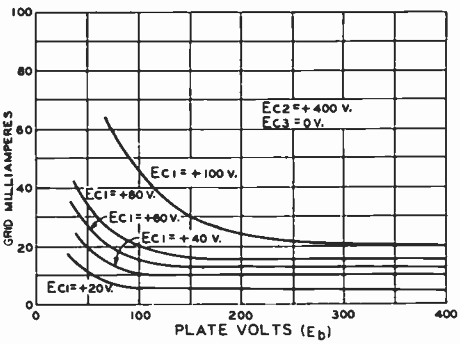


Figure 13

E_{g1} VERSUS E_p CHARACTERISTICS OF 813 TUBE

$$X_c = \frac{R_L}{Q} = \frac{5870}{12} = 490 \text{ ohms}$$

10. For an operating frequency of 4.0 MHz, the effective resonant capacitance is:

$$C = \frac{10^6}{6.28 \times 4.0 \times 490} = 81 \text{ pf}$$

11. The inductance required to resonate at 4.0 MHz with this value of capacitance is:

$$L = \frac{490}{6.28 \times 4.0} = 19.5 \text{ microhenrys}$$

Grid-Circuit Considerations

1. The maximum positive grid potential is 60 volts and the peak r-f grid voltage is 105 volts. Required peak driving power is 0.525 watt. The equivalent grid resistance of this stage is:

$$r_g = \frac{(e_{g\,max})^2}{2 \times P_d} = \frac{105^2}{2 \times 0.525} = 10{,}000 \text{ ohms}$$

2. As in the case of the class-B audio amplifier the grid resistance of the linear amplifier varies from infinity to a low value when maximum grid current is drawn. To decrease the effect of this resistance excursion, a swamping resistor should be placed across the grid-tank circuit. The value of the resistor should be dropped until a shortage of driving power begins to be noticed. For this example, a resistor of 3000 ohms is used. The grid circuit load for no grid current is now 3000 ohms instead of infinity, and drops to 2300 ohms when maximum grid current is drawn.
3. A circuit Q of 15 is chosen for the grid tank. The capacitive reactance required is:

$$X_C = \frac{2300}{15} = 154 \text{ ohms}$$

4. At 4.0 MHz the effective capacitance is:

$$C = \frac{10^6}{6.28 \times 4.0 \times 154} = 259 \text{ pf}$$

5. The inductive reactance required to resonate the grid circuit at 4.0 MHz is:

$$L = \frac{154}{6.28 \times 4.0} = 6.1 \text{ microhenrys}$$

6. By substituting the loaded-grid resistance figure in the formula in the first paragraph, the peak grid driving power is now found to be approximately 2.4 watts.

Screen-Circuit Considerations By reference to the plate characteristic curve of the 813 tube, it can be seen that at a minimum plate potential of 420 volts, and a maximum plate current of 0.6 ampere, the screen current will be approximately 30 milliamperes, dropping to one or two milliamperes in the quiescent state. It is necessary to use a well-regulated screen supply to hold the screen voltage at the correct potential over this range of current excursion. The use of an electronically regulated screen supply is recommended.

7-5 Grounded-Grid and Cathode-Follower R-F Power Amplifier Circuits

The r-f power amplifier discussions of Sections 7-3 and 7-4 have been based on the assumption that a conventional grounded-cathode or cathode-return type of amplifier was in question. It is possible, however, as in the case of a-f and low-level r-f amplifiers to use circuits in which electrodes other than the cathode are returned to ground insofar as the signal potential is concerned. Both the plate-return or cathode-follower amplifier and the *grid-return* or *grounded-grid* amplifier are effective in certain circuit applications as tuned r-f power amplifiers.

Disadvantages of Grounded-Cathode Amplifiers An undesirable aspect of the operation of cathode-return r-f power amplifiers using triode tubes is that such amplifiers must be neutralized. Principles and methods of neutralizing r-f power amplifiers are discussed in the chapter *Generation of R-F Energy*. As the frequency of operation of an amplifier is increased the stage becomes more and more difficult to neutralize due to inductance in the grid and cathode leads of the tube and in the leads to the neutralizing capacitor. In other words the bandwidth of neutralization decreases as the presence of the neutralizing capacitor adds additional undesirable capacitive loading to the grid and plate tank circuits of the tube or tubes. To look at the problem in another way, an amplifier that may be perfectly neutralized at a frequency of 30 MHz may be completely out of neutralization at a frequency of 120 MHz. Therefore, if there are circuits in both the grid and plate circuits which offer appreciable impedance at this high frequency it is quite possible that the stage may develop a parasitic oscillation in the vicinity of 120 MHz.

Grounded-Grid R-F Amplifiers This condition of restricted-range neutralization of r-f power amplifiers can be greatly alleviated through the use of a *cathode-driven* or *grounded-grid* r-f stage. The grounded-grid amplifier has the following advantages:

1. The output and input capacitances of a stage are reduced to approximately one-half the value which would be obtained if the same tube or tubes were operated as a conventional neutralized amplifier.
2. The tendency toward parasitic oscillations in such a stage is greatly reduced since the shielding effect of the control grid between the filament and the plate is effective over a broad range of frequencies.
3. The feedback capacitance within the stage is the plate-to-cathode capacitance which is ordinarily very much less than the grid-to-plate capacitance. Hence neutralization is ordinarily not required in the high frequency region. If neutralization is required the neutalizing capacitors are very small in value and are cross-connected between plates and cathodes in a push-pull stage, or between the opposite end of a split plate tank and the cathode in a single-ended stage.

The disadvantages of a grounded-grid amplifier are:

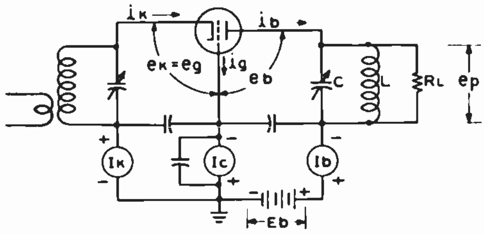
1. A large amount of excitation energy is required. However, only the normal amount of energy is lost in the grid circuit of the amplifier tube; most additional energy over this amount is deliver to the load circuit as useful output.
2. The cathode of a grounded-grid amplifier stage is above r-f ground. This means that the cathode must be fed through a suitable impedance from the filament supply, or the filament transformer must be of the low capacitance type and adequately insulated for the r-f voltage which will be present.
3. A grounded-grid r-f amplifier cannot be plate modulated 100 percent unless the output of the exciting stage is modulated also. Approximately 70-per cent modulation of the exciter stage, while the final stage is modulated 100 percent, is recommended. However the grounded-grid r-f amplifier is quite satisfactory as a class-B linear r-f amplifier for single-sideband or conventional amplitude-modulated waves or as an amplifier for a straight c-w or f-m signal.

Figure 14 shows a simplified representation of a grounded-grid zero-bias triode r-f power amplifier stage. The relationships between input and output power and the peak fundamental components of electrode voltages and currents are given below the drawing. The calculation of the complete operating conditions for a grounded-grid amplifier stage is somewhat more complex than that for a conventional amplifier because the input circuit of the tube is in series with the output circuit as far as the load is concerned. The primary result of this effect is, as stated before, that considerably more power is required from the driver stage. The normal power gain for a g-g stage is from 3 to 15 depending on the grid-circuit conditions chosen for the output stage. The higher the grid bias and grid swing required on the output stage, the higher will be the requirement from the driver.

Calculation of Operating Conditions of Grounded-Grid R-F Amplifers

It is most convenient to determine the operating conditions for a class-B or class-C grounded-grid r-f power amplifier in a two-step process. The first step is to determine the plate-circuit and grid-circuit operating conditions of the tube as though it were to operate as a conventional grid-driven amplifier. The second step is to then add in the additional conditions imposed on the original data by the fact that the stage is to operate as a grounded-grid amplifier. This step is the addition of the portion of the drive power contributed by the conversion of drive power to plate output power. This portion of the drive power is referred to as *converted drive power,* or *feedthrough power.* The latter term is misleading, as this portion of drive power does not appear in the plate load circuit of the cathode-driven stage until after it is converted to a *varying-d.c.* plate potential effectively in series with the main amplifier power supply. The converted drive power serves a useful function in linear amplifier service because it swamps out the undesirable effects of nonlinear grid loading and presents a reasonably constant load to the exciter.

Special constant-current curves are often



$$\text{PEP POWER TO LOAD} = \frac{(e_{b\,MIN} + e_{g\,MAX}) \times i_{1\,MAX}}{2}$$

$$\text{PEP POWER DELIVERED BY OUTPUT TUBE} = \frac{e_{b\,MIN} \times i_{1\,MAX}}{2}$$

$$\text{PEP DRIVE POWER} = \frac{e_{g\,MAX} \times i_{1\,MAX}}{2} + 0.9\,(e_{g\,MAX} \times I_c)$$

$$Z_K \cong \frac{e_{g\,MAX}}{i_{1\,MAX} + 1.5 \times I_c}$$

$$R_L \cong \frac{E_b}{1.8 \times I_b}$$

Figure 14

ZERO-BIAS GROUNDED-GRID AMPLIFIER

The equations in the above figure give the relationships between the output power, drive power, feedthrough power, and input and output impedances expressed in terms of the various voltages and currents of the stage.

used for grounded-grid operation wherein the grid drive voltage is expressed as the *cathode-to-grid voltage* and is negative in sign. It must be remembered, however, that a negative cathode voltage is equal to a positive grid voltage, and normal constant-current curves may also be employed for cathode-driven computations.

For the first step in the calculations, the procedure given in Section 7-3 is used. For this example, a 3-1000Z "zero bias" triode is chosen, operating at 3000 plate volts at 2000 watts PEP input in class-B service. Computations are as follows:

3-1000Z at 3000 volts class-B

1,2,3. $E_b = 3000$
$P_i = 2000$ watts PEP
Let $N_p = 65\%$, an average value for class-B mode
$P_o = 2000 \times 0.65 = 1300$ W PEP
$\mu = 200$

4. $$I_b = \frac{2000}{3000} = 0.67 \text{ amp}$$

5. Approx. $i_{b\,max} = 3.1\ I_b$ (for $N_p = 0.65$) $= 3.1 \times 0.65 = 2.08$ amperes
6. Locate the point on the constant-current chart where the constant-current line corresponding to the appropriate value of $i_{b\,max}$ determined in step 5 inflects sharply upward. Approximate $e_{b\,min} = 500$ volts.
7. $e_{p\,min} = 3000 - 500 = 2500$ volts.
8. $$\frac{i_{1\,max}}{I_b} = \frac{2 \times 0.65 \times 3000}{2500} = 1.56$$
9. $$\frac{i_{b\,max}}{I_b} = 3.13 \text{ (from figure 9).}$$
10. $i_{b\,max} = 3.13 \times 0.67 = 2.1$ amps.

This agrees closely with the approximation made in Step 5.

11. Read the values maximum cathode-to-filament voltage (e_k) and peak grid current ($i_{g\,max}$) from the constant-current chart for the values of $e_{b\,min}$ and $i_{b\,max}$ found in steps 6 and 10 respectively.

$$e_k = -88$$
$$i_{g\,max} = 0.8 \text{ amp}$$

12. $$\cos\frac{\theta_b}{2} = 2.32\ (1.56 - 1.57) = 0$$
(Conduction angle is approximately 180° and cos 180° = 0)

13. $E_c = 0$
14. $e_{k\,max} = -88$ volts
15-17. For zero bias class-B mode, $I_c \cong 0.25\ i_{g\,max}$. $I_c \cong 0.25 \times 0.8 = 0.2$ amp. (200 ma)
18. $p_d = 0.9 \times |88| \times 0.2 = 15.8$ watts PEP
19. $p_g = 15.8$ watts PEP
20. $i_{1\,max} =$ (Ratio of step 8) $\times I_b$
$i_{1\,max} = 1.56 \times 0.67 = 1.06$ amp

$$P_o\ (\text{PEP}) = \frac{1.06 \times 2500}{2}$$
$$= 1325 \text{ watts.}$$

21. $$R_L \cong \frac{3000}{1.8 \times 0.67} = 2500 \text{ ohms}$$

22. Total peak drive power,

$$p_k = \frac{e_k \times i_{1\,max}}{2} + p_d$$

$$p_k = \frac{88 \times 1.06}{2} + 15.8 \cong 61 \text{ watts PEP}$$

23. Total power output of the stage is equal to 1325 watts (contributed by 3-1000Z) plus that portion of drive power contributed by the conversion of drive power to plate output power. This is approximately equal to the first term of the equation of step 22.

$$P_o\ (\text{PEP})\ \text{total} = 1325 + 44$$
$$= 1369 \text{ watts}$$

24. Cathode driving impedance of the grounded grid stage is:

$$Z_k \cong \frac{e_k}{i_{1\,max} + 1.5 \times I_c}$$

$$Z_k \cong \frac{88}{1.06 + 0.3} = 64 \text{ ohms}$$

A summary of the typical operating parameters for the 3-1000 Z at $E_b = 3000$ are

D-c Plate Voltage	3000
Zero-Signal Plate Current (from constant-current chart)	180 ma

Max. Signal (PEP) Plate Current	670 ma
Max. Signal (PEP) Grid Current	200 ma
Max Signal (PEP) Drive Power	61 watts
Max. Signal (PEP) Power Input	2000 watts
Max. Signal (PEP) Power Output (including feedthrough power)	1369 watts
Plate Load Impedance	2500 ohms
Cathode Driving Impedance	64 ohms

Cathode Tank of G-G or C-F Power Amplifier The cathode tank circuit for either a grounded-grid or cathode-follower r-f power amplifier may be a conventional tank circuit if the filament transformer for the stage is of the low-capacitance high-voltage type. Conventional filament transformers, however, will not operate with the high values of r-f voltage present in such a circuit. If a conventional filament transformer is to be used, the cathode tank coil may consist of two parallel heavy conductors (to carry the high filament current) bypassed at both the ground end and at the tube socket. The tuning capacitor is then placed between filament and ground. It is possible in certain cases to use two r-f chokes of special design to feed the filament current to the tubes, with a conventional tank circuit between filament and ground. Coaxial lines also may be used to serve both as cathode tank and filament feed to the tubes for vhf and uhf work.

Control-Grid Dissipation in Grounded-Grid Stages Tetrode tubes may be operated as grounded-grid (cathode-driven) amplifiers by tying the grid and screen together and operating the tube as a high-μ triode (figure 15). Combined grid and screen current, however, is a function of tube geometry and may reach destructive values under conditions of full excitation. Proper division of excitation between grid and screen should be as the ratio of the screen-to-grid amplification, which is approximately 5 for tubes such as the 4-250A, 4-400A, etc. The proper ratio of grid/screen excitation may be achieved by tapping the grid at some point on the input circuit, as shown. Grid dissipation is reduced, but the over-all level of excitation is increased about 30% over the value required for simple grounded-grid operation.

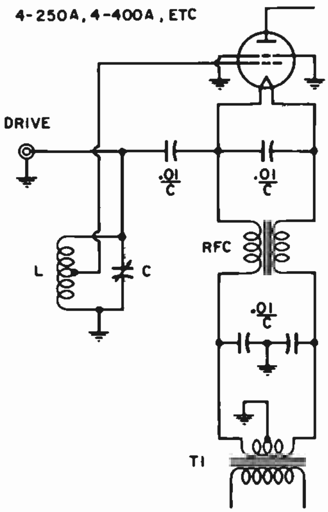


Figure 15

TAPPED INPUT CIRCUIT REDUCES EXCESSIVE GRID DISSIPATION IN G-G CRCUIT

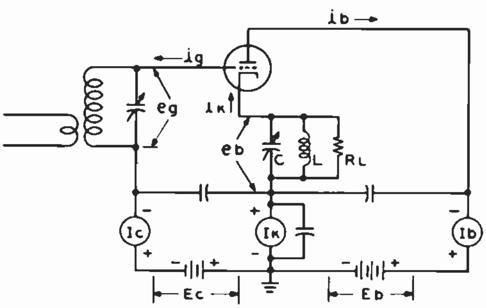
C = 20 pf per meter wavelength
RFC = Dual-winding on 1/2-inch diameter, 3 1/2-inch long ferrite rod. Q-1 material. (Indiana General).

Plate-Return or Cathode-Follower R-F Power Amplifier Circuit diagram, electrode potentials and currents, and operating conditions for a cathode-follower r-f power amplifier are given in figure 16. This circuit can be used, in addition to the grounded-grid circuit just discussed, as an r-f amplifier with a triode tube and no additional neutralization circuit. However, the circuit will oscillate if the impedance from cathode to ground is allowed to become capacitive rather than inductive or resistive with respect to the operating frequency. The circuit is not recommended except for vhf or uhf work with coaxial lines as tuned circuits since the peak grid swing required on the r-f amplifier stage is approximately equal to the plate voltage on the amplifier tube if high-efficiency operation is desired. This means, of course, that the grid tank must be able to withstand slightly more peak voltage than the plate tank. Such a stage may not be plate modulated unless the driver stage is modulated the same percentage as the final amplifier. However, such a stage may be used as an amplifier of modulated waves (class-B linear) or as a c-w or f-m amplifier.

The design of such an amplifier stage is essentially the same as the design of a grounded-grid amplifier stage as far as the first step is concerned. Then, for the second step the operating conditions given in figure 16 are applied to the data obtained in the first step.

7-6 Class-AB$_1$ Radio-Frequency Power Amplifiers

Class-AB$_1$ r-f amplifiers operate under such conditions of bias and excitation that grid current does not flow over any portion of the input cycle. This is desirable, since distortion caused by grid-current loading is absent, and also because the stage is capable of high power gain. Stage efficiency is about 60 percent when a plate-current conduction



POWER OUTPUT TO LOAD = $\frac{e_{b\,MIN}\,(i_{1\,MAX} + 1.8\,I_c)}{2}$

POWER DELIVERED BY OUTPUT TUBE = $\frac{e_{b\,MIN} \times i_{1\,MAX}}{2}$

DRIVE POWER = $\frac{(e_{gMP} + e_{b\,MIN}) \times 1.8\,I_c}{2}$

$$Z_G \cong \frac{e_{g\,MAX}}{1.8\,I_c}$$

$$R_L \cong \frac{E_b}{1.8 \times I_K}$$

Figure 16

CATHODE-FOLLOWER R-F POWER AMPLIFIER

The equations show the relationship between the tube potentials and currents and the input and output power of the stage. The approximate input and output load impedances are also given.

angle of 210° is chosen, as compared to 65 percent for class-B operation.

The level of static (quiescent) plate current for *lowest distortion* is quite high for class-AB$_1$ tetrode operation. This value is determined by the tube characteristics, and is not greatly affected by the circuit parameters or operating voltages. The maximum d-c potential is therefore limited by the static dissipation of the tube, since the resting plate current figure is fixed. The static plate current of a tetrode tube varies as the 3/2 power of the screen voltage. For example, raising the screen voltage from 300 to 500 volts will double the plate current. The optimum static plate current for minimum distortion is also doubled, since the shape of the E_c-I_b curve does not change.

In actual practice, somewhat lower static plate current than optimum may be employed without raising the distortion appreciably, and values of static plate current of 0.6 to 0.8 of optimum may be safely used, depending on the amount of nonlinearity that can be tolerated.

As with the class-B linear stage, the minimum plate voltage swing ($e_{b\,min}$) of the class-AB$_1$ amplifier must be kept above the d-c screen potential to prevent operation in the nonlinear portion of the characteristic curve. A *low value* of screen voltage allows greater r-f plate voltage swing, resulting in improvement in plate efficiency of the tube. A balance between plate dissipation, plate efficiency, and plate-voltage swing must be achieved for best linearity of the amplifier.

The S-Curve The perfect linear amplifier delivers a signal that is a replica of the input signal. Inspection of the plate-characteristic curve of a typical tube will disclose the tube linearity under class-AB$_1$ operating conditions (figure 17). The curve is usually of exponential shape, and the signal distortion is held to a small value by operating the tube well below its maximum output, and centering operation over the most linear portion of the characteristic curve.

The relationship between exciting voltage in a class-AB$_1$ amplifier and the r-f plate-circuit voltage is shown in figure 18. With a small value of static plate current the lower portion of the line is curved. Maximum undistorted output is limited by the point on the line (A) where the instantaneous plate voltage drops down to the screen voltage. This "hook" in the line is caused by current diverted from the plate to the grid and

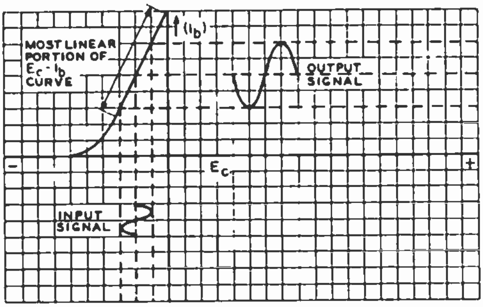


Figure 17

E_g-I_p CURVE

Amplifier operation is confined to the most linear portion of the characteristic curve.

screen elements of the tube. The characteristic plot of the usual linear amplifier takes the shape of an S-curve. The lower portion of the curve is straightened out by using the proper value of static plate current, and the upper portion of the curve is avoided by limiting minimum plate voltage swing to a point substantially above the value of the screen voltage.

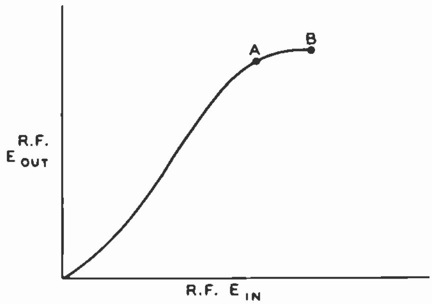


Figure 18

LINEARITY CURVE OF TYPICAL TETRODE AMPLIFIER

At point A the instantaneous plate voltage is swinging down to the value of the screen voltage. At point B it is swinging well below the screen and is approaching the point where saturation, or plate-current limiting takes place.

Operating Parameters for the Class-AB$_1$ Linear Amplifier The approximate operating parameters may be obtained from the constant-current curves (E_c-E_b) or the E_c-I_b curves of the tube in question (figure 19). The following example will make use of the latter information, although equivalent results may be obtained from constant-current curves. An operating load line is first approximated. One end of the load line is determined by the d-c operating voltage of the tube, and the required static plate current. As a starting point, let the product of the plate voltage and current approximate the plate dissipation of the tube. Assuming a 4-400A tetrode is used, this end of the load line will fall on point A (figure 19). Plate power dissipation is 360 watts (3000V at 120 ma). The opposite end of the load line will fall on a point determined by the minimum instantaneous plate voltage, and by the maximum instantaneous plate current. The minimum plate voltage, for best linearity should be considerably higher than the screen voltage. In this case, the screen voltage is 500, so the minimum plate voltage excursion should be limited to 600 volts. Class-AB$_1$ operation implies no grid current, therefore the load line cannot cross the E_c = 0 line. At the point $e_{b\,min}$ = 600, E_c = 0, the maximum instantaneous plate current is 580 ma (Point B).

Each point at which the load line crosses a grid-voltage axis may be taken as a point for construction of the E_c-I_b curve, just as was done in figure 22, chapter 6. A constructed curve shows that the approximate static bias voltage is −74 volts, which checks closely with point A of figure 19. In actual practice, the bias voltage is set to hold the actual dissipation slightly below the maximum limit of the tube.

The single tone PEP power output is:

$$P_o = \frac{(E_b - e_{b\,min}) \times i_{b\,max}}{4} =$$

$$\frac{(3000 - 600) \times 0.58}{4} = 348 \text{ watts}$$

The plate current conduction angle efficiency factor for this class of operation is 0.73, and the actual plate circuit efficiency is:

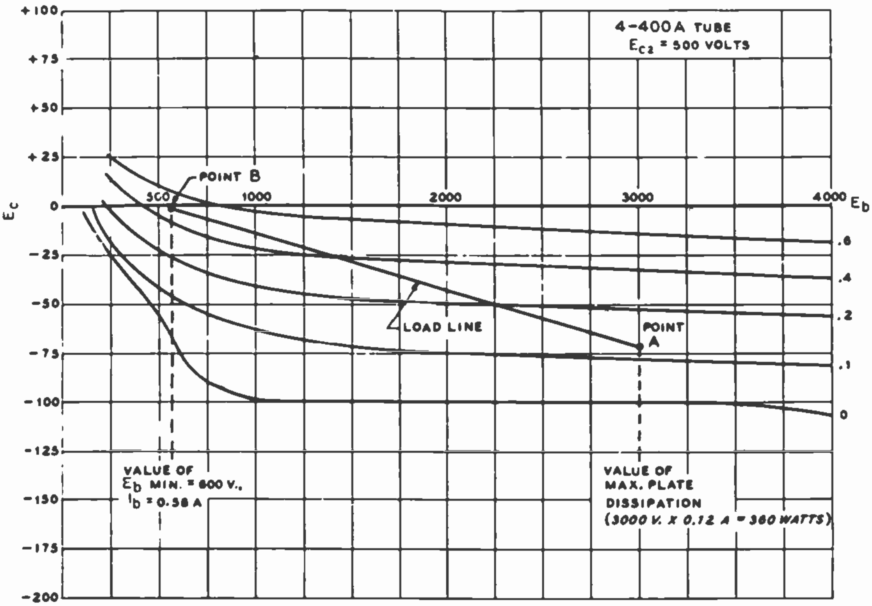


Figure 19

OPERATING PARAMETERS FOR TETRODE LINEAR AMPLIFIER ARE OBTAINED FROM CONSTANT-CURRENT CURVES.

$$N_p = \frac{E_b - e_{b\,min}}{E_b} \times 0.73 = 58.4\%$$

The peak power input to the stage is therefore:

$$\frac{P_o}{N_p} \times 100 = \frac{348}{58.4} = 595 \text{ watts PEP}$$

The peak plate dissipation is:

$$595 - 348 = 247 \text{ watts}$$

(Note: A 4-250A may thus be used in lieu of the 4-400A as peak plate dissipation is less than 250 watts, provided resting plate current is lowered to 70 ma.)

It can be seen that the limiting factor for either the 4-250A or 4-400A is the static plate dissipation, which is quite a bit higher than the operating dissipation level. It is possible, at the expense of a higher level of distortion, to drop the static plate dissipation and to increase the screen voltage to obtain greater power output. If the screen voltage is set at 800, and the bias increased sufficiently to drop the static plate current to 70 ma, the single-toned d-c plate current may rise to 300 ma, for a power input of 900 watts. The plate circuit efficiency is 55.6 percent, and the power output is 500 watts.

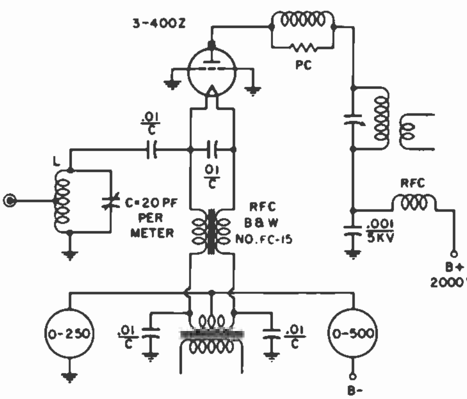


Figure 20

SIMPLE GROUNDED-GRID LINEAR AMPLIFIER

Tuned cathode (L—C) is required to prevent distortion of driving-signal waveform.

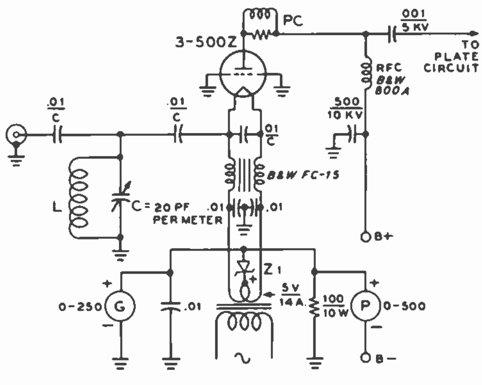


Figure 21

ZENER-DIODE BIAS FOR GROUNDED-GRID STAGE

The resting plate current of a grounded stage may be reduced by inclusion of a Zener diode in the filament return circuit. At a plate potential of 3250 volts, for example, a Zener bias of 4.7 volts reduces the resting plate current of the 3-500Z from 160 to approximately 90 milliamperes. A 1N4551 Zener may be used, bolted to the chassis for a heat sink.

Static plate dissipation is 210 watts, within the rating of either tube.

At a screen potential of 500 volts, the maximum screen current is less than 1 ma, and under certain loading conditions may be negative. When the screen potential is raised to 800 volts maximum screen current is 18 ma. The performance of the tube depends on the voltage fields set up in the tube by the cathode, control grid, screen grid, and plate. The quantity of current, flowing in the screen circuit is only incidental to the fact that the screen is maintained at a positive potential with respect to the electron stream surrounding it.

The tube will perform as expected as long as the screen current, in either direction, does not create undesirable changes in the screen voltage, or cause excessive screen dissipation. Good regulation of the screen supply is therefore required. Screen dissipation is highly responsive to plate loading conditions and the plate circuit should always be adjusted so as to keep the screen current below the maximum dissipation level as established by the applied voltage.

7-7 Grounded-Grid Linear Amplifiers

The popularity of grounded-grid (cathode-

R-F LINEAR AMPLIFIER SERVICE FOR SSB AND CW

CATHODE-DRIVEN (GROUNDED-GRID)
CLASS-B MODE

TUBE	PLATE VOLTAGE Eb	FIL V/A	APPROX. ZERO SIG. PLATE Ibo CURRENT	MAX. SIG. PLATE Ib CURRENT	MAX. SIG. GRID Ic1 CURRENT	DRIVING IMPEDAN. Rk	PLATE LOAD RL IMPEDAN.	MAX. SIG. DRIVING POWER W.	PEP PLATE INPUT POWER W.	USEFUL OUTPUT POWER W.	AVERAGE PLATE DISSIPAT. Pd	APPROX. 3d ORDER IMD·Db
811A	1250 1700	7.5/4	18 30	175 160	28 28	320	3600 5200	12 15	220 270	135 175	70 85	−33 −28
572B / T-160L	2400	7.5/4	20	250	45	215	4500	30	600	350	160	−28
813	2000 2500	10/5	20 30	200	50 50	270	5000 7000	10 11	400 500	270 350	130 150	−30 −33
3-400Z / 8163	2000 2500 3000	5/14.5	62 73 100	400 400 333	148 142 120	120	2750 3450 4750	— — 32	800 1000 1000	445 600 655	355 400 345	−40 −35 −32
3-500Z	2000 2500 3000	5/14.5	95 130 160	400 400 370	130 120 115	115	2750 3450 5000	— — 30	800 1000 1100	500 600 750	300 400 350	−38 −33 −30
3-1000Z / 8164	2500 3000 3500	7.5/21	162 175 200	800 670 750	270 220 245	65 65 65	1800 2400 2600	95 65 65	2000 2000 2600	1250 1250 1770	750 750 830	−38 −35 −30
3CX1000A7 / 8283	2500 3000	5/30	200 310	875 800	590 320	41 42	1100 1670	78 67	2200 2400	1000 1200	1000 1000	−32 −32
4-125A	2000 2500 3000	5/6.5	10 15 20	105 110 115	55 55 55	340 340 340	10 500 13 500 15 700	16 16 16	210 275 345	145 190 240	65 85 100	— — —
4-400A	2000 2500 3000	5/14.5	55 60 70	265 270 330	100 100 108	160 150 140	3950 4500 5600	38 39 40	530 675 990	325 435 572	200 225 390	— — −30
4-1000A	3000 4000 5000	7.5/21	60 90 120	700 675 540	200 200 115	104 106 110	2450 2450 5500	130 105 70	2100 2700 2700	1475 1870 1900	750 730 700	−34 −34 —

Figure 22

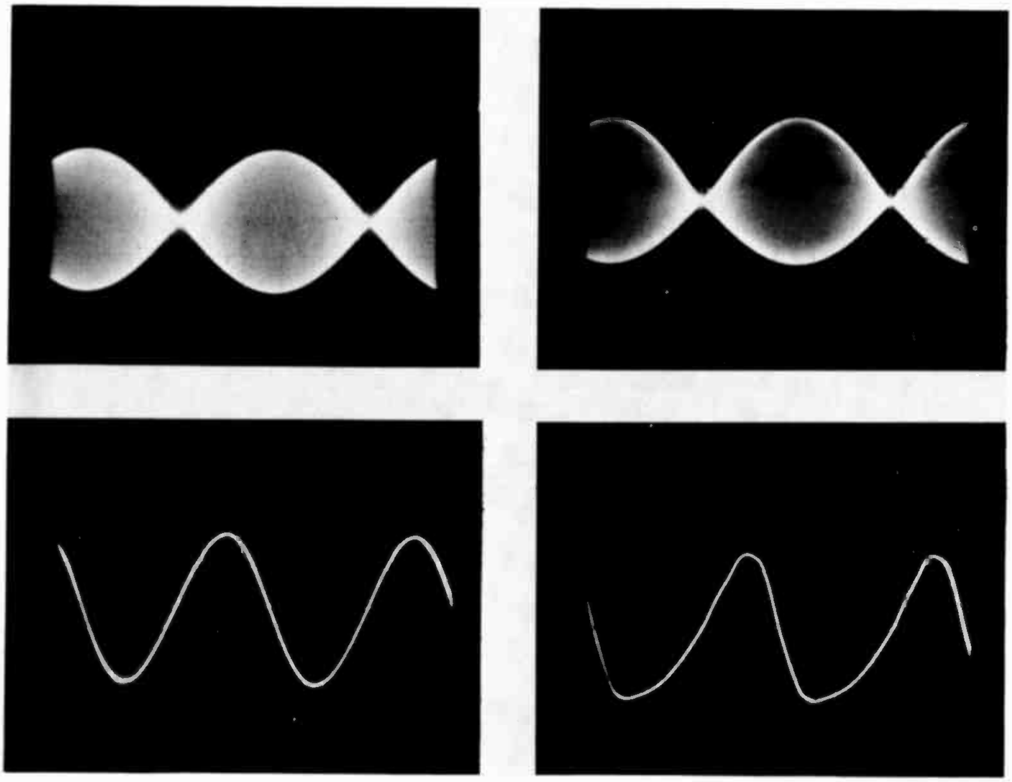


Figure 23

Waveform distortion caused by half-cycle loading at cathode of grounded-grid amplifier may be observed (right) whereas undistorted waveform is observed with tuned cathode circuit (left). Two-tone tests at 2.0 MHz proved the necessity of using a cathode tank circuit for lowest inter-modulation distortion.

driven) linear amplifiers for SSB service is unique in the Amateur Service. Elimination of costly and bulky bias and screen power supplies make the "g-g" amplifier an economical and relatively light-weight power unit.

A typical grounded-grid amplifier is shown in figure 20. The driving signal is applied between the grid and the cathode, with the grid held at r-f ground potential. The control grid serves as a shield between the cathode and the plate, thus making neutralization unnecessary at medium and high frequencies. High-μ triodes and triode-connected tetrodes may be used in this configuration. Care must be taken to monitor the #1-grid current of the tetrode tubes as it may run abnormally high in some types (4X150A family) and damage to the tube may possibly result unless a protective circuit of the form shown in figure 21 is used.

"Zero-bias" triodes (811-A, 3-400Z and 3-1000Z) and certain triode-connected tetrodes (813 and 4-400A, for example) require no bias supply and good linearity may be achieved with a minimum of circuit components. An improvement of the order of 5 to 10 decibels in intermodulation distortion may be gained by operating such tubes in the grounded-grid mode in contrast to the same tubes operated in class-AB_1, grid-driven mode. The improvement in the distortion figure varies from tube type to tube type, but all so-called "grounded-grid" triodes and triode-connected tetrodes show some degree of improvement in distortion figure when cathode-driven as opposed to grid-driven service.

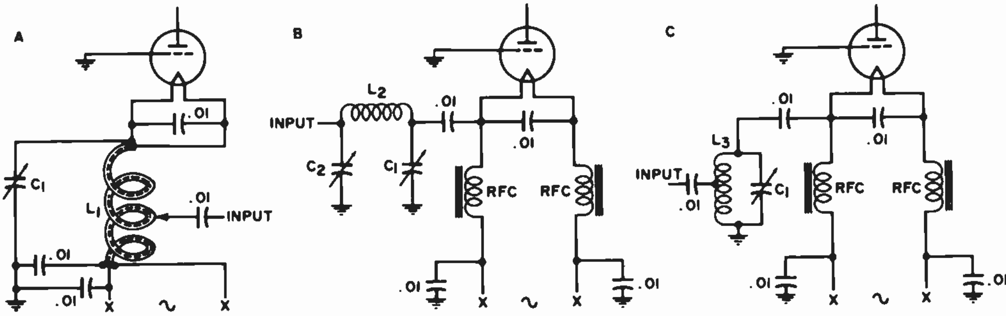


Figure 24

Tuned cathode network for cathode-driven circuit may take form of bifilar coil (A), pi-network (B), or shunt LC circuit (C). Circuit Q of at least 2 is recommended. Capacitor C_1 may be a 3-gang broadcast-type unit. Coils L_1, L_2, or L_3 are adjusted to resonate to the operating frequency with C_1 set to approximately 13 pf-per meter wavelength. Capacitor C_2 is approximately 1.5 times the value of C_1. The input taps on coils L_1 and L_3, or the capacitance of C_2 are adjusted for minimum SWR on coaxial line to the exciter.

Cathode-Driven High-μ Triodes High-μ triode tubes may be used to advantage in cathode-drive (grounded-grid) service. The inherent shielding of a high-μ tube is better than that of a low-μ tube and the former provides better gain per stage and requires less drive than the latter because of less feedthrough power. Resistive loading of the input or driving circuit is not required because of the constant feedthrough power load on the exciter as long as sufficient Q exists in the cathode tank circuit. Low-μ triodes, on the other hand, require extremely large driving signals when operated in the cathode-driven configuration, and stage gain is relatively small. In addition, shielding between the input and output circuits is poor compared to that existing in high-μ triodes.

Bias Supplies for G-G Amplifiers Medium-μ triode tubes that require grid bias may be used in cathode-driven service if the grid is suitably bypassed to ground and placed at the proper negative d-c potential. Bias supplies for such circuits, however, must be capable of good voltage regulation under conditions of grid current so that the d-c bias value does not vary with the amplitude of the grid current of the stage. Suitable bias supplies for this mode of operation are shown in the *Power Supply* chapter of this Handbook. *Zener bias* (figure 21) may be used for low values of bias voltage. Approximate values of bias voltage for linear amplifier service data may be obtained from the audio data found in most tube manuals, usually stated for push-pull class-AB_1 or AB_2 operation. As the tube "doesn't know" whether it is being driven by an audio signal or an r-f signal, the audio parameters may be used for linear service, but the stated d-c currents should be divided by two for a single tube, since the audio data is usually given for two tubes. Grounded-grid operating data for popular triode and tetrode tubes is given in figure 22.

The Tuned Cathode Circuit Input waveform distortion may be observed at the cathode of a grounded-grid linear amplifier as the result of grid- and plate-current loading of the input circuit on alternate half-cycles by the single-ended stage (figure 23). The driving source thus "sees" a very low value of load impedance over a portion of the r-f cycle and an extremely high impedance over the remaining portion of the cycle. Unless the output voltage regulation of the r-f source is very good, the portion of the wave on the loaded part of the cycle will be degraded. This waveform distortion contributes to intermodulation distortion and also may cause TVI difficulties as a result of the harmonic content of the wave. Use of a tuned cathode circuit in the grounded-grid stage will preserve the waveform as shown in the photographs. The tuned-cathode circuit need have only a Q of

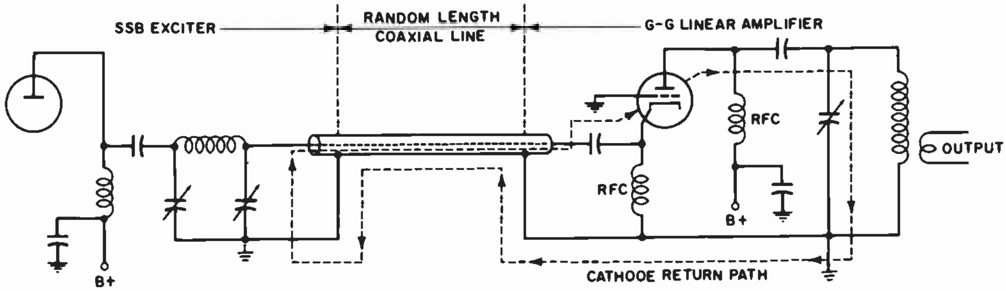


Figure 25

Untuned cathode circuit of grounded-grid amplifier offers high-impedance path to the r-f current flowing between plate and cathode of the amplifier tube. The alternative path is via the interconnecting coaxial line and tank circuit of the exciter. Waveform distortion of the driving signal and high intermodulation distortion may result from use of alternative input circuit.

2 or more to do the job, and should be resonated to the operating frequency of the amplifier. Various versions of cathode tank circuits are shown in figure 24.

In addition to reduction of waveform distortion, the tuned-cathode circuit provides a short r-f return path for plate current pulses from plate to cathode (figure 25). When the tuned circuit is not used, the r-f return path is via the outer shield of the coaxial line, through the output capacitor of the exciter plate-tank circuit and back to the cathode of the linear amplifier tube via the center conductor of the coaxial line. This random, uncontrolled path varies with the length of interconnecting coaxial line, and permits the outer shield of the line to be "hot" compared to r-f ground.

7-8 Intermodulation Distortion

If the output signal of a linear amplifier is an exact replica of the exciting signal there will be no distortion of the original signal and no distortion products will be generated in the amplifier. Amplitude distortion of the signal exists when the output signal is not strictly proportional to the driving signal and such a change in magnitude may result in *intermodulation distortion* (IMD). IMD occurs in any nonlinear device driven by a complex signal having more than one frequency. A voice signal (made-up of a multiplicity of tones) will become blurred or distorted by IMD when amplified by a nonlinear device. As practical linear amplifiers have some degree of IMD (depending on design and operating parameters) this disagreeable form of distortion exists to a greater or lesser extent on most SSB signals.

A standard test to determine the degree of IMD is the *two-tone test,* wherein two radio-frequency signals of equal amplitude are applied to the linear equipment, and the resulting output signal is examined for spurious signals, or unwanted products. These unwanted signals fall in the fundamental-signal region and in the various harmonic regions of the amplifier. Signals falling outside the fundamental-frequency region are termed *even-order products*, and may be attenuated by high-Q tuned circuits in the amplifier. The spurious products falling close to the fundamental-frequency region are termed *odd-order products*. These unwanted products cannot be removed from the wanted signal by tuned circuits and show up on the signal as "splatter," which can cause severe interference to communication in an adjacent channel. Nonlinear operation of a so-called "linear" amplifier will generate these unwanted products. Amateur practice calls for suppression of these spurious products to better than 30 decibels below peak power level of one tone of a two-tone test signal. Commercial practice demands suppression to be better than 40 decibels below this peak level.

Additional data on IMD and two-tone test techniques is given in chapter 16.

CHAPTER EIGHT

Special Circuitry for Vacuum Tubes and Semiconductor Devices

A whole new concept of vacuum-tube and semiconductor applications has been developed in recent years. No longer are these devices chained to the field of radio or wire communication. This chapter is devoted to some of the more common circuits encountered in computer technology and in industrial and military applications.

8-1 Limiting Circuits

The term *limiting* refers to the removal or suppression, by electronic means, of the extremities of an electronic signal. Circuits which perform this function are referred to as *limiters* or *clippers*. Limiters are useful in waveshaping circuits where it is desirable to square off the extremities of the applied signal. A sine wave may be applied to a limiter circuit to produce a rectangular wave. A peaked wave may be applied to a limiter circuit to eliminate either the positive or negative peaks from the output. Limiter circuits are employed in f-m receivers where it is necessary to limit the amplitude of the signal applied to the detector. Limiters may be used to reduce automobile ignition noise in short-wave receivers, or to maintain a high average level of modulation in a transmitter. They may also be used as protective devices to limit input signals to special circuits.

Diode Limiters The characteristics of a diode are such that the device conducts only when the anode is at a positive potential with respect to the cathode. A positive potential may be placed on the cathode, but the diode will not conduct until the voltage on the anode rises above an equally positive value. As the anode becomes more positive with respect to the cathode, the diode conducts and passes that portion of the wave which is more positive than the cathode voltage. Diodes may be used as either series or parallel limiters, as shown in figure 1. A diode may be so biased that only a certain portion of the positive or negative cycle is removed.

Peak Limiting A peak clipper consisting of two diode limiters may be used to limit the amplitude of an a-c signal to a predetermined value to provide a high average signal level. Limiters of this general type are useful in transmitters to provide a high level of modulation without danger of overmodulation. An effective limiter for this service is the *series-diode-gate clipper* (figure 2). The signal to be clipped is coupled to the clipper through capacitor C_1. R_1 and R_2 are the clipper input and output load resistors. The clipper anodes are tied together and connected to the clipping level control (R_4) through series resistor R_3. The level control acts as a voltage divider between the anode supply and ground. The exact level at which clipping will occur is set by R_4, which controls the positive potential applied to the anodes of the diodes.

Under static conditions, a d-c voltage is obtained through R_4 and applied through R_3 to both anodes of the clipping devices. Current flows through R_3 and R_4, dividing through the two diodes and the two load

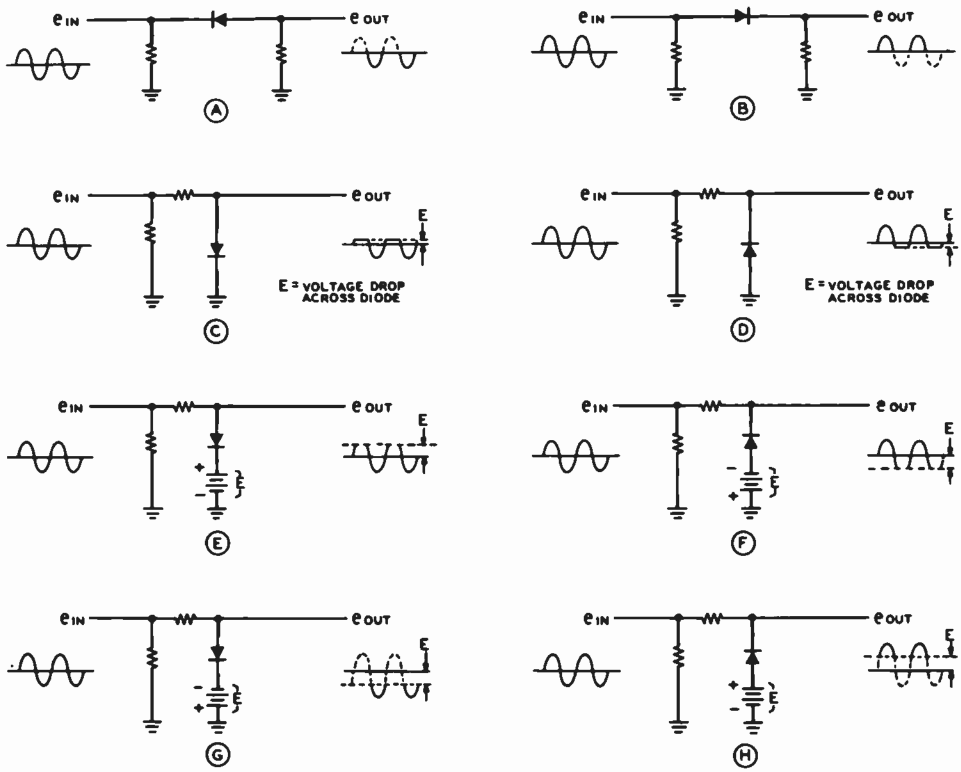


Figure 1

VARIOUS DIODE LIMITING CIRCUITS

Series diodes limiting positive and negative peaks are shown in A and B. Parallel diodes limiting positive and negative peaks are shown in C and D. Parallel diodes limiting above and below ground are shown in E and F. Parallel-diode limiters which pass negative and positive peaks are shown in G and H.

resistors. All points of the clipper circuit are maintained at a positive potential above ground. The voltage drop between the anode and the cathode of each diode is very small compared to the drop across the 300K resistor in series with the anodes. The anode and cathode of each diode are therefore maintained at approximately equal potentials as long as no diode current flows. Signal clipping does not occur until the peak signal input voltage reaches a value greater than the static voltages at the anodes of the diodes.

Assume that R_4 has been set to a point that will give 4 volts at the anodes of the diodes. When the peak signal voltage is less than 4 volts, both the diodes conduct at all times. As long as the diodes conduct, their resistance is very low compared with resistor R_3. Whenever a voltage change occurs across input resistor R_1, the voltage at all of the circuit elements increases or decreases by the same amount as the input voltage changes, and the voltage drop across R_3 changes by an equal amount. Thus, as long as the peak signal voltage is less than 4 volts, the diodes act as conductors and the output voltage of the device follows all voltage fluctuations at the input cathode.

If, under static conditions, 4 volts appear at the anodes, then twice this voltage (8 volts) will be present if one of the diode circuits is opened, thus removing its d-c load from the circuit. As long as only one of the diodes continues to conduct, the voltage at the diode anodes cannot rise above twice the voltage selected by R_1. In this example, the voltage cannot rise above 8 volts. Now, if the input signal applied through C_1 is increased to any peak value between zero and

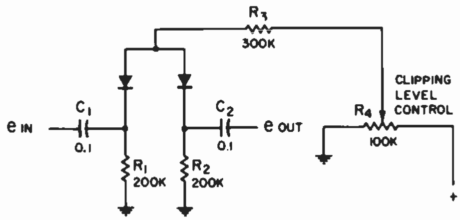


Figure 2

THE SERIES-DIODE GATE CLIPPER FOR PEAK LIMITING

+4 volts, the cathode of the first diode will increase in voltage by the same amount to the proper value between 4 and 8 volts. The other elements will assume the same potential as the first cathode. However, the anodes cannot increase more than 4 volts above their original 4-volt static level condition. When the input signal voltage rises to more than +4 volts, the cathode potential increases to more than 8 volts. Since the diode circuit potential remains at 8 volts, the first diode ceases to conduct until the signal voltage across R_1 drops below 4 volts.

When the input voltage swings in a negative-going direction, it will subtract from the 4-volt drop across R_1 and decreases the voltage on the cathode of the input diode by an amount equal to the input voltage. The anodes and the output cathode will follow the voltage level at the input cathode so long as the input voltage does not swing below −4 volts. If the input voltage does not change more than 4 volts in a negative direction, the diode anodes will also become negative. The potential at the output cathode will follow the input cathode voltage and decrease from its normal value of 4 volts until it reaches zero potential. As the input cathode decreases to less than zero, the anodes will follow the change. However, the output cathode, grounded through R_2, will stop at zero potential as the anode becomes negative. Conduction through the second diode is impossible under these conditions. The output cathode remains at zero potential until the voltage at the input cathode swings back to zero.

The voltage developed across output resistor R_2 follows the input voltage variations as long as the input voltage does not swing to a peak value greater than the static voltage at the diode anodes, which is determined by R_1. Effective clipping may thus be obtained at any desired level.

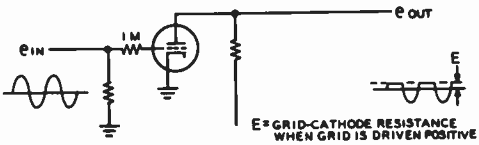


Figure 3

GRID LIMITING CIRCUIT

The square-topped audio waves generated by this clipper are high in harmonic content, but these higher-order harmonics may be greatly reduced by a low-level speech filter.

Grid Limiters A triode grid limiter is shown in figure 3. On positive peaks of the input signal, the triode grid attempts to swing positive, and the grid-cathode resistance drops to about 1000 ohms or so. The voltage drop across the series grid resistor (usually of the order of 1 megohm) is large compared to the grid-cathode drop, and the resulting limiting action removes the top part of the positive input wave.

8-2 Clamping Circuits

A circuit which holds either amplitude extreme of a waveform to a given reference level of potential is called a *clamping circuit* or a *d-c restorer*. Clamping circuits are used after RC-coupling circuits where the waveform swing is required to be either above or below the reference voltage, instead of alternating on both sides of it (figure 4). Clamping circuits are usually encountered in oscilloscope sweep circuits. If the sweep voltage does not always start from the same reference point, the trace on the screen does not begin at the same point on the screen each time the sweep is repeated and therefore is "jittery." If a clamping circuit is placed between the sweep amplifier and the deflection element, the start of the sweep can be regulated by adjusting the d-c voltage applied to the clamping tube (figure 5).

8-3 Multivibrators

The *multivibrator*, or *relaxation oscillator*,

(A) POSITIVE CLAMPING CIRCUIT (B) NEGATIVE CLAMPING CIRCUIT

Figure 4

SIMPLE POSITIVE AND NEGATIVE CLAMPING CIRCUITS

is used for the generation of nonsinusoidal waveforms. The output is rich in harmonics, but the inherent frequency stability is poor. The multivibrator may be stabilized by the introduction of synchronizing voltages of harmonic or subharmonic frequency.

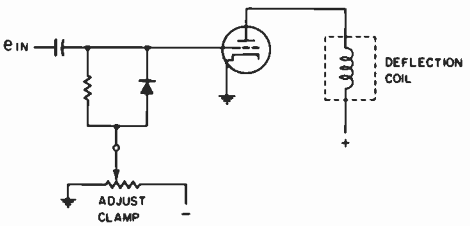


Figure 5

NEGATIVE CLAMPING CIRCUIT EMPLOYED IN ELECTROMAGNETIC SWEEP SYSTEM

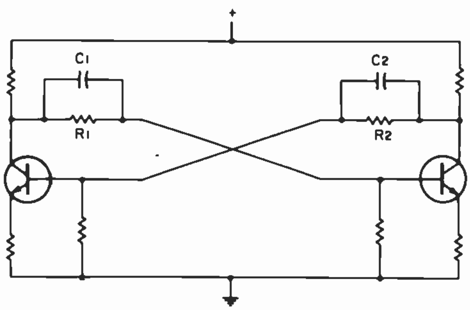
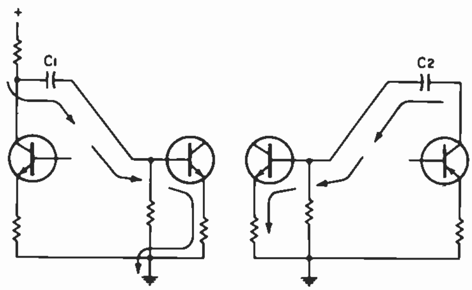


Figure 6

BASIC MULTIVIBRATOR CIRCUITS

In its simplest form, the multivibrator is a simple two-stage RC-coupled amplifier with the output of the second stage coupled through a capacitor to the input of the first stage, as shown in figure 6. Since the output of the second stage is of the proper polarity to reinforce the input signal applied to the first, oscillations can readily take place, started by thermal-agitation and miscellaneous noise. Oscillation is maintained by the process of building up and discharging the store of energy in the coupling capacitors of the two stages. The charging and discharging paths are shown in figure 7. Various types of multivibrators are shown in figure 8.



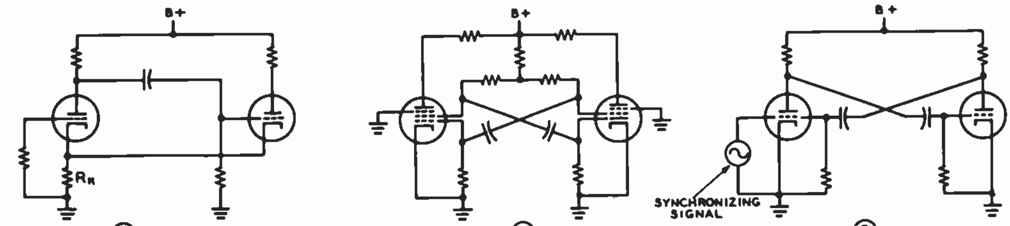
C1 CHARGE PATH C2 DISCHARGE PATH

Figure 7

THE CHARGE AND DISCHARGE PATHS IN THE FREE-RUNNING MULTIVIBRATOR OF FIGURE 6

The output of a multivibrator may be used as a source of square waves, as an electronic switch, or as a means of obtaining frequency division. Submultiple frequencies as low as one-tenth of the injected synchronizing frequency may easily be obtained.

The Eccles-Jordan Circuit The *Eccles-Jordan* trigger circuit is shown in figure 9A. This is not a true multivibrator, but rather a circuit that possesses two conditions of stable equilibrium. One condition is when V_1 is conducting and V_2 is cutoff; the other when V_2 is conducting and V_1 is cutoff. The circuit remains in one or the other of these two stable conditions with no change in operating potentials until some external action occurs which causes the nonconducting tube to conduct.



Ⓐ DIRECT-COUPLED CATHODE MULTIVIBRATOR

Ⓑ ELECTRON-COUPLED MULTIVIBRATOR

Ⓒ MULTIVIBRATOR WITH SINE-WAVE SYNCHRONIZING SIGNAL APPLIED TO ONE TUBE

Figure 8

VARIOUS TYPES OF MULTIVIBRATOR CIRCUITS

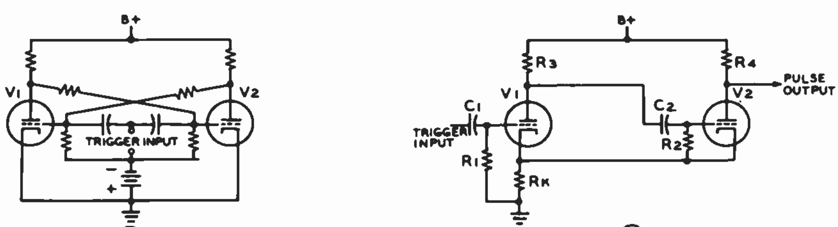
The tubes then reverse their functions and remain in the new condition as long as no plate current flows in the cut-off tube. This type of circuit is known as a *flip-flop* circuit.

Figure 9B illustrates a modified Eccles-Jordan circuit which accomplishes a complete cycle when triggered by a positive pulse. Such a circuit is called a *one-shot* multivibrator. For initial action, V_1 is cut off and V_2 is conducting. A large positive pulse applied to the grid of V_1 causes this tube to conduct, and the voltage at its plate decreases by virtue of the IR drop through R_3. Capacitor C_2 is charged rapidly by this abrupt change in V_1 plate voltage, and V_2 becomes cut off while V_1 conducts. This condition exists until C_2 discharges, allowing V_2 to conduct, raising the cathode bias of V_1 until it is once again cut off.

A direct–cathode-coupled multivibrator is shown in figure 8A. R_K is a common cathode resistor for the two tubes, and coupling takes place across this resistor. It is impossible for a tube in this circuit to completely cut off the other tube, and a circuit of this type is called a *free-running* multivibrator in which the condition of one tube temporarily cuts off the other.

8-4 The Blocking Oscillator

A *blocking oscillator* is any oscillator which cuts itself off after one or more cycles caused by the accumulation of a negative charge on the grid capacitor. This negative charge may gradually be drained off through the grid resistor of the tube, allowing the circuit to oscillate once again. The process is repeated and the tube becomes an intermittent oscillator. The rate of such an occurrence is determined by the RC time constant of the grid circuit. A *single-swing blocking oscillator* is shown in figure 10, wherein the tube is cut off before the completion of one cycle. The tube produces single pulses of energy, the time between the pulses being regulated by the discharge time of the grid RC network. The *self-pulsing* blocking oscillator is shown in figure 11, and



Ⓐ BASIC ECCLES-JORDAN TRIGGER CIRCUIT

Ⓑ ONE-SHOT MULTIVIBRATOR

Figure 9

ECCLES-JORDAN MULTIVIBRATOR CIRCUITS

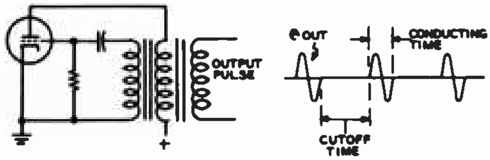


Figure 10

SINGLE-SWING BLOCKING OSCILLATOR

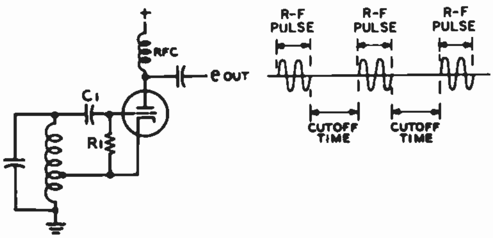


Figure 11

HARTLEY OSCILLATOR USED AS BLOCKING OSCILLATOR BY PROPER CHOICE OF R_1 C_1

is used to produce pulses of r-f energy, the number of pulses being determined by the timing network in the grid circuit of the oscillator. The rate at which these pulses occur is know as the *pulse-repetition frequency*, or *p.r.f.*

8-5 Counting Circuits

A *counting circuit*, or *frequency divider*, is one which receives uniform pulses (representing units to be counted) and produces a voltage that is proportional to the frequency of the pulses. A counting circuit may be used in conjunction with a blocking oscillator to produce a trigger pulse which is a submultiple of the frequency of the applied pulse. Either positive or negative pulses may

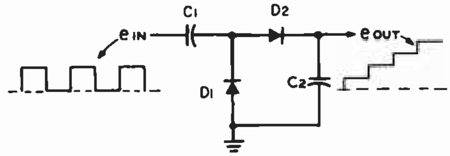


Figure 13

STEP-BY-STEP COUNTING CIRCUIT

be counted. A positive counting circuit is shown in figure 12A, and a negative counting circuit is shown in figure 12B. The positive counter allows a certain amount of current to flow through R_1 each time a pulse is applied to C_1.

The positive pulse charges capacitor C_1 and makes the anode of diode 2 positive with respect to its cathode. Diode 2 conducts until the exciting pulse passes. Capacitor C_1 is then discharged by diode 1, and the circuit is ready to accept another pulse. The average current flowing through R_1 increases as the pulse-repetition frequency increases, and decreases as the p.r.f. decreases.

By reversing the diode connections, as shown in figure 12B, the circuit is made to respond to negative pulses. In this circuit, an increase in the p.r.f. causes a decrease in the average current flowing through R_1, which is opposite to the effect in the positive counter.

A *step-counter* is similar to the circuits discussed, except that a capacitor which is large compared to C_1 replaces the diode load resistor. The charge of this capacitor is increased during the time of each pulse, producing a step voltage across the output (figure 13). A blocking oscillator may be connected to a step counter, as shown in figure 14. The oscillator is triggered into operation when the voltage across C_2 reaches a point sufficiently positive to raise the grid of V

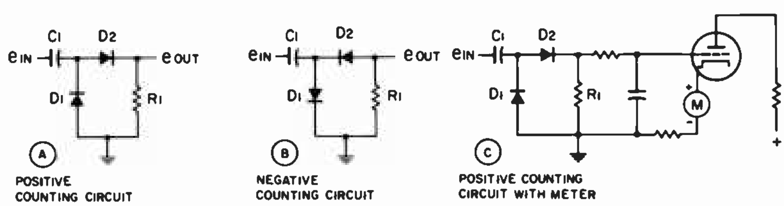


Figure 12

POSITIVE AND NEGATIVE COUNTING CIRCUITS

above cutoff. Circuit parameters may be chosen so that a count division up to 1/20 may be obtained with reliability.

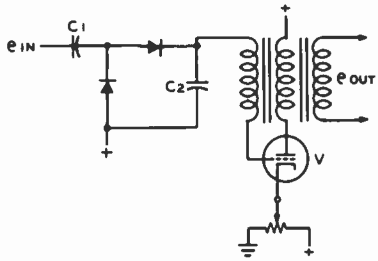


Figure 14

THE STEP-BY-STEP COUNTER USED TO TRIGGER A BLOCKING OSCILLATOR. THE BLOCKING OSCILLATOR SERVES AS A FREQUENCY DIVIDER.

8-6 Resistance-Capacitance Oscillators

In an *RC oscillator,* the frequency is determined by a resistance capacitance network that provides regenerative coupling between the output and input of a feedback amplifier. No use is made of a tank circuit consisting of inductance and capacitance to control the frequency of oscillation.

The *Wien-Bridge* oscillator employs a *Wien network* in the RC feedback circuit and is shown in figure 15. Tube V_1 is the oscillator tube, and tube V_2 is an amplifier and phase-inverter tube. Since the feedback voltage through C_1 produced by V_2 is in phase with the input circuit of V_1 at all frequencies, oscillation is maintained by voltages of any frequency that exist in the circuit. The bridge circuit is used, then, to eliminate feedback voltages of all frequencies except the single frequency desired at the output of the oscillator. The bridge allows a voltage of only one frequency to be effective in the circuit because of the degeneration and phase shift provided by this circuit. The frequency at which oscillation occurs is:

$$f = \frac{1}{2\pi R_1 C_1}$$

when,

$R_1 \times C_1$ equals $R_2 \times C_2$

A lamp (L_P) is used for the cathode resistor

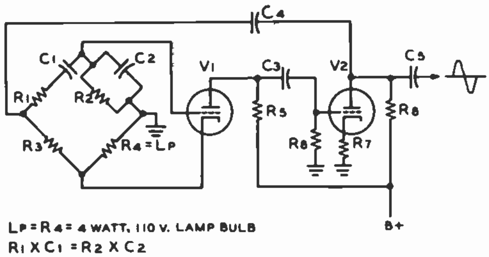


Figure 15

THE WIEN-BRIDGE AUDIO OSCILLATOR

of V_1 as a thermal stabilizer of the oscillator amplitude. The variation of the resistance with respect to the current of the lamp bulb holds the oscillator output voltage at a nearly constant amplitude.

The phase-shift oscillator shown in figure 16 is a single-tube oscillator using a four mesh phase-shift network. Each section of

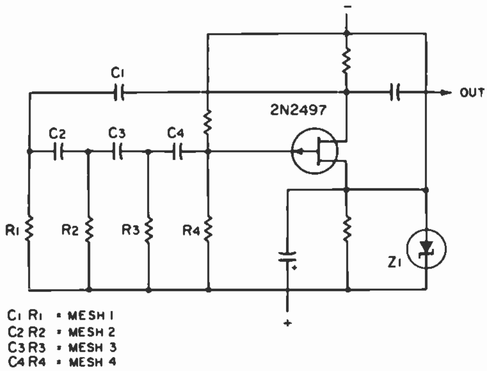


Figure 16

THE PHASE-SHIFT OSCILLATOR

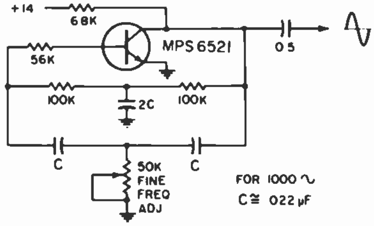


Figure 17

THE BRIDGE-TYPE PHASE-SHIFT OSCILLATOR

the network produces a phase shift in proportion to the frequency of the signal that passes through it. For oscillations to be produced, the signal through the network must be shifted 180°. Four successive phase shifts of 45° accomplish this, and the frequency of oscillation is determined by this phase shift.

A high-μ triode or a pentode may be used in this circuit. In order to increase the frequency of oscillation, either the resistance or the capacitance must be decreased by an appropriate amount.

A *bridge-type phase-shift oscillator* is shown in figure 17. The bridge is so proportioned that only at one frequency is the phase shift through the bridge equal to 180°. Voltages of other frequencies are fed back to the amplifying device out of phase with the existing input signal, and are cancelled by being amplified out of phase.

The *Bridge-T oscillator* developed by the National Bureau of Standards consists of a two-stage amplifier having two feedback loops, as shown in figure 18. Loop 1 consists of a regenerative cathode-to-cathode loop, consisting of L_{p1} and C_3. The bulb regulates the positive feedback, and tends to stabilize the output of the oscillator, much as in the manner of the Wien circuit. Loop 2 consists of a grid-cathode degenerative circuit, containing the Bridge-T.

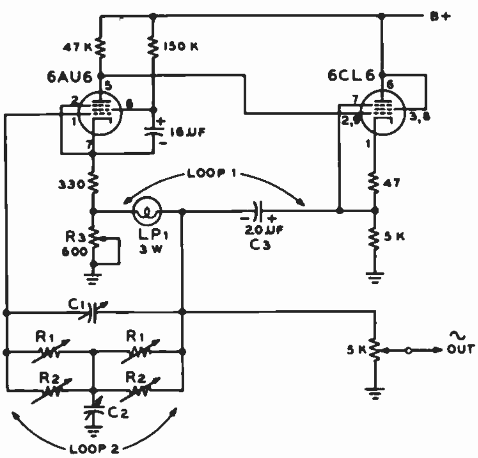


Figure 18

THE NBS BRIDGE-T OSCILLATOR CIRCUIT EMPLOYS TWO FEEDBACK LOOPS. LOOP 1 IS REGENERATIVE, LOOP 2 IS DEGENERATIVE

Oscillation will occur at the null frequency of the bridge, at which frequency the bridge allows minimum degeneration in loop 2 (figure 19).

8-7 Feedback

Feedback amplifiers have been discussed in Chapter 6, page 6-25 of this Handbook. A more general use of feedback is in automatic control and regulating systems.

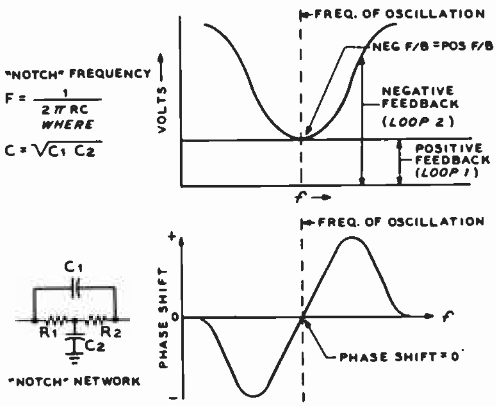


Figure 19

BRIDGE-T FEEDBACK LOOP CIRCUITS

Oscillation will occur at the null frequency of the bridge, at which frequency the bridge allows minimum degeneration in loop 2.

Mechanical feedback has been used for many years in such forms as engine-speed governors and servo steering engines on ships.

A simple feedback system for temperature control is shown in figure 20. This is a *cause-and-effect system*. The furnace (F) raises the room temperature (T) to a predetermined value at which point the sensing thermostat (TH) reduces the fuel flow to the furnace. When the room temperature drops below the predetermined value the fuel flow is increased by the thermostat control. An interdependent control system is created by this arrangement: the room temperature depends on the thermostat action, and the thermostat action depends on the room temperature. This sequence of events may be termed a *closed-loop feedback system*.

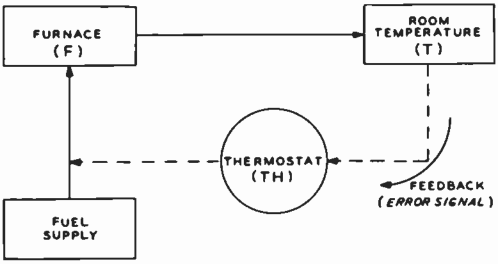


Figure 20

SIMPLE CLOSED-LOOP FEEDBACK SYSTEM

Room temperature (T) controls fuel supply to furnace (F) by feedback loop through thermostat (TH) control.

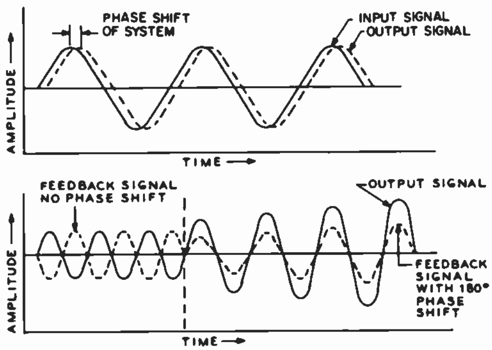


Figure 21

PHASE SHIFT OF ERROR SIGNAL MAY CAUSE OSCILLATION IN CLOSED LOOP SYSTEM

To prevent oscillation, the gain of the feedback loop must be less than unity when the phase shift of the system reaches 180 degrees.

Error Cancellation A feedback control system is dependent on a degree of error in the output signal, since this error component is used to bring about the correction. This component is called the *error signal*. The error, or deviation from the desired signal is passed through the feedback loop to cause an adjustment to reduce the value of the error signal. Care must be taken in the design of the feedback loop to reduce over-control tendencies wherein the correction signal would carry the system past the point of correct operation. Under certain circumstances the new error signal would cause the feedback control to overcorrect in the opposite direction, resulting in *hunting* or oscillation of the closed-loop system about the correct operating point.

Negative-feedback control would tend to damp out spurious system oscillation if it were not for the time lag or phase shift in the system. If the over-all phase shift is equal to one-half cycle of the operating frequency of the system, the feedback will maintain a steady state of oscillation when the circuit gain is sufficiently high (figure 21). In order to prevent oscillation, the gain figure of the feedback loop must be less than unity when the phase shift of the system reaches 180 degrees. In an ideal control system the gain of the loop would be constant throughout the operating range of the device, and would drop rapidly outside the range to reduce the bandwidth of the control system to a minimum.

The time lag in a closed-loop system may be reduced by using electronic circuits in place of mechanical devices, or by the use of special circuit elements having a *phase-lead* characteristic. Such devices make use of the properties of a capacitor, wherein the current leads the voltage applied to it.

CHAPTER NINE

Single-Sideband Transmission and Reception

Single-sideband (SSB) communication is a unique, sophisticated information transmission system well suited for wire and radio services. Although known in theory for several decades, "sideband" was sparingly used in commercial service for a number of years, and only in the last decade has it achieved popularity and general acceptance in the Amateur Service. Economical in cost, sparing of valuable spectrum space, and usable under the most trying propagation conditions, SSB is the stepping stone to a future era of better and more reliable rapid h-f communication.

9-1 The SSB System

Single sideband is a recent attempt to translate human intelligence into electrical impulses capable of being economically transmitted over great distances. The general flow of information in a communication system includes a *source*, followed by a *translator* which propagates the intelligence through a conducting *medium*. A second translator is used to extract the intelligence conveyed by the medium and to make it available in a usable form. The vocal chords, vibrations in the atmosphere, and the ear drum accomplish this sequence of events for sound; the light source, the "ether," and the human eye provide the same sequence for sight.

Experiments before the turn of the century proved the existence of electromagnetic waves which could be propagated and put to use for transmission of information. When voice transmission via radio waves was successfully accomplished *circa* 1907, the concept of carrier waves and sidebands was unknown, although it was understood that "a channel separation high compared with the pitch of the sound waves transmitted" was required. An implication that a *transmission band* of frequencies was involved was apparently not grasped at the time, and the idea that intelligence could be transmitted by a single carrier wave of constant frequency and varying amplitude persisted until about 1921 at which time the sideband concept had been established by a series of discoveries, experiments, and inventions.

Early SSB experiments with single-sideband transmission were conducted by the telephone industry which was interested in transmitting electrical impulses corresponding to the human voice over long-distance telephone circuits. Since the transmission properties of wire and cable deteriorate rapidly with cable length and increasing frequency, a means of frequency conservation was desired which would permit the "stacking" of different voices in an electromagnetic package so that many voices could be sent over a single circuit. The voice impulses were mainly concentrated in the band 300—3,000 Hz and the problem at hand was to translate this voice band to a higher

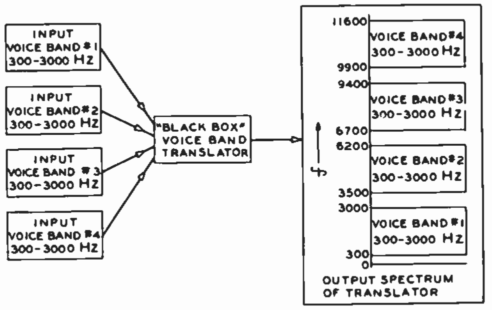


Figure 1

THE "BLACK BOX" VOICE BAND TRANSLATOR

A simple device for "stacking" voice bands in an electromagnetic "package" for transmitting many voices over a single circuit cannot be built as it is impossible to translate a band of frequencies directly to another band. Translation must be accomplished by an indirect method, making use of an auxiliary carrier wave and a mixing process termed "modulation."

band of frequencies (15,300—18,000 Hz, for example) for transmission on the telephone circuit, then to reverse the translation process at the receiving terminal to recover the original band of frequencies. Experiments proved, however, that a simple and economical apparatus for translation of the voice frequencies from one band to another was not forthcoming. No device could be built that would do the job that looked so simple when sketched on paper (figure 1). It proved possible, however, to generate a continuous electrical signal at some high frequency (15,000 Hz, for example) and to impress the voice impulses on this signal. For convenience, the continuous signal was termed the *carrier wave*, as it was assumed to "carry" the intelligence in some way or other. A suitable device at the receiving terminal detected the intelligence on the carrier, recovering the original speech frequencies impressed on the carrier at the transmitter. Mathematical analysis of this process (called *modulation*) showed that the carrier remained unchanged and additional frequencies were created lying on either side of the carrier, spaced from it by a frequency proportional to the modulation frequency (figure 2). These additional frequencies were termed *sidebands* and conclusive evidence of separate sidebands was achieved in 1915 by the use of electric filters that separated sidebands and carriers, proving their individuality.

The sideband theory was of little more than passing interest to radio engineers, but it was a matter of considerable importance to the telephone industry. The carrier wave was useless except as an operator necessary to generate and then upon which to "hang" the two sidebands, both of which carried the same information (figure 3). For economic reasons and spectrum conservation it was desirable to remove one sideband and the carrier from the translator, passing only one sideband through the conducting medium. At the receiver, a locally generated carrier wave of the correct frequency and amplitude was combined with the incoming single-sideband signal. The resulting output was a reproduction of the signal impressed on the translator. Commercial wire telephone systems using this technique were placed in operation in 1918 and the first h-f SSB telephone link was activated in 1927.

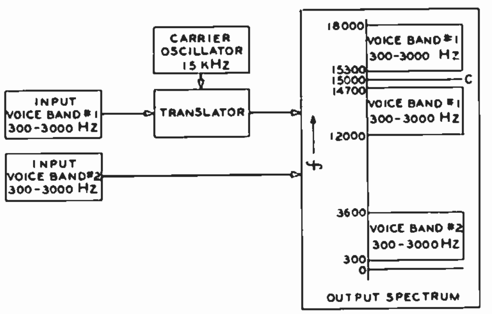


Figure 2

THE TRANSLATOR MIXER

Voice band #1 is impressed on a carrier signal in a translator (mixer) stage. Voice band #2 is unchanged. The output spectrum of the device shows that two voice bands are available, one "stacked" above the other in frequency. Addition of other translators will permit additional voice bands to be "stacked" in the frequency region between 3600 Hz and 12,000 Hz. The voice packages thus created could be sent over a single circuit. Note that the translation process creates two symmetrical voice bands from the original #1 signal, spaced each side of a carrier frequency between the bands. Elimination of carrier signal and one voice band would permit addition of another signal in this portion of the spectrum.

Practical Application of SSB The spectrum waste arising from a frequency translation process utilizing simple amplitude modulation could be eliminated by suppression of one sideband and the carrier, and the transmission of only the remaining sideband. To date, no method exists to directly generate an SSB signal. All translation techniques involve the use of a carrier wave, and the resulting signal includes the original carrier and two auxiliary sidebands. Elimination of the unwanted signal components was (and still is) a complex and sophisticated task.

The post-World War II acceptance of SSB transmission for military and commercial circuits has stimulated research and development in this field and has contributed to a heightened interest in the technique by the radio amateur. Mass production of sharp-cutoff filters and stable translation oscillators, plus the use of advanced and simplified circuitry has brought SSB close to the point of obsoleting simple amplitude-modulation transmission on the high-frequency amateur bands. Undoubtedly, in the years ahead, further design refinements and technical advances will make the use of SSB even more advantageous to all concerned with transmission of intelligence by electrical means.

The popularity of SSB for general amateur use has been brought about as this technique has consistently proved to allow more reliable communication over a greater range than has amplitude modulation. It has greater ability to pierce interference, static, and man-made noise than has amplitude

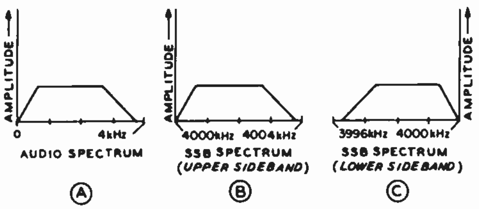


Figure 3

RELATIONSHIP OF AUDIO AND SSB SPECTRUMS

The single-sideband components are the same as the original audio components except that the frequency of each is raised by the frequency of the carrier. The relative amplitude of the various components remains the same.

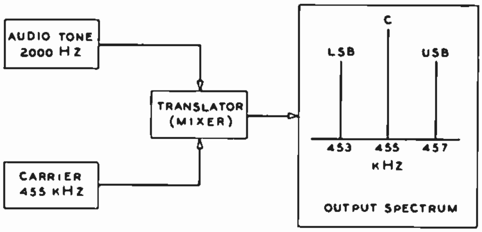


Figure 4

THE TRANSLATOR SPECTRUM

The SSB signal is an audio signal raised (mixed, or translated) to the desired radio frequency. A 455-kHz carrier signal upon which is impressed a 2-kHz audio tone in a translator stage will possess two sidebands, separated from the carrier frequency by the frequency of the tone. The carrier has been generated by the separate oscillator and the two adjacent signals (sidebands) are a product of the mixing process taking place between the audio signal and the carrier. The output spectrum pictured is of a double sideband, with carrier. To produce an SSB signal, it is necessary to eliminate the carrier and one sideband.

modulation and is inherently resistant to propagation abnormalities that render a.m. completely useless. In addition, the annoying interference caused by heterodynes between a-m carriers is completely missing in SSB service. Finally, the cost of high-power SSB equipment has dropped sufficiently to compete with a-m equipment of the same or greater power capability.

Single sideband is now well established in the field of amateur radio as more and more amateurs turn toward this natural means of communication as they discover for themselves the benefits and advantages SSB can offer.

Basic SSB A single-sideband signal can be best be described as an audio signal *raised* (or translated) to the desired radio frequency. The translation process may not result in the inversion of the audio-frequency components in the signal, depending on the sideband selected (figure 4). For example, a single audio tone of 2000 Hz is to be translated into an SSB signal in the 455-kHz region. The tone is amplified and applied to one input of a translator stage (usually termed a *balanced modulator*). A radio-frequency *carrier* is applied to the other input terminal of the modulator. For this example, the frequency of the carrier is 455 kHz. The translation process takes place in the balanced modulator; creating two *sidebands* positioned each side of the carrier, and separated from it by the modulation frequency. Thus, at least four signals are flowing within the modulator: the 2000-Hz (2-kHz) *audio signal,* the *lower sideband* (455 − 2 = 453 kHz), the *carrier* (455 kHz), and the *upper sideband* (455 + 2 = 457 kHz). The carrier, of course, has been generated by the separate local oscillator, and the two sidebands are a product of the mixing process taking place between the audio signal and the carrier.

The balanced modulator is usually designed to balance (or cancel) the carrier signal to a large degree, leaving only the two sidebands and the audio signal to appear in the output circuit. Some modulators also balance out the audio signal. Part of the job of creating an SSB signal has now been accomplished. The high-frequency components of the output signal of the balanced modulator comprise a *double-sideband, suppressed-carrier signal.* The remaining step to create an SSB signal is to eliminate one of the sidebands and to reduce to minor proportions any vestige of carrier permitted to pass through the balanced-modulator stage. A *sideband filter* accomplishes this last step. At the output of the filter is the desired SSB signal. The passband of the filter should be just wide enough to pass the intelligence without passing the carrier wave or the unwanted sideband. For voice communication, such filters usually pass a band of radio frequencies about 2 or 3 kHz wide.

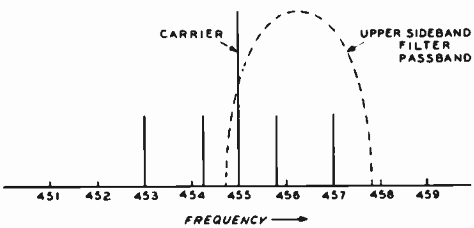


Figure 5

THE SSB SIGNAL

The SSB signal may be generated by passing a double-sideband-with-carrier signal through a filter which removes one sideband and partially suppresses the carrier. In this example, a two-tone audio signal (700 and 2000 Hz) is mixed with a 455-kHz carrier signal. The output signal from the mixer, or modulator, contains four sideband frequencies: 453 kHz, 454.3 kHz, 455.7 kHz, and 457 kHz, in addition to the carrier at 455 kHz. Additional carrier suppression may be obtained by the use of a balanced modulator.

The unwanted carrier and sideband that are eliminated by the filter and balanced modulator are actually absorbed by the filter and modulator and converted to heat. In order to hold the cost and size of the filter to a reasonable figure, it is necessary that the above process take place at a relatively low signal level, of the order of a volt or two, so that power dissipation is low.

The SSB Spectrum A single audio tone in a perfect SSB system remains a simple sine wave at all points in the system and cannot be distinguished from a c-w signal generated by more conventional means. A voice signal, on the other hand, is a complex band of audio components having many frequencies of varying amplitudes. A simple and useful compromise signal for testing SSB equipment is the *two-tone* signal, composed of two equal and separate sine waves separated a very small percentage in frequency. If two audio tones are applied to the input circuit of the SSB exciter previously discussed, the output of the 455-kHz balanced modulator will contain *four* sideband frequencies (figure 5). Assume the audio tones are 700 and 2000 Hz. The output frequencies of the balanced modulator will be: 453 kHz, 454.3 kHz, 455 kHz (the partially suppressed carrier), 455.7 kHz and 457 kHz. The two lower frequencies represent the lower sideband, and the two higher frequencies represent the upper sideband. With a properly designed filter following the balanced modulator, both the frequencies in one sideband and the remainder of the carrier will be almost completely eliminated. If the filter completely eliminates the lower sideband and the carrier, the output of the exciter will be two radio frequencies at 455.7 kHz and 457 kHz. An observer examining these r-f signals could not tell if the signals were generated by two oscillators operating at the observed frequencies, or if the two signals were the result of two audio tones applied to an SSB exciter.

The waveform of the SSB signal changes drastically as the number of audio tones is increased, as shown in figure 6. A single-tone waveform is shown in illustration A and is simply a single, steady sine-wave r-f output. A signal composed of two audio tones is shown in illustration B. The two radio-frequency signals are separated by the difference in frequency between the audio tones and beat together to give the SSB envelope shown. The figure has the shape of half-sine waves, and from one null to the next represents one full cycle of the difference frequency. If one tone has twice the amplitude of the other, the envelope shape is as shown in illustration C. The SSB envelope of three equal tones of equal frequency spacings and at one particular phase relationship is shown in illustration D. Illustration E shows the SSB envelope of four equal tones having equal frequency spacings and at one particular phase relationship. Finally, illustration F shows the SSB envelope of a square wave having an infinite number of odd harmonics. A pure square wave requires infinite bandwidth, so in theory the SSB envelope requires infinite amplitude. This emphasizes the point that the SSB envelope shape may not be the same as the original audio waveshape, and usually bears no similarity to it. This is because the percentage difference between the radio frequencies is small, even though one audio tone may be several times the other in terms of frequency. Because of nonlinearity and phase shift in the practical SSB transmitter, the peak amplitude of a transmitted square wave is not so great as predicted by theory through the addition of the harmonic coefficients, making it impossible to faithfully reproduce a square wave. Speech processing in the form of heavy audio *clipping* therefore is of limited value in SSB because the SSB r-f envelopes are so different from the audio envelopes. A heavily clipped wave approaches a square wave which will have the tendency to exhibit the high amplitude peaks shown in illustration 6F, a waveform the SSB transmitter is theoretically unable to transmit.

The Received SSB Signal In summary, if an *audio spectrum* containing many different tones (the human voice, for example) is applied to the SSB exciter, an *r-f spectrum* is generated that corres-

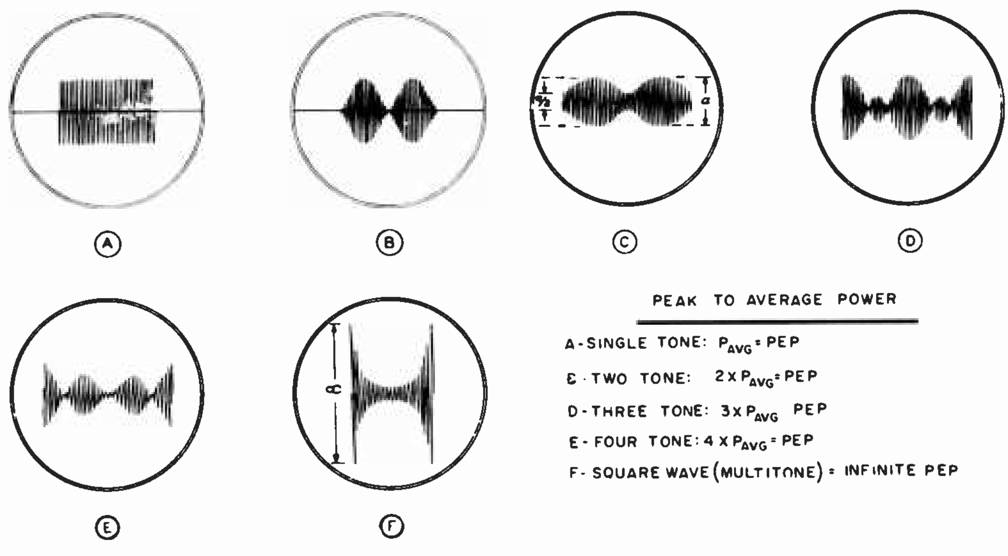


Figure 6

SSB WAVEFORMS

The waveform of the SSB signal changes with the nature of the modulating signal, and the envelope shape of the SSB wave may not be the same as the original audio waveshape. The peak power in the SSB wave is a direct function of the r-f waveform, as shown here Peak and average power in the SSB wave will be discussed later in this chapter.

ponds to the audio tones. If the audio spectrum encompasses the range of 300—3000 Hz, the output of the 455-kHz balanced modulator will be 452 to 454.7 kHz (the lower sideband), 455 kHz (the partially suppressed carrier), and 455.3 to 458 kHz (the upper sideband). An "upper-sideband" type filter having a passband of 455.3 to 458 kHz will substantially eliminate the residual carrier and lower sideband.

Listening to the output of the SSB exciter on a typical a-m receiver will divulge a series of unintelligible sounds having no apparent relation to the original speech impressed on the SSB exciter. (A low-pitched voice can be read with difficulty as the syllabic content is preserved and is apparent). Injection in the receiver of a local carrier frequency of 455 kHz (corresponding to the suppressed carrier eliminated in the exciter) will produce intelligible speech that is a replica of the original voice frequencies.

In order to transmit simple double sideband with carrier (amplitude modulation) with this SSB exciter, it is only necessary to bypass the sideband filter and unbalance the balanced modulator. The resulting a-m signal with carrier may be intelligible on the ordinary receiver without the necessity of local-oscillator injection, the latter function being fulfilled by the transmitted carrier, if it has sufficient strength relative to the sidebands.

SSB Power Rating The SSB transmitter is usually rated at *peak envelope input or output power.* Peak envelope power (PEP) is the root-mean-square (rms) power generated at the peak of the modulation envelope. With either a two-equal-tone test signal or a single-tone test signal, the following equations approximate the relationships between single-tone and two-tone meter readings, peak envelope power, and average power for class-B or class-AB linear amplifier operation:

Single tone:

D-C Plate Current (Meter Reading):

$$I_b = \frac{i_{pm}}{\pi}$$

Plate Input (Watts):

$$P_{in} = \frac{i_{pm} \times E_b}{\pi}$$

Average Output Watts and PEP:

$$P_o = \frac{i_{pm} \times e_p}{4}$$

Plate Efficiency:

$$N_p = \frac{\pi \times e_p}{4 \times E_b}$$

Two equal tones:

D-C Plate Current (Meter Reading):

$$I_b = \frac{2 \times i_{pm}}{\pi^2}$$

Plate Input (Watts):

$$P_{in} = \frac{2 \times i_{pm} \times E_b}{\pi^2}$$

Average Output Watts:

$$P_o = \frac{i_{pm} \times e_p}{8}$$

PEP Output Watts:

$$P_o = \frac{i_{pm} \times e_p}{4}$$

Plate Efficiency:

$$N_p = \left(\frac{\pi}{4}\right)^2 \times \frac{e_p}{E_b}$$

where,

i_{pm} equals peak of the plate-current pulse,
e_p equals peak value of plate-voltage swing,
π equals 3.14,
E_b equals d-c plate voltage,
N_p equals efficiency in percent.

"Average" Speech Section 97.67 of the Amateur Radio Service Rules of the FCC indicates that the average power input of an SSB transmitter in the amateur service shall not exceed one kilowatt on modulation peaks, as indicated by a plate-current meter having a time constant of not more than 0.25 second. It is common practice among amateurs to define this as equivalent to a *peak envelope power* input of two kilowatts. This is convenient, since a two-tone test signal having a peak-to-average power ratio of two to one can thereby be employed for tuneup and adjustment purposes with the reasonable assumption that the SSB equipment will be properly adjusted for one kilowatt average power voice operation.

It is difficult to determine the ratio of peak to average power in the human voice, as the range of intensity of speech sounds may vary as much as 40 decibels. "Average" speech seems to have an intensity range of about 20 decibels and a ratio of instantaneous peak-to-average power of about 14 decibels for 99 percent of the time of speech. Speech processing (clipping or compression) may alter this figure, bringing the peak to average power ratio closer to unity. In any event, adjustment of the amateur SSB transmitter to achieve a peak power input of twice the average power input level has proven by experience to allow sufficient peak-power capability to cover the majority of cases. In those situations where the peak capability of the equipment is exceeded at an average-power input level of one kilowatt, the average-power level must be reduced to conform with the maximum capability of the transmitter. In any case, the use of an oscilloscope is mandatory to determine the peak-power capability of an SSB transmitter.

Power Advantage of SSB over AM Single sideband is a very efficient form of voice communication by radio. The amount of radio-frequency spectrum occupied can be no greater than the frequency range of the audio or speech signal transmitted, whereas other forms of radio transmission require from two to several times as much spectrum space. The r-f power in the transmitted SSB signal is directly proportional to the power in the original audio signal and no strong carrier is transmitted. Except for a weak pilot carrier present in some commercial usage, there is no r-f output when there is no audio input.

The power output rating of an SSB transmitter is given in terms of *peak envelope power* (PEP). This may be defined as the rms power at the crest of the modulation envelope. The peak envelope power of a conventional amplitude-modulated signal at 100% modulation is four times the carrier power. The average power input to an SSB transmitter is therefore a very small fraction of the power input to a conventional amplitude-modulated transmitter of the same power rating.

Single sideband is well suited for long-range communications because of its spectrum and power economy and because it is less susceptible to the effects of selective fading and interference than amplitude modulation. The principal advantages of SSB arise from the elimination of the high-energy carrier and from further reduction in sideband power permitted by the improved performance of SSB under unfavorable propagation conditions.

In the presence of narrow-band manmade interference, the narrower bandwidth of SSB reduces the probability of destructive interference. A statistical study of the distribution of signals on the air versus the signal strength shows that the probability of successful communication will be the same if the SSB power is equal to one-half the power of one of the two a-m sidebands. Thus SSB can give from 0 to 9 db improvement under various conditions when the *total* sideband power is equal in SSB and regular amplitude

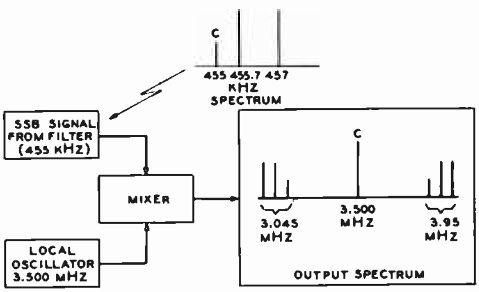


Figure 7

SSB FREQUENCY TRANSLATION

The SSB signal may be translated higher in frequency in the same manner the voice signals are translated to a radio-frequency spectrum. In this example, the 455-kHz two-tone, suppressed-carrier signal is translated (mixed) with a 3.5-MHz oscillator to provide two new sidebands, one at 3.045 MHz and the other at 3.95 MHz. If the 3.95-MHz signal is desired, filter circuits may be used to eliminate the unwanted 3.045-MHz sideband and the 3.5-MHz carrier signal from the local oscillator. The 3.95-MHz signal may now be shifted in frequency by changing the frequency of the local oscillator.

modulation. In general, it may be assumed that 3 db of the possible 9 db advantage will be realized on the average contact. In this case, the SSB power required for equivalent performance is equal to the power in one of the a-m sidebands. For example, this would rate a 100-watt SSB and a 400-watt (carrier) a-m transmitter as having equal performance. It should be noted that in this comparison it is assumed that the receiver bandwidth is just sufficient to accept the transmitted intelligence in each case.

To help evaluate other methods of comparison the following points should be considered. In conventional amplitude modulation two sidebands are transmitted, each having a peak envelope power equal to ¼ carrier power. For example, a 100-watt a-m signal will have 25-watt peak envelope power in each sideband, or a total of 50 watts. When the receiver detects this signal, the voltages of the two sidebands are added in the detector. Thus the detector output voltage is equivalent to that of a 100-watt SSB signal. This method of comparison says that a 100-watt SSB transmitter is just equivalent to a 100-watt a-m transmitter. This assumption is valid only when the receiver bandwidth used for SSB is the same as that required for amplitude modulation (e.g., 6 kHz), when there is no noise or interference other than broadband noise, and if the a-m signal is not degraded by propagation. By using half the bandwidth for SSB reception (e.g., 3 kHz) the noise is reduced 3 db so the 100-watt SSB signal becomes equivalent to a 200-watt carrier a-m signal. It is also possible for the a-m signal to be degraded another 3 db on the average due to narrow-band interference and poor propagation conditions, giving a possible 4 to 1 power advantage to the SSB signal.

It should be noted that 3 db signal-to-noise ratio is lost when receiving only one sideband of an a-m signal. The narrower receiving bandwidth reduces the noise by 3 db but the 6 db advantage of coherent detection is lost, leaving a net loss of 3 db. Poor propagation will degrade this "one-sideband" reception of an a-m signal less than double-sideband reception, however. Also under severe narrow-band interference conditions (e.g., an adjacent strong signal) the ability to reject all interference on one side of the carrier is a great advantage.

Advantage of SSB with Selective Fading On long-distance communication circuits using amplitude modulation, selective fading often causes severe distortion and at times makes the signal unintelligible. When one sideband is weaker than the other, distortion results; but when the carrier becomes weak and the sidebands are strong, the distortion is extremely severe and the signal may sound like "monkey chatter." This is because a carrier of at least twice the amplitude of either sideband is necessary to demodulate the signal properly. This can be overcome by using *exalted-carrier reception* in which the carrier is amplified separately and then reinserted before the signal is demodulated or detected. This is a great help, but the reinserted carrier must be very close to the same phase as the original carrier. For example, if the reinserted carrier were 90 degrees from the original source, the a-m signal would be converted to phase modulation and the usual a-m detector would deliver no output.

The phase of the reinserted carrier is of no importance in SSB reception and by using a strong reinserted carrier, exalted-carrier reception is in effect realized. Selective fading with one sideband simply changes the amplitude and the frequency response of the system and very seldom causes the signal to become unintelligible. Thus the receiving techniques used with SSB are those which inherently greatly minimize distortion due to selective fading.

SSB Amplification and Frequency Changing The single-sideband signal appearing at the output of the filter must be amplified to a sufficiently strong level for practical use. The amplifying stage must have low distortion and the output signal must be a faithful replica of the input signal. An amplifier meeting these requirements is called a *linear amplifier*. Any deviation from amplitude linearity produces signal distortion and spurious products which rapidly degrade the SSB signal. It is therefore impossible to pass the SSB signal through frequency doublers or class-C amplifiers without creating severe distortion, because these are inherently nonlinear devices. Linear amplifier stages must be used, and if a change of frequency of the

SSB signal is desired, it must be heterodyned to the new frequency by means of a mixer stage and another local oscillator (figure 7). The resulting signal may be vfo controlled by varying the frequency of the local oscillator, but the frequency at which the SSB signal is generated is held constant. Thus by means of linear amplifiers and mixer stages, a low frequency SSB signal may be amplified and converted to any other frequency desirable for communication purposes.

9-2 A Basic Single-Sideband Transmitter

The general outline of a practical SSB transmitter suitable for high-frequency operation can be assembled from the preceding information. A block diagram of such a unit is shown in figure 8. The transmitter consists of a speech amplifier, a carrier oscillator, a balanced modulator, a sideband filter, a high-frequency mixer stage and conversion oscillator, and a linear amplifier having a high-Q tuned output circuit. Incidental equipment such as power supplies and metering circuits are also necessary. Many variations of this basic block diagram are possible.

The Speech Amplifier—A typical speech amplifier consists of a microphone which converts the voice into electrical signals in the audio band, followed by one or more stages of voltage amplification. No appreciable audio power output is required making the audio system of the SSB transmitter quite different from that of the usual a-m transmitter, which requires an audio power level equal to one-half the class C amplifier power input. Included in the speech system is a *speech level* (audio volume) *control* and additional stages to allow *automatic voice operation* (VOX) of the equipment.

The Carrier Oscillator—A highly stable *r-f oscillator* (often crystal-controlled) is used to generate the carrier signal required in the mixing process. The choice of carrier frequency is determined by the design of the sideband filter, and frequencies in the range of 250 kHz to 20 MHz are common. Power output is low and frequency stability is a prime necessity in this circuit.

The Balanced Modulator—The balanced modulator translates the audio frequencies supplied by the speech amplifier into r-f sidebands adjacent to the carrier generated by the carrier oscillator. In addition, the balanced modulator partially rejects the carrier which has no further use after the mixing process is completed. A *carrier-balance (null) control* is an integral part of this circuit and is adjusted for optimum carrier suppression.

The Sideband Filter—Selection of one of the two sidebands at the output of the balanced modulator is the function of the filter. A practical filter may consist of small

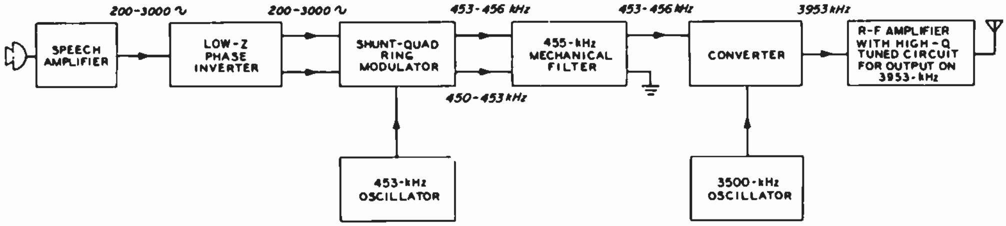


Figure 8

BLOCK DIAGRAM OF FILTER-TYPE SSB TRANSMITTER

Voice frequencies in the range of 200 to 3000 Hz are amplified and fed to a balanced modulator. Depending on the choice of frequency of the local oscillator, either the upper or lower sideband may be passed through to the mechanical filter. The carrier has, to some extent, been reduced by the balanced modulator. Additional carrier rejection is afforded by the filter. The SSB signal at the output of the filter is translated directly to a higher operating frequency. Suitable tuned circuits follow the converter stage to eliminate the conversion oscillator signal and the image signal.

tuned LC circuits, or it may consist of mechanical resonators made of quartz or steel. A representative passband for a sideband filter is shown in figure 9. The filter must provide a sharp cutoff between the wanted sideband and the carrier, as well as rejection of the unwanted sideband.

The Converter (Mixer) Stage and Conversion Oscillator—It is usually necessary to obtain an SSB signal at a frequency other than that of the sideband filter passband. Frequency conversion is accomplished in the same manner the voice frequencies were translated to the filter frequency region; that is, by the use of a converter stage and conversion oscillator. The process carried out in this step may be referred to as *translation*, *mixing*, *heterodyning*, or *converting*. For this example, it is desired to convert a 455-kHz SSB signal to 3.95 MHz. The operation takes place in a second balanced-modulator circuit. One input is the 455-kHz SSB signal, and the other input signal is from an oscillator operating on 3.500 MHz. The output of the second mixer is a partially suppressed carrier (3.500 MHz), the lower sideband in the 3.045-MHz range (3.500 − 0.455 = 3.045 MHz), and the upper sideband in the 3.95-MHz range (3.500 + 0.455 = 3.95 MHz).

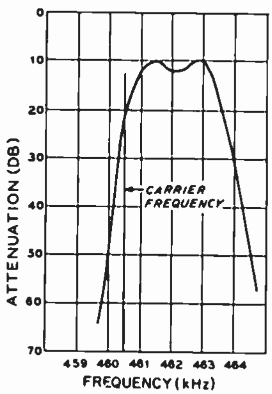


Figure 9

PASSBAND OF CRYSTAL LATTICE FILTER

A 460.5-kHz crystal-lattice filter composed of eight crystals has an excellent passband for voice waveforms. Carrier rejection is about −20 decibels, and unwanted sideband rejection is better than −35 decibels. Passband is essentially flat up to 463 kHz, providing an audio passband of about 300 to 2500 Hz.

The upper sideband is the desired one, so a simple auxiliary *image filter* is used to separate it from the unwanted sideband and the partially suppressed carrier. In most cases, this filter consists of the two or three parallel-tuned circuits normally associated with the following amplifier stages tuned to 3.95 MHz.

The Linear Amplifier—The output of the last mixer stage is usually of the order of a few milliwatts and must be amplified to a usable level in one or more *linear amplifier* stages. For lowest distortion, the output of the linear amplifier should be a nearly exact reproduction of its input signal. Any amplitude nonlinearity in the amplifier not only will produce undesirable distortion within the SSB signal, but will also produce annoying spurious products in adjacent channels. Distortion may be held to a low value by the proper choice of tubes, their operating voltages and driving-circuit considerations, and by the use of external negative feedback, as discussed in Chapter Twelve.

9-3 Selective Tuned Circuits

The selectivity requirements of the tuned circuits following a mixer stage often become quite severe. For example, using an input signal at 250 kHz and a conversion injection frequency of 4000 kHz the desired output may be 4250 kHz. Passing the 4250-kHz signal and the associated sidebands without attenuation and realizing 100 db of attenuation at 4000 kHz (which is only 250 kHz away) is a practical example. Adding the requirement that this selective circuit must tune from 2250 to 4250 kHz further complicates the basic requirement. The best solution is to cascade a number of tuned circuits. Since a large number of such circuits may be required, the most practical solution is to use permeability tuning, with the circuits tracked together. An example of such circuitry is found in the *Collins* 32S sideband transmitter.

If an amplifier tube is placed between each tuned circuit, the over-all response will be the sum of one stage multipled by the number of stages (assuming identical tuned circuits). Figure 10 is a chart which may be used to determine the number of tuned circuits required for a certain degree of attenuation at some nearby frequency. The Q of the circuits is assumed to be 50, which is normally realized in small permeability-tuned coils. The number of tuned circuits with a Q of 50 required for providing 100 db of attenuation at 4000 kHz while passing 4250 kHz may be found as follows:

$$\Delta f \text{ is } 4250 - 4000 = 250 \text{ kHz}$$

where,

f_r is the resonant frequency (4250 kHz),

and,

$$\frac{\Delta f}{f_r} = \frac{250}{4250} = 0.059$$

The point on the chart where .059 intersects 100 db is between the curves for 6 and 7 tuned circuits, so 7 tuned circuits are required.

Another point which must be considered in practice is the tuning and tracking error of the circuits. For example, if the circuits were actually tuned to 4220 kHz instead of 4250 kHz, the $\frac{\Delta f}{f_r}$ would be $\frac{220}{4220}$ or 0.0522. Checking the curves shows that 7 circuits would just barely provide 100 db of attenuation. This illustrates the need for very accurate tuning and tracking in circuits having high attenuation properties.

Coupled Tuned Circuits When as many as 7 tuned circuits are required for proper attenuation, it is not necessary to have the gain that 6 isolating amplifier tubes would provide. Several vacuum tubes can be eliminated by using two or three coupled circuits between the amplifiers. With a coefficient of coupling between circuits 0.5 of critical coupling, the over-all response is very nearly the same as isolated circuits. The gain through a pair of circuits having 0.5 coupling is only eight-tenths that of two critically coupled circuits, however. If critical coupling is used between two tuned circuits, the nose of the response curve is broadened and about 6 db is lost on the skirts of each pair of critically coupled circuits. In some cases it may be necessary to broaden the nose of the response curve to avoid adversely affecting the frequency response of the desired passband. Another tuned circuit may be required to make up for the loss of attenuation on the skirts of critically coupled circuits. In some cases it may be necessary to broaden the nose of the response curve to avoid adversely affecting the frequency response of the desired passband. Another tuned circuit may be required to make up for the loss of attenuation on the critically coupled circuits.

Frequency-Conversion Problems The example in the previous section shows the difficult selectivity problem encountered when strong undesired signals appear near the desired frequency. A high-frequency SSB transmitter may be required to operate at any carrier frequency in the range of 1.7 to 30 MHz. The problem is to find a practical and economical means of heterodyning the generated SSB frequency to any carrier frequency in this range. There are many modulation products in the output of the mixer and a frequency scheme must be found that will not have undesired output of appreciable amplitude at or near the desired signal. When tuning across a frequency range some products may "cross over" the desired frequency. These undesired crossover frequencies should be at least 60 db below the desired signal to meet modern standards. The amplitude of the undesired products depends on the particular characteristics of the mixer and the particular order of the product. In general, most products of the 7th order and higher will be at least 60 db down. Thus any crossover frequency lower than the 7th order must be avoided since there is no way of attenuating them if they appear within the desired passband. The book *Single Sideband Principles and Circuits* by Pappenfus, McGraw Hill Book Co., Inc., N. Y., covers the subject of spurious products and incorporates a "mix selector" chart that is useful in determining spurious products for various different mixing schemes.

In general, for most applications when the intelligence-bearing frequency is lower than the conversion frequency, it is desirable that the ratio of the two frequencies be between

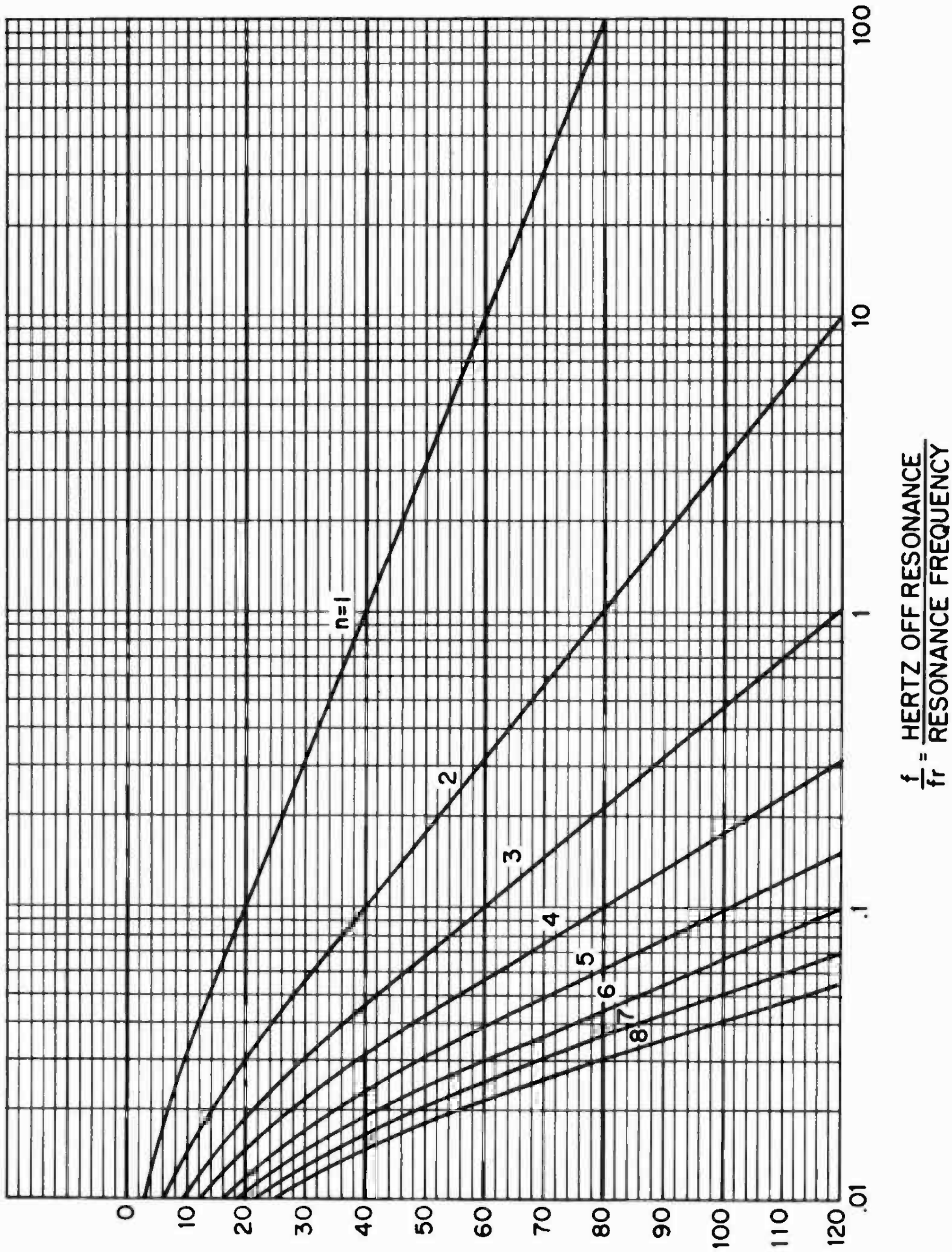


Figure 10

RESPONSE OF "N" NUMBER OF TUNED CIRCUITS, ASSUMING EACH CIRCUIT Q IS 50

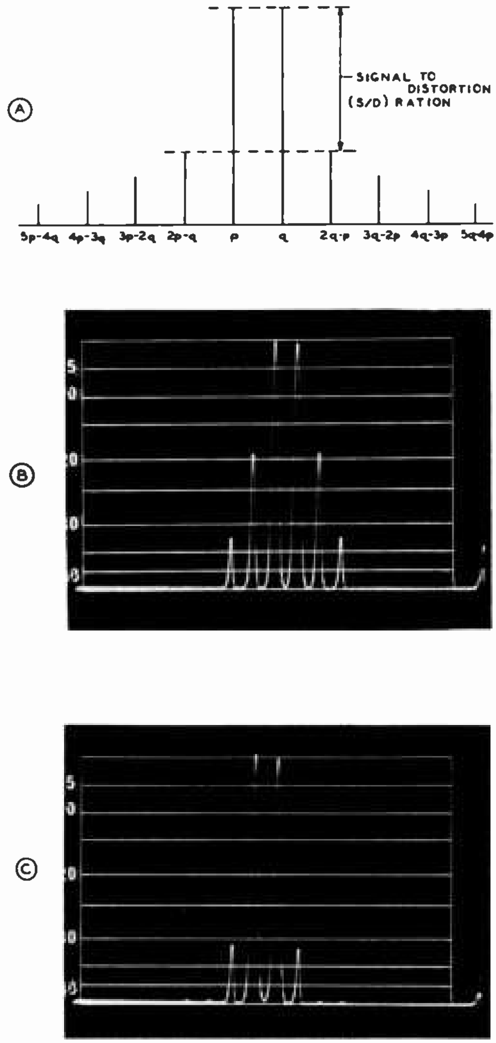


Figure 11

A shows SSB distortion products pictured up to ninth order. B shows SSB distortion products as seen on a panoramic analyzer. Third-order products are 19 decibels below two-tone test signal and fifth-order products are 32 decibels below the test signal. C illustrates that third-order products are about 31 decibels below test signal and higher-order products are better than 40 decibels down from test signal.

5 to 1 and 10 to 1. This a compromise between avoiding low-order harmonics of this signal input appearing in the output, and minimizing the selectivity requirements of the circuits following the mixer stage.

9-4 Distortion Products Due to Nonlinearity of R-F Amplifiers

When the SSB envelope of a *voice* or *multi-tone* signal is distorted, a great many new frequencies are generated. These represent all of the possible combinations of the sum and difference frequencies of all harmonics of the original frequencies. For purposes of test and analysis, a *two-tone* test signal (two equal-amplitude tones) is used as the SSB source. Since the SSB radio-frequency amplifiers use tank circuits, all distortion products are filtered out except those which lie close to the desired frequencies. These are all odd-order products; third order, fifth order, etc. The third-order products are $2p-q$ and $2q-p$ where p and q represent the two SSB r-f tone frequencies. The fifth order products are $3p-2q$ and $3q-2p$. These and some higher order products are shown in figure 11 A, B, and C. It should be noted that the frequency spacings are always equal to the difference frequency of the two original tones. Thus when an SSB amplifier is badly overloaded, these spurious frequencies can extend far outside the original channel width and cause an unintelligible "splatter" type of interference in adjacent channels. This is usually of far more importance than the distortion of the original tones with regard to intelligibility or fidelity. To avoid interference in another channel, these distortion products should be down at least 30 db below the adjacent channel signal. Using a two-tone test, the distortion is given as the ratio of the amplitude of one test tone to the amplitude of a third-order product. This is called the *signal-to-distortion ratio* (S/D) and is usually given in decibels. The use of

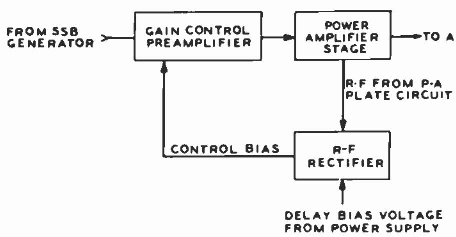


Figure 12

BLOCK DIAGRAM OF AUTOMATIC LOAD CONTROL (A.L.C.) SYSTEM

feedback r-f amplifiers make S/D ratios of greater than 40 db possible and practical.

Vacuum-Tube Nonlinearity Distortion products caused by amplifier departure from a linear condition are termed *intermodulation products* and the distortion is termed *intermodulation distortion.* This distortion can be caused by nonlinearity of amplifier gain or phase shift with respect to input level, and only appears when a multitone signal is used to drive the linear amplifier. This is the case for a voice signal which is composed of many tones, and intermodulation distortion will show up as a "gravelly" tone on the voice and will create interference to signals on adjacent channels. The main source of intermodulation distortion in a linear amplifier is the vacuum tube or transistor as these components have inherently nonlinear characteristics. Maximum linearity may be achieved by proper choice of tube or transistor and their operating conditions.

A practical test of linearity is to employ a two-tone, low-distortion signal to drive the tube or transistor and to use a spectrum analyzer to display a sample of the output spectrum on an oscilloscope (figure 11). The test signal, along with spurious intermodulation products may be seen on the screen, separated on the horizontal axis by the difference in frequency between the two tones. A reading is made by comparing the amplitude of a specific intermodulation product with the amplitude of the test signal. For convenience, the ratio between one of the test signals and one of the intermodulation products is read as a power ratio expressed in decibels below the test signal level. Measurements made on a number of power tubes have shown typical intermodulation distortion levels in the range of −20 to −40 decibels below one tone of a two-tone test signal.

The present state of the art in commercial and military SSB equipment calls for third-order intermodulation products better than −40 to −60 decibels below one tone of a two-tone test signal. Amateur requirements are less strict, running as low as −20 decibels, and may be justified on an economic basis since signal distortion, at least to the listener, is a highly subjective thing. To date, the use of inexpensive TV-type sweep tubes as linear amplifiers in amateur SSB gear has been acceptable, regardless of the rather high level of distortion inherent in these tube types.

9-5 Speech Processing

Several means may be used to keep the amplitude of distortion products down to acceptable levels and yet provide a high average degree of modulation. One method is to design the amplifier for excellent linearity over the expected amplitude or power range. A second method is to employ *audio processing* to insure that high amplitude peaks are suitably restricted before they cause trouble. The third method is to limit the amplitude of the SSB envelope by employing an r-f driven source of processing.

It should be noted that the r-f waveshapes of the SSB signal are always sine waves because the tank circuits make them so. It is the change in gain or phase with signal level in an amplifier that distorts the SSB envelope and generates unwanted distortion products. A processing system may be used to limit the input signal to an amplifier to prevent a change in gain level caused by excessive input level.

The processing system is adjusted so the power amplifier is operating near its maximum power capability and at the same time is protected from being over-driven.

In amplitude-modulated systems it is common to use speech compressors and speech clipping systems to perform this function. These methods are not equally useful in SSB. The reason for this is that the SSB envelope is different from the audio envelope and the SSB peaks do not necessarily correspond with the audio peaks as explained earlier in this chapter. For this reason an r-f processor of some sort located between the SSB generator and the power amplifier is most effective because it is controlled by SSB envelope peaks rather than audio peaks.

Speech Processing Techniques Because of the relatively high peak-to-average ratio of the human voice, it is not the most effective waveform for maximum utilization of SSB equipment (figure 13). The "talk power" or effectivity of

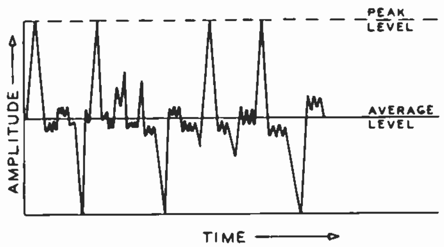


Figure 13

TYPICAL VOICE WAVEFORM

High peak-to-average power ratio of the human voice may be modified by distorting the waveform. The technique which provides the greatest increase in average power with the least amount of distortion will provide the greatest intelligibility of the received signal.

the average voice does not take advantage of the maximum capability of the SSB transmitter as well as it could if the peak-to-average ratio were not so high.

The peak-to-average ratio of voice waveforms may be modified by distorting the signal and the most effective means of increasing talk power is that technique which provides the greatest increase in average power with the least amount of distortion. Increasing the average power by boosting microphone gain causes linear amplifier limiting (*flat-topping*) to occur with consequent broadening of the transmitter bandwidth in the form of distortion products caused by nonlinear operation.

Speech processing may be judged on a basis of enhancing speech *intelligibility*, expressed as the signal-to-noise ratio at the receiver in terms of the average sideband power at the receiver compared to the noise power at the receiver (the sum of the distortion products of the speech processing system and the average noise power from other sources). A practical evaluation of processing techniques is to observe the improvement in speech intelligibility in the presence of noise having the same peak power as the unprocessed signal. Thus, the signal-to-noise ratio with processing may be compared to the signal-to-noise ratio of the unprocessed signal.

Several techniques are available to increase the average power in the modulation envelope while effectively limiting the envelope at the peak power capability point. These techniques include *a-f clipping, r-f envelope clipping, a-f compression,* and *r-f compression.*

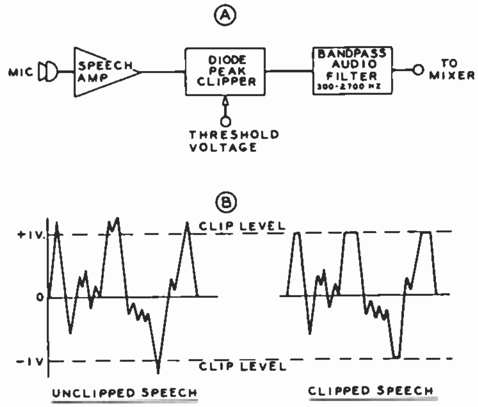


Figure 14

BLOCK DIAGRAM OF AUDIO PEAK CLIPPER

Diode-gate clipper (A) limits amplitude of positive and negative peaks. Clipping level (B) may be adjusted by varying threshold-control voltage. Clipper is followed by bandpass filter which limits audio signal to less than 3 kHz bandwidth for typical voice waveforms.

A-F Clipping Simple audio peak clippers, or limiters, may be of the diode-gate type discussed in Chapter 8, Section 1 (figure 1). The clipped signal is bandwidth limited to frequencies between approximately 300 to 2700 Hz by an audio filter (figure 14). Harmonics up to the ninth are present in the clipped and filtered wave and such a waveform has a peak-to-average ratio of about 4.3 decibels. The addition of 15 decibels of clipping to such a waveform provides an increase in speech *intelligibility* of about 4 decibels.

R-F Envelope Clipping An r-f peak clipper may be placed in the i-f portion of the SSB transmitter (figure 15). It is followed by an r-f filter to remove the r-f harmonics and out-of-band intermodulation products. With 15 decibels of clipping, an increase in speech *intelligibility* of nearly 8 decibels may be achieved. Generally speaking, the distortion produced by

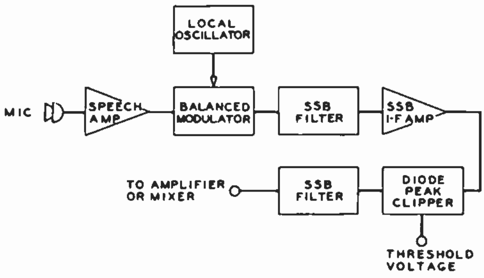


Figure 15

BLOCK DIAGRAM OF R-F ENVELOPE CLIPPER

An r-f clipper may be placed in the i-f portion of the SSB transmitter to limit amplitude of SSB signal. The clipper is followed by an r-f filter to remove harmonics and out-of-band products caused by clipping action. Clipping level is controlled by threshold voltage.

r-f envelope clipping is less objectionable than that caused by an equivalent amount of a-f clipping.

A-F Compression An audio-derived compressor is a form of automatic variable-gain-control amplifier whose output bears some consistent relationship to the input and which is controlled by a feedback loop which samples the output of the compressor (figure 16). The sample signal is rectified and the resulting control voltage is applied to a preceding gain-controlled stage. The time constants of this form of circuit are slow in order to prevent oscillation and distortion. Typically, *attack time* is in the region of 10 milliseconds and *release time* is in the order of 300 milliseconds, or more. A compression level as high as 15 decibels may be used, but the increase in speech intelligibility is minor unless sophisticated circuits having compression capability of up to 40 decibels and incorporating a fast time constant are used.

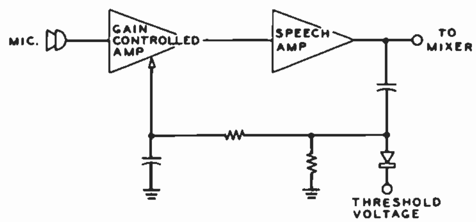


Figure 16

BLOCK DIAGRAM OF A-F COMPRESSOR

The audio-derived signal compressor is a form of automatic gain control. Control signal is taken from output of the compressor, is rectified, and the resulting control voltage is fed back to a low-level gain-controlled stage. Time constants of the control circuit are chosen in order to prevent oscillation and distortion.

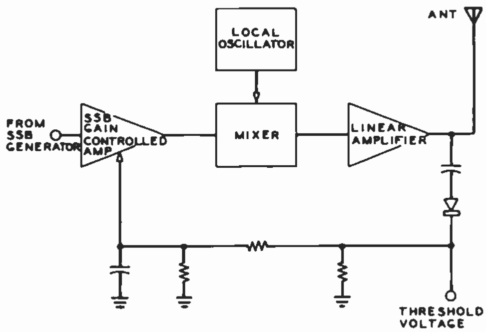


Figure 17

BLOCK DIAGRAM OF R-F COMPRESSOR

R-f compression (automatic load control) is similar to automatic gain control circuit of a receiver. Control voltage is obtained from rectified output signal of final linear amplifier stage and is applied to low level gain-controlled stage. Threshold bias is set so that no gain reduction takes place until output signal is nearly up to the maximum linear signal capability of the amplifier.

R-F Compression R-f compression (often termed *automatic load control*, or *ALC*) may take the form shown in figure 17. Operation is very similar to the i-f stage of a receiver having automatic gain control. Control voltage is obtained from the amplifier output circuit and a large delay (threshold) bias is used so that no gain reduction takes place until the output signal is nearly up to the maximum linear signal capability of the amplifier. At this level, the rectified output signal overcomes the delay bias and the gain of the preamplifier is reduced rapidly with increasing signal level. Peak r-f compression levels of up to 15 decibels are commonly used in SSB service, providing an increase in average-to-peak power of up to 5 decibels. Speech *intelligibility* may be improved only by about one decibel by such a technique.

A Comparison of Processing Techniques Outboard speech-processing adapters incorporated into existing equipment are becoming quite popular, but should be viewed with caution, since the equipment in question may have inherent limitations that preclude the use of a driving signal having a high average-to-peak ratio. Excessive dissipation levels may be reached in amplifier tubes, or low-level stages may be overloaded by the intemperate use of speech processing equipment. In any case, the output spectrum of the transmitter should be carefully examined for out-of-passband emissions.

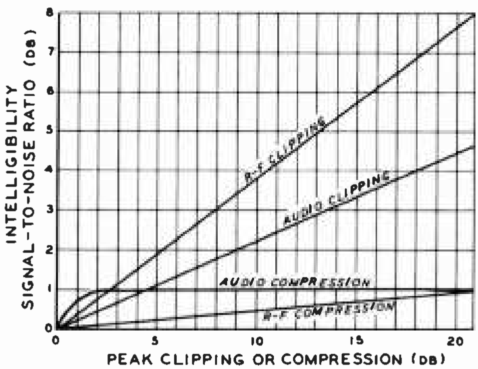


Figure 18

COMPARISON OF SPEECH-PROCESSING TECHNIQUES

In terms of over-all speech intelligibility, r-f clipping has an advantage of several decibels over other systems. R-f clipping up to 10 decibels or so may be used with many SSB transmitters without objectionable distortion. Use of add-on speech processing of any type should be done with caution since the user has no knowledge of limitations of the transmitter, which may preclude drastic changes in peak-to-average ratio of driving signal.

Figure 18 shows a comparison of the four different methods of speech processing used in SSB work. R-f envelope clipping has an advantage of several decibels over the other systems. All techniques increase transmitted average-to-peak power to a degree, thereby improving the over-all speech intelligibility. Use of two speech-processing systems, however, is not directly additive, and only the larger improvement factor should be considered.

Power-Supply Requirements The power load of an SSB transmitter can fluctuate between the zero-signal value and that required for maximum signal power output. For a class-B stage, this may represent a current ratio of 10 to 1, or more. The rate and amount of current fluctuation are related to the envelope of the SSB signal and the frequency components in the supply current variation may be much lower and higher than the frequency components of the driving signal. For voice modulation, supply current fluctuations corresponding to syllabic variations may be as low as 20 Hz and high-order distortion products of nonlinear stages may produce fluctuations higher than 3000 Hz. The power supply for an SSB transmitter, therefore, must have good *dynamic regulation*, or the ability to absorb a sudden change in the load without an abrupt voltage change. The most effective means of achieving good dynamic regulation in the supply is to have sufficient filter capacity in the supply to overcome sudden current peaks caused by abrupt changes of signal level. At the same time, *static regulation* of the supply may be enhanced by reducing voltage drops in the power transformer, rectifier, and filter choke, and by controlling transformer leakage reactance.

9-6 SSB Reception

Single-sideband reception may be considered the reverse of the process used in SSB transmission. The received SSB signal is amplified, translated downward in frequency, further amplified and converted into a replica of the original audio frequencies. The SSB receiver is invariably a superheterodyne in order to achieve high sensitivity and selectivity.

To recover the intelligence from the SSB signal, it is necessary to restore the carrier in such a way as to have the same relationship with the sideband components as the original carrier generated in the SSB exciter. To achieve this, it is important that the receiver oscillators have good frequency accuracy and stability.

To take advantage of the narrow bandwidth occupied by the SSB signal, selectivity characteristics of the receiver must be held to narrow limits. Excessive receiver bandwidth degrades the signal by passing unnecessary interference and noise.

SSB Receivers In a conventional a-m receiver, the audio intelligence is recovered from the radio signal by an envelope amplitude detector, such as a diode rectifier. This technique may be used to recover the audio signal from an SSB transmission provided the amplitude of the local carrier generated by the beat oscillator is sufficiently high to hold audio distortion at a reasonable low level. Better performance with respect to distortion may be achieved if a *product detector* is used to recover the audio signal.

The characteristics of the *automatic volume control* (or *automatic gain control*) system of an SSB receiver differ from those of a conventional a-m receiver. In the latter, the agc voltage is derived by rectifying the received carrier, as the carrier is relatively constant and does not vary rapidly in amplitude. The agc system can therefore have a rather long time constant so that an S-meter may be used to indicate relative carrier amplitude.

In an SSB receiver, however, the signal level varies over a large range at a syllabic rate and a fast time-constant agc system is required to prevent receiver overload on initial bursts of a received signal. To prevent background noise from receiving full amplification when the SSB signal is weak or absent, a relatively slow agc release time is required.

The agc system, moreover, must be isolated from the local-oscillator voltage to prevent rectification of the oscillator voltage from placing an undesired no-signal static bias voltage on the agc line of the receiver.

Thus, the SSB receiver differs from the a-m receiver in that it requires a higher order of oscillator stability and i-f bandwidth, a more sophisticated agc system, and the capability of receiving signals over a very wide range of strength without overload or cross modulation. In addition, the tuning rate of the SSB receiver should be substantially less than that of an a-m receiver; generally speaking, tuning rates of 25 to 100 kHz per dial revolution are common in modern SSB receivers.

Because of variations in the propagation path, transmitter power, and distance between stations, the input signal to an SSB receiver can vary over a range of 120 decibels or so. The receiver requires, therefore, a large dynamic range of signal-handling capability and an enhanced degree of gain-adjusting capability.

SSB Receiver Circuitry For minimum spurious response it is desired to have good selectivity ahead of the amplifier stages in the SSB receiver. This is possible to a degree, provided circuit simplicity and receiver sensitivity are not sacrificed. For the case when sensitivity is not important, an attenuator may be placed in the receiver input circuit to reduce the amplitude of strong, nearby signals (figure 18). To further reduce the generation of cross-modulation interference, it is necessary to

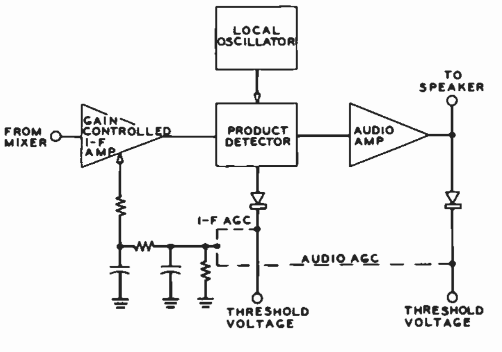


Figure 19

BLOCK DIAGRAM OF AUTOMATIC GAIN CONTROL SYSTEM

Audio or i-f derived control signal is applied to low-level gain-controlled i-f amplifier in typical SSB receiver. A.g.c. system reduces the gain of controlled stage(s) on signal peaks to prevent receiver overload. Control voltage must be derived from the modulation envelope, since carrier is not transmitted with voice SSB signal.

carefully select the tube type used in the r-f amplifier stage to determine if it will retain its linearity with the application of agc-bias control voltage. Suitable r-f stage circuits are shown in the *Radio Receiver Fundamentals* chapter of this Handbook.

Avoidance of images and spurious responses is a main problem in the design of SSB receiver mixers. Due to the presence of harmonics in the mixer/oscillator signal and nonlinearity in the mixer, higher-order products are generated in addition to the desired mixing product. These undesired products vary in frequency as the oscillator is tuned and may fall within the received passband, creating *crossovers*, or *birdies* (spurious beat-notes which tune faster than the normal tuning rate).

Choice of an intermediate frequency low with respect to signal frequency minimizes the probability of strong birdie signals within the receiver passband. The low intermediate frequency, however, may lead to image problems at the higher received frequencies.

The twin problems of images and crossovers can be resolved through the use of double conversion. The first (high) conversion provides adequate image rejection and the second (low) conversion may be adjusted so as to reduce crossover points to a minimum. In addition, double conversion allows the use of a crystal-controlled oscillator for the first converter stage, which can provide a higher order of stability than a tunable oscillator. The oscillator for the lower mixer stage may be made tunable, covering only a single frequency range, eliminating some of the mechanical and electrical factors contributing to receiver instability.

The bandwidth of the low-frequency i-f system determines the over-all selectivity of the SSB receiver. For SSB voice reception, the optimum bandwidth at the 6-db point is about 2 kHz to 3 kHz. It is good practice to place the selective filter in the circuit ahead of the i-f amplifier stages so that strong adjacent-channel signals are attenuated before they drive the amplifier tubes into the overload region. In addition to the sideband filter, additional tuned circuits are usually provided to improve over-all receiver selectivity, especially at frequencies which are down the skirt of the selectivity curve. Some types of SSB filters have spurious responses outside the passband which can be suppressed in this manner.

Desensitization, Intermodulation, and Crossmodulation When a receiver is tuned to a weak signal with a strong signal close to the received frequency, an apparent decrease in receiver gain may be noted. This loss of gain is called *desensitization* or *blocking*. It commonly occurs when the unwanted signal voltage is sufficient to overcome the operating bias of an amplifier or mixer stage, driving the stage into a nonlinear condition. Rectified signal current may be coupled back into the gain-control system, reducing over-all gain and increasing signal distortion.

Amplifier and mixer stages using transistors and vacuum tubes may generate in-band spurious products resulting from beats between the components of the desired signal in the receiver, or between two received signals. This class of distortion is termed *intermodulation distortion* and is evident in a nonlinear device driven by a complex signal having more than one frequency, such as the human voice.

Intermodulation occurs at any signal level and spurious products are developed by this action. For example, assume a signal is on 900 kHz and a second signal is on 1.5 Mhz. The receiver is tuned to the 80-meter band. Intermodulation distortion within the receiver can result in a spurious signal appearing at 3.9 MHz as a result of mixing in a nonlinear stage. The product mix is: $(2 \times 1.5) + 0.9 = 3.9$ MHz.

This particular spurious signal (often termed a *spur*) is a result of a harmonic of the 1.5-MHz signal being produced in the receiver and beating against the incoming 0.9-MHz signal. Other spurious signals, composed of the sums and differences and harmonics of the fundamental signals exist in addition to the one at 3.9 MHz. Some of these products fall at: 0.3, 1.8, 2.1, 2.7, 3.0, 3.3, and 4.5 MHz. Other spurs may be generated by higher order linearities. Thus, two signals passed through a nonlinear device can create a whole range of unwanted signals. Since the radio spectrum is crowded with numerous strong signals, all of which can create spurious intermodulation products simultaneously in varying degrees of severity, it is important that high-Q circuits or a number of tuned circuits be used in the front-end of a receiver to prevent out-of-band signals from entering the receiver. In addition, the optimum choice of transistor or tube must be made for each receiver stage, and its correct operating point established.

Crossmodulation is the transfer of intelligence from an unwanted strong signal to a wanted weak one. Thus, if a receiver is tuned to a wanted signal at 3.9 MHz and a strong unwanted signal is at 3.8 MHz, the modulation on the second signal may be imposed on the wanted signal, even though the second signal is well outside the i-f passband of the receiver. Multiple signals, moreover, can produce multiple crossmodulation effects. Crossmodulation can be minimized by optimum selection of amplifying and mixing devices and by careful selection of signal levels and operating voltages in the various receiver stages.

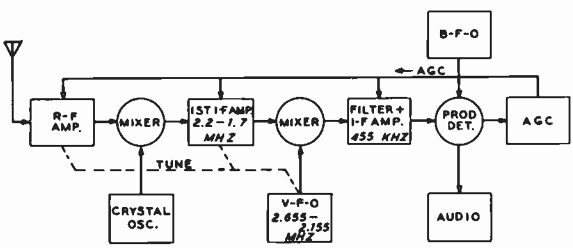
Intermodulation, crossmodulation and desensitization can all occur simultaneously in a receiver and the over-all effect is a loss in intelligibility and signal-to-noise ratio of the desired signal. These receiver faults may be ascertained by injecting test signals of various frequencies and amplitudes into the receiver, a stage at a time.

Generally speaking, field-effect transistors and remote-cutoff vacuum tubes exhibit a significant improvement in linearity and provide enhanced rejection to these unwanted effects as opposed to bipolar transistors, which have a lower linearity figure than the other devices.

Automatic Gain Control and Signal Demodulation The function of an *automatic gain control* system is to reduce the gain of the controlled stages on signal peaks to prevent receiver overload and hold constant audio output. Since the carrier is not transmitted in SSB, the receiver agc system must obtain its signal voltage from the modulation envelope. The agc voltage may be derived either from the i-f signal or the audio signal (figure 19). Audio-derived agc has the advantage of easier isolation from the local carrier voltage, but the i-f system will function on both SSB and a-m signals in a satisfactory manner.

Product detectors are preferred for SSB reception because they minimize intermodulation distortion products in the audio signal and, in addition, do not require a large local-oscillator voltage. The product detector also affords a high degree of isolation between the carrier oscillator and the agc circuit. The undesired mixing products present in the output circuit of the detector may be suppressed by a low-pass filter placed in the audio line.

A Representative SSB Receiver A typical SSB receiver is made up of circuits resembling those discussed in the previous section. To achieve both high stability and good image rejection, many amateur SSB receivers are double-conversion types, such as outlined in figure 20. An accurate, stable low-frequency tunable oscillator is employed, together with a standard 455-kHz i-f channel and a crystal or



TUNING RANGE (MHz)	CRYSTAL (MHz)
3.5-4.0	5.7
7.0-7.5	9.0
14.0-14.5	16.2
21.0-21.5	23.2
28.5-29.0	30.7

Figure 20

DOUBLE CONVERSION SSB RECEIVER

Typical double-conversion SSB receiver employs tunable first i-f and crystal-controlled local oscillator, with tunable oscillator and fixed-frequency i-f amplifier and sideband filter. This receiver tunes selected 500-kHz segments of the h-f spectrum. Additional conversion crystals are required for complete coverage of the 10-meter band.

mechanical SSB filter. The frequency coverage of the vfo may be as high as 500 kHz to cover all of the low-frequency amateur bands, or it may be restricted to only 100 kHz or so, necessitating the use of a multiplicity of crystals in the first conversion oscillator to achieve complete band coverage. A tunable first i-f stage covering the required passband may be ganged with the variable-frequency oscillator and with the r-f amplifier tuning circuits. The high-frequency tuning range is chosen by the appropriate high-frequency crystal.

To permit sideband selection, the bfo may be tuned to either side of the i-f passband. Proper tuning is acomplished by ear, the setting of the bfo on the filter passband slope may be quickly accomplished by experience and by recognition of the proper voice tones.

In addition to the special circuitry covered in this chapter, SSB receivers make full use of the general receiver design information given in this Handbook.

9-7 The SSB Transceiver

The SSB *transceiver* is a unit in which the functions of transmission and reception are combined, allowing single-channel semiduplex operation at a substantial reduction in cost and complexity along with greatly increased ease of operation. The transceiver is especially popular for mobile operation where a savings in size, weight, and power consumption are important. Dual usage of components and stages in the SSB transceiver permits a large reduction in the number of circuit elements and facilitates tuning to the common frequency desired for two-way communication.

Figure 21 shows a basic filter-type transceiver circuit. Common mixer frequencies are used in each mode and the high-frequency vfo is used to tune both transmit and receive channels to the same operating frequency. In addition, a common i-f system and sideband filter are used.

The transceiver is commonly switched from receive to transmit by a multiple-contact relay which transfers the antenna and removes blocking bias from the activated stages. Transceivers are ideal for net operation since the correct frequency may be ascertained by tuning the received signal to make the voice intelligible and pleasing. With practice, the SSB transceiver may be adjusted to a predetermined frequency with an error of 100 Hz or less by this simple procedure.

Single-Band Transceivers An important development is the single-band transceiver, a simplified circuit designed for operation over one narrow frequency band. Various designs have been made available for the 50-MHz band as well as the popular h-f amateur bands. Commercial transceiver designs are usually operated on crystal-controlled channels in the h-f and vhf spectrum using a crystal synthesizer for channel control. Elaborate synthesizers permit selection of discrete operating frequencies as closely separated as 100 Hz. Some units include a *clarifier* control which permits a slight frequency adjustment to place the unit exactly on the chosen operating channel.

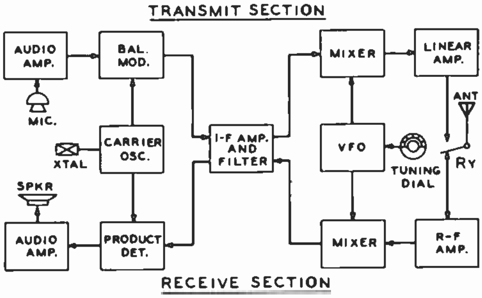


Figure 21

THE SSB TRANSCEIVER

Common carrier oscillator, i-f amplifier/filter, and vfo are used in transceiver, designed to communicate on a single frequency selected by proper vfo setting. Transfer from receive to transmit is carried out by relays and by application of blocking voltage to unused tubes.

CHAPTER TEN

Communication Receiver Fundamentals

Part I—The H-F Receiver

Communication receivers vary widely in their cost, complexity and design, depending on the intended application and various economic factors. A receiver designed for amateur radio use must provide maximum intelligibility from signals varying widely in received strength, and which often have interfering signals in adjacent channels, or directly on the received channel. The practical receiver should permit reception of continuous wave (c-w), amplitude-modulated (a-m) and single-sideband (SSB) signals. Specialized receivers (or receiver adapters) are often used for reception of narrow band f-m (NBFM), radio teletype (RTTY), slow scan television (SSTV) and facsimile (FAX) signals.

The desired signal may vary in strength from a fraction of a microvolt to several volts at the input terminals of the receiver. Many extraneous strong signals must be rejected by the receiver in order to receive a signal often having a widely different level than the rejected signals.

The modern receiver, in addition, must have a high order of electrical and mechanical stability, and its tuning rate should be slow enough to facilitate the exact tuning of c-w and SSB signals. Finally, the receiver should be rugged and reliable as well as easy to service, maintain, and repair. All of these widely differing requirements demand a measure of compromise in receiver design in order to achieve a reasonable degree of flexibility.

Modern receivers utilizing either solid-state devices or vacuum tubes can readily meet most of these requirements. The present design trend is toward use of semiconductors and field-effect devices, particularly in manufactured receivers because of performance, cost, and assembly considerations. The solid-state receivers, however, tend to be more complex than their vacuum-tube counterparts and often do not offer as high a dynamic signal range (the ability to cope with both very strong and very weak signals) as do receivers utilizing vacuum tubes.

10-1 Types of Receivers

All receivers are *detectors* or *demodulators* which are devices for removing the modulation (intelligence) carried by the incoming signal. Figure 1 illustrates an elementary receiver wherein the induced voltage from the signal is diode rectified into a varying direct current. The current is passed through earphones which reproduce the modulation placed on the radio wave.

The Autodyne Detector Since a c-w signal consists of an unmodulated carrier interrupted by dots and dashes, it is apparent that such a signal would not be made audible by detection alone. Some means must be provided whereby an audible tone is heard when the carrier is received, the tone stopping when the carrier is interrupted. Audible detection may be accomplished by generating a local carrier of a slightly different frequency and mixing it with the incoming signal in the detector stage to form

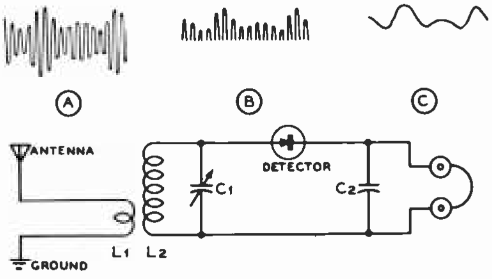


Figure 1

ELEMENTARY FORM OF RECEIVER

This is the basis of the "crystal set" type of receiver. The tank circuit (L_2-C_1) is tuned to the frequency it is desired to receive. The bypass capacitor across the phones should have a low reactance to the carrier frequency being received, but a high reactance to the modulation on the received radio signal.

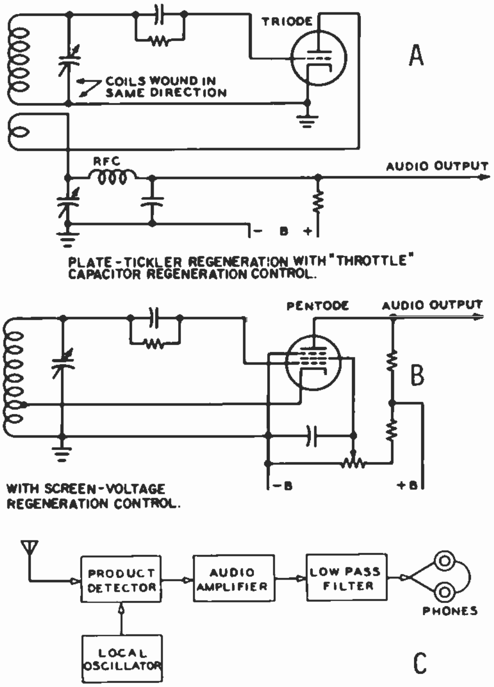


Figure 2

REGENERATIVE DETECTOR CIRCUITS

Regenerative detectors are seldom used at the present time due to their poor selectivity. Detector may be used for c-w or SSB when adjusted for oscillation or for a-m phone when set just below point of oscillation. Direct conversion receiver uses separate heterodyne oscillator to produce audio beat note signal. Passband is restricted by use of audio filter.

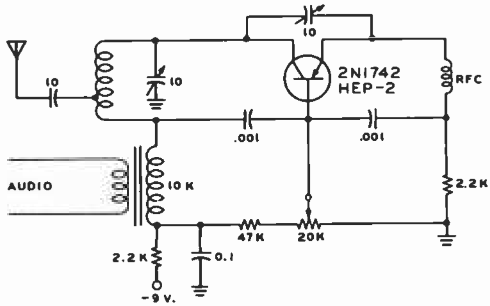


Figure 3

SUPERREGENERATIVE DETECTOR FOR VHF

A self-quenched superregenerative detector is capable of giving good sensitivity in the vhf range, but has relatively poor selectivity. Such a circuit should be preceded by an r-f stage to suppress radiation from the oscillating detector.

a beat note. The difference frequency, or *heterodyne*, exists only when both the incoming signal and the locally generated signal are present in the mixer. The mixer (or detector) may be made to supply the beating signal, as shown in the *autodyne detector* circuits of figure 2. A variation of the autodyne detector makes use of a separate oscillator and is termed a *direct conversion* receiver. A product detector may be used and signal selectivity is obtained at audio frequencies through the use of a low-pass audio filter.

The Superregenerative Detector At ultrahigh frequencies, when it is desired to keep weight and cost at a minimum a special form of the regenerative receiver known as the *superregenerator* is often used for radiotelephony reception. The superregenerator is essentially a regenerative receiver with a means provided to throw the detector rapidly in and out of oscillation. The frequency at which the detector is made to go in and out of oscillation varies with the frequency to be received, but is usually between 20,000 and 500,000 times a second. This superregenerative action considerably increases the sensitivity of the oscillating detector so that the usual *background hiss* is greatly amplified when no signal is being received.

The simplest type of superregenerative detector circuit is arranged so as to produce its own interruption frequency oscillation, without the aid of a separate stage. The detector tube or transistor damps (or quenches) itself out of signal-frequency oscillation at a high rate by virtue of the use of a high value of grid resistor and proper size blocking and coupling capacitors, in conjunction with an excess of feedback. A representative self-quenched superregenerative detector circuit is shown in figure 3.

The optimum quenching frequency is a function of the signal frequency. As the operating frequency goes up, so does the optimum quenching frequency. When the quench frequency is too low, maximum sensitivity is not obtained. When it is too high, both sensitivity and selectivity suffer. In fact, the optimum quench frequency for an operating frequency below 15 MHz is in the audible range. This makes the superregenerator impractical for use on the lower frequencies.

The selectivity of a superregenerator is rather poor compared to a superheterodyne, but is suprisingly good for so simple a receiver when figured on a percentage basis rather than absolute kHz bandwidth.

Superregenerative receivers radiate a strong, broad, and rough signal. For this reason, it is necessary in most applications to employ a radio-frequency amplifier stage ahead of the detector, with thorough shielding throughout the receiver.

10-2 Receiver Performance Requirements

Receiver performance may be defined in terms of *sensitivity, selectivity, spurious response, tuning rate,* and *dynamic signal range* (discussed in Chapter 9, section 6). Other factors may enter into receiver specifications, but these properties are of the greatest interest to the radio amateur. A well designed communication receiver must be able to receive all modes of emission used on the amateur bands while meeting minimum levels of performance in these important areas of operation.

Sensitivity The *sensitivity* of a high-frequency receiver may be defined as the ability of the receiver to detect a weak signal through the general noise level of the receiving system. Specifically, it is the input level to the receiver in microvolts required to give a signal-plus-noise output of some ratio above the noise output of the receiver. A perfect "noiseless" receiver would generate no internal noise and the minimum detectable signal would be limited only by the thermal noise in the antenna system and the external noise (or "r-f smog") about the receiving location. Below 30 MHz or so, external noise, rather than internal receiver noise, is the limiting factor in weak signal reception.

A modern h-f communication receiver, generally speaking, should have a c-w signal selectivity of better than one microvolt to provide better than 20 db signal-plus-noise to noise ratio in a passband of less than 1000 Hz.

The sensitivity of any receiver may be increased by reducing the bandwidth of response, within the limits imposed by the mode of transmission being received. The absolute sensitivity of the receiver may also be defined, independent of receiver band-

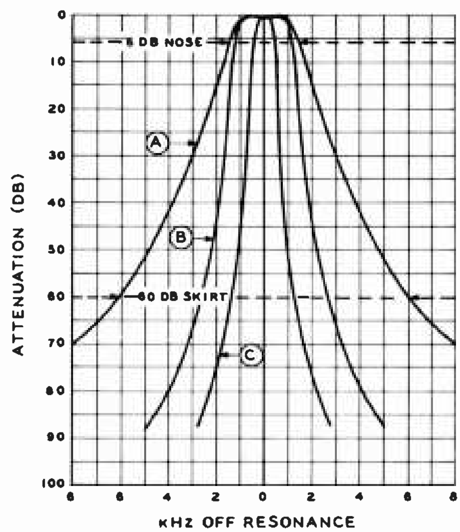


Figure 4

SKIRT SELECTIVITY

Receiver bandwidth is determined by selectivity of i-f system. Curve A shows typical response for reception of double-sideband, amplitude-modulated signal. SSB reception on a good communication receiver is shown by curve B. C-w selectivity is shown by curve C. Strong-signal selectivity is determined by bandwidth at 60-db skirt points.

width, in terms of *noise factor*, as discussed in Section 10-4.

Selectivity The *selectivity* of a communications receiver is the ability of the receiver to separate signals on closely adjacent frequencies. Ultimate selectivity is determined by the bandwidth of intelligence being received. For reception of double-sideband a-m signals, a bandwidth of about 5 kHz is required. SSB reception bandwidth may be as little as 2 kHz for voice reception. For c-w reception, bandwidths less than 100 Hz are often employed. As circuit bandwidth is reduced, transmitter and receiver stability requirements become more strict and practical bandwidths in receivers may often have to be greater than the theoretical minimum requirement to compensate for frequency drift of the equipment.

Receiver bandwidth may be defined in terms of *skirt selectivity*, or the degree of attenuation to a signal received at some frequency removed from the center frequency of reception. The bandwidth is taken as the width of the over-all resonance curve of the receiver at the 6-db nose, as shown in figure 4.

Stability The ability of a receiver to remain tuned to a chosen frequency is a measure of the *stability* of the receiver. Environmental changes such as variations in temperature, supply voltage, humidity and mechanical shock or vibration tend to alter the receiver characteristics over a period of time. Most receivers, to a greater or lesser degree, have a steady frequency variation known as *warm-up drift* which occurs during the first minutes of operation. Once the receiver components have reached operating temperature, the drift settles down, or subsides. *Long-term drift* may be apparent over a period of days, weeks or even years as components age or gradually shift in characteristics due to heat cycling or usage. Many receivers include a high-stability calibration oscillator to provide *marker signals* at known frequencies to allow rapid frequency calibration of the receiver dial. Typical short-term receiver drift is shown in figure 5.

Spurious Responses The mark of a good communication receiver is its ability to reject spurious signals outside of the passband of the receiver. Spurious responses such as *images* and *birdies* may interfere with the received signal, although the interfering signal causing unwanted response may be many hundreds of kilohertz removed from the signal frequency (Chapter 9, Section 6). All superheterodyne receivers suffer from image response which becomes progressively more important as the signal frequency is raised. Careful system design of the receiver and choice of i-f and mixing frequencies can reduce images and birdies to a minimum. Generally speaking, a good communication receiver should have an image rejection of greater than 40 decibels at its highest operating frequency, and the majority of birdie signals generated by mixing products within the receiver should be reduced by the same amount below a one microvolt signal.

Tuning Rate A good communication receiver should have a slow *tuning rate*. That is, each revolution of the tuning control should represent only a moderate frequency change when compared to the bandwidth of reception. SSB receivers often have a tuning rate of 100 to 500 kHz per tuning dial revolution. Receivers intended for c-w reception may have a tuning rate as low as 5 kHz per dial revolution. The tuning rate may be determined mechanically by means of a step-down gear train or rim-drive mechanism placed between the tuning dial

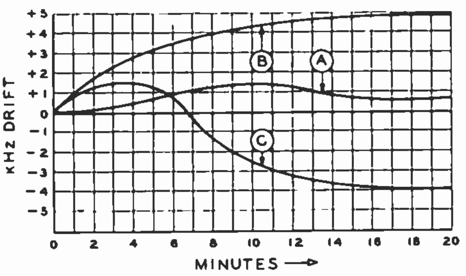


Figure 5

RECEIVER FREQUENCY STABILITY

Frequency drift of receiver depends on electrical and mechanical stability of tuned circuits. Temperature compensation (A) reduces warmup drift to a minimum. No compensation may result in long term, continual drift (B) and overcompensation can show as reversal of drift (C). Frequency compensation may be achieved by use of special capacitors having controlled temperature characteristics in critical circuits and by temperature stabilization of oscillator circuitry.

and the tuning control of the receiver. In some instances, electrical *bandspread* (see Section 10-5) may be employed. Regardless of the technique used, the tuning mechanism should have a smooth action and be free of mechanical or electrical backlash.

The Superheterodyne Receiver By changing the frequency of a received signal to a lower, fixed, *intermediate frequency* before ultimate detection, high gain and selectivity may be obtained with a good order of stability. A receiver that performs this frequency changing (heterodyning) process is termed a *superheterodyne* or *superhet* receiver. A block diagram of a typical superhet receiver is shown in figure 6.

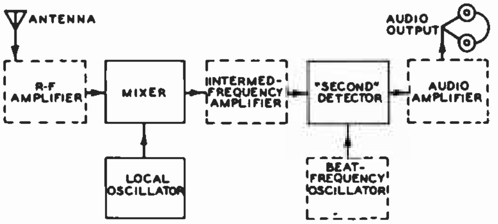


Figure 6

ESSENTIAL UNITS OF A SUPERHETERODYNE RECEIVER

The basic portions of the receiver are shown in solid blocks. Practicable receivers employ the dotted blocks and also usually include such additional circuits as a noise limiter, an agc circuit, and a bandpass filter in the i-f amplifier.

The incoming signal is applied to a *mixer* consisting of a nonlinear impedance such as a vacuum tube, transistor, or diode. The signal is mixed with a locally generated variable-frequency signal, with the result that a third signal bearing all the modulation applied to the original signal but of a frequency equal to the difference between the local oscillator and the incoming signal frequency appears in the mixer output circuit. The output from the mixer is fed into a fixed-tuned intermediate-frequency amplifier, wherein it is amplified, detected, and passed on to an audio amplifier.

Although the mixing process is inherently noisy, this disadvantage can be overcome by including a radio-frequency amplifier stage ahead of the mixer, if necessary.

Advantages of the Superheterodyne The advantages of superheterodyne reception are directly attributable to the use of the fixed-tuned *intermediate-frequency* (i-f) *amplifier*. Since all signals are converted to the intermediate frequency, this section of the receiver may be designed for optimum selectivity and high amplification. High amplification is easily obtained in the intermediate-frequency amplifier, since it operates at a relatively low frequency, where conventional pentode-type tubes and transistors give adequate voltage gain.

While the regenerative receiver may be suitable for c-w reception, and the superregenerative receiver provides inexpensive vhf reception, the over-all advantages of the superhet circuit have made it the universal choice for general communications reception at all frequencies, from the very low frequencies well up into the uhf range. Various advantages and shortcomings of the superhet receiver will be discussed at length in the following sections of this Chapter.

10-3 The Superheterodyne Receiver

While superhet receivers are the universally accepted type of circuitry for serious radio reception at most commonly used frequencies, the device has practical disadvantages that should be recognized. The greatest handicap of this type of receiver is its susceptibility to various forms of spurious response and the complexity of proper adjustment to reduce this response. Proper circuit design will alleviate some of these problems.

Choice of Intermediate Frequency The choice of a frequency for the i-f amplifier involves several considerations. One of these considerations concerns selectivity—the lower the intermediate frequency the greater the obtainable selectivity. On the other hand, a rather high intermediate frequency is desirable from the standpoint of *image* elimination, and also for the reception of signals from television and f-m transmitters both of which occupy a rather wide band of frequencies, making a broad selectivity characteristic desirable.

Images are a peculiarity common to all superheterodyne receivers, and for this reason they are given a detailed discussion later in this chapter.

While intermediate frequencies as low as 50 kHz are used where extreme selectivity is a requirement, and frequencies of 60 MHz and above are used in some specialized forms of receivers, many communication receivers use intermediate frequencies near 455 or 1600 kHz. Some receivers make use of high-frequency crystal-lattice filters in the i-f amplifier and use an intermediate frequency as high as 5 MHz or 9MHz to gain image rejection. Entertainment receivers normally use an intermediate frequency centered about 455 kHz, while many automobile receivers use a frequency of 262 kHz. The standard frequency for the i-f channel of f-m receivers is 10.7 MHz, whereas the majority of television receivers use an i-f which covers the band between 41 and 46 MHz.

Arithmetical Selectivity Aside from allowing the use of fixed-tuned bandpass amplifier stages, the superheterodyne has an overwhelming advantage over the tuned radio frequency (trf) type of receiver because of what it commonly known as *arithmetical selectivity*.

This can best be illustrated by considering two receivers, one of the trf type and one of the superheterodyne type, both attempting to receive a desired signal at 10,000 kHz and eliminate a strong interfering signal at 10,010 kHz. In the trf receiver, separating these two signals in the tuning circuits is practically impossible, since they differ in frequency by only 0.1 percent. However, in a superheterodyne with an intermediate frequency of, for example, 1000 kHz, the desired signal will be converted to a frequency of 1000 kHz and the interfering signal will be converted to a frequency of 1010 kHz, both signals appearing at the input of the i-f amplifier. In this case, the two signals may be separated much more readily, since they differ by 1 percent, or 10 times as much as in the first case.

Double-conversion receivers make use of two or more separate cascaded i-f amplifiers working at different frequencies. These receivers will be discussed later in this chapter.

Images There always are *two* signal frequencies which will combine with a given frequency to produce the same difference frequency. For example: assume a superheterodyne with its oscillator operating on a higher frequency than the signal (which is common practice in many superheterodynes) tuned to receive a signal at 14,100 kHz. Assuming an i-f amplifier frequency of 450 kHz, the mixer input circuit will be tuned to 14,100 kHz, and the oscillator to 14,100 plus 450, or 14,550 kHz. Now, a *strong* signal at the oscillator frequency plus the intermediate frequency (14,550 plus 450, or 15,000 kHz) will also give a difference frequency of 450 kHz in the mixer output and will be heard also. Note that the image is always *twice* the intermediate frequency away from the desired signal. Images cause *repeat points* on the tuning dial.

The only way that the image could be eliminated in this particular case would be to make the selectivity of the mixer input circuit, and any circuits preceding it, great enough so that the 15,000-kHz signal never reaches the mixer input circuit in sufficient amplitude to produce interference.

For any particular intermediate frequency, image interference troubles become increasingly greater as the frequency (to which the signal-frequency portion of the receiver is tuned) is increased. This is due to the fact that the percentage difference between the desired frequency and the image frequency decreases as the receiver is tuned to a higher

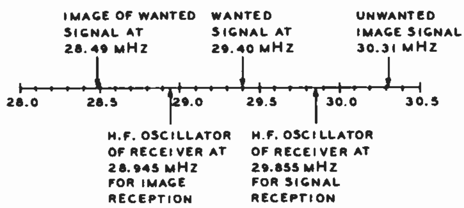


Figure 7

IMAGE SIGNAL

Relation between image signal and wanted signal when receiver local oscillator operates on high-frequency side of wanted signal. Image of 29.40 MHz signal appears at 28.49 MHz when 455 kHz i-f system is used. Unwanted signal at 30.31 MHz appears as image signal when receiver is tuned to desired signal at 29.40 MHz. Conditions are reversed for operation of oscillator on low-frequency side of signal.

frequency. The ratio of strength between a signal at the image frequency and a signal at the frequency to which the receiver is tuned producing equal output is known as the *image ratio.* The higher this ratio is, the better the receiver will be in regard to image interference troubles.

With but a single tuned circuit between the mixer grid and the antenna, and with 400- to 500-kHz i-f amplifiers, image ratios of 60 db and over are easily obtainable up to frequencies around 2000 kHz. Above this frequency, greater selectivity in the mixer grid circuit through the use of additional tuned circuits between the mixer and the antenna is necessary if a good image ratio is to be maintained.

Image signal reception can be confusing, especially in SSB reception, when an image signal may appear on the opposite sideband and tune "in the wrong direction" as compared to normal signals. Figure 7 illustrates the relationship between image signals when the receiver local oscillator operates on the high-frequency side of the received signal. The conditions are reversed for oscillator operation on the low-frequency side of the received signal. For reasons of economy and maximum oscillator stability, many receivers employ "low-side" oscillator operation on all but the highest frequency bands, where "high-side" operation may be occasionally used.

Double Conversion As previously mentioned, the use of a higher intermediate frequency will also improve the image ratio, at the expense of i-f selectivity, by placing the desired signal and the image farther apart. To give both good image ratio at the higher frequencies and good selectivity in the i-f amplifier, a system known as *double conversion* is sometimes employed. In this system, the incoming signal is first converted to a rather high intermediate frequency, and then amplified and again converted, this time to a much lower frequency. The first intermediate frequency supplies the necessary wide separation between the image and the desired signal, while the second one supplies the bulk of the i-f selectivity.

The double-conversion system, as illustrated in figure 8, is receiving two general

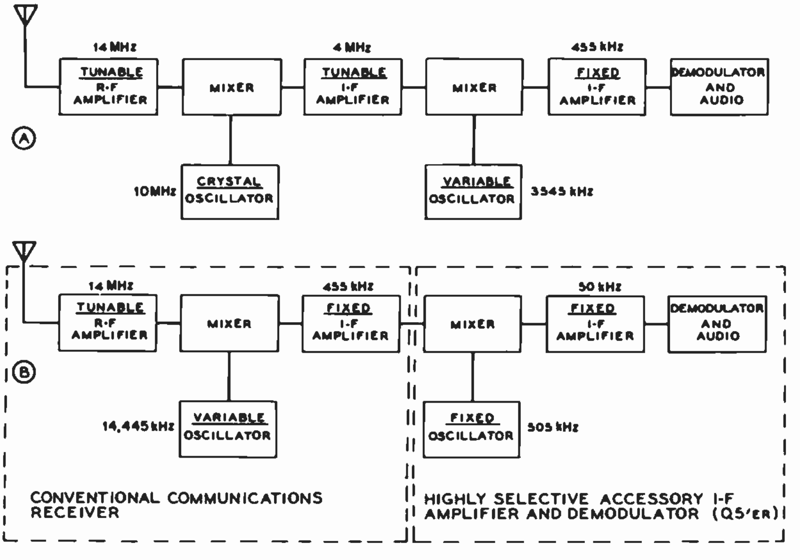


Figure 8

TYPICAL DOUBLE-CONVERSION SUPERHETERODYNE RECEIVERS

Illustrated at A is the basic circuit of a commercial double-conversion superheterodyne receiver. At B is illustrated the application of an accessory sharp i-f channel for obtaining improved selectivity from a conventional communications receiver through the use of the double-conversion superheterodyne principle.

types of application at the present time. The first application is for the purpose of attaining extremely good stability in a communications receiver through the use of crystal control of the first oscillator. In such an arrangement the first oscillator is crystal controlled and is followed by a tunable i-f amplifier which then is followed by a mixer stage and a fixed-tuned i-f amplifier on a much lower frequency. Through such a circuit arrangement the stability of the complete receiver is equal to the stability of the oscillator which feeds the second mixer, while the selectivity is determined by the bandwidth of the second fixed i-f amplifier.

The second common application of the double-conversion principle is for the purpose of obtaining a very high degree of selectivity in the complete communications receiver. In this type of application, as illustrated in figure 8B, a conventional communications receiver is modified in such a manner that its normal i-f amplifier (which usually is in the 450- to 915-kHz range) instead of being fed to a demodulator and then to the audio system, is alternatively fed to a fixed-tuned mixer stage and then into a much lower intermediate-frequency amplifier before the signal is demodulated and fed to the audio system. The accessory i-f amplifier system normally is operated on a frequency of 175 kHz, 85 kHz, or 50 kHz.

Some specialized high-frequency receivers make use of an intermediate frequency *above* the tuning range of the receiver (30 MHz, for example). The extremely high i-f permits a high order of attenuation of image responses and allows continuous tuning of all frequencies up to 30 MHz or so. In other designs, dual conversion is accomplished with the use of but a single local oscillator, with the injection frequency chosen so that oscillator drift is automatically eliminated.

In all double-conversion receivers, the problem of spurious responses is aggravated because of the multiple-frequency signals existing within the receiver circuitry. Careful shielding and filtering of power leads must be incorporated in a receiver of this type if birdies and spurious signals are to be avoided.

Audio Circuitry The communication receiver has no need to reproduce audio frequencies outside of the required communication passband. The high-frequency response of such a receiver is usually limited by the selective i-f passband. For voice reception, the lower audio frequencies are also attenuated in order to make speech crisp and clear. An audio passband of about 200 to 2000 Hz is all that is normally required for good SSB reception of speech. For c-w reception, the audio passband can be narrowed further by peaking the response to a frequency span of about 700 to 1000 Hz. High-Q audio filters may be used in the communication receiver to shape the audio response to the desired characteristic. In addition, audio or i-f filters may be added to either provide a special, narrow response characteristic, or a sharp rejection notch to eliminate heterodynes or objectionable interference.

Control Circuitry Under normal circumstances, the communication receiver is disabled during periods of transmission. A standby control may take the form of a switch or circuit that removes high voltage from certain tubes or transistors in the receiver. Alternatively, the bias level applied to the r-f and i-f stages may be substantially increased during standby periods to greatly reduce receiver gain. This will permit use of the receiver as a monitoring device during periods of transmission. In all cases, the input circuitry of the receiver must be protected from the relatively strong r-f field generated by the transmitter. Receiver control circuitry may be actuated by the transmitter control devices through the use of suitable interconnecting relay circuits (VOX), as discussed in Chapter 18 of this Handbook.

Receiver Power Supplies Communications receivers are generally designed to operate from a 117-234 volt, 50- to 60-Hz power source, with the possible addition of auxiliary circuitry to permit operation from a 12-volt automotive electrical system. The majority of receivers incorporate the power supply on the receiver chassis and thus must accommodate the heat the power supply generates during operation. Silicon diodes are to be preferred for power rectifiers as opposed to vacuum-tube rectifiers because of the lesser heat radiation of the solid-state devices. In some instances, voltage regulation circuits or devices are added to the supply to stabilize the voltages applied to critical oscillator circuits. In all instances, the primary circuit of a well designed communications receiver is fused to protect the equipment from overload and the complete receiver is designed and built to protect the operator from accidental shock.

10-4 Noise and Spurious Products

Because of noise sources within the receiver, it is impossible to increase receiver gain and sensitivity without limit. All amplifying and mixing stages contribute to noise generation with varying degrees. The figure of merit for receiver sensitivity is expressed as a ratio called *noise factor* which is independent of input impedance and bandwidth, but not of source impedance. Noise factor is a measure of the degradation of signal-to-noise ratio of a signal as it is processed by the receiver.

Noise Factor The limiting condition for sensitivity in any receiver is the thermal noise generated in the antenna and in the first tuned circuit. However, with proper coupling between the antenna and the input element of the amplifying device, through the first tuned circuit, the noise contribution of the first tuned circuit can be made quite small. Unfortunately, though, the major noise contribution in a properly designed receiver is that of the first tube or transistor. The noise contribution due to electron flow and due to losses in the tube can be lumped into an equivalent value of resistance which, if placed in the grid circuit of a perfect tube having the same gain but no noise would give the same noise voltage output in the plate load. The equivalent noise resistance of tubes such as the 6BA6, 6DC6, etc., runs from 500 to 1000 ohms. Very high G_m tubes such as the 6BZ6 and 6EH7 have equivalent noise resistances as low as 300 to 700 ohms. The lower the value of equivalent noise resistance, the lower will be the noise output under a fixed set of conditions.

The equivalent noise resistance of a tube must not be confused with the actual input loading resistance of a tube. For highest signal-to-noise ratio in an amplifier the input loading resistance should be as high as possible so that the amount of voltage that can be developed from grid to ground by the antenna energy will be as high as possible. The equivalent noise resistance should be as low as possible so that the noise generated by this resistance will be lower than that attributable to the antenna and first tuned circuit, and the losses in the first tuned circuit should be as low as possible.

The absolute sensitivity of receivers has been designated in recent years in government and commercial work by the noise factor. The noise factor is the ratio of noise output of a "perfect" receiver having a given amount of gain with a dummy antenna matched to its input, to the noise output of the receiver having the same amount of gain with an injected signal, and the dummy antenna matched to its input. Although a perfect receiver is not a physically realizable thing, the noise factor of a receiver under measurement can be determined by calculation from the amount of additional noise (from a temperature-limited diode or other calibrated noise generator) required to increase the noise-power output of a receiver by a predetermined amount.

The noise factor expression of a transistor is derived from basic physical principles and is rather complex and of limited usefulness, since many of the parameters of the system are not specified by transistor manufacturers. The noise factor, therefore, is best determined by direct measurement.

The interplay between the sensitivity, noise figure, and audio bandwidth (between the 6-db points) is illustrated by the nomograph of figure 9. The graph is based on the noise figure equation given in Section 10-14, assuming an antenna input impedance of 50 ohms and "room temperature" of 80.5°F.

Tube Input Loading As has been mentioned in a previous paragraph, greatest gain in a vacuum-tube receiver is obtained when the antenna is matched, through the r-f coupling transformer, to the input resistance of the r-f tube. However, the higher the ratio of tube input resistance to equivalent noise resistance of the tube the higher will be the signal-to-noise ratio of the

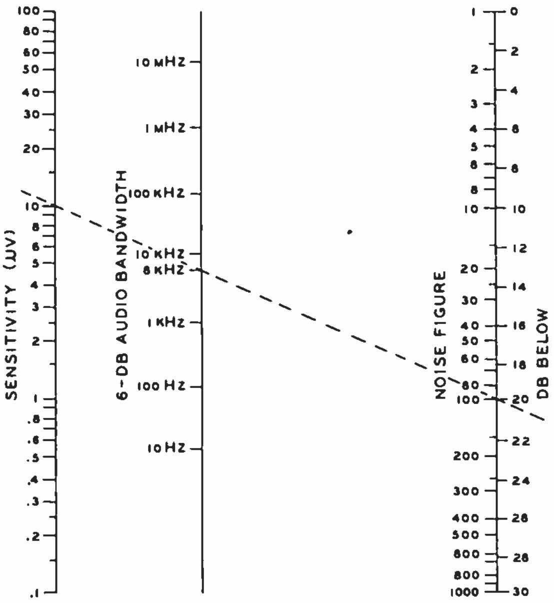


Figure 9

NOISE-FIGURE NOMOGRAPH

To find the noise figure of a receiver, a line extended between sensitivity and audio bandwidth points will intersect noise-figure line at right. Dashed line shows bandwidth of 6 kHz and sensitivity of 10 microvolts gives a noise figure of 100, or 20 db.

stage—and of course, the better will be the noise factor of the over-all receiver. The input resistance of a tube is very high at frequencies in the broadcast band and gradually decreases as the frequency increases. Tube input resistance of conventional tube types begins to become an important factor at frequencies of about 25 MHz and above. At frequencies above about 100 MHz the use of conventional tube types becomes impractical since the input resistance of the tube has become so much lower than the equivalent noise resistance that it is impossible to attain reasonable signal-to-noise ratio on any but very strong signals.

The lowering of the effective input resistance of a vacuum tube at higher frequencies is brought about by a number of factors. The first, and most obvious, is the fact that the *dielectric loss* in the internal insulators, and in the base and press of the tube increases with frequency. The second factor is due to the fact that a finite *transit time* is required for an electron to move from the space charge in the vicinity of the cathode, pass between the grid wires, and travel on to the plate. The fact that the electrostatic effect of the grid on the moving electron acts over an appreciable portion of a cycle at these high frequencies causes a current flow in the grid circuit which appears to the input circuit feeding the grid as a resistance. The decrease in input resistance of a tube due to electron transit time varies as the square of the frequency. The undesirable effect of transit time can be reduced in certain cases by the use of higher plate voltages. Transit time varies inversely as the square root of the applied plate voltage.

Cathode lead inductance is an additional cause of reduced input resistance at high frequencies. This effect has been reduced in certain tubes such as the 6EA5 and the 6BC5 by providing two cathode leads on the tube base. One cathode lead should be connected to the input circuit of the tube and the other lead should be connected to the bypass capacitor for the plate return of the tube.

Plate-Circuit Coupling For the purpose of this section, it will be considered that the function of the plate load circuit of a tuned vacuum-tube amplifier is to deliver energy to the next stage with the greatest efficiency over the required band of frequencies. Figure 10 shows three methods of interstage coupling for tuned r-f voltage amplifiers. As the coefficient of coupling between the circuits of figures 10B and 10C is increased the bandwidth becomes greater but the response over the band becomes progressively more double-humped. The response over the band is the flattest when the Q's of

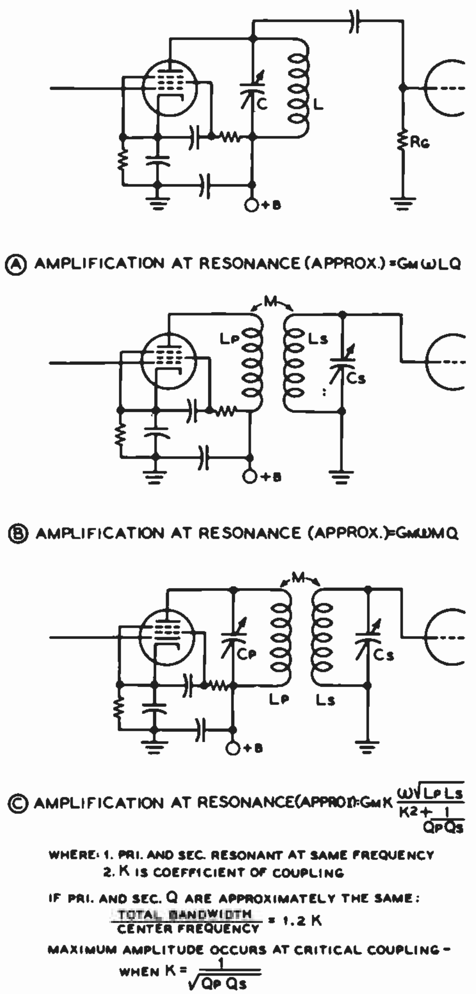


Figure 10

Gain equations for pentode r-f amplifier stages operating into a tuned load

primary and secondary are approximately the same and the value of each Q is equal to $1.75/k$.

Spurious Products It is common practice to control the gain of a succession of r-f or i-f amplifier stages by varying the average bias on their control grids. However, as the bias is raised above the operating value on a conventional sharp-cutoff tube the tube becomes increasingly nonlinear in operation as cutoff of plate current is approached. The effect of such nonlinearity is to cause cross-modulation between strong signals which appear on the grid of the tube. When a tube operating in such a manner is in one of the first stages of a receiver a number of signals are appearing on its grid simultaneously and cross-modulation between them will take place. The result of this effect is to produce a large number of spurious signals in the output of the receiver—in most cases these signals will carry the modulation of both the carriers which have been cross-modulated to produce the spurious signal.

The undesirable effect of cross-modulation can be eliminated in most cases and greatly reduced in the balance through the use of a variable-μ tube in all stages which have avc voltage or other large negative bias applied to their grids. The variable-μ tube has a characteristic which causes the cutoff of plate current to be gradual with an increase in grid bias, and the reduction in plate current is accompanied by a decrease in the effective amplification factor of the tube. Variable-μ tubes ordinarily have somewhat reduced G_m as compared to a sharp-cutoff tube of the same group. Hence the sharp-cutoff tube will perform best in stages to which avc voltage is not applied.

If the desired signal is strong enough, an attenuator pad may be placed between the receiver and the antenna, thus reducing the level of the undesired signal before it does harm. Remote cutoff tubes are advantageous for reducing cross-modulation from strong off-frequency signals.

Cross-modulation is a serious problem in transistorized receivers as most transistors have a very limited dynamic range, the FET types being the best in this respect, although not equal to remote-cutoff tubes.

Mixer Noise Mixer noise of the shot-effect type, which is evidenced by a hiss in the audio output of the receiver, is caused by small irregularities in the current in the mixer stage and will mask weak signals. Noise of an identical nature is generated in an amplifier stage, but due to the fact that the conductance in the mixer stage is considerably lower than in an amplifier stage using the same device, the proportion of inherent noise present in a mixer usually is considerably greater than in an amplifier stage using a comparable device.

Although this noise cannot be eliminated, its effects can be greatly minimized by placing sufficient signal-frequency amplification having a high signal-to-noise ratio ahead of the mixer. This remedy causes the signal output from the mixer to be large in proportion to the noise generated in the mixer stage. Increasing the gain *after* the mixer will be of no advantage in eliminating mixer noise difficulties; greater selectivity after the mixer will help to a certain extent, but cannot be carried too far, since this type of selectivity decreases the i-f bandpass and if carried too far will not pass the sidebands that are an essential part of a voice-modulated signal.

Triode Mixers A triode having a high transconductance is the *quietest* mixer tube, exhibiting somewhat less gain but a better signal-to-noise ratio than a comparable multigrid mixer tube. However, below 30 MHz it is possible to construct a receiver that will get down to the atmospheric noise level without resorting to a triode mixer. The additional difficulties experienced in avoiding *pulling*, undesirable feedback, etc., when using a triode with control-grid injection tend to make multigrid tubes the popular choice for this application on the lower frequencies.

Injection Voltage The amplitude of the injection voltage will affect the conversion transconductance of the mixer. and therefore should be made optimum if maximum signal-to-noise ratio is desired. If fixed bias is employed on the injection element, the optimum injection voltage is quite critical. If cathode or base bias is used, the optimum voltage is not so critical; and if grid-leak bias is employed, the optimum injection voltage is not at all critical—just so it is adequate. Typical optimum injection voltages will run from 1 to 3 volts for proper mixing action.

10-5 R-F Amplifier Stages

Since the necessary tuned circuits between the mixer stage and the antenna can be combined with solid-state devices or tubes to form r-f amplifier stages, the reduction of the effects of mixer noise and enhancement of the image ratio can be accomplished in the input section of the receiver. The tuned input circuits, moreover, provide protection against unwanted signal response but, unfortunately, may increase the susceptibility of the receiver to cross-modulation, blocking, and desensitization because of the enhanced gain level of the received signals. In all cases, receiver gain (and particularly front-end gain) should be limited to that amount necessary to only override mixer noise. Excess receiver gain usually creates more problems than it solves.

If the r-f amplifier stage has its own tuning control, it is often known as a *preselector* Some preselectors employ regeneration to boost signal gain and selectivity at the expense of the signal-to-noise ratio, which usually is degraded in such a circuit.

Generally speaking, atmospheric and man-made noise below about 30 MHz is so high

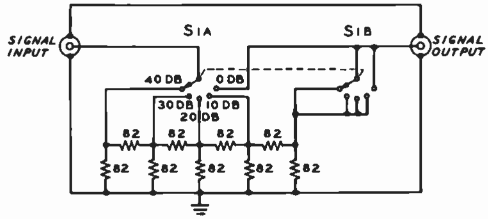


Figure 11

INPUT ATTENUATOR

Communication receiver may be protected from front-end overload and crosstalk by addition of variable attenuator between antenna and input of receiver. Attenuator is built in a small aluminum utility box using ¼-watt composition resistors mounted directly on the switch deck. Attenuator is useful up to 30 MHz or so.

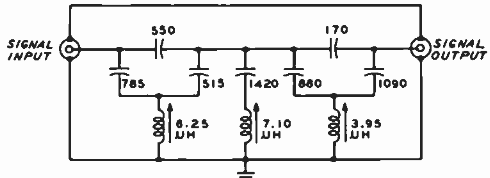


Figure 12

HIGH-PASS INPUT FILTER

High-pass filter reduces crossmodulation and intermodulation from local broadcast stations. At 1.6 MHz, response is down about 40 db. High-Q inductors are wound on Ferroxcube 4C4 pot cores for optimum performance.

that receiver sensitivity and signal-to-noise ratio is not a serious problem. Above 30 MHz or so, noise generated within the receiver is usually greater than the noise received on the antenna. Vhf and uhf r-f amplifiers will be discussed in Section II of this Chapter.

Experience has shown that about an 8-db noise figure is adequate for weak-signal reception under most circumstances below 30 MHz. Interference immunity is very important below 30 MHz because of the widespread use of high-power transmitters and high-gain antennas and large-signal handling ability is usually more important to the h-f communicator than is extreme weak-signal reception.

To minimize receiver overload from strong local signals, a variable attenuator such as the type shown in figure 11 may be placed in the receiver input circuit. The attenuation can be varied from zero to 40 decibels in 10-decibel steps and the unit is useful in dropping the signal level of strong, local transmitters.

A high-pass filter is shown in figure 12 which eliminates cross modulation and intermodulation from local broadcast stations. Both of these devices provide good front-end rejection for unwanted signals.

If the circuit Q is known for the tuned circuits in the r-f stage of the receiver, the image-rejection capability may be determined with the aid of the universal selectivity curve shown in figure 13. The operating Q of the coupled input circuit may be taken as about sixty percent of the unloaded Q and the Q of the output circuit may be estimated to be about eighty percent of the unloaded Q, for frequencies below 30 MHz.

Solid-State R-F Amplifiers Typical common solid-state r-f amplifiers are shown in figure 14. A *common base amplifier* is shown in illustration A. To overcome the possibility of oscillation at the higher frequencies, an external neutralizing circuit may be added, which consists of a neutralizing capacitor placed between the collector and the lower end of the input circuit, which is lifted above ground. If the external feedback circuit cancels both resistive and reactive changes in the input circuit due to voltage feedback, the amplifier is considered to be *unilateralized*. If only the

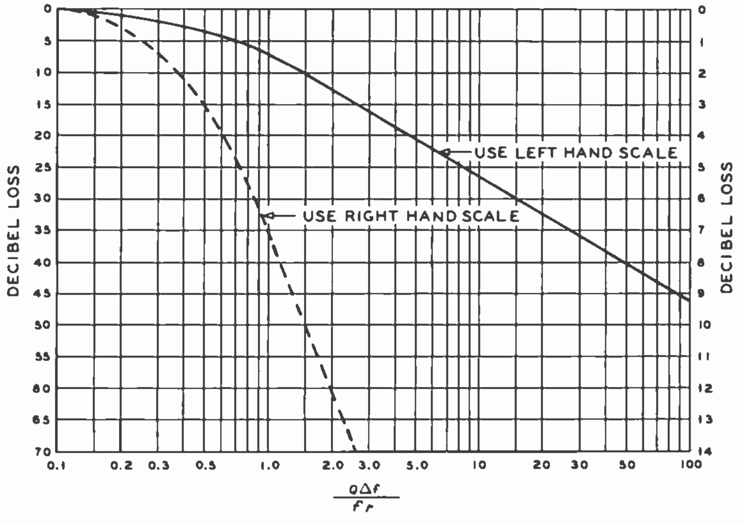


Figure 13

UNIVERSAL SELECTIVITY CURVE

Image rejection capability may be determined with aid of universal curves. Selectivity required to adequately suppress the various spurious signals is provided by tuned input circuits. The number of circuits required depends upon Q, frequency, and attenuation desired. These curves are for a single tuned circuit.

reactive changes in the input circuit are cancelled, the amplifier is considered to be neutralized. Neutralization, then, is a special case of unilateralization. Modern silicon NPN epitaxial planar type transistors are designed for vhf use up to 470 MHz and many have sufficiently low feedback capacitance so that neutralization is unnecessary.

The *common-emitter amplifier* (figure 14B) corresponds to the grounded-cathode vacuum-tube circuit and provides the highest power gain of common transistor circuitry. As the phase of the output signal is opposite to that of the driving signal, the feedback from output to input circuit is essentially negative.

Field-effect transistors may be used in *common-source, common-gate,* or *common-drain* configurations. The common-source arrangement (figure 14C) is most frequently used as it provides high input impedance and medium-to-high output impedance. The first neutralized transistor drives the second connected in common-gate configuration which is used to transform from a low or medium input impedance to a high output impedance. The relatively low voltage gain of the second stage makes dual neutralization unnecessary in most cases. The two FET transistors are arranged in a cascode amplifier circuit, with the first stage inductively neutralized by coil L_N. FET amplifiers of this type have been used to provide low-noise reception at frequencies in excess of 500 MHz. A single gate MOSFET amplifier is shown in figure 14D.

A dual-gate diode protected MOSFET r-f amplifier is shown in figure 15A. The signal input is coupled to gate 1 and the output signal is taken from the drain. Gain control is applied to gate 2 and a d-c sensing current may be taken from the source to be applied to the S-meter circuit, if desired. With proper intrastage shielding, no neutralization of this circuit is required in the h-f region.

An integrated circuit may be used as an r-f amplifier (figure 15B). It is connected as a differential amplifier and provides high gain, good stability and improved agc characteristic as compared to a bipolar device.

A dual-gate MOSFET device is shown in figure 15C and will be more fully discussed in the vhf section of this chapter.

Vacuum-Tube R-F Amplifiers A typical h-f vacuum-tube amplifier circuit is shown in figure 16. A high-gain pen-

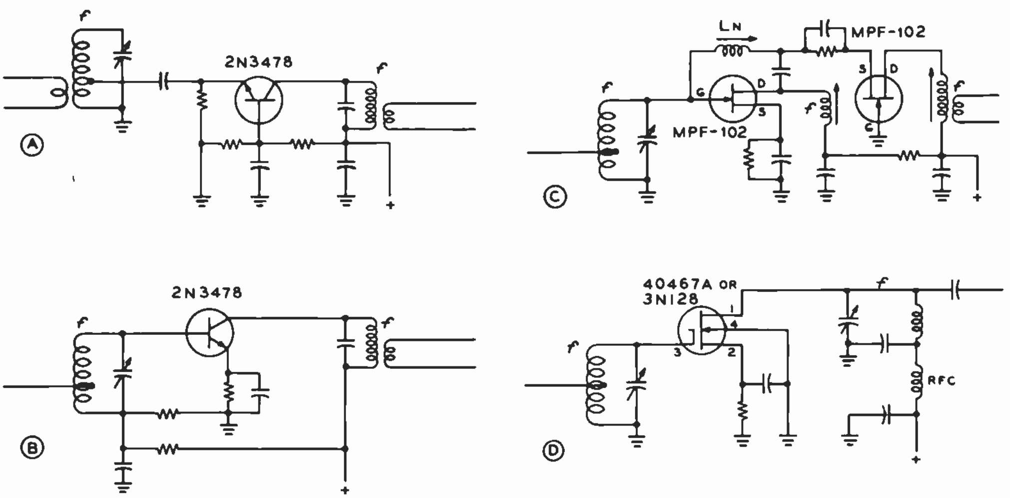


Figure 14
HIGH-FREQUENCY TRANSISTOR R-F STAGES

A—Common-base amplifier. B—Common-emitter amplifier. C—Cascode amplifier using FET transistors in cascode circuit. D—Single-gate MOSFET amplifier.

tode such as a 6BA6 or 6BZ6 may be used with the input circuit connected between grid and cathode. The output signal is taken from the plate circuit. Modern pentode tubes provide very high gain, combined with low grid-to-plate capacitance, and usually do not require neutralization. Remote-cutoff tubes are most often used in r-f amplifier stages because of their superior large-signal handling capability and their good agc characteristics.

With tight coupling to a low-impedance (50-ohm) antenna circuit, the grid circuit of the pentode r-f amplifier is often made adjustable from the receiver panel to allow compensation for reactive antenna terminations. Some r-f pentodes have twin cathode leads, one for input and the other for output terminations, in order to reduce intrastage coupling via a common cathode lead. Tubes having a common cathode lead should employ "common point" bypassing, such as shown in figure 17.

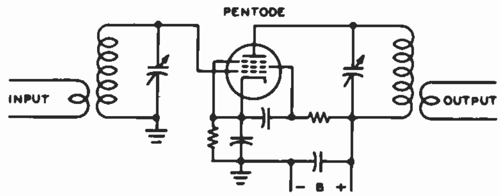


Figure 16

TYPICAL PENTODE R-F AMPLIFIER STAGE

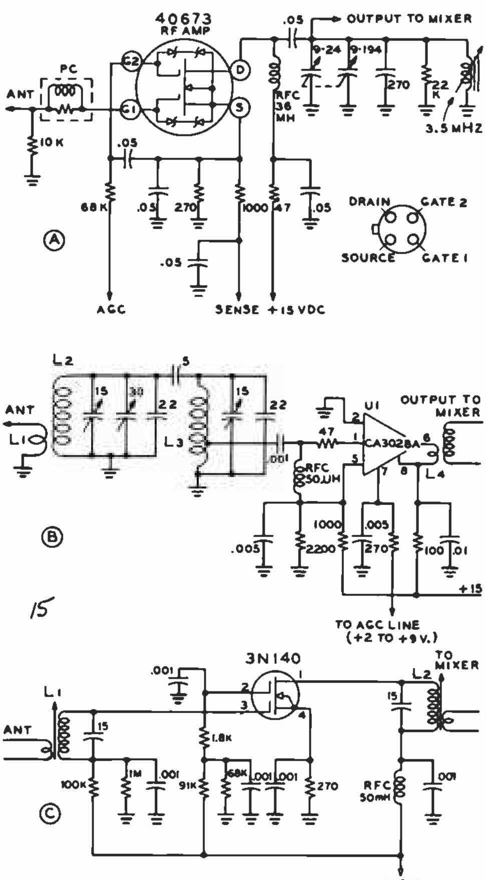


Figure 15

SOLID-STATE R-F AMPLIFIER STAGES

A—Dual-gate, diode-protected MOSFET amplifier. B—Integrated circuit differential amplifier with double-tuned input circuit. C—Dual-gate MOSFET amplifier.

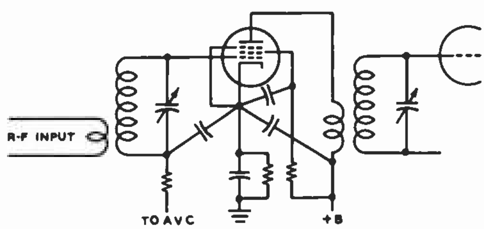


Figure 17

ILLUSTRATING "COMMON-POINT" BYPASSING

To reduce the detrimental effects of cathode circuit inductance in vhf stages, all bypass capacitors should be returned to the cathode terminal at the socket. Tubes with two cathode leads can give improved performance if the grid return is made to one cathode terminal while the plate and screen bypass returns are made to the cathode terminal which is connected to the suppressor within the tube.

Tube input loading, as discussed earlier, influences the gain and selectivity of the r-f amplifier stage to a great degree.

Shown in figure 18 are four types of triode r-f amplifier stages that are useful in the h-f and vhf range up to several hundred MHz. A low noise, grounded-grid amplifier is shown in illustration A. This stage provides medium gain with good intermodulation characteristics and is often used before a low-noise mixer stage. It may be fed directly from a low-impedance transmission line. The *cathode-coupled* circuit of illustra-

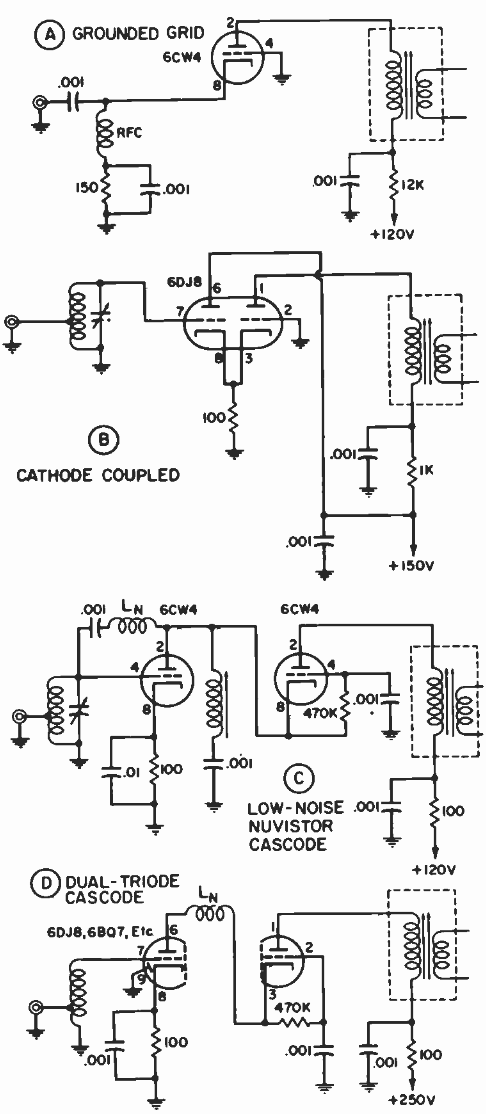


Figure 18

TYPICAL TRIODE VHF R-F AMPLIFIER STAGES

Triode r-f stages contribute the least amount of noise output for a given signal level, hence their frequent use in the vhf range.

tion B provides somewhat more gain than the circuit of illustration A, but an input matching circuit is required.

The effective gain of this circuit is somewhat reduced when it is being used to amplify a broad band of frequencies since the effective G_m of the cathode-coupled dual tubes is somewhat less than half the G_m of either of the two tubes taken alone.

The Cascode Amplifier The *cascode* r-f amplifier is a low-noise circuit employing a grounded-cathode triode driving a grounded-grid triode, as shown in figure 18C. The stage gain of such a circuit is about equal to that of a pentode tube, while the noise figure remains at the low level of a triode tube. Neutralization of the first triode tube is usually unnecessary below 50 MHz. Above this frequency, a definite improvement in the noise figure may be obtained through the use of neutralization. The neutralizing coil (L_N) should resonate at the operating frequency with the grid-plate capacitance of the first triode tube.

The TV-type double triodes such as the 6DJ8 (and older style 6BQ7 and 6BZ7) may be used to good advantage up to 144 MHz or so.

The 6CW4 *nuvistor* is also used in the circuits of figure 18.

Signal-Frequency Circuits The signal-frequency tuned circuits in high-frequency superheterodyne receivers consist of coils of either the solenoid or universal-wound (air or powdered-iron core) type shunted by variable capacitors. It is in these tuned circuits that the causes of success or failure of a receiver often lie. The universal-wound type coils usually are used at frequencies below 2000 kHz; above this frequency the single-layer solenoid type of coil is more satisfactory.

Impedance and Q The two factors of greatest significance in determining the gain-per-stage and selectivity, respectively, of a tuned amplifier are tuned-circuit impedance and tuned-circuit Q. Since the resistance of modern capacitors is low at ordinary frequencies, the resistance usually can be considered to be concentrated in the coil. The resistance to be considered in making Q determinations is the r-f resistance, not the d-c resistance of the wire in the coil. The latter ordinarily is low enough that it may be neglected. The increase in r-f resistance over d-c resistance primarily is due to skin effect and is influenced by such factors as wire size and type, and the proximity of metallic objects or poor insulators, such as coil forms with high losses. Higher values of Q lead to better selectivity and increased r-f voltage across the tuned circuit. The increase in voltage is due to an increase in the circuit impedance with the higher values of Q.

Frequently it is possible to secure an increase in impedance in a resonant circuit (and consequently an increase in gain from an amplifier stage) by increasing the reactance through the use of larger coils and smaller tuning capacitors (higher LC ratio).

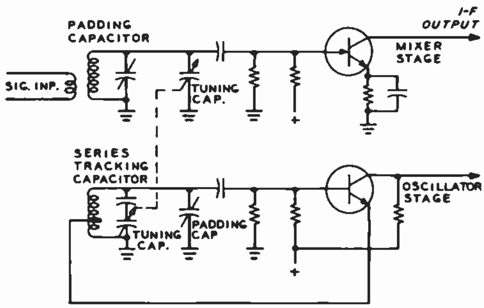


Figure 19

SERIES TRACKING EMPLOYED IN THE H-F OSCILLATOR OF A SUPERHETERODYNE

The series tracking capacitor permits the use of identical gangs in a ganged capacitor, since the tracking capacitor slows down the rate of frequency change in the oscillator so that a constant difference in frequency between the oscillator and the r-f stage (equal to the i-f amplifier frequency) may be maintained.

Superheterodyne Tracking Because the tunable local oscillator in a superheterodyne operates "offset" from the other front-end circuits, it is often necessary to make special provisions to allow the oscillator to track when similar tuning capacitor sections are ganged. The usual method of obtaining good tracking is to operate the oscillator on the high-frequency side of the mixer and use a *series tracking capacitor* to retard the tuning rate of the oscillator. The oscillator tuning rate must be slower because it covers a smaller range than does the mixer when both are expressed as a percentage of frequency. At frequencies above 7000 kHz and with ordinary intermediate frequencies, the difference in percentage between the two tuning ranges is so small that it may be disregarded in receivers designed to cover only a small range, such as an amateur band.

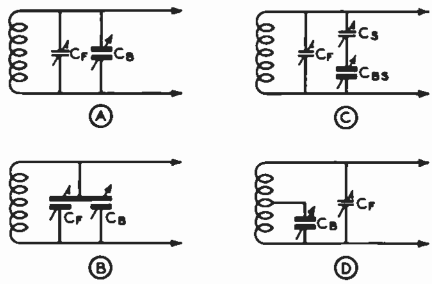


Figure 20

BANDSPREAD CIRCUITS

Parallel bandspread is illustrated at (A) and (B), series bandspread at (C), and tapped-coil band-spread at (D).

A mixer- and oscillator-tuning arrangement in which a series tracking capacitor is provided is shown in figure 19. The value of the tracking capacitor varies considerably with different intermediate frequencies and tuning ranges, capacitances as low as 100 pf being used at the lower tuning-range frequencies, and values up to .01 μfd being used at the higher frequencies.

Superheterodyne receivers designed to cover only a single frequency range, such as the standard broadcast band, sometimes obtain tracking between the oscillator and the r-f circuits by cutting the variable plates of the oscillator tuning section to a different shape than those used to tune the r-f stage. In receivers using large tuning capacitors to cover the shortwave spectrum with a minimum of coils, tuning is likely to be quite difficult, owing to the large frequency range covered by a small rotation of the variable capacitors. To alleviate this condition, some method of slowing down the tuning rate, or *bandspreading*, must be used as shown in figure 20.

Types of Bandspread Bandspreading systems are of two general types: electrical and mechanical. Mechanical systems are exemplified by high-ratio dials in which the tuning capacitors rotate much more slowly than the dial knob. In this system, there is often a separate scale or pointer either connected or geared to the dial knob to facilitate accurate dial readings. However, there is a practical limit to the amount of mechanical bandspread which can be obtained in a dial and capacitor before the speed-reduction unit and capacitor bearings become prohibitively expensive. Hence, most receivers employ a combination of electrical and mechanical bandspread. In such a system, a moderate reduction in the tuning rate is obtained in the dial, and the rest of the reduction obtained by *electrical bandspreading.*

Stray Circuit Capacitance In this book and in other radio literature, mention is sometimes made of *stray* or *circuit capacitance.* This capacitance is in the usual sense defined as the capacitance remaining across a coil when all the tuning, bandspread, and padding capacitors across the circuit are at their minimum capacitance setting.

Circuit capacitance can be attributed to two general sources. One source is that due to the input and output capacitance of the tube when its cathode is heated. The input capacitance varies somewhat from the static value when the tube is in actual operation. Such factors as plate load impedance, grid bias, and frequency will cause a change in input capacitance. However, in all except the extremely high-transconductance tubes, the published measured input capacitance is reasonably close to the effective value when the tube is used within its recommended frequency range. But in the high-transconductance types the effective capacitance will vary considerably from the published figures as operating conditions are changed.

The second source of circuit capacitance, and that which is more easily controllable, is that contributed by the minimum capacitance of the variable capacitors across the circuit and that due to capacitance between the wiring and ground. In well-designed high-frequency receivers, every effort is made to keep this portion of the circuit capacitance at a minimum since a large capacitance reduces the tuning range available with a given coil and prevents a good LC ratio, and consequently a high-impedance tuned circuit, from being obtained.

A good percentage of stray circuit capacitance is due also to distributed capacitance of the coil and capacitance between wiring points and chassis.

10-6 Mixer Stages

The mixer, or frequency-converter stage of a superhet receiver translates the received signal to the intermediate frequency by means of a modulation process similar to that employed in transmitters (figure 21). The signal and local-oscillator voltages appearing in the output circuit of the mixer are rejected by selective circuits and only the mixer product at the intermediate frequency is accepted.

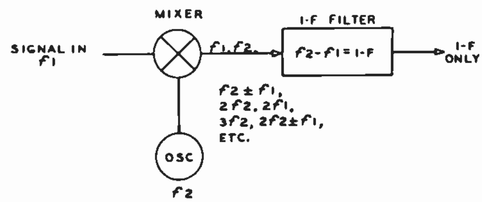


Figure 21

RECEIVER MIXER STAGE

Received signal is translated to intermediate frequency by the mixer stage. Signal and local-oscillator voltages and various mixer products are rejected by selective circuits in i-f amplifier and only the mixer product at the intermediate frequency is accepted.

Any nonlinear circuit element will act as a mixer, with the injection frequencies and sum and difference frequencies appearing in the output circuit. Thus any diode, vacuum tube, or solid-state device may be used as a mixer. High-frequency communication receivers often use special multigrid mixer tubes because of their high conversion gain and good order of isolation between the signal and local-oscillator circuits. Triode mixers, are also used for their low noise figure, but are more susceptible to interlocking adjustments between the two input circuits.

Figure 22 illustrates some of the more common vacuum-tube and transistor mixer circuits.

The *pentagrid converter* tube is shown in figure 22A. Tubes of this type are good conversion devices at medium frequencies, although their performance drops off as the frequency of operation is raised. Their use is practical up to 50 MHz or so. Electrically, grids 2 and 4 shield the signal grid from the oscillator section and also act as an anode for the electron-coupled oscillator portion

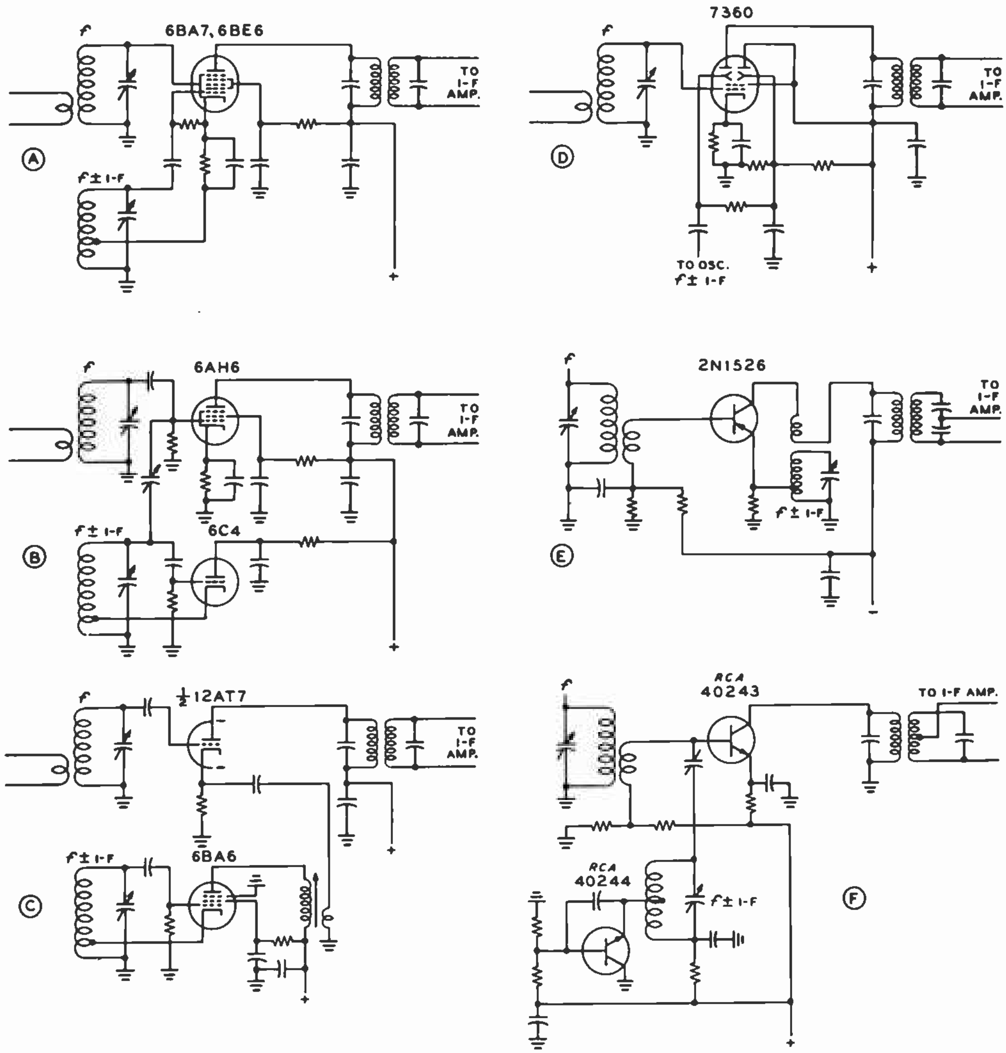


Figure 22

TYPICAL FREQUENCY-CONVERTER (MIXER) STAGES

A—Pentagrid converter. B—Pentode mixer with grid injection. C—Triode mixer with cathode injection. D—Beam-deflection mixer. E—Transistor self-oscillating (autodyne) mixer. F—Transistor mixer with base injection.

of the tube which is composed of grid 1 and the cathode. The pentagrid converter is characterized by an equivalent noise resistance of several hundred thousand ohms, consequently it must be preceded by an r-f stage having a fairly high gain figure if a low noise factor is desired in the receiver.

A second frequency-conversion technique utilizes a separate oscillator tube and a pentode mixer (figure 22B). The local oscillator voltage is applied to the mixer control grid by capacitive or inductive coupling, or a combination of the two. Tubes containing electrically independent oscillator and mixer units in the same envelope, such as the 6U8A, 6KZ8, and 6EA8 are designed especially for this application and find use in TV tuners and f-m receivers. Another version of this circuit is shown in figure 22C utilizing a low-noise triode mixer with cathode injection from a separate electron-coupled oscillator. This circuit has a wide dynamic range and is capable of mixing relatively strong signals while retaining a low level of intermodulation distortion. Tube types 6DJ8 and 6CW4 are often used for h-f and vhf mixer operation.

A beam-deflection tube (7360) may be used as a mixer in the circuit of figure 22D, providing low noise figure and high mixer gain. The incoming signal is applied to control grid 1 and the mixing signal from the separate local oscillator is applied to the deflection plates. The electron stream is modulated by the received signal and then switched from one collector anode to the other by the local oscillator switching voltage. The i-f output signal is taken from one deflection anode of the 7360 in the usual manner, although a push-pull output configuration may be used for improved local-oscillator rejection at the intermediate frequency.

Inexpensive transistor radios make use of an *autodyne mixer* such as shown in figure 22E. The oscillator circuit is placed in the emitter lead, with inductive feedback to the collector circuit. In the h-f range stable mixer operation is more readily obtained when a separate transistor is used for the oscillator function. In the latter arrangement, the oscillator voltage is injected in the mixer stage by inductive coupling to the emitter, or a combination of inductive and capacitive coupling to emitter or base may be used (figure 22F).

Mixers employing control-grid injection of the local mixing signal (figure 22B, for example) should be preceded by an r-f stage if local oscillator spurious radiation is to be held to a minimum.

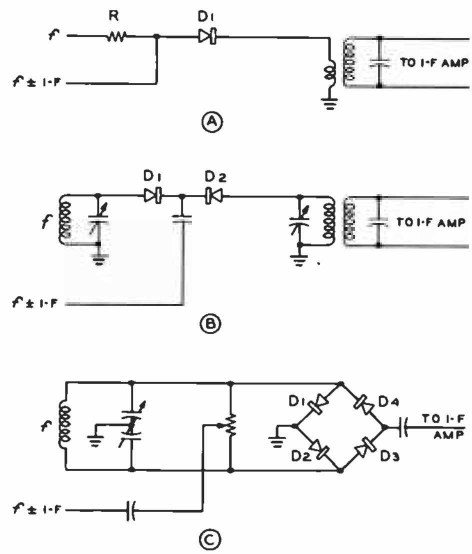


Figure 23

DIODE MIXER STAGES

A—Single diode mixer. B—Double diode mixer with input signal applied in series and mixing signal applied in parallel to diodes. C—Balanced ring modulator with carrier balance controls.

Diode Mixers Typical diode mixers are shown in figure 23. A simple (and inefficient) single-diode circuit is shown in illustration A. The input signal is attenuated below the local-oscillator signal by resistor R to provide low-distortion mixing action. A double-diode mixer is shown at B, the mixing signal being applied in parallel to the diode cathode terminals. The input signal is applied in series with the two diodes. Mixing produces a product of the signals, instead of sums and differences, and this circuit is often termed a *product mixer.*

One form of balanced ring diode modulator is shown in illustration C. The input signal is fed into the ring in push-pull mode and the output signal is taken in the parallel mode. The local oscillator is applied in the parallel mode. Various versions of diode mixers are used as SSB modulators and demodulators in amateur and commercial equipment, as well as mixers in vhf and uhf receivers.

FET and MOSFET Mixers Typical FET mixer circuits are shown in figure 24. These circuits are preferred over bipolar mixer circuits because the dynamic characteristics of bipolar transistors prevent them from handling high signal levels without severe intermodulation distortion. Illustration A shows a junction FET with signal and oscillator frequencies applied to the gate. Source injection is shown at B. Both circuits can handle high input signal levels without overloading.

A dual gate MOSFET is shown in a typical mixer circuit in figure 25A. The unit shown has no internal chip protection and great care must be taken during installation to prevent the thin dielectric material of the gate from being punctured by static electricity. All leads should be shorted together until after the device is connected in the circuit. The MOSFET should be handled by its case and it should never be inserted or removed from a circuit when operating voltages are applied.

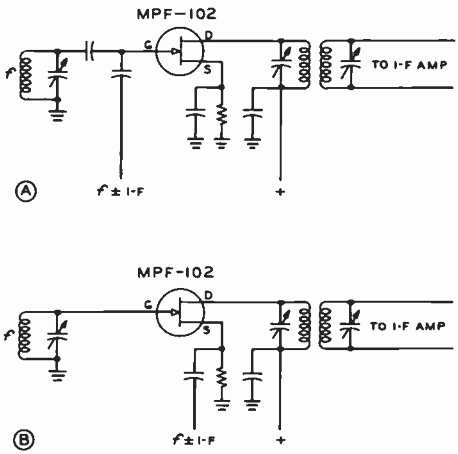


Figure 24

TYPICAL FET MIXER STAGES

A—Junction FET mixer with gate injection. B—JFET mixer with source injection.

The dual gate MOSFET shown in illustration B has internal protection diodes that allow it to be handled with ordinary care. Both circuits offer high conversion gain, relative immunity from cross modulation, and do not load the local oscillator heavily.

10-7 The Local Oscillator

The exact frequency of reception of a superheterodyne receiver is governed by the frequency of the mixing oscillator or oscillators. The overall stability of the receiver, moreover, is determined by the frequency

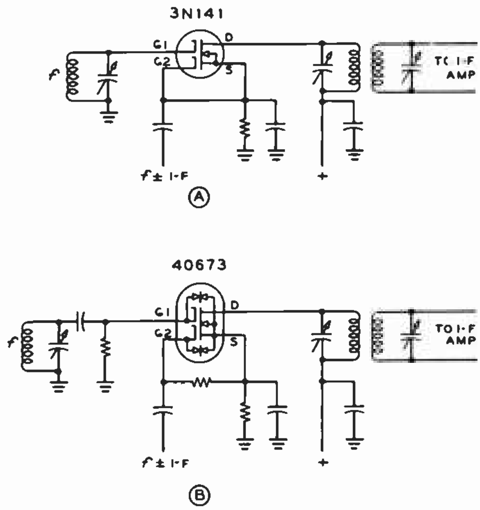


Figure 25

TYPICAL MOSFET MIXER STAGES

A—Dual-gate MOSFET mixer. B—Dual-gate MOSFET with diode protection. Both circuits offer high conversion gain and relative immunity from crossmodulation.

stability of the oscillator. The frequency accuracy for SSB reception is rather precise when compared with most other communication systems. A frequency error of, say, 50 Hz in carrier reinsertion results in noticeable voice distortion, and intelligibility is impaired when the frequency error is 150 Hz or greater.

Oscillator stability should be relatively immune to mechanical shock and temperature rise of the receiver. A tunable oscillator should have good resetability and tuning should be smooth and accurate. Construction should be sturdy, with short, heavy interconnecting leads between components, that resist vibration. Variable capacitors should be mounted so that no strain exists on the bearings and the capacitors should be selected to have good, low-inductance wiping contacts that will resist aging.

The oscillator coil should be preferably wound on a ceramic form and the winding should be locked in position for maximum stability. Variable inductors with movable cores should be avoided if possible, because of possible movement of the core under vibration.

Oscillator Circuits Mixer tubes having internal oscillator sections were shown in figure 22. Several separate oscillator circuits are shown in figure 26. A typical vacuum-tube oscillator is shown in illustration A. This is a Hartley, grounded-plate circuit. Feedback is obtained via a cathode tap on the grid coil. Mixing voltage is taken from the grid of the tube, or may be taken from the cathode for better circuit isolation. At the higher frequencies, the cathode-filament capacitance of the tube may introduce 60-Hz frequency modulation into the oscillator signal via filament-cathode leakage and a plate coil feedback circuit is to be preferred, as shown in illustration B. Other oscillator circuits shown in the chapter, "Generation of R-f Energy" may be used for local oscillators in receivers.

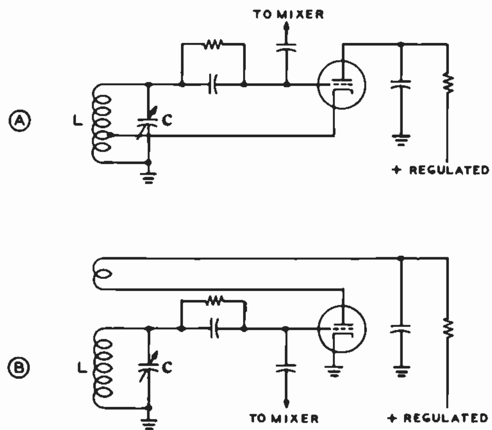


Figure 26

TYPICAL VACUUM-TUBE OSCILLATOR CIRCUITS

A—Hot cathode circuit with plate at r-f ground potential. B—Plate feedback circuit with cathode at ground potential.

Solid-State Oscillators Transistor local oscillator circuitry is employed in most modern SSB receivers. A bipolar circuit is shown in figure 27A. The base element is near r-f ground potential and feedback is between the collector and the emitter. A JFET oscillator circuit (B) and a MOSFET circuit (C) are shown for comparison. The diode placed between gate and ground limits the level of gate bias to improve oscillator stability.

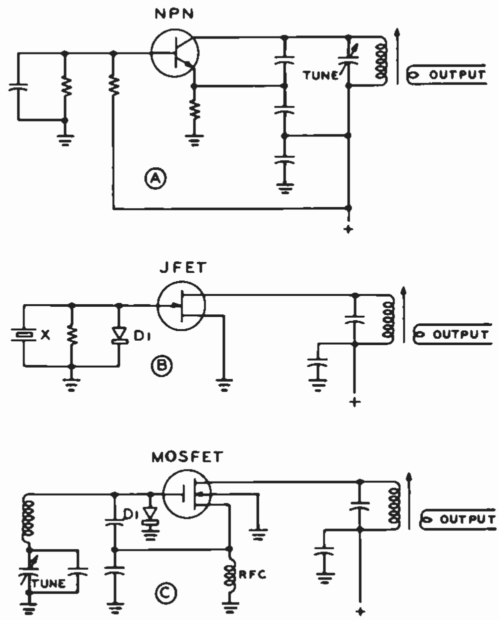


Figure 27

TYPICAL SOLID-STATE OCSILLATOR CIRCUITS

A—Bipolar transistor with emitter feedback from collector. B—JFET crystal oscillator. C—MOSFET oscillator. Diode D_1 between gate and ground limits level of gate bias to improve oscillator stability.

Because of the nonlinear change in the collector-base capacitance during oscillator operation, most transistor oscillators exhibit a high level of harmonic energy. A low-pass filter may be required after the oscillator to minimize spurious response in the receiver caused by mixing between unwanted signals and oscillator harmonics. In addition, one or more buffer stages may be required between oscillator and mixer to prevent the mixer from "pulling' the oscillator frequency when the strength of the incoming signal varies up and down.

The Frequency Synthesizer A higher order of accuracy of frequency control for both receiver and transmitter may be achieved by crystal control of the various conversion oscillators. Multiple-frequency operation, however, call for an uneconomical and bulky number of crystals. These problems are solved by the use of a *frequency synthesizer* (figure 28). This is a device in which the harmonics and subharmonics of one or more oscillators are mixed to provide a multiplicity of output frequencies, all of which are harmonically related to a subharmonic of the master oscillator. A discussion of the frequency synthesizer is included in Chapter 11, "The Generation of Radio-Frequency Energy."

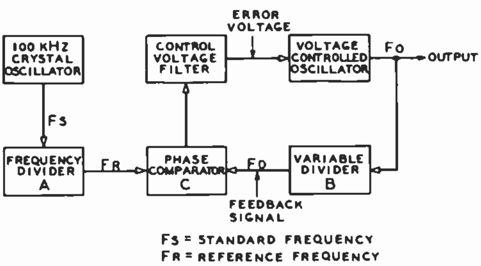


Figure 28

FREQUENCY SYNTHESIZER

Subharmonics (F_R) of crystal oscillator are compared with divided signal (F_D) of voltage-controlled variable oscillator. Error signal corrects frequency of voltage-controlled oscillator.

10-8 The I-F Amplifier

The main voltage gain of a superhet receiver is achieved in the i-f amplifier stages. Intermediate-frequency amplifiers commonly employ bandpass circuits which can be arranged for any degree of selectivity, depending on the ultimate application of the amplifier. I-f amplifier circuitry is very similar to those circuits discussed for r-f amplifiers earlier in this chapter and the stage gain of the i-f chain may be controlled by an automatic gain control circuit actuated by the received signal.

Choice of Intermediate Frequency The intermediate frequency used is a compromise between high gain, good selectivity, and image rejection. The lower the frequency, the higher will be the gain and selectivity, and the lower the image rejection of the particular receiver. Conversely, the higher the i-f, the lower the gain and selectivity will be and the higher the image rejection. By traditional usage and international agreement, the most commonly used intermediate frequencies are 262 kHz, 455 kHz, and 1600 kHz for communication and entertainment receivers. Some sideband equipments make use of crystal-filter i-f systems in the 5-MHz to 9-MHz range and vhf equipment may have intermediate frequencies as high as 50 MHz. When a high value of i-f is employed, it is common technique to convert the signal a second time to a lower intermediate frequency in order to pick up gain and selectivity that cannot be economically achieved in the higher i-f.

I-F Transformers Intermediate-frequency transformers commonly consist of two or more tuned circuits cuits together. Some representative arrangements are shown in figure 29. The circuit shown at A is the conventional i-f transformer, with the coupling (M) between the tuned circuits being provided by inductive coupling from one coil to the other. As the coupling is increased, the selectivity curve becomes less peaked, and when a condition known as *critical coupling* is reached, the top of the curve begins to flatten out. When the coupling is increased still more, a dip occurs in the top of the curve.

The windings for this type of i-f transformer, as well as most others, nearly always consist of small, flat universal-wound pies mounted either on a piece of dowel to provide an air core or on powdered iron for *iron-core* i-f transformers. The iron-core transformers generally have somewhat more gain and better selectivity than equivalent air-core units.

The circuits shown at figure 29B and C are quite similar. Their only difference is the type of mutual coupling used, an inductance being used at B and a capacitance at C. The operation of both circuits is similar. Three resonant circuits are formed by the components. In B, for example, one resonant circuit is formed by L_1, C_1, C_2, and L_2 all in series. The frequency of this resonant circuit is just the same as that of a single one of the coils and capacitors, since the coils and ca-

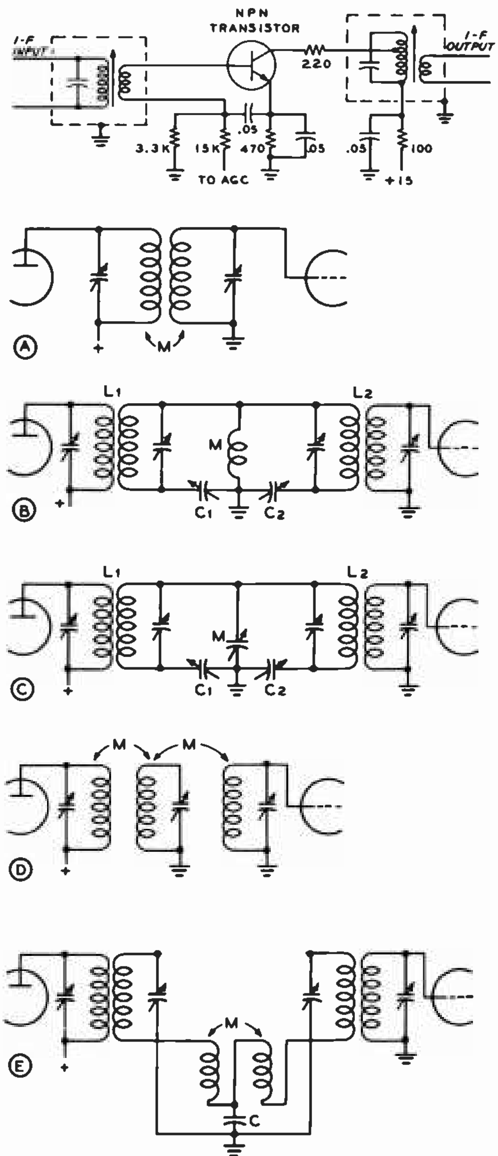


Figure 29

I-F AMPLIFIER COUPLING ARRANGEMENTS

The interstage coupling arrangements illustrated above give a better shape factor (more straight-sided selectivity curve) than would the same number of tuned circuits coupled by means of tubes.

pacitors are similar in both sides of the circuit, and the resonant frequency of the two capacitors and the two coils all in series is the same as that of a single coil and capacitor. The second resonant frequency of the complete circuit is determined by the characteristics of each half of the circuit containing the mutual coupling device. In B, this second frequency will be lower than the first, since the resonant frequency of L_1, C_1, and inductance M; or L_2, C_2, and M is lower than that of a single coil and capacitor, due to the inductance of M being added to the circuit.

The opposite effect takes place in figure 29C, where the common coupling impedance is a capacitor. Thus, at C the second resonant frequency is higher than the first. In either case, however, the circuit has two resonant frequencies, resulting in a flat topped selectivity curve. The width of the top of the curve is controlled by the reactance of the mutual coupling component. As this reactance is increased (inductance made greater, capacitance made smaller), the two resonant frequencies become further apart and the curve is broadened.

In the circuit of figure 29D, there is inductive coupling between the center coil and each of the outer coils. The result of this arrangement is that the center coil acts as a sharply tuned coupler between the other two. A signal somewhat off the resonant frequency of the transformer will not induce as much current in the center coil as will a signal of the correct frequency. When a smaller current is induced in the center coil, it in turn transfers a still smaller current to the output coil. The effective coupling between the outer coils increases as the resonant frequency is approached, and remains nearly constant over a small range and then decreases again as the resonant band is passed.

Another very satisfactory bandpass arrangement, which gives a very straight-sided, flat-topped curve, is the negative mutual arrangement shown in figure 29E. Energy is transferred between the input and output circuits in this arrangement by both the negative mutual coils (M) and the common capacitive reactance (C). The negative mutual coils are interwound on the same form, and connected *backward*.

Shape Factor It is obvious that to accept a single sideband the i-f amplifier must pass not a single frequency but a band of frequencies. The width of this passband, usually 2 kHz to 3 kHz in a good communication receiver, is known as the *passband*, and is arbitrarily taken as the width between the two frequencies at which the response is attenuated 6 db, or is "6 db down." However, it is apparent that to discriminate against an interfering signal which is stronger than the desired signal, much more than 6 db attenuation is required. The attenuation commonly chosen to indicate adequate discrimination against an interfering signal is 60 db.

It is apparent that it is desirable to have the bandwidth at 60 db down as narrow as possible, but it must be done without making the passband (6-db points) too narrow for satisfactory reception of the desired signal. The figure of merit used to show the ratio of bandwidth at 6 db down to that at 60 db down is designated as *shape factor*. The ideal i-f curve (a rectangle), would have a shape factor of 1.0. The i-f shape factor in typical communications receivers runs from 2.0 to 5.5.

The most economical method of obtaining a low shape factor for a given number of tuned circuits is to employ them in pairs, as in figure 29A, adjusted to *critical coupling* (the value at which two resonance points just begin to become apparent). If this gives too sharp a *nose* or passband, then coils of lower Q should be employed, with the coupling maintained at the critical value. As the Q is lowered, closer coupling will be required for critical coupling.

Conversely if the passband is too broad, coils of higher Q should be employed, the coupling being maintained at critical. If the passband is made more narrow by using looser coupling instead of raising the Q and maintaining critical coupling, the shape factor will not be as good.

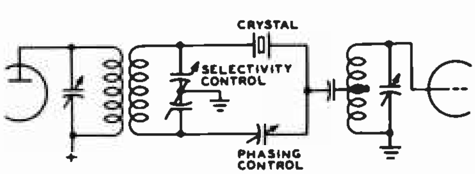


Figure 30

TYPICAL CRYSTAL FILTER CIRCUIT

The *passband* will not be much narrower for several pairs of identical, critically coupled tuned circuits than for a single pair. However, the *shape factor* will be greatly improved as each additional pair is added, up to about 5 pairs, beyond which the improvement for each additional pair is not significant. The passband of a typical communication receiver is shown in figure 4.

Miller Effect As mentioned previously, the dynamic input capacitance of a tube varies slightly with bias. As agc voltage normally is applied to i-f tubes, the effective grid-cathode capacitance varies as the signal strength varies, which produces the same effect as slight detuning of the i-f transformer. This effect is known as *Miller effect*, and can be minimized to the extent that it is not troublesome either by using a fairly low LC ratio in the transformers or by incorporating a small amount of degenerative feedback, the latter being most easily accomplished by leaving part of the cathode resistor unbypassed for radio frequencies.

The passband of an intermediate-frequency amplifier may be made very narrow through the use of a *piezoelectric crystal* filter employed as a series-resonant circuit in a bridge arrangement known as a *crystal filter*. The shape factor is quite poor, as would be expected when the selectivity is obtained from the equivalent of a single tuned circuit, but the very narrow passband obtainable as a result of the extremely high Q of the crystal makes the crystal filter useful for c-w telegraphy reception. The passband of a 455-kHz crystal filter may be made as narrow as 50 Hz while the narrowest passband that can be obtained with a 455-kHz tuned circuit of practical dimensions is about 5 kHz. A crystal filter for c-w operation is shown in figure 30.

The Crystal Filter It is necessary to balance out the capacitance across the crystal holder to prevent bypassing around the crystal undesired signals off the crystal resonant frequency. The balancing is done by a *phasing* circuit which takes out-of-phase voltage from a balanced input circuit and passes it to the output side of the crystal in proper phase to neutralize that passed through the holder capacitance.

Rejection Notch The single crystal filter of figure 30 has both a resonant (series-resonant) and an antiresonant (parallel-resonant) frequency—the impedance of the crystal being quite low at the former frequency, and quite high at the latter frequency. The antiresonant frequency is just slightly higher than the resonant frequency, the difference depending on the effective shunt capacitance of the filter crystal and holder. As adjustment of the phasing capacitor controls the effective shunt capacitance of the crystal, it is possible to vary the antiresonant frequency of the crystal slightly without unbalancing the circuit sufficiently to let undesired signals *leak through* the shunt capacitance in appreciable amplitude. At the exact antiresonant frequency of the crystal the attenuation is exceedingly high because of the high impedance of the crystal at this frequency. This is called the *rejection notch,* and can be utilized to virtually eliminate the heterodyne image or *repeat tuning* of c-w signals. The beat-frequency oscillator can be so adjusted and the phasing capacitor so adjusted that the desired beat note is of such a pitch that the image (the same audio note on the other side of zero beat) falls in the rejection notch and is inaudible. The receiver then is said to be adjusted for *single-signal* operation (figure 31).

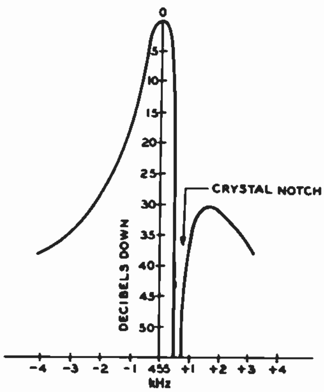


Figure 31

I-F PASSBAND OF TYPICAL CRYSTAL FILTER COMMUNICATION RECEIVER

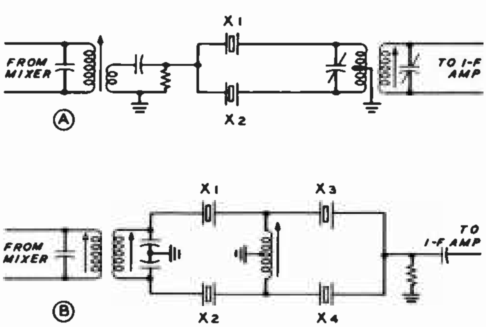


Figure 32

BANDPASS CRYSTAL FILTERS

A—Dual crystal filter. B—Multiple crystal filter improves passband response.

Bandpass Crystal Filters The sharply peaked response of the single-crystal filter is adequate for c-w reception but has a poor shape factor for voice reception. A *bandpass filter,* which passes a band of frequencies, is much more suitable for this mode. Typically, a good bandpass filter for SSB reception might have a passband of about 2 kHz or so at 6 db down, and perhaps 8 to 10 kHz at 60 db down. Typical crystal bandpass filters are shown in figure 32. A simple filter utilizing two crystals is shown in illustration A. The series resonance of the crystals differs by an amount equal to the desired bandwidth. To improve the shape factor of the passband, additional crystals may be added to the filter as shown at B. Provided there is no leakage of signal around the filter, extremely good shape factors can be achieved with relatively inexpensive crystal filters, operating at a center frequency as high as 50 MHz. Vhf filters, moreover, have been used in commercial and military communication systems.

The Mechanical Filter The *mechanical filter* is an electromechanical bandpass device about half the size of a cigarette package. As shown in figure 33, it consists of an input transducer, a resonant mechanical section comprised of a number of metal discs, and an output transducer.

The frequency characteristics of the reso-

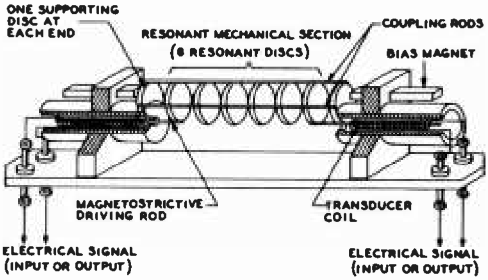


Figure 33

MECHANICAL FILTER FUNCTIONAL DIAGRAM

nant mechanical section provide the almost rectangular selectivity curves shown in figure 34. The input and output transducers serve only as electrical-to-mechanical coupling devices and do not affect the selectivity characteristics which are determined by the metal discs. An electrical signal applied to the input terminals is converted into a mechanical vibration at the input transducer by means of *magnetostriction*. This mechan-

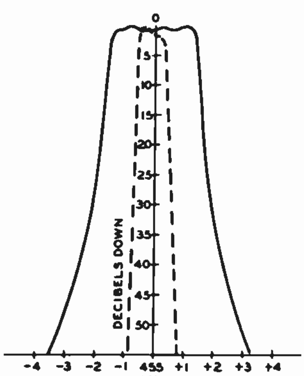


Figure 34

Selectivity curves of 455-kHz mechanical filters with nominal 0.8-kHz (dotted line) and 3.1-kHz (solid line) bandwidth at -6 db.

ical vibration travels through the resonant mechanical section to the output transducer, where it is converted by magnetostriction to an electrical signal which appears at the output terminals.

In order to provide the most efficient electromechanical coupling, a small magnet in the mounting above each transducer applies a magnetic bias to the nickel transducer core. The electrical impulses then add to or subtract from this magnetic bias, causing vibration of the filter elements which corresponds to the exciting signal. There is no mechanical motion except for the imperceptible vibration of the metal discs.

Magnetostrictively driven mechanical filters have several advantages over electrical equivalents. In the region from 100 kHz to 500 kHz, the mechanical elements are extremely small, and a mechanical filter having better selectivity than the best of conventional i-f systems may be enclosed in a package smaller than one i-f transformer.

The frequency characteristics of the mechanical filter are permanent, and no adjustment is required or is possible. The filter is enclosed in a hermetically sealed case.

In order to realize full benefit from the mechanical filter's selectivity characteristics, it is necessary to provide shielding between the external input and output circuits, capable of reducing transfer of energy external to the filter by a minimum value of 100 db. If the input circuit is allowed to couple energy into the output circuit external to the filter, the excellent skirt selectivity will deteriorate and the passband characteristics will be distorted.

As with almost any mechanically resonant circuit, elements of the mechanical filter have multiple resonances. These result in spurious modes of transmission through the filter and produce minor passbands at frequencies outside the primary passband. Design of the filter reduces these subbands to a low level and removes them from the immediate area of the major passband. Two conventional i-f transformers supply increased attenuation to these spurious responses, and are sufficient to reduce them to an insignificant level.

Diode Filter Switching Two filters of different bandwidths are commonly used for SSB and c-w reception. Mechanical switching of such filters may lead to unwanted coupling between input and output, thus seriously degrading the shape factor of the filter. By using diode-controlled switching (figure 35), the switching components may be placed close to the filter terminals, thus offering a minimum of deterioration in isolation between ports. The diodes are triggered by a panel switch, and the appropriate diode pair places the desired filter in the i-f signal path. Operation of switch S_1 forward-biases a pair of diodes at a time and reverse-biases the other pair, allowing one filter to function at a time.

The Transfilter A small mechanical resonator (*transfilter*) may be used in place of an i-f transformer in transistor i-f circuits (figure 36A). A second transfilter resonator may be substituted for the conventional emitter bypass capacitor to enhance i-f selectivity. Transfilters may also be employed in the high-Q oscillator tuned circuits. The passband of a single transfilter i-f stage with emitter resonator is shown in figure 36B.

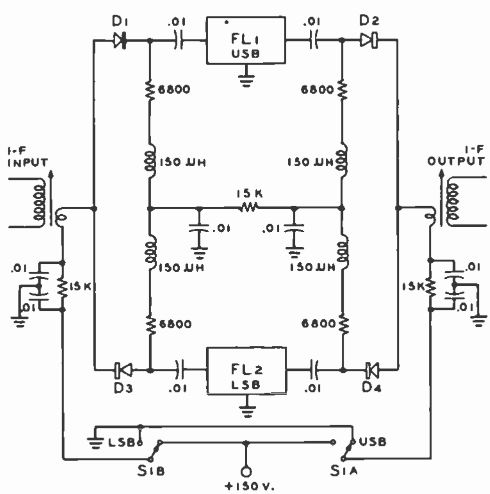


Figure 35

DIODE FILTER SWITCHING

Diode-controlled switching reduces unwanted coupling between input and output circuits of filters, thus preserving shape factor of the filter. Appropriate diode pairs are triggered by panel switch (S_1). One diode pair is forward-biased at a time, allowing proper filter to function.

Bilateral Amplifier A *bilateral amplifier* is one that amplifies in two signal directions (figure 37). Such a stage is useful in SSB transceivers wherein r-f and i-f stages function in both receive and transmit modes. During the receive function, the bilateral amplifier passes the signal from the mixer to the balanced modulator and during transmit it passes the signal in the opposite direction —from the balanced modulator to the mixer. The same tuned circuits are used for both transmitting and receiving. The various injection oscillators operate continuously, supplying the local mixing signals to the proper mixer stages.

In the circuit shown, the amplifier operates in the common-emitter configuration. In the receive mode, the 33K base-bias resistor is returned to the receiver cutoff-bias control line, disabling transistor Q_1.

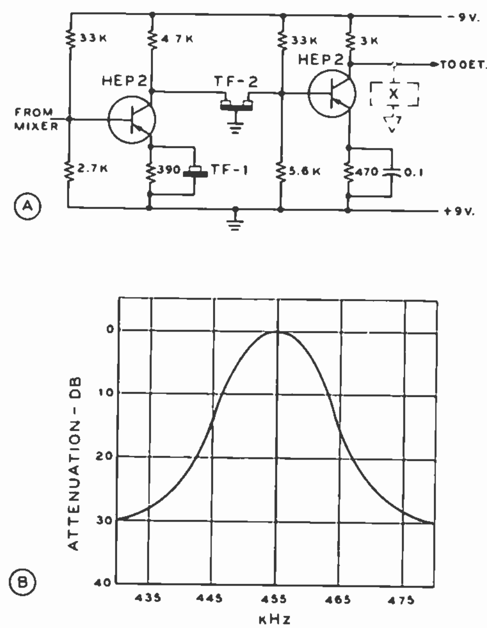


Figure 36

MECHANICAL RESONATOR USED AS I-F FILTER

A—Transistorized i-f amplifier using Transfilters (TF-1, TF-2). Addition of second Transfilter (X) will sharpen selectivity. B—Passband of single Transfilter i-f stage with emitter resonator.

The 15K base-bias resistor of transistor Q_2 is returned to the transmitter bias-control circuit, which is at ground potential when the VOX relay is actuated. Thus, in the receive mode, a signal appearing at the receiver i-f transformer (T_2) will be amplified by transistor Q_2 and delivered to the i-f transformer (T_1). When the VOX circuit is activated to the transmit mode, the two bias-control lines are inverted in polarity so that transistor Q_2 is cut off and Q_1 is able to conduct. Therefore, a signal

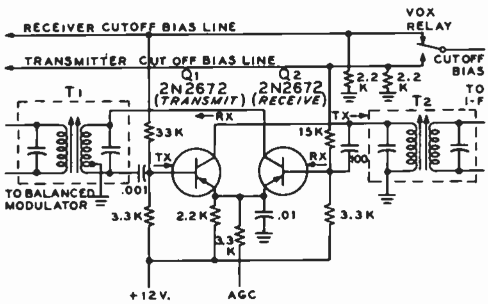


Figure 37

BILATERAL I-F AMPLIFIER FOR TRANSCEIVER

Bilateral i-f amplifier stage functions in both receive and transmit modes in SSB transceiver. Cutoff-bias lines transfer operation from transistor Q_1 to transistor Q_2 as VOX relay is actuated. Common-emitter stages are used with base-bias control.

appearing at transformer T_1 is amplified by Q_1 and impressed on transformer T_2. Unilateral stages that are not required on either transmit or receive may be turned off by returning their base-bias resistors to an appropriate cutoff-bias control line.

10-9 The Beat-Frequency Oscillator

The *beat-frequency oscillator* (*bfo*) or *carrier-injection oscillator* is a necessary adjunct to the communication receiver for the reception of c-w or SSB signals.

The oscillator is coupled into or just ahead of the second detector circuit and supplies a signal of nearly the same frequency as that of the desired signal from the i-f amplifier. If the i-f amplifier is tuned to 455 kHz, for example, the bfo is tuned to approximately 454 or 456 kHz to produce an audible (1000-Hz) c-w beat note in the output of the second detector of the receiver. The carrier signal itself is, of course, inaudible. The bfo is not used for a-m reception, except as an aid in searching for weak stations.

Care must be taken with the bfo to prevent harmonics of the oscillator from being picked up at multiples of the bfo frequency. The complete bfo together with the coupling circuits to the second detector, should be thoroughly shielded to prevent pickup of the bfo harmonics by the input circuitry of the receiver. The local h-f oscillator circuits shown in Section 10-7 may be used for beat-frequency oscillators, as can the various oscillator circuits shown in the chapter "Generation of R-F Energy."

Many modern SSB receivers employ separate crystal-controlled beat-frequency oscillators to provide upper- and lower-sideband reception (figure 38). D-c switching is used in this particular circuit which is preferable to crystal switching. A buffer stage isolates the oscillators from the load, while increasing the bfo voltage to the proper level for the detector stage. The crystals are placed on the correct frequencies by means of the trimming capacitors.

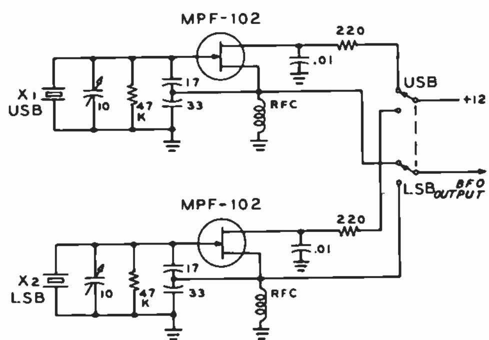


Figure 38

DUAL BFOs FOR USB AND LSB RECEPTION

10-10 Detectors and Demodulators

Conventional detectors for amplitude-modulated signals are shown in figure 39. The *grid-leak detector* (A) is capable of excellent signal gain but has poor strong-signal capability and is little used except in portable receivers. The diode detector (B) functions directly on a-m signals and may be used for SSB reception in conjunction with a carrier oscillator. The diode detector allows a simple method of obtaining automatic gain control to be used. The diode, however, loads the tuned circuit and thus reduces the selectivity of the i-f system to a degree. Special i-f transformers are used for the purpose of providing a low-impedance input circuit to a diode detector. To minimize audio distortion on a-m signals

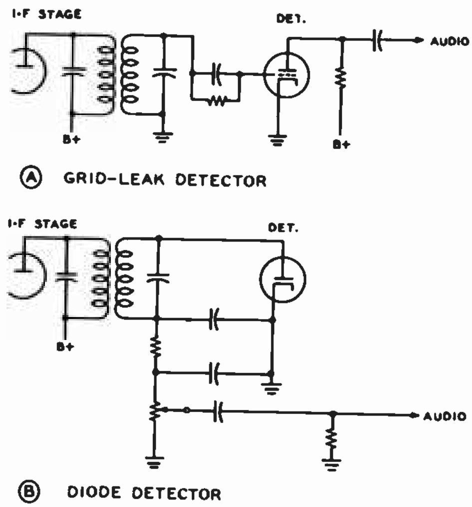


Figure 39

TYPICAL GRID-LEAK AND DIODE DETECTOR CIRCUITS

having a high percentage of modulation, the capacitance across the diode load resistor should be as low as possible.

Referring to figure 40, a dual-diode tube is used as a combination diode detector and agc (automatic gain control) rectifier. The left-hand diode operates as a simple rectifier in the manner described earlier in this Handbook. Audio voltage, superimposed on a d-c voltage, appears across the 500,000-ohm potentiometer (the volume control) and the 100-pf capacitor, and is passed on to the audio amplifier. The right-hand diode receives signal voltage directly from the primary of the last i-f, amplifier, and acts as the agc rectifier. The pulsating d-c voltage across the 1-megohm agc-diode load resistor is filtered by a 500,000-ohm resistor and a .05-μfd capacitor, and is applied as bias to the grids of the r-f and i-f amplifier tubes; an increase or decrease in signal strength will cause a corresponding increase or decrease in agc bias voltage, and thus the gain of the receiver is automatically adjusted to compensate for changes in signal strength.

A-C Loading of Second Detector By disassociating the agc and detecting functions through the use of separate diodes, as shown, most of the ill effects of *a-c shunt loading* on the detector diode are avoided. This type of loading causes serious distortion, and the additional components required to eliminate it are well worth their cost. Even with the circuit shown, a-c loading can occur unless a *very high* (5 megohms, or more) value of grid resistor is used in the following audio amplifier stage.

AGC in BFO Equipped Receivers With a simple diode detector of this type having a beat-frequency oscillator for the reception of c-w or SSB signals, the use of an agc system such as shown in figure 40 can result in a great loss in sensitivity when the bfo is switched on. This is because the beat-oscillator output acts exactly like a strong received signal, and causes the agc circuit to put high bias on the r-f and i-f stages, thus greatly reducing the receiver's sensitivity. Due to the above effect, it is necessary to either isolate the agc voltage or make the agc circuit inoperative when the bfo is being used. The simplest method of eliminating the agc action is to short the agc line to ground when the bfo is turned on. This results in no agc action for SSB reception.

The Plate and Infinite-Impedance Detectors The plate and infinite-impedance detectors (figure 41) offer low loading and good detection efficiency. The infinite-impedance detector, in particular has constant negative current feedback across the large cathode resistor, reducing detection distortion to a low value. Bfo injection may be fed to either detector through a small capacitance coupled to the grid of the tube.

SSB Demodulators The *product detector* is a *linear demodulator in* which two signals are multiplied together to produce a resultant output audio signal. Product detectors are preferred over other detectors for SSB reception because they minimize intermodulation distortion products in the audio output signal and do not require excessively large local carrier voltage. A simple double-diode product detector is shown in figure 42A. This circuit has good large-signal handling capability and may be used with an inexpensive high impedance i-f transformer.

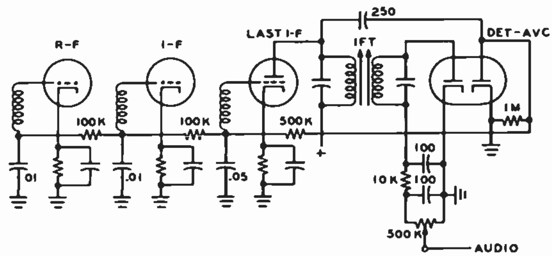


Figure 40

AGC CIRCUIT FOR A-M RECEPTION

Right-hand diode rectifies carrier and delivers d-c control voltage to r-f and i-f amplifier stages. Control voltage is negative and reduces stage gain in proportion to input signal.

A diode ring demodulator is shown in figure 42B. This demodulator provides better low-signal response than the double-diode demodulator and provides a substantial degree of carrier cancellation. The i-f signal is applied to the ring demodulator in push-pull and the local carrier is applied in a parallel mode, where it is rejected by the push-pull output configuration.

A simple transistor sideband demodulator is shown in figure 42C. The transistor is heavily reverse-biased to a class-C condition and the two input signals are mixed in the base circuit. The audio product of mixing is taken from the collector circuit.

A source-follower product detector employing two JFETs is shown in figure 42D. Its vacuum-tube counterpart will be recognized in figure 43A. The two gates provide high-impedance input for both the i-f signal and the carrier oscillator, while providing good isolation between the two signals. Both intermodulation distortion and conversion gain are low in this circuit.

A dual-gate MOSFET is used as a product detector in figure 42E. Various MOSFETs, designed for mixer applications, provide a wide dynamic operating range which permit them to handle large signal levels.

Good isolation between i-f signal and carrier signal may be obtained with simple vacuum-tube product detector circuits, A single-triode product detector is shown in figure 42F. The tube is cathode-biased into the nonlinear operating region and the demodulated signal is taken from the plate circuit through a simple r-f network that filters out the unwanted r-f mixing products.

A dual-triode demodulator circuit (similar to the JFET circuit shown in figure 42D) provides excellent isolation and low intermodulation distortion (figure 43A). The SSB signal from the i-f amplifier is applied to a cathode follower stage that effectively isolates the signal source from the mixing circuit. The carrier signal is fed to the mixing tube and is amplified. The signals mix within the tube and the product output is taken from the plate circuit of the mixer.

A multigrid converter tube may be used as a product detector, with one section of the tube serving as the carrier oscillator (figure 43B). An input attenuator is used to reduce the i-f signal to the proper level for mixing. The audio product is taken from the plate circuit through a low-pass filter network.

The 7360 beam-deflection tube makes a good balanced demodulator or modulator. (figure 44) For demodulator service, the i-f signal is applied to the beam-deflection plates in push pull and the carrier signal is applied to the control grid. The demodulated audio output is taken off in push pull by grounding

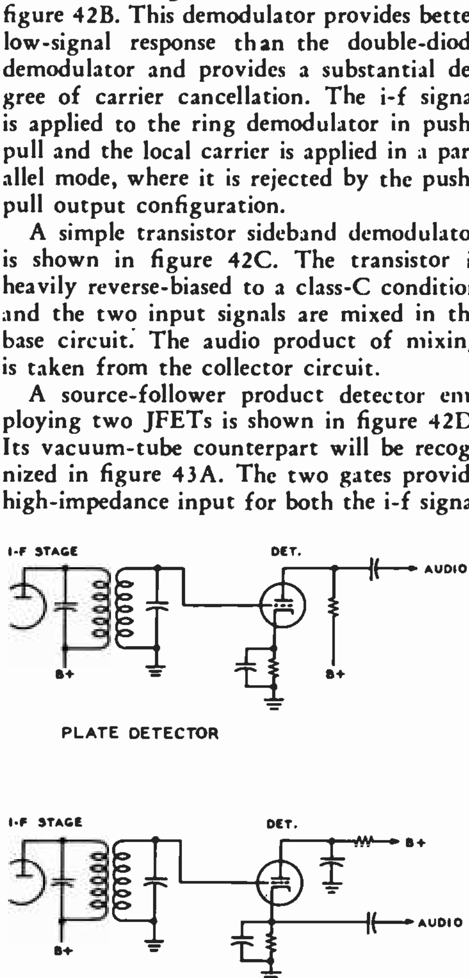


PLATE DETECTOR

INFINITE-IMPEDANCE DETECTOR

Figure 41

TYPICAL PLATE AND INFINITE-IMPEDANCE DETECTOR CIRCUITS

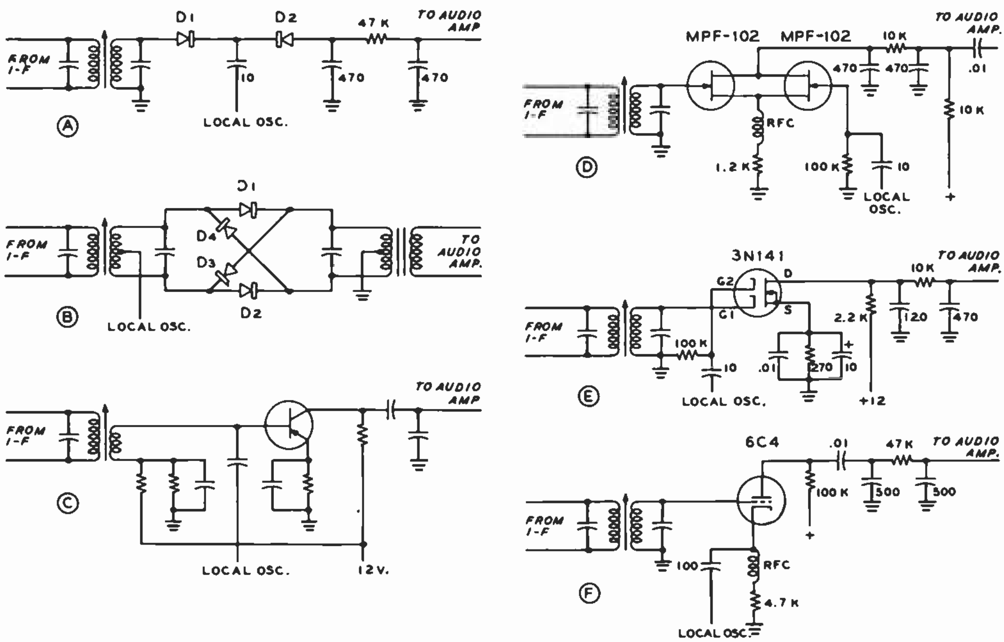


Figure 42

SSB DEMODULATORS

A—Double diode product detector. Simple RC filter is used in audio circuit to remove r-f products from output. B—Diode ring demodulator. C—Bipolar transistor demodulator. Input and local oscillator are mixed in base circuit. D—Source follower demodulator using two JFETs. E—Dual gate MOSFET product detector. F—Cathode-biased triode product detector.

one output plate for audio and taking the desired signal from the other plate.

Sideband Detectors in General Any sideband modulator can be altered to become a demodulator by feeding in carrier and a sideband signal instead of a carrier and audio signal and changing appropriate r-f transformers to audio transformers. Generally speaking, the magnitude of the carrier signal should be from 10 to 20 times as strong as the sideband signal for lowest intermodulation distortion and highest signal overload capability. All signal components other than the desired audio signal must be filtered from the output section of the demodulator if good performance is to be achieved. Carrier injection level should be adjusted for minimum intermodulation distortion on large signals, however, care must be taken to prevent the carrier signal from reaching the i-f stages of the receiver by radiation and conduction along circuit wiring. Excessive carrier signal may also cause overloading or desensitization of the audio section of the receiver and also cripple the agc action. Stray coupling from the carrier oscillator to other portions of the receiver circuitry, then, must be carefully controlled.

10-11 Automatic Gain Control

Modern communication receivers include a control loop to automatically adjust the r-f and i-f gain level. The loop holds the receiver output substantially constant despite changes in input signal level. This system is termed *automatic gain control* (*agc*). Conventional a-m automatic volume control systems are generally not usable for SSB

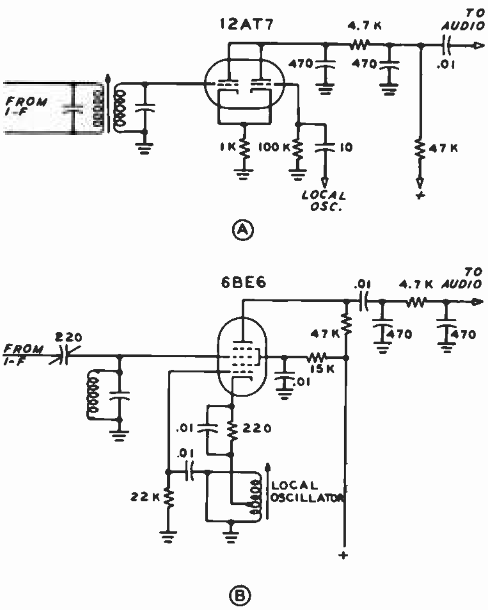


Figure 43

TYPICAL VACUUM-TUBE DEMODULATORS

A—Dual triode product detector provides low intermodulation distortion at high signal level. B—Multigrid product detector and local oscillator.

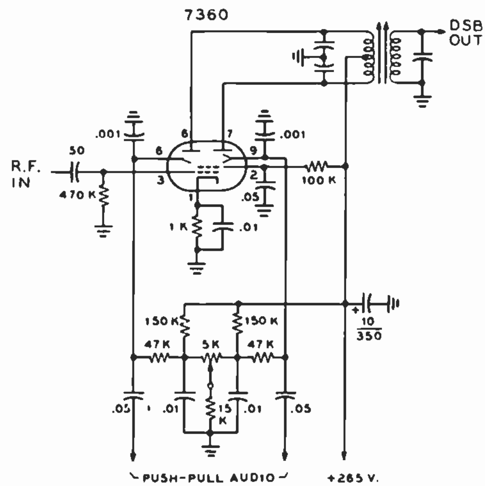


Figure 44

7360 BALANCED-MODULATOR CIRCUIT

7360 may be used as balanced modulator or product detector. When used as demodulator, i-f signal is applied in push-pull to beam-deflection plates and local oscillator is applied to control grid. Audio is taken from plate circuit.

since they operate on the level of the carrier, which is suppressed in SSB. A system must be used which obtains its information directly from the modulation envelope of the incoming signal. The control voltage derived from the agc detector is applied to a variable gain element in the receiver, usually in the r-f and i-f chain.

For optimum SSB reception, the control voltage must be applied rapidly to the variable element to avoid transient overload at the beginning portion of each word, otherwise an annoying *agc thump* will be apparent at the start of the first syllable. As the syllabic envelope of the SSB signal is a replica of the original audio signal, the agc voltage must rise rapidly with the start of the syllable and then hold at a value corresponding to the average of the syllable undulations of the signal over an extended period of seconds. Too-rapid variations of the agc voltage with respect to syllabic peaks may bring up background noise in an objectionable manner termed *agc pumping*. The ideal agc action, then, exhibits a fast-attack, slow-decay time constant. Circuits having a charge time of 50 to 200 milliseconds and a discharge time of 0.5 to 3 seconds have proven successful.

A simple i-f operated agc circuit with the control voltage applied to remote cutoff pentodes in the r-f and i-f systems is shown in figure 45.

The i-f signal may be used to control the agc system in a solid state receiver, as shown in figure 46. An IC is used as an amplifier to provide gain and isolation. The resulting signal is rectified and further amplified by cascaded d-c amplifiers Q_1 and Q_2. Transistor Q_1 is forward-biased by the agc voltage to provide a voltage drop across the collector load resistor. This voltage biases Q_2 more heavily in the forward direction when a large signal arrives and increases the voltage drop across the emitter resistor. This voltage varies in accord with the strength of the incoming signal and changes the bias voltage in accord with the strength of the incoming signal and changes the bias voltage on various

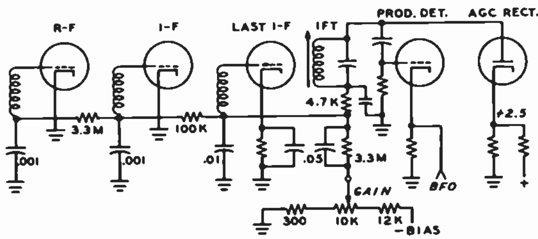


Figure 45

I-F OPERATED AGC SYSTEM

Product detector and agc system. Bfo voltage is isolated from agc system so that rectified oscillator voltage does not actuate agc loop. Initial gain level is set by gain potentiometer.

signal stages. The agc characteristic is determined by the agc time constant, R_1, R_2, C_1.

Audio Derived AGC Since agc voltage follows the average SSB syllabic undulation of speech, it is possible to derive the agc voltage from the audio system of the receiver as shown in figure 47. A portion of the audio signal is rectified and returned to the controlled stages after passing through a combination filter and delay network. A second audio-derived circuit is shown in figure 48A. Transistor Q_1 is operated without base bias so that no output is obtained until the input signal exceeds a critical peak level (0.6 volt), enough to turn on the transistor. Once this level is reached, very little additional voltage is needed to achieve full output from the agc rectifier. This results in a very flat agc characteristic.

A different audio-derived agc circuit is shown in figure 48B. A JFET serves as a source follower from the audio line, driving the gain control transistor (Q_1). The no-signal voltage at the base of Q_1 is about 0.4 volt, rising to about 0.55 volt before gain reduction starts.

Signal-Strength Indicators Visual means for determining the relative strength of the received signal may be provided by a tuning indicator, or S-meter. A d-c milliammeter may be connected in a Wheatstone bridge circuit in the i-f system of a receiver, as shown in figure 49A. The d-c plate resistance of the tube serves as one leg of the bridge, with resistors for the other three legs. A change in plate current, due to the action of the agc voltage, will be indicated on the instrument as a result of the consequent bridge unbalance. Sensitivity of the circuit is determined by resistor R.

An electronic "eye" tube, such as the 6FG6 may also be used as a signal-strength meter, as shown in figure 49B. A solid-state

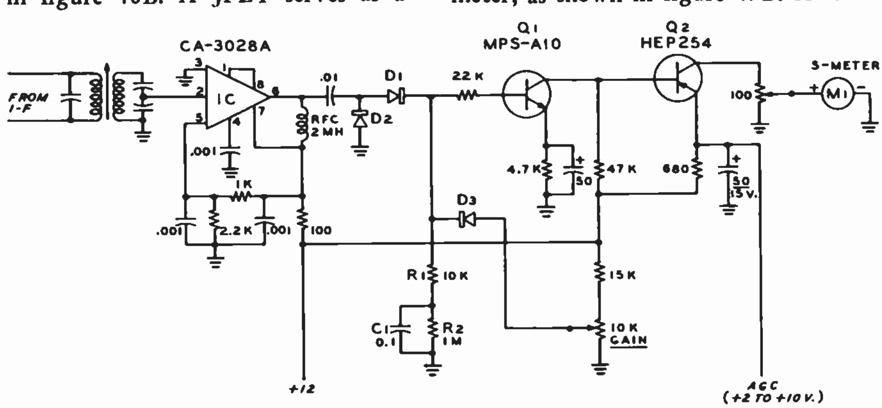


Figure 46

SOLID-STATE AGC SYSTEM

IC amplifier stage provides gain and isolation for i-f signal applied to diode rectifier (D_1, D_2) and cascaded d-c amplifiers, Q_1 and Q_2. Agc signal is taken from emitter circuit of Q_2. Signal-strength meter (M_1) is placed in collector circuit. Agc gain is controlled by the base-bias potentiometer in the Q_1 base circuit.

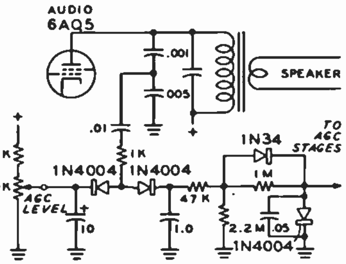
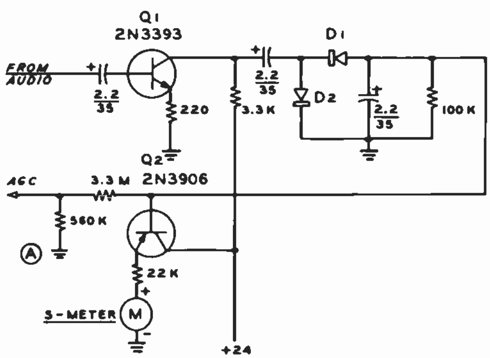


Figure 47

AUDIO-DERIVED AGC CIRCUIT

Agc level is set by bias potentiometer and drive signal is taken from plate circuit of audio output stage. System provides fast-attack, slow-release response.



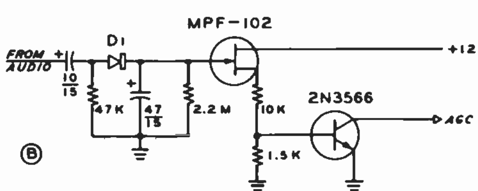


Figure 48

TYPICAL AGC CIRCUITS

A—Transistor Q_1 is operated without base bias so that no output is obtained until the input signal exceeds a critical peak level (0.6 volt), enough to turn on Q_1. Audio voltage is rectified and applied to agc system. Q_2 serves as signal meter amplifier with milliammeter in emitter circuit. B—JFET serves as a source follower driving the gain control transistor.

S-meter circuit which monitors agc voltage is shown in figure 49C. The collector current

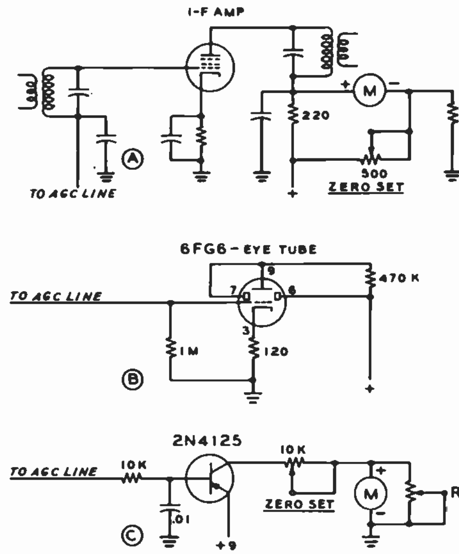


Figure 49

SIGNAL-STRENGTH INDICATORS

Gain-controlled stages of receiver provide variable voltage for signal-strength indication. A—Vacuum-tube i-f stage uses milliammeter in bridge circuit. B—Type 6FG6 indicator tubes registers agc voltage. C—Transistorized high-impedance voltmeter measures average agc voltage.

of the transistor rises as the negative agc voltage is increased and this current causes the meter reading to increase in accord with the agc voltage.

S-Meters in General The calibration of an S-meter in the great majority of communication receivers varies with the band of reception and receiver gain. The actual reading, therefore, should be taken as a relative indication of received signal strength, rather than as an accurate measurement. Some manufacturers establish an S-9 meter indication as equivalent to an input signal to the receiver of 50 microvolts, but even this level changes between like receivers of the manufacturer. It should be remembered, therefore, that the typical S-meter is merely a high-impedance voltmeter that reads the average agc voltage of the receiver, which may vary widely as receiver gain varies.

Part II—VHF and UHF Receivers

Vhf and uhf receiver design follows the same general philosophy discussed in the first part of this chapter, but with important consequences dictated by the pecularities of radio propagation at frequencies above 30 MHz. The outstanding factor in vhf uhf reception, as compared to reception at the lower frequencies, is that ultimate system sensitivity is primarily limited by receiver noise, rather than by noise external to the receiver. It is therefore possible to realize superior performance in terms of usable signal-to-noise ratio and sensitivity in a typical vhf/uhf system as opposed to an h-f system, in which external atmospheric noise and man-made interference ("r-f smog") makes such receiver attributes relatively useless.

Vhf/uhf receivers are externally limited in sensitivity only by extraterrestrial (galactic) noise and some forms of man-made noise. Sophisticated receivers for this portion of the spectrum can reach the galactic noise level while rejecting man-made noise to a great degree. The state-of-the-art receiver noise figure is approximately as shown in figure 1.

10-12 VHF/UHF Noise Sources

External noise may be composed of atmospheric noise, galactic (cosmic) noise, and man-made noise as shown in figure 2. Above 30 MHz or so, external noise drops to a level that makes receiver noise of paramount importance. The development of low-noise vhf/uhf receivers is a continuing task as this portion of the spectrum becomes of greater and greater importance to the modern world.

Atmospheric noise is due mainly to lightning discharges in the atmosphere which are propagated worldwide by ionospheric reflection. The noise varies inversely with frequency, being greatest at the lower frequencies and least at the higher frequencies. It also varies in intensity with time of day, weather, season of the year, and geographical location. It is particularly severe in the tropical areas of the world during the rainy seasons.

Galactic noise is caused by disturbances that originate outside the earth's atmosphere. The primary sources of such noise are the sun and a large number of "radio stars" dis-

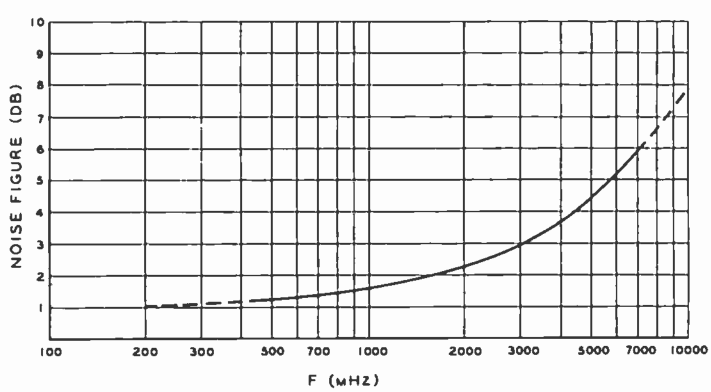


Figure 1

REPRESENTATIVE RECEIVER NOISE FIGURE

State-of-the-art receiver noise figure rises from about 1.2 decibels at 450 MHz to near 6 decibels at 7000 MHz for specialized solid-state devices operating at room temperature.

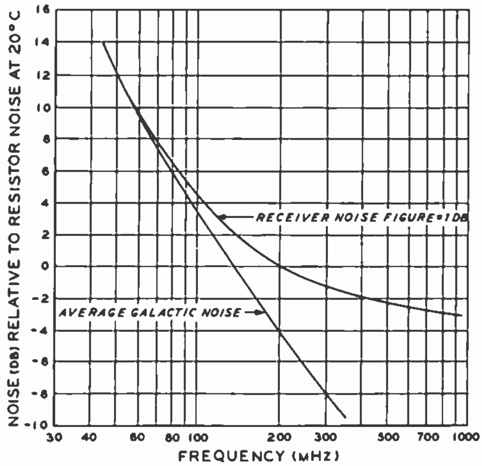


Figure 2

AVERAGE GALACTIC (COSMIC) NOISE LEVEL

Atmospheric noise predominates below 30 MHz. Galactic noise drops with increasing frequency, reaching low values at uhf. Receiver with 1-dB noise figure would have ultimate capability shown by top curve. Reduction of receiver noise figure becomes increasingly important for weak-signal reception above 100 MHz.

tributed principally along the galactic plane. Galactic noise is largely blocked out by atmospheric noise at frequencies below approximately 20 MHz.

Man-made noise tends to decrease with increasing frequency, although it may peak at some discrete frequency, depending on the electrical characteristics of the noise source. It can be caused by electrical appliances of all types, television receivers, ignition systems, motors, and erratic radiation of high-frequency components from power lines. Propagation is by direct transmission over power lines and by radiation, induction, and occasionally by ionospheric reflection.

Thermal noise, or *Johnson noise*, is caused by the thermal agitation of electrons and pervades nature. It is only at absolute zero that such motion ceases. As the temperature of a conductor rises above absolute zero, the random motion of free electrons increases and this motion corresponds to a minute electric current flowing in the conductor. This "white noise" is generated over a wide

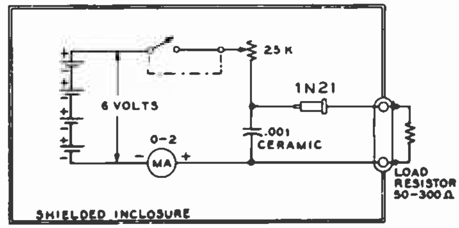


Figure 3

A SILICON DIODE NOISE GENERATOR

band of frequencies and the portion of it falling within the passband of a receiving system will contribute to the noise output of the system. Limiting system bandwidth, therefore, will tend to limit the thermal noise. Thermal noise takes place in the receiving antenna, the feedline, and the receiver itself, the noise level of the input stage of the receiver being particularly critical as to system performance.

10-13 Noise and Receiver Circuitry

Input Circuit Considerations Since the full amplification of a receiver follows the first tuned circuit, the operating conditions existing in that circuit and in its coupling to the antenna on one side and to the input of the first amplifier stage on the other are of greater importance in determining the signal-to-noise ratio of the receiver on weak signals.

First Tuned Circuit It is obvious that the highest ratio of signal to noise be impressed on the input element of the first r-f amplifier device. Attaining the optimum ratio is a complex problem since noise will be generated in the antenna due to its equivalent radiation resistance (this noise is in addition to any noise of atmospheric origin) and in the first tuned circuit due to its equivalent coupled resistance at resonance. The *noise voltage* generated due to antenna radiation resistance and to equivalent tuned-circuit resistance is similar to that generated in a resistor due to thermal agitation and is expressed by the following equation:

$$E_n = (4kTR\Delta f)^{1/2}$$

where,

E_n = rms value of noise voltage over the interval Δf,

k = Boltzman's constant (1.38×10^{-23} joule per °K),

T = Absolute temperature °K,

R = Resistive component of impedance across which thermal noise is developed,

Δf = Frequency band across which voltage is measured.

In the above equation Δf is essentially the frequency band passed by the intermediate-frequency amplifier of the receiver under consideration. This equation can be greatly simplified for the conditions normally encountered in communications work. If we assume the following conditions: $T = 300°$ K or 27° C or 80.5° F, room temperature; $\Delta f = 8000$ Hertz (the average passband of an a-m communications receiver or speech amplifier), the equation reduces to: $E_n = 0.0115 \sqrt{R}$ microvolts. Accordingly, the thermal agitation voltage appearing in the center of a half-wave antenna (assuming effective temperature to be 300°K) having a radiation resistance of 73 ohms is approximately 0.096 microvolts. Also, the thermal-agitation voltage appearing across a 500,-000-ohm input resistor in the first stage of a speech amplifier is approximately 8 microvolts under the conditions cited above. Further, the voltage due to thermal agitation being impressed on the input network of the first r-f stage in a receiver by a first tuned circuit whose resonant resistance is 50,000 ohms, is approximately 2.5 microvolts. Suffice to say, however, that the value of thermal-agitation voltage appearing across the first tuned circuit when the antenna is properly coupled to this circuit will be very much less than this value.

It is common practice to match the impedance of the antenna transmission line to the input impedance of the amplifying device of the first r-f amplifier stage in a receiver. This is the condition of antenna coupling which gives maximum gain in the receiver. However, when vhf tubes and transistors are used at frequencies somewhat less than their maximum capabilities, a significant improvement in *signal-to-noise* ratio can be attained by *increasing* the coupling between the antenna and first tuned circuit to a value greater than that which gives greatest signal amplitude out of the receiver. In other words, in the vhf bands, it is possible to attain somewhat improved signal-to-noise ratio by increasing antenna coupling to the point where the gain of the receiver is slightly reduced.

It is always possible, in addition, to obtain improved signal-to-noise ratio in a vhf receiver through the use of devices which have improved input-impedance characteristics at the frequency in question over conventional types.

Noise Figure Expressed in decibels, the *noise figure* of a receiver is:

$$f_N = 10 \log_{10} \frac{N_2}{N_1}$$

where,

N_1 and N_2 are the noise power figures in watts and represent the output from an actual receiver, (N_2) at 290° K (63°F), divided by the noise power output from an ideal receiver (N_1) at the same temperature.

The *noise power* is a function of the noise voltage (E_n) and is expressed as:

$$N = kBT$$

where,

k = Boltzman's constant,

B = Noise bandwidth in Hz,

T = 290° K.

The noise figure of a receiver may be ascertained by direct measurement with a *noise generator*. The receiver input is terminated with a resistor and wideband random noise, generated by thermal agitation in a suitable generator, is injected into the input circuit of the receiver. The power output of the receiver is measured with no noise input and the generator output is then increased until the receiver noise output is doubled. the noise figure of the receiver is a function of these two levels, and may be computed from these measurements. The circuit of a simple noise generator is shown in figure 3.

Vacuum Tubes in VHF/UHF Receivers The vacuum tube has been eclipsed for low-noise reception above 30 MHz by solid-state devices. Because of the hot filament within the tube, thermal agitation and noise level are excessive for weak-signal reception. Vacuum-tube noise is composed of *shot noise* (electron noise), *partition noise* (noise caused by a random division of space current between the elements of the tube), and *induced grid noise* caused by fluctuations in cathode current passing the grid element. The summation of these noises is expressed as the *equivalent noise resistance* of the vacuum tube. In addition to noise, most vacuum tubes have comparatively high input and output capacitances and a low input impedance, all of which inhibit the design of high-Q, high-impedance tuned circuits above 50 MHz or so.

Semiconductors in VHF/UHF Receivers Great advances have been made in recent years in both bipolar and field-effect devices and these improved units have preempted the vacuum tube in vhf/uhf operation in low-noise receiver circuitry. While the bipolar transistor exhibits circuit loading due to low input impedance and often has characteristics that vary widely with temperature, these problems are being overcome by new design and production techniques. The field-effect device, on the other hand, exhibits an input impedance equal to, or better, than vacuum tubes in the vhf/uhf region.

The better solid-state devices are superior to vacuum tubes as far as good noise factor is concerned and noise figures of 3 db or better are possible up to 1000 MHz or so with selected transistors and field-effect devices.

10-14 VHF Receiver Circuitry

Vhf r-f receiver circuitry resembles the configurations discussed for h-f receivers to a great degree. Solid-state r-f circuits specifically designed for efficient vhf operation are discussed in this section and they may be compared against the circuitry shown earlier in this chapter.

The common-base (or gate) r-f amplifier circuit (figure 4) is often used with bipolar devices in the Vhf range since it is stable and requires no neutralization. Either PNP or NPN transistors may be used, with due attention paid to supply polarity. The input signal is fed to the emitter (sink); the base (gate) is at r-f ground potential; and the output signal is taken from the collector (drain) circuit. Stage gain is low and two or more stages are often cascaded to provide sufficient signal level to overcome mixer noise. The input impedance of the common-base circuit is low and this configuration does not offer as much r-f selectivity as does the common-emitter (sink) circuit of figure 5. This circuit often requires neutralization, accomplished by feeding energy back from the output to the input circuit in proper amplitude and phase so as to cancel the effects of spurious signal feedthrough in and around the device. Tuning and neutralization are interlocking adjustments.

The cascode amplifier (figure 6) is a series-connected, grounded-emitter (sink), grounded-base (gate) circuit. Neutraliza-

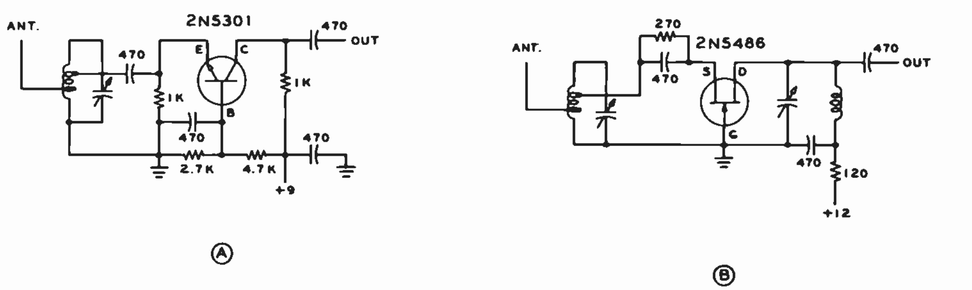


Figure 4

COMMON-BASE (GATE) R-F AMPLIFIER

Input signal is applied to emitter (A) or sink (B) and output signal is taken from collector (A) or drain (B). Stage gain and input impedance are both low in this configuration.

tion, while not always necessary, may be employed to achieve lowest noise figure.

A neutralized, IGFET vhf amplifier stage is shown in figure 7A. Protective diodes D_1 and D_2 (discussed in the next section) are used in the input circuit. A dual-gate, diode-protected MOSFET is employed in the amplifier circuit of figure 7B. Input and output points are tapped down the tuned circuits to reduce stage gain and to remove the necessity for neutralization, which otherwise may be necessary.

Special vacuum tubes, such as high-gain TV pentodes and low-noise triodes may be used in these typical vhf circuits and are often used in simple converters designed for 6 and 2 meters. Typical circuits suitable for this use are shown in the first part of this chapter.

To optimize the noise figure of all of these circuits, the input coupling, bias level, and neutralizing adjustment (if any) are made with a weak signal source used for alignment. Adjustment is not complicated provided proper vhf construction techniques and shielding are used in construction of the amplifier.

Amplifier Protection Vhf solid-state devices are vulnerable to burnout by accidental application of high input signal voltage to the receiver. Reverse-connected diodes (either silicon or germanium) placed across the input circuit will limit maximum signal voltage to a few tenths of a volt, providing automatic protection against damaging overload. In particular, the protection diodes will absorb r-f energy that leaks around an antenna changeover relay, or that is received from a nearby transmitter.

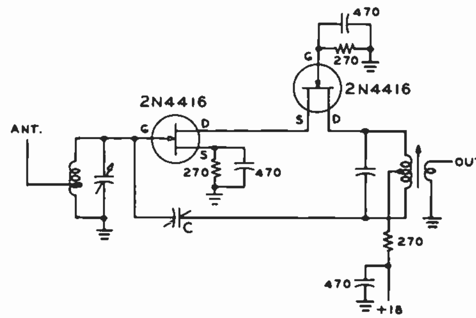


Figure 6

CASCODE R-F AMPLIFIER

Two FET devices are series-connected, the first being driven at the gate and the second at the sink. Bipolar transistors or tubes are used in a similar arrangement. Neutralization is required to achieve highest over-all gain and optimum noise figure.

VHF/UHF Mixers Conventional multielement vacuum-tube mixers are occasionally used in the lower portion of the vhf spectrum because of their high-signal overload capability, giving away to solid-state mixers as the noise factor of the

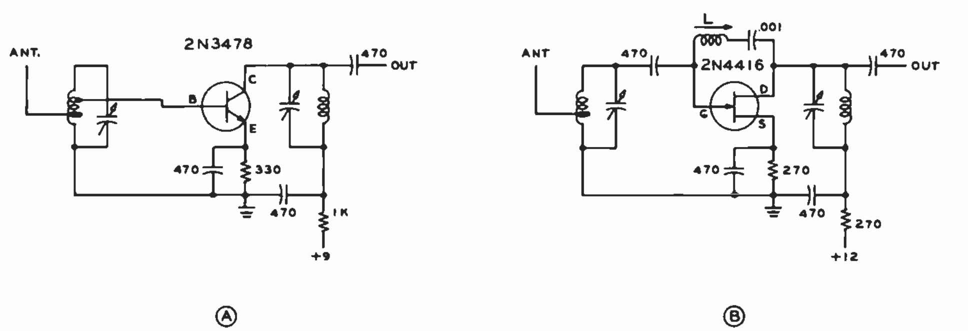


Figure 5

COMMON-EMITTER (SINK) R-F AMPLIFIER

Input signal is applied to base (A) or gate (B) and output signal is taken from collector (A) or drain (B). Stage gain is high and neutralization is often required to cancel signal feed-through, as shown in (B).

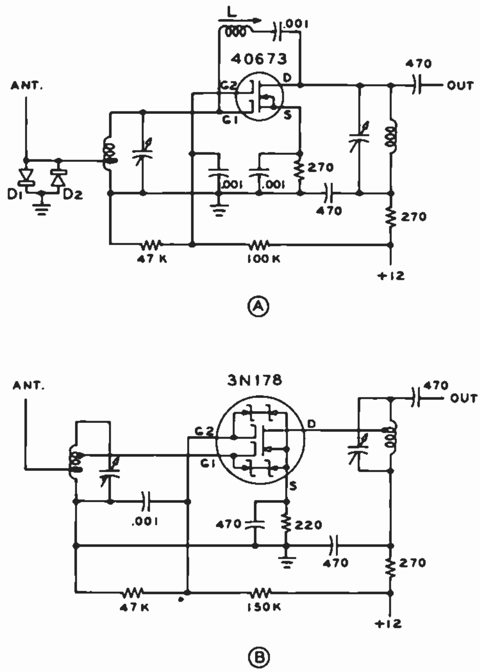


Figure 7

FETs IN VHF CIRCUITRY

A—Neutralized IGFET using 1N100 protective diodes in input (gate) circuit. B—Dual-gate, self-protected MOSFET circuit. Neutralization may be required for maximum stage gain and optimum noise figure.

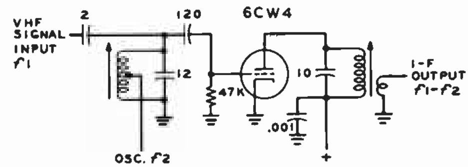


Figure 8

LOW-NOISE TRIODE MIXER USEFUL UP TO 250 MHz OR SO WHEN PRECEDED BY LOW-NOISE R-F AMPLIFIER

vacuum tube deteriorates rapidly with increasing frequency. Low-noise triode mixers (figure 8) are useful up to 250 MHz or so, when preceded by a high-gain, low-noise r-f amplifier chain.

As the noise figure of the solid-state device and the vacuum tube falls off above a few thousand MHz, the usefulness of the r-f amplifier stage becomes marginal and it becomes practical to couple the antenna circuit directly to the mixer stage, following the mixer with a low-noise, high-gain i-f amplifier. The mixer, thus, becomes the dominant stage in determining receiver noise figure.

Various diodes are available for use as mixers and the *hot-carrier diode* serves as a low-noise mixer for applications up to and including the uhf region (figure 9). This device (also known as a *Schottky-barrier* diode) is a planar version of a conventional point-contact microwave mixer diode. The hot-carrier diode has closely matched transfer characteristics from unit to unit and a high front-to-back ratio. In addition, it provides extremely fast switching speed combined with low internal noise figure. Input and output impedances are low, but overall conversion efficiency is high.

10-15 I-F Strips and Conversion Oscillators

To combine good image rejection with a high order of selectivity, double frequency conversion is normally used for vhf/uhf small-signal reception. The first intermediate frequency is usually rather high to provide adequate rejection of image signals and the second is low to provide good selectivity. Care must be used in choosing the first intermediate frequency or image problems will arise from signals in the 80- to 130-MHz range, which includes high power f-m transmitters and strong aircraft signals.

It is common practice to construct the r-f amplifier and first conversion circuits in a separate *converter* unit, the low output of which is fed into an h-f communications receiver which serves as the low-frequency i-f strip. Choice of the first i-f channel is important, since many vhf/uhf converters provide scant selectivity at the received frequency, having bandwidths measured in hundreds of MHz. If the image ratio is unity, the image signal may be as strong as the wanted signal and the noise figure of the receiving system is degraded by 3 decibels,

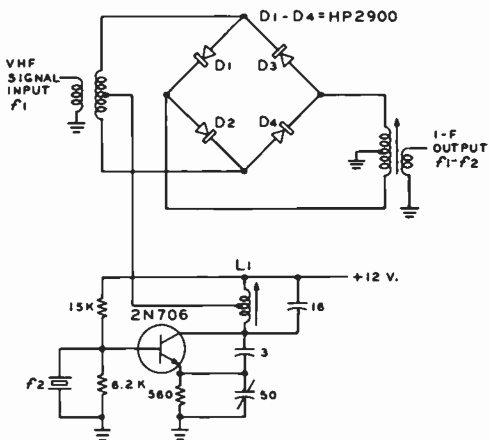


Figure 9

HOT-CARRIER DIODE MIXER

Schottky-barrier diode is a planar version of a conventional point-contact microwave mixer diode having closely matched transfer characteristics from unit to unit and high front-to-back ratio. It provides extremely fast switching time combined with low internal noise figure.

regardless of the noise figure of the converter. The first i-f channel, and the r-f selectivity of the converter should therefore be sufficiently high so that images are not a problem. Generally speaking a first i-f channel of 15 MHz to 30 MHz is suitable for 144-MHz and 220-MHz reception and a frequency in the region of 144 MHz is often used as the first i-f channel for 432-MHz (and higher) reception.

In addition to attention to image problems, care must be taken to ensure that the harmonics of the local oscillator of the communications receiver used for the i-f strip do not fall within the input passband of the converter. Attention should also be given to the input circuit shielding of the communications receiver to prevent breakthrough of strong h-f signals falling within the first i-f passband. Unwanted h-f signals may also enter the receiver via the speaker wires or the power cord.

Spurious signals and unwanted "birdies" can be reduced to a minimum by using the highest practical injection frequency for the local oscillator. Most first conversion oscillators in vhf receiving systems are crystal controlled and high-overtone crystals are to be preferred as contrasted to lower-frequency crystals and a multiplier string. Unwanted harmonics generated by a multiplier string must be prevented from reaching the mixer stage by means of a high-Q trap circuit in order to avoid unwanted mixing action between received signals and the various harmonics.

When low-frequency conversion crystals are employed, the use of multiple tuned intermediate circuits in the multiplier string is suggested, as shown in figure 10. A simple diode multiplier may also be used in place of a tube or transistor, as shown in figure 11.

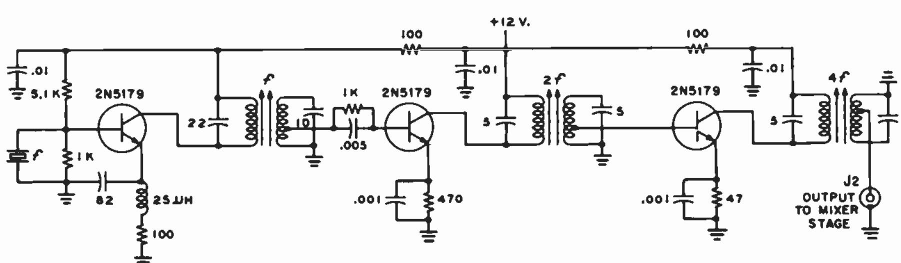


Figure 10

LOCAL-OSCILLATOR "STRING" FOR VHF RECEIVER

Multiple tuned high-Q circuits between stages prevent unwanted harmonics of oscillator from reaching the mixer stage. Fundamental oscillator signal and 3rd and 5th harmonics could produce spurious responses in receiver unless suitably attenuated.

10-16 Special Consideration in UHF Receiver Design for VHF Receivers

As one advances higher into the uhf region, the physical dimension of an electrical half wavelength of radio energy begins to assume the proportions of some of the components that make up the circuitry of the receiver. At 1000 MHz, for example, a half wavelength is about six inches, and the uhf converter itself become an appreciable fraction of a wavelength long. Components, moreover, are fractions of a wavelength long and their physical size, shape, and inherent capacitance and inductance become critical portions of the circuitry.

At increasingly higher frequencies, it becomes progressively more difficult to obtain a satisfactory amount of selectivity and impedance from an ordinary coil and capacitor used as a resonant circuit. On the other hand, quarter-wavelength sections of parallel conductors or concentric transmission line are not only more efficient but also approach practical dimensions.

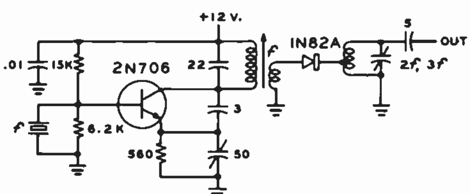


Figure 11

DIODE MULTIPLIER FOR LOCAL OSCILLATOR INJECTION AT A HIGH HARMONIC

One or more tuned circuits or traps are used after diode multiplier to attenuate unwanted harmonics of local oscillator.

Tuning Short Lines Transistors and tuning capacitors connected to the open end of a transmission line provide a capacitance that makes the resonant length less than a quarter wavelength. The amount of shortening for a specified capacitive reactance is determined by the surge impedance of the line section. It is given by the equation for resonance:

$$\frac{1}{2\pi fC} = Z_0 \tan l$$

π equals 3.1416,
f equals the frequency,
C equals the capacitance,
Z_0 equals the surge impedance of the line,
$\tan l$ equals the tangent of the electrical length in degrees.

The capacitive reactance of the capacitance across the end is $1/(2\pi fC)$ ohms. For resonance, this must equal the surge impedance of the line times the tangent of its electrical length (in degrees, where 90° equals a quarter wave). It will be seen that twice the capacitance will resonate a line if its surge impedance is halved; also that a given capacitance has twice the loading effect when the frequency is doubled.

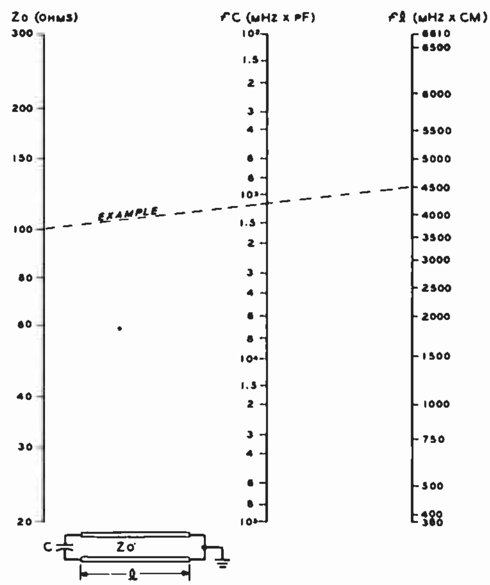


Figure 12

LINE RESONATOR CHART

Capacitance-loaded resonant line is used in vhf and uhf ranges in place of typical coil-capacitor tank circuit. In the example shown a 100-ohm line is to be used as a resonator at 150 MHz. The line is 30 cm. long. The product f × l (150 × 30)=4500 and f × c is read at 1200. Capacitance is found to be 8 pF by dividing 100 by 50.

Calculations for capacity-loaded line resonators may be simplified by the use of the nomograph shown in figure 12 that establishes capacitance and line length as products of frequency. The *fC* ordinate is equal to frequency in MHz times capacitance in pF, and the *fl* ordinate is equal to frequency in MHz times line length in centimeters.

Coupling Into Lines and Coaxial Circuits Either inductive or capacitive coupling may be used with transmission-line and coaxial circuits. Inductive coupling is accomplished by means of a variable loop or tap at a low-impedance point in the circuit whereas capacitive coupling is done at a high-impedance point (figure 13). The area of the loop or capacitor plate and spacing from the line determine the impedance matching and loading of the circuit.

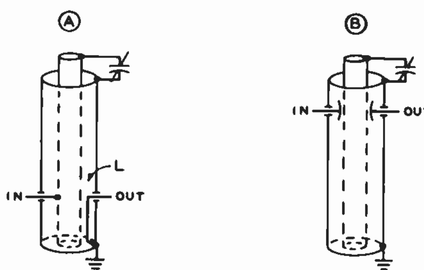


Figure 13

COUPLING IN AND OUT OF COAXIAL RESONATOR

A—Input line is tapped on center conductor and output line is inductively coupled to resonator.

B—Input and output lines are capacitively coupled at high impedance end of center conductor.

Resonant Cavities A *cavity* is a closed resonant chamber made of metal. The cavity, having both inductance and capacitance, supersedes coil-capacitor and capacitance loaded transmission-line tuned circuits at extremely high frequencies where conventional L and C components, of even the most refined design, prove impractical because of the tiny electrical and physical dimensions they must have. Microwave cavities have high Q factors and are superior to conventional tuned circuits. They may be employed in the manner of an absorption wavemeter or as the tuned circuit in other r-f test instruments, and in microwave transmitters and receivers.

Resonant cavities usually are closed on all sides and all of their walls are made of conducting material. However, in some forms, small openings are present for the purpose of excitation (figure 14).

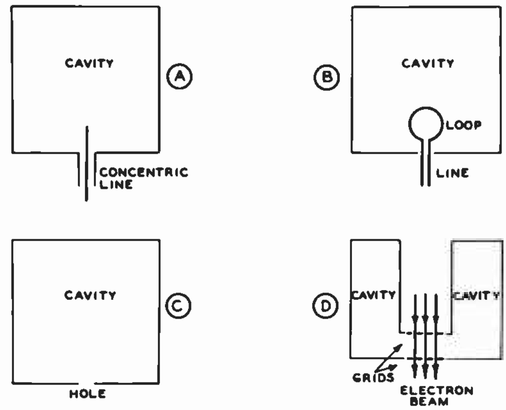


Figure 14

METHODS OF EXCITING A RESONANT CAVITY

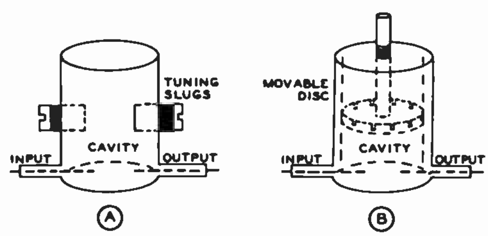


Figure 15

TUNING METHODS FOR CYLINDRICAL RESONANT CAVITIES

Cavities have been produced in several shapes including the plain sphere, dimpled sphere, sphere with re-entrant cones of various sorts, cylinder, prism (including cube), ellipsoid, ellipsoid - hyperboloid, doughnut-shape, and various re-entrant types. In appearance, they resemble in their simpler forms metal boxes or cans.

The cavity actually is a linear circuit, but one which is superior to a conventional coaxial resonator in the uhf range. The cavity resonates in much the same manner as does a barrel or a closed room with reflecting walls.

Because electromagnetic energy (and the associated electrostatic energy) oscillates to and fro inside them in one mode or another, resonant cavities resemble waveguides. The mode of operation in a cavity is affected by the manner in which microwave energy is injected. A cavity will resonate to a large number of frequencies, each being associated with a particular mode or standing-wave pattern. The lowest mode (lowest frequency of operation) of a cavity resonator normally is the one used.

The resonant frequency of a cavity may be varied, if desired, by means of movable plungers or plugs, as shown in figure 15A, or a movable metal disc (figure 15B). A cavity that is too small for a given wavelength will not oscillate.

The resonant frequencies of simple spherical, cylindrical, and cubical cavities may be calculated simply for one particular mode. Wavelength and cavity dimensions (in centimeters) are related by the following simple resonance formulas:

for cylinder $\lambda_r = 2.6 \times$ radius;
for cube $\lambda_r = 2.83 \times$ half of 1 side;
for sphere $\lambda_r = 2.28 \times$ radius.

Butterfly Circuit Unlike the cavity resonator, which in its conventional form is a device which can tune over a relatively narrow band, the *butterfly circuit* is a tunable resonator which permits coverage of a fairly wide uhf band. The butterfly circuit is very similar to a conventional coil/variable-capacitor combination, except that both inductance and capacitance are provided by what appears to be a variable capacitor alone. The Q of this device is somewhat less than that of a concentric-line tuned circuit but is entirely adequate for numerous applications.

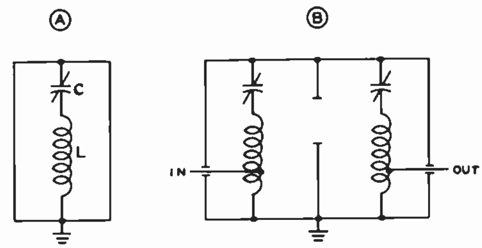


Figure 17

VHF HELICAL RESONATOR

A—High-Q modified cavity consists of inductor placed within metal enclosure. B—Double-coupled resonator. Coupling is achieved via aperture cut in shield between cavities.

Figure 16A shows construction of a single butterfly section. The butterfly-shaped rotor, from which the device derives its name, turns in relation to the unconvention-

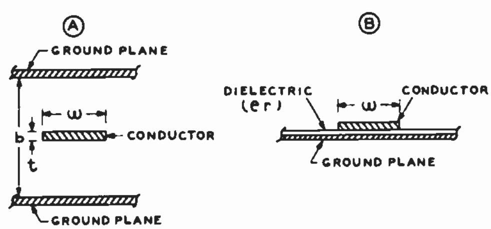


Figure 18

STRIP-LINE CIRCUIT

The strip-line circuit is a flat conductor placed between ground planes (A). Characteristic impedance is a function of plane spacing (b), conductor width (w) and conductor thickness (t). Microstrip line (B) is a version of the strip line adapted for circuit-board techniques. Copper surfaces of board form ground plane and line. Impedance of the line is a function of dielectric properties (e_r) and width (w) of strip.

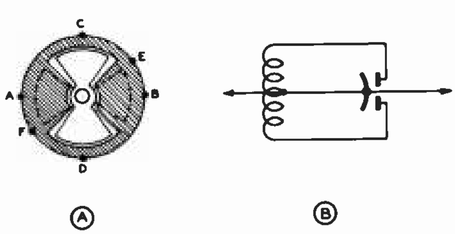


Figure 16

THE BUTTERFLY RESONANT CIRCUIT

Shown at A is the physical appearance of the butterfly circuit as used in the vhf and lower uhf range. B shows an electrical representation of the circuit.

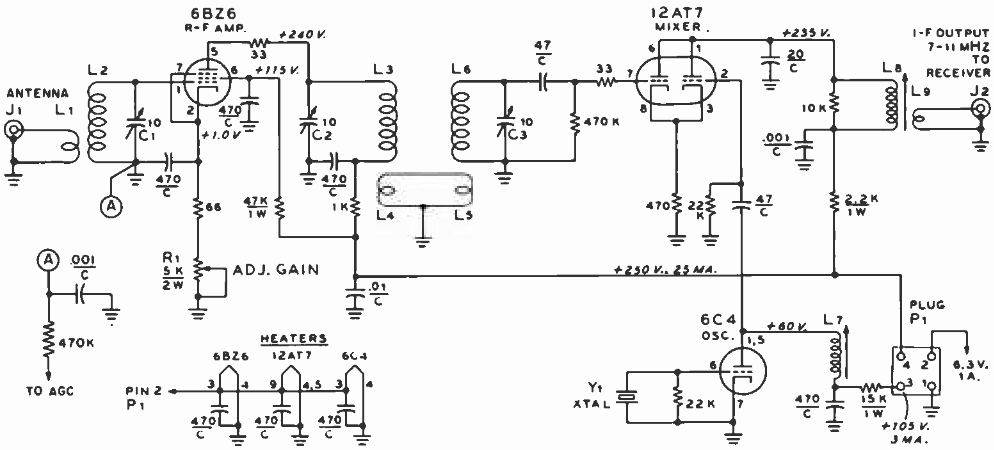


Figure 19

SCHEMATIC OF GENERAL PURPOSE HIGH-FREQUENCY CONVERTER

C_1, C_2, C_3—10-pf miniature capacitors (Johnson 160-107 or equiv.)
J_1, J_2—Coaxial receptacle, BNC type UG-625/u
L_1-L_9—See coil table
P_1—4-pin chassis-mounting plug. (Cinch-Jones P-304AB or equiv.)
All resistors ½ watt unless otherwise noted.

al stator. The two groups of stator "fins" or sectors are, in effect, joined together by a semicircular metal band, integral with the sectors, which provides the circuit inductance. When the rotor is set to fill the loop opening (the position in which it is shown in figure 16A), the circuit inductance and capacitance are reduced to minimum. When the rotor occupies the position indicated by the dotted lines, the inductance and capacitance are at maximum. The tuning range of practical butterfly circuits is in the ratio of 1.5:1 to 3.5:1.

Direct circuit connections may be made to points A and B. If balanced operation is desired, either point C or D will provide the electrical midpoint. Coupling may be effected by means of a small single-turn loop placed near point E or F. The butterfly thus permits continuous variation of both capacitance and inductance, as indicated by the equivalent circuit in figure 16B, while at the same time eliminating all pigtails and wiping contacts.

Butterfly circuits have been applied specifically to oscillators for transmitters, superheterodyne receivers, and heterodyne frequency meters in the 100- to 1000-MHz frequency range.

Helical Resonators A *helical resonator* is a modified cavity configuration often used in the 30- to 800-MHz range and consists of an inductor placed within a metal cavity. It is less expensive, smaller, and lighter than an equivalent cavity resonator for the lower portion of the vhf/uhf range (figure 17).

Helical resonators tuned to the same frequency band can be cascaded to produce a compact, bandpass vhf filter of high unloaded Q. Inter-resonator coupling is provided typically by capacitance or mutual coupling.

Design data for helical resonators may be found in the February, 1966 issue of *Micro-Waves* magazine, a Hayden publication.

The Strip-Line Circuit The *strip-line* circuit is another modification of the resonant cavity, making use of a flat strip of metal placed within a square or rectangular cavity or between two ground planes. This design is very useful in the vhf region, particularly for high-power amplifier stages (figure 18A).

A strip-line circuit may be tuned by a movable disc capacitor placed at the high-impedance end of the line. The line, more-

TABLE 1

COIL TABLE FOR GENERAL-PURPOSE H-F CONVERTER

Band	Coil L_2 L_3, L_6	Approx. Ind. (μH)	Coil L_1, L_4, L_5
6	8 to 10 turns #16e., ⅝" diam. × 1" long (Air-Dux 516-T)	0.7 μH	2 turns hookup wire at cold end
10	14 turns #18e., ⅝" diam. × ½" long (Air-Dux 532-T)	2.2 μH	2 turns hookup wire at cold end
15	24 turns #18e., ½" diam. × ¾" long (Air-Dux 432-T)	4 μH	3 turns hookup wire at cold end

Note: Adjacent cold ends of L_3 and L_6 are ¾" apart.

L_8—40 turns #28e., wire closewound on ⅜" diam. slug-tuned form. (J. W. Miller 42A-OOOCBI)
L_9—3 turns hookup wire over B-plus end of L_8.

OSCILLATOR DATA

BAND	CRYSTAL (MHz)	COIL L_7
6	43 MHz	10 turns #28e., closewound on ½" diam. slug-tuned form (1.8μH). (J. W. Miller 41A-OOCBI)
10	21 MHz	Same as for 6 meters, with 50-pf variable mica capacitor in parallel
15	14 MHz	Same as for 6 meters, with 100-pf variable mica capacitor in parallel

over, may consist of a "sandwich" of two plates, separated by insulating material, thus permitting operating voltage to be applied to the tube or device, yet isolating the tuning capacitor from the d-c voltage.

The *microstrip line* is a version of the strip line adapted to circuit-board techniques (figure 18B). A dual copper surface board is used, one face being the ground plane, the board forming the dielectric of the line and the opposite face being etched to form the strip line. Design data on strip and microstrip lines may be found in *Reference Data For Radio Engineers*, Howard W. Sams & Co., Inc.

10-17 Representative VHF Converter Circuits

Shown in this section are representative circuits of converters for the various vhf bands. The state of the art embracing vhf solid-state devices advance rapidly and transistors and FETS used today may become obsolete tomorrow. The circuit designs shown, however, reflect modern concepts in vhf circuitry and are adaptable to other devices than those shown, with appropriate voltage changes.

A tube converter is shown for the 6-meter

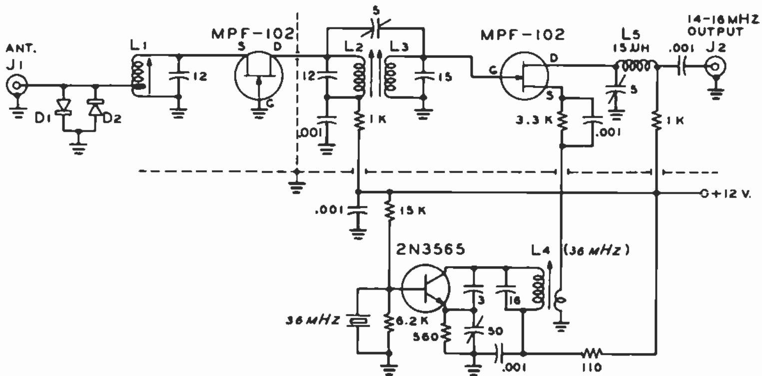


Figure 20

FET CONVERTER FOR 50 MHz

L_1, L_2—10 turns #20 e. on ¼" diameter slug-tuned form, spaced ⅜" long (approx. 0.7 μH). Use J. W. Miller 4501. Coils L_2 and L_3 mounted with 1" spacing, center to center.
L_4—Same as L_1 with 1-turn loop of hookup wire at "cold" end
L_5—J. W. Miller 20A-155-RBI D_1, D_2—1N100 or 1N34A

band that has high overload capability and is recommended for use in congested, high-signal areas. Also shown are various solid-state designs for the 50-, 144-, and 432-MHz bands.

Except as otherwise indicated, decimal values of capacitances are in microfarads, all other values are in picofarads. Resistances are in ohms or kilohms (K) and are ½-watt values unless otherwise indicated. Tuned circuits are approximate value and are grid-dipped to frequency. Bypass capacitors are either feedthrough types, or equivalent low-inductance units suitable for vhf operation.

An "Antioverload" 50-MHz Converter The problem of overload and crossmodulation is acute in metropolitan areas, particularly on 50 MHz. This converter provides overload protection to unwanted signals as strong as 100,000 microvolts removed from the wanted signal by only 75 kHz. The noise figure is better than 5 decibels. The schematic is shown in figure 19 and coil data for operation on the 6- , 10- , and 15 meter bands is given in Table I.

The converter is built on a 4" × 6" copper-laminate (two sides) phenolic board and a 4" × 6" × 2" aluminum chassis box serves as a support and shield. All components except the gain control (R_1) are mounted on the board. A small shield is placed across the r-f amplifier socket to prevent oscillation and instability.

Interstage coils L_5 and L_6 lie along the same axis and are spaced about ¾" apart, with the "cold" ends facing each other. With all circuits peaked for maximum signal the link coils are adjusted for minimum signal consistent with good reception and the prevailing state of nearby strong signals. For use in the lower 1 MHz of the band, all circuits may be peaked to 50.5 MHz.

A FET Converter for 50 MHz The FET converter shown in figure 20 provides good overload and crossmodulation characteristics and a noise figure better than the tube converter discussed in the previous paragraph. Reverse-connected diodes across the input protect the FET devices from transient voltages.

The converter is built upon a 3" × 5" copper-laminate (two sides) circuit board which is mounted on the top of an aluminum box which serves as a shield. Small pieces of board are soldered to the "chassis" board to provide interstage shields between the various circuits.

Initial alignment may be made with a local

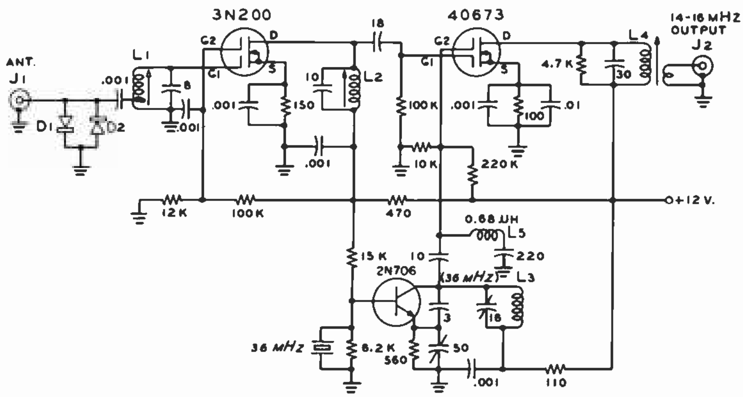


Figure 21

MOSFET CONVERTER FOR 50 MHz

D_1, D_2—1N100 or 1N34A
L_1—11 turns #20 e. on 1/4" diameter slug-tuned form, spaced 3/8" long (approx. 0.9 µH). Use J. W. Miller 4502
L_2—As above, but 9 turns. Use J.W. Miller 4501
L_3—1.5 µH. Use J. W. Miller 4503
L_4—2.5 µH. Use J. W. Miller 4503. Link 3 turns of hookup wire
L_5—0.68 µH. Use J. W. Miller 9250-681

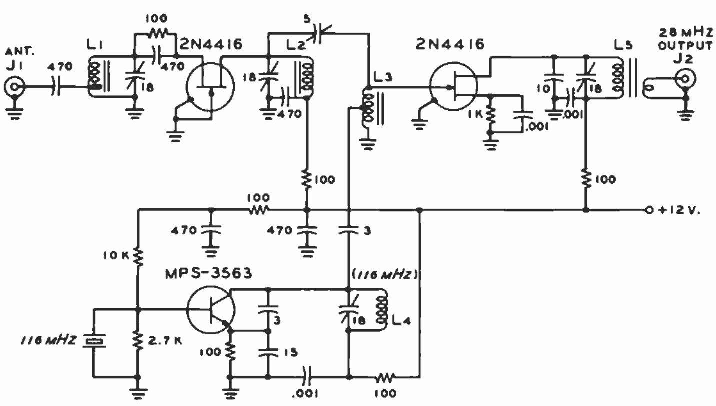


Figure 22

JFET CONVERTER FOR 144 MHz

L_1, L_3—8 turns #22 e. tapped 3 turns from "cold" end. Wound on Micrometals T30-0 miniature ferrite toroid core
L_2—As above, untapped
L_4—3 turns #18 e., 3/8" diam., 3/8-inch long
L_5—21 turns #28 e. wound on Micrometals T-37-10 core. Secondary 3 turns #28 e. over "cold" end of primary winding.

signal, peaking all coils for maximum response. Input coil L_1 should then be adjusted for best signal-to-noise ratio on a weak signal. The 5-pF coupling capacitor between the interstage coils should be set at minimum capacitance consistent with the bandwidth response desired.

A MOSFET Converter for 50 MHz This converter features an internally protected MOSFET device in the r-f stage, eliminating most of the former problems of premature burnout common to unprotected devices. Sensitivity, noise figure, and resistance to crossmodulation are good and

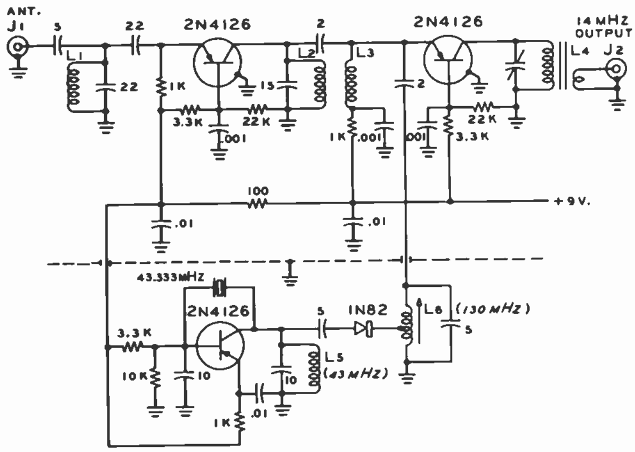


Figure 23

UTILITY CONVERTER FOR 144 MHz

L_1—4 turns #20 e., 3/16-inch i.d., 5/16-inch long
L_2—5 turns, as above
L_3—15 turns #20 e., 3/16-inch i.d., 5/16-inch long
L_4—1.2 μH. Use J. W. Miller 4502. Link 3 turns of hookup wire
L_5—6 turns #16 e., 1/4-inch i.d., 3/8-inch long

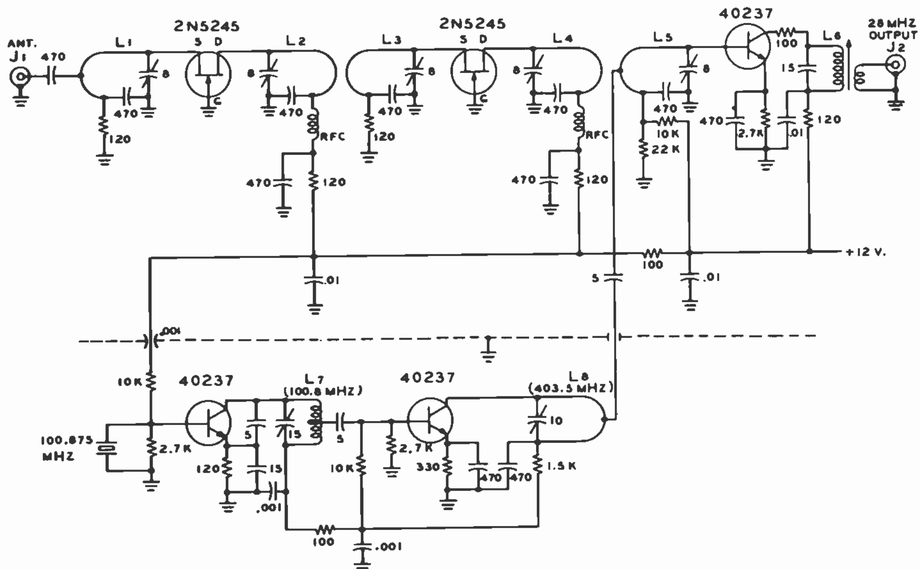


Figure 24

LOW NOISE 432 MHz CONVERTER

L_1, L_2, L_3, L_4, L_8—Copper strap 2 1/8" long × 1/4" wide bent into U-shaped loop about 1 3/8" high
L_5—As above, but 1 7/8" long
L_6—18 turns #24 e., on 3/16" diam. slug tuned form. Link 3 turns of hookup wire
L_7—5 turns #20 e., 1/4" diam., 1/2" long. Tap 1 turn from top

a trade off between noise figure and overload capability may be achieved by varying the 12K resistor connected between gate 2 and ground of the r-f stage. A higher value will provide a better noise figure and a lower value will provide greater overload protection (figure 21).

The converter is built upon a 2½" × 4" copper-laminate (two sides) board which is mounted to the open side of an alumnium chassis used for shielding and support. Placement of components is not critical. A small shield made of circuit-board material placed between coils L_1 and L_2 may be necessary to improve amplifier stability.

A JFET Converter for 144 MHz This general purpose 2-meter converter is ideal for general operation, combining good noise figure with excellent overload capability. It is a good beginners project as the circuit is simple and easy to get working.

The converter may be built upon a 3" × 5" copper-laminate (two sides) circuit board and mounted on the top of an aluminum box to serve as a shield. No internal shields are required. The use of a high-overtone crystal eliminates the bothersome "birdie" problem, common with many converters using lower-frequency oscillator injection (figure 22).

Tuned circuits are peaked on a local signal and then the input circuit is readjusted for best signal-to-noise ratio on a weak signal. Capacitance coupling between L_2 and L_3 should be the minimum value for good signal response.

A General Purpose Converter for 144 MHz This simple converter uses PNP transistors in a proven circuit. Components are mounted on a 4" × 2" copper-laminate (two sides) board which may be placed within an aluminum box, or mounted to an aluminum chassis to provide protection and shielding. Placement of parts is not critical. Coils are air-wound and soldered at one end directly to the copper board. Coils L_4 and L_5 are at right angles to each other coupled by the 2-pF capacitor. A shield is placed between the oscillator and multiplier circuits and the r-f amplifier and mixer to reduce unwanted coupling, as shown in the schematic of figure 23.

The converter is aligned on a local signal and the input circuit peaked for best signal-to-noise ratio on a weak signal.

A Low Noise Converter for 432 MHz This inexpensive and easily adjusted converter provides a noise figure better than 4 decibels at 432 MHz. Two 2N5245 FET devices are used as cascade grounded gate amplifiers, followed by a 40237 mixer stage. Oscillator injection is at 403.5 MHz for a 28-MHz i-f system, as shown in figure 24.

The converter may be built on a 5" × 7" copper-laminate (two sides) circuit board with a shield separating the local oscillator chain from the r-f signal stages.

The 470-pF capacitors in the tuned circuits are vhf button-mica units soldered directly to small holes drilled in the circuit board. The various vhf coils are hairpin loops made of ¼-inch wide, 20-gauge flashing copper and are mounted in place between the mica capacitors and the piston-type variable capacitors. Coil pairs L_2, L_3 and L_4, L_5 are mounted parallel to each other, with the center line of the inductors about ¾ inch apart. The hairpin coils are mounted vertically with respect to the surface of the board. Coil inductance is critical, and the circuits may be grid-dipped to about 440 MHz with the transistors out of the circuit.

As in the case of the other converters, preliminary alignment is done with a local signal and fine alignment made with a weak signal, or noise generator for best signal-to-noise ratio.

CHAPTER ELEVEN

Generation and Amplification of Radio-Frequency Energy

Part I - H-F Circuits

A radio communication or broadcast transmitter consists of a source of radio frequency power, or *carrier;* a system for *modulating* the carrier whereby voice or telegraph keying or other modulation is superimposed upon it; and an antenna system, including feedline, for *radiating* the intelligence-carrying radio-frequency power. The power supply employed to convert primary power to the various voltages required by the r-f and modulator portions of the transmitter may also be considered part of the transmitter.

Modulation usually is accomplished by varying either the amplitude or the frequency of the radio-frequency carrier in accord with the components of intelligence to be transmitted or by generation of an SSB signal.

Radiotelegraph keying normally is accomplished either by interrupting, shifting the frequency of, or superimposing an audio tone on the radio-frequency carrier in accordance with the intelligence to be transmitted.

The complexity of the radio-frequency generating portion of the transmitter is dependent on the power, order of stability, and frequency desired. An oscillator feeding an antenna directly is the simplest form of radio-frequency generator. A modern high-frequency transmitter, on the other hand, is a very complex generator. Such equipment comprises a very stable crystal-controlled or synthesized oscillator to stabilize the output frequency, a series of frequency multipliers, or mixers, one or more amplifier stages to increase the power up to the level which is desired for feeding the antenna system, and a filter system for keeping the harmonic energy generated in the transmitter from being fed to the antenna system.

11-1 Self-Controlled Oscillators

The amplifying properties of a three- (or more) element vacuum tube, a bipolar transistor, or an FET give them the ability to generate an alternating current of a frequency determined by auxiliary components associated with them. Such circuits are termed *oscillators.* To generate a-c power with an amplifier, a portion of the output power must be returned or fed back to the input in phase with the starting power (figure 1). The power delivered to the load will be the output power less the feedback power.

Initial Oscillation Oscillation may be initially caused in a transistor or tube circuit by external triggering, or by self-excitation. In the latter case, at the moment the d-c power is applied, the energy level does not instantly reach maximum but, instead, gradually approaches it. Oscillations build up to a point limited by the normal operation of the amplifier, the feedback energy, and the nonlinear condition of the circuit. Practical oscillator circuits employ a variety of feedback paths, and some of the most useful ones are shown in figure 2. Either tubes, transistors, or FETs may be used in these circuits.

The oscillator is commonly described in terms of the feedback circuit. The *Hartley* oscillator (figure 2A) employs a tapped inductor in the resonant circuit to develop the proper phase relationship for the feedback voltage, while the *Colpitts* oscillator derives the exciting voltage by means of a capacitive voltage divider. The *Clapp* circuit (figure 2C) employs a series-tuned tank circuit, shunted by a large capacitive voltage divider (C_1-C_2).

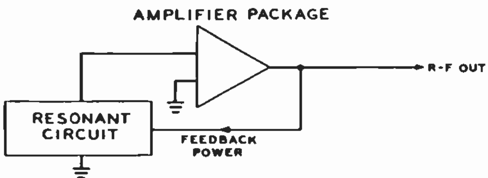


Figure 1

THREE TERMINAL OSCILLATOR

A portion of the output of a three-terminal amplifier is fed back to the input in proper phase and amplitude with the starting power which is generated initially by thermal noise. Power delivered to the load is output power less feedback power. Resonant circuit in input determines frequency of oscillation.

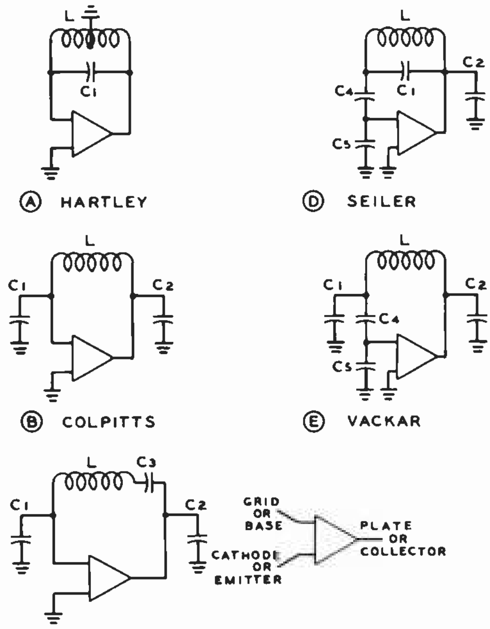


Figure 2

COMMON TYPES OF SELF-EXCITED OSCILLATORS

The circuits are named after the inventors and are based on variations in the method of coupling and introducing feedback into oscillator tank circuit. A—Hartley circuit with inductive feedback. B—Colpitts circuit with capacitive feedback. C—Clapp circuit with capacitive feedback plus series-tuned tank. D—Seiler circuit with capacitive feedback and separate parallel-tuned tank circuit. E—Vackar circuit with capacitive feedback plus parallel-tuned tank circuit. Circuits may be used with either solid-state devices or vacuum tubes by adjustment of feedback amplitude and applied potentials.

The *Seiler* and *Vackar* circuits employ a voltage divider (C_1-C_5) to establish the correct feedback level for proper operation. At resonance, all circuits are versions of a pi-network in one way or another, the tuning scheme and feedback path being different for the various configurations. Vacuum-tube versions of these circuits are shown in figure 3.

When plate voltage is applied to the Hartley circuit (figure 3A), the sudden flow of plate current accompanying the application of plate voltage will cause an electromagnetic field to be set up about the coil, resulting in a potential drop across the turns of the coil. Due to the inductive coupling between the portion of the coil in which the plate current is flowing and the grid portion, a potential will be induced in the grid portion.

Since the cathode tap is between the grid and plate ends of the coil, the induced grid voltage acts in such a manner as to increase further the plate current to the tube. This action will continue for a short period of time determined by the inductance and capacitance of the tuned circuit, until the *flywheel effect* of the tuned circuit causes this action to come to a maximum and then to reverse itself. The plate current then decreases (the magnetic field around the coil also decreasing) until a minimum is reached, when the action starts again in the original direction and at a greater amplitude than before. The amplitude of these oscillations, the frequency of which is determined by the coil-capacitor circuit, will increase in a very short period of time to a limit determined by the plate voltage of the oscillator tube.

The Colpitts Figure 3B shows a version of the *Colpitts* oscillator. It can be seen that this is essentially the same circuit as the Hartley except that the ratio

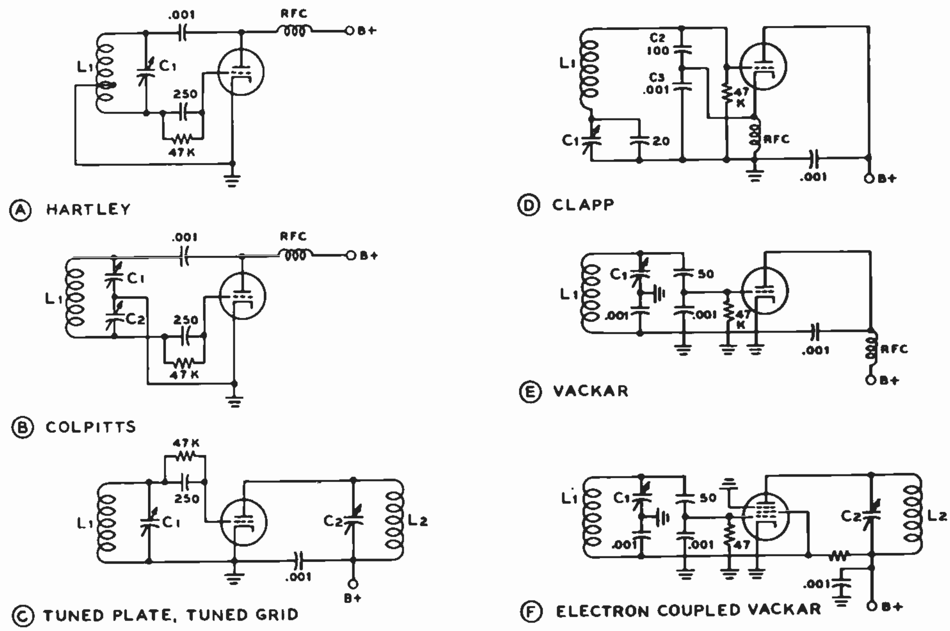


Figure 3

VACUUM-TUBE SELF-EXCITED OSCILLATORS

A—Shunt-fed Hartley. B—Shunt-fed Colpitts. C—Series-fed oscillator with feedback accomplished through plate-to-grid interelectrode capacitance. D—Clapp oscillator employs series-resonant tuned circuit. Capacitor C_1 is of the order of 50 pf. E—Vackar circuit is variation of Clapp circuit having improved tuning range and more constant output. F—Electron-coupled oscillator using screen element of tube as the plate of the oscillator.

of a pair of capacitances in series determines the effective cathode tap, instead of actually using a tap on the tank coil. Also, the net capacitance of these two capacitors comprises the tank capacitance of the tuned circuit. This oscillator circuit is somewhat less susceptible to parasitic (spurious) oscillations than the Hartley.

For best operation of the Hartley and Colpitts oscillators, the voltage from grid to cathode, determined by the tap on the coil or the setting of the two capacitors, normally should be from 1/3 to 1/5 that appearing between plate and cathode.

The T.P.T.G. The *tuned-plate tuned-grid* oscillator illustrated at (C) has a tank circuit in both the plate and grid circuits. The feedback of energy from the plate to the grid circuits is accomplished by the plate - to - grid interelectrode capacitance within the tube. The necessary phase reversal in feedback voltage is provided by tuning the grid tank capacitor to the low side of the desired frequency and the plate capacitor to the high side. A broadly resonant coil may be substituted for the grid tank to form the *T.N.T.* (tuned-not-tuned) oscillator.

Electron-Coupled Oscillators In any of the oscillator circuits just described it is possible to take energy from the oscillator circuit by coupling an external load to the tank circuit. Since the tank circuit determines the frequency of oscillation of the tube, any variations in the conditions of the external circuit will be coupled back into the frequency-determining portion of the oscillator. These variations will result in frequency instability.

The frequency-determining portion of an oscillator may be coupled to the load circuit only by an electron stream, as illustrated in (F) of figure 3. When it is considered that the screen of the tube acts as the plate to the oscil!ator circuit, the plate merely acting as a coupler to the load, then

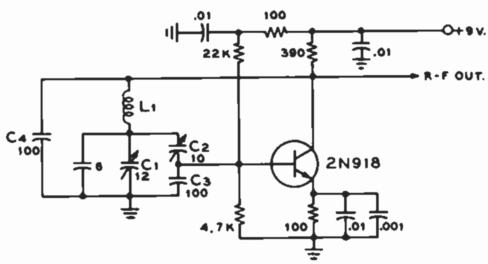


Figure 4

TRANSISTORIZED VACKAR OSCILLATOR

Thirty-MHz oscillator for vhf frequency control. Coil L_1 is 1.5 µH, wound on a ceramic form. Capacitor C_2 is adjusted for optimum drive level.

the similarity between the cathode-grid-screen circuit of these oscillators and the cathode-grid-plate circuits of the corresponding prototype can be seen.

The *electron-coupled* oscillator has good stability with respect to load and voltage variation. Load variations have a relatively small effect on the frequency, since the only coupling between the oscillating circuit and the load is through the electron stream flowing through the other elements to the plate The plate is electrostatically shielded from the oscillating portion by the bypassed screen.

The Clapp Oscillator The *Clapp* oscillator differs from the previous circuits in that it employs a series-resonant circuit while in all the more common oscillator circuits the frequency-controlling circuit is parallel resonant.

The Clapp oscillator operates in the following manner: at the resonant frequency of the oscillator tuned circuit (L_1, C_1) the impedance of this circuit is at minimum (since it operates in series resonance) and maximum current flows through it. Note however, that C_2 and C_3 also are included within the current path for the series-resonant circuit, so that at the frequency of resonance an appreciable voltage drop appears across these capacitors. The voltage drop appearing across C_2 is applied to the grid of the oscillator tube as excitation, while the amplified output of the oscillator tube appears across C_3 as the driving power to keep the circuit in oscillation.

Capacitors C_2 and C_3 should be made as large in value as possible, while still permitting the circuit to oscillate over the full tuning range of C_1. The larger these capacitors are made, the smaller will be the coupling between the oscillating circuit and the tube, and consequently the better will be oscillator stability with respect to tube variations. High-G_m tubes such as the 6AH6, 5763, and 6CB6 will permit the use of larger values of capacitance at C_2 and C_3 than will more conventional tubes such as the 6BA6, 6AQ5, and such types. In general it may be said that the reactance of capacitors C_2 and C_3 should be on the order of 40 to 120 ohms at the operating frequency of the oscillator—with the lower values of reactance going with high-G_m tubes and the higher values being necessary to permit oscillation with tubes having G_m in the range of 2000 micromhos.

It will be found that the Clapp oscillator will have a tendency to vary in power output over the frequency range of tuning capacitor C_1. The output will be greatest where C_1 is at its largest setting, and will tend to fall off with C_1 at minimum capacitance. In fact, if capacitors C_2 and C_3 have too large a value the circuit will stop oscillating near the *minimum* capacitance setting of capacitor C_1.

Hence it will be necessary to use a slightly *smaller* value of capacitance at C_2 and C_3 (to provide an increase in the capacitive reactance at this point), or else the frequency range of the oscillator must be restricted by paralleling a fixed capacitor across C_1 so that its effective capacitance at minimum setting will be increased to a value which will sustain oscillation.

The Vackar Oscillator The *Vackar* oscillator is a variation of the basic Clapp circuit which has improved tuning range and relatively constant output combined with good stability with respect to a varying load. A practical Vackar circuit designed for 30 MHz is shown in figure 4. With the constants shown, the range is from 26.9 to 34.7 MHz, with an output amplitude change of less than -1.5 db relative to the lower frequency. Capacitor C_1 tunes the circuit while capacitor C_2 is adjusted for optimum drive level such that the transistor is not driven to cutoff or saturation.

The output level, when properly adjusted, is about 4 volts peak-to-peak for a 9-volt supply. The emitter-bias resistor is bypassed for r-f and audio frequencies to eliminate a tendency for the circuit to oscillate at a parasitic frequency that is low in comparison to the working frequency. The value of capacitors C_3 and C_4 are approximately:

$$C\ (\text{pf}) = \frac{3000}{f\ (\text{MHz})}$$

The frequency of oscillation is approximately:

$$f_{(osc)} = \frac{1}{2\pi \sqrt{L\ (C_1 + C_2)}}$$

The Seiler Oscillator The *Seiler* oscillator is another variation of the Clapp circuit, permitting one end of the tank coil to be at ground potential, and exhibiting slightly less loading of the tuned circuit than either the Vackar or the Clapp configuration. The large capacitors placed across the amplifying tube or transistor tend to swamp out any reactive changes in the active device and also limits the harmonic output, thereby enhancing frequency stability. A Seiler oscillator designed for SSB service is shown in figure 5.

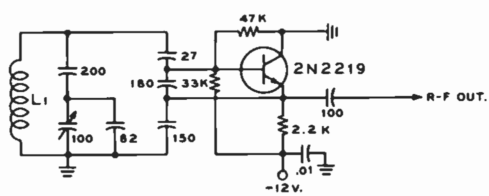


Figure 5

TRANSISTORIZED SEILER OSCILLATOR

Seiler oscillator is a variation of the Clapp circuit which permits one end of the tank coil to be at ground potential. Coil is 22 turns # 16e., 1¼" diam., close wound for range of 5.0-5.6 MHz.

Negative-Resistance Oscillators *Negative - resistance* oscillators often are used when unusually high frequency stability is desired, as in a frequency meter. The *dynatron* of a few years ago and the newer *transitron* are examples of oscillator circuits which make use of the negative-resistance characteristic between different elements in some multigrid tubes.

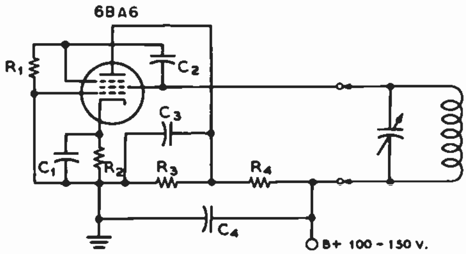
One version of the transitron circuit uses a pentode tube with the suppressor element coupled to the screen. The negative resistance is obtained from a combination of secondary emission and interelectrode coupling. A representative transitron circuit is shown in figure 6A.

The chief distinction between a conventional *negative-grid* oscillator and a *negative-resistance* oscillator is that in the former the tank circuit must act as a phase inverter in order to permit the amplification of the tube to act as a negative resistance, while in the latter the tube acts as its own phase inverter (figure 6B). Thus a negative-resistance oscillator requires only an untapped coil and a single capacitor as the frequency-determining tank circuit, and is classed as a *two-terminal oscillator*. In fact, the time constant of an RC circuit may be used as the frequency-determining element and such an oscillator is rather widely used as a tunable audio-frequency oscillator.

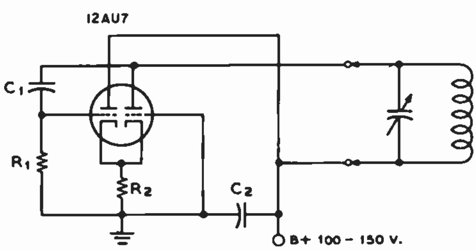
The Franklin Oscillator The *Franklin* oscillator makes use of two cascaded tubes to obtain the negative-resistance effect (figure 7). The tubes may be either a pair of triodes, tetrodes, or pentodes; a dual triode; or a combination of a triode and a multigrid tube. The chief advantage of this oscillator circuit is that the frequency-determining tank only has two terminals, and one side of the circuit is grounded.

The second tube acts as a phase inverter to give an effect similar to that obtained with the dynatron or transitron, except that the effective transconductance is much higher. If the tuned circuit is omitted or is replaced by a resistor, the circuit becomes a *relaxation* oscillator or a *multivibrator*.

Oscillator Stability The oscillator providing minimum coupling between the active device and the tuned circuit has proven to have the highest degree of stability. However, this inherently good stability is with respect to tube or transistor variations; instability of the tuned circuit with respect to vibration or temperature will of course have as much effect on the frequency



Ⓐ TRANSITRON OSCILLATOR



Ⓑ CATHODE COUPLED OSCILLATOR

Figure 6

TWO-TERMINAL OSCILLATOR CIRCUITS

Both circuits may be used for an audio oscillator or for frequencies into the vhf range simply by placing a tank circuit tuned to the proper frequency where indicated on the drawing. Recommended values for the components are given below for both oscillators.

TRANSITRON OSCILLATOR
C_1—0.01-μfd mica for r.f. 10-μfd elect. for a.f.
C_2—0.00005-μfd mica for r.f. 0.1-μfd paper for a.f.
C_3—0.003-μfd mica for r.f. 0.5-μfd paper for a.f.
C_4—0.01-μfd mica for r.f. 8-μfd elect. for a.f.
R_1—220K ½-watt carbon
R_2—1800 ohms ½-watt carbon
R_3—22K 2-watt carbon
R_4—22K 2-watt carbon

CATHODE-COUPLED OSCILLATOR
C_1—0.00005-μfd mica for r.f. 0.1-μfd paper for audio
C_2—0.003-μfd mica for r.f. 8-μfd elect. for audio
R_1—47K ½-watt carbon
R_2—1K 1-watt carbon

of oscillation as with any other type of oscillator circuit. Solid mechanical construction of the components of the oscillating circuit, along with a small negative-coefficient compensating capacitor included as an element of the tuned circuit, usually will afford an adequate degree of oscillator stability.

VFO Transmitter Controls When used to control the frequency of a transmitter in which there are stringent limitations on frequency tolerance, several precautions are taken to ensure that a variable-frequency oscillator will stay on frequency. The oscillator is fed from a voltage-regulated power supply, uses a well-designed and temperature-compensated tank circuit, is of rugged mechanical construction to avoid the effects of shock and vibration, is protected against excessive changes in ambient room temperature, and is isolated from feedback or stray coupling from other portions of the transmitter by shielding, filtering of voltage supply leads, and incorporation of one or more buffer-amplifier stages. In a high-power transmitter a small amount of stray coupling from the final amplifier to the oscillator can produce appreciable degradation of the oscillator stability if both are on the same frequency. Therefore, the oscillator usually is operated on a subharmonic or image of the transmitter output frequency, with one or more frequency multipliers or mixers between the oscillator and final amplifier.

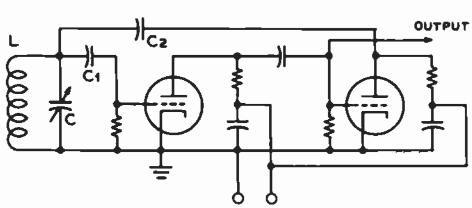


Figure 7

THE FRANKLIN OSCILLATOR CIRCUIT

A separate phase-inverter tube is used in th oscillator to feed a portion of the output ba to the input in the proper phase to sustain oscillation. The values of C_1 and C_2 should be as small as will permit oscillations to be sustained over the desired frequency range.

11-2 Quartz-Crystal Oscillators

Quartz is a naturally occurring crystal having a structure such that when plates

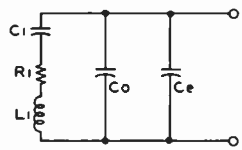


Figure 8

EQUIVALENT CIRCUIT OF A QUARTZ PLATE

The equivalent series-resonant circuit of the crystal itself is at the left, with shunt capacitance of electrodes and holder (C_o) and capacitance between electrodes with quartz as the dielectric (C_e) at right. The composite circuit may exhibit both series resonance and parallel resonance (antiresonance), the separation in frequency between the two modes being very small and determined largely by the ratio of series capacitance (C_1) to shunt capacitance.

are cut in certain definite relationships to the crystallographic axes, these plates will show the *piezoelectric effect*. That is, the plates will be deformed in the influence of an electric field, and, conversely, when such a plate is deformed in any way a potential difference will appear on its opposite sides.

A quartz-crystal plate has several mechanical resonances. Some of them are at very-high frequencies because of the stiffness of the material. Having mechanical resonance, like a tuning fork, the crystal will vibrate at a frequency depending on the dimensions, the method of electrical excitation, and crystallographic orientation. Because of the piezoelectric properties, it is possible to cut a quartz plate which, when provided with suitable electrodes, will have the characteristics of a resonant circuit having a very high LC ratio. The circuit Q of a crystal is many times higher than can be obtained with conventional inductors and capacitors of any size. The Q of crystals ranges from 10,000 to several million.

The equivalent electrical circuits of a quartz-crystal plate are shown in figure 8. The shunt capacitance of the electrodes and holder is represented by C_o, and the capacitance between the electrodes with quartz as the dielectric is C_e. The series capacitance (C_1) represents the motional elasticity of the quartz, while the inductance (L_1) is a function of the mass. The series resistance (R_1) represents the sum of the crystal losses, including friction, acoustic loading, and power transmitted to the mounting structure.

Practical Quartz Crystals While quartz, tourmaline, Rochelle salts, ADP, and EDT crystals all exhibit the piezoelectric effect, only quartz has a low temperature coefficient and exhibits chemical and mechanical stability. The greater part of the raw quartz used today for frequency control is man-made rather than natural and crystal blanks are produced in large quantities at low prices. The crystal blank is cut from a billet of quartz at a predetermined orientation with respect to the optical and electrical axes, the orientation determining the activity, temperature coefficient, thickness coefficient, and other characteristics of the crystal.

The crystal blank is rough-ground almost to frequency, the frequency increasing in inverse ratio to the oscillating dimensions (usually the thickness, but often the length). It is then finished to exact frequency by careful lapping, by etching, or by plating. Care is taken to stabilize the crystal so frequency and activity will not change with time.

Unplated crystals are mounted in pressure holders, in which an air gap exists between the crystal and electrodes. Only the corners of the crystal are clamped. At frequencies requiring a low ratio of length to thickness (usually below 2 MHz or so) a "free" air gap is required because even the corners of the crystal move.

Control of the orientation of the blank when cut from the quartz billet determines the characteristics of the crystal. The *turning point* (point of zero temperature coefficient) may be adjusted to room temperature, usually taken as 20° C. A graph of the normal frequency ranges of popular crystal cuts is shown in figure 9. For frequencies between 550 kHz and 55 MHz, the AT-cut crystal is now widely used. A large quantity of BT-cut crystals in the range of 6 MHz to 12 MHz exists as surplus stock from World War II. These crystals are mounted in the obsolete FT-243 style holder. The AT-cut, however, is now used because modern techniques allow it to be produced cheaply, and in quantity.

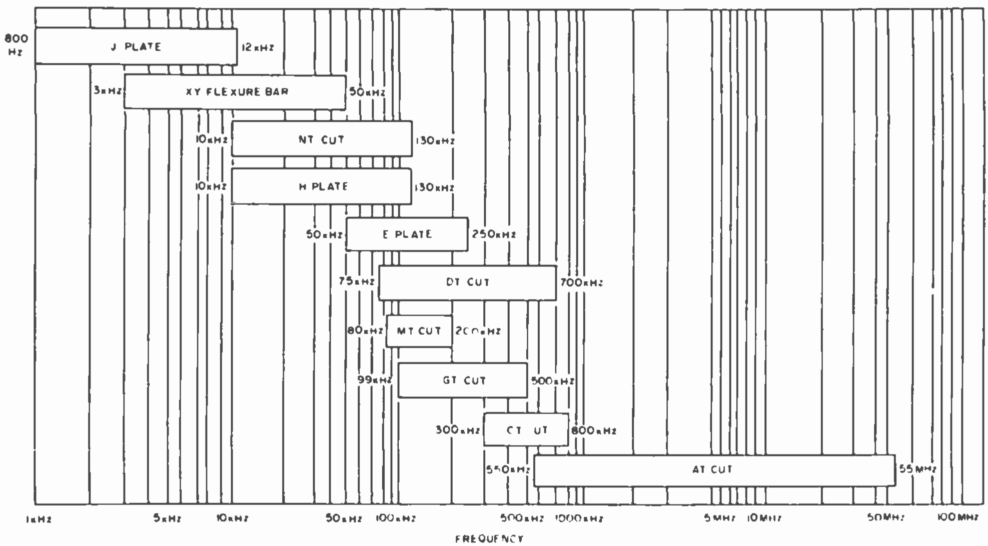


Figure 9

FREQUENCY RANGE OF CRYSTAL CUTS

Crystal Holders Crystals are normally purchased ready-mounted. Modern high-frequency crystals are mounted within metal holders, hermetically sealed with glass insulation and a metal-to-glass bond. Older crystal types make use of a phenolic holder sealed with a metal plate and a rubber gasket. A summary of crystal holders and crystal types is given in figure 10.

Precision crystals for calibrating equipment are vacuum-sealed in a glass envelope. Special vacuum-sealed crystals having a relatively constant temperature coefficient are used in high-stability frequency standards in place of the near-obsolete and expensive temperature-controlled "crystal oven."

Overtone-cut Crystals Just as a vibrating string can be made to vibrate on its overtone frequencies, a quartz crystal will exhibit mechanical resonance (and therefore electrical resonance) at overtones of its fundamental frequency. (The terms *overtone* and *harmonic* should not be used interchangeably. The overtone is a mechanical phenomenon and its frequency differs from the harmonic by virtue of the mechanical loading of the crystal. The harmonic is an electrical phenomenon and is an exact multiple of the fundamental frequency.)

By grinding the crystal especially for overtone operation, it is possible to enhance its operation as an overtone resonator. AT-cut crystals designed for optimum overtone operation on the 3rd, 5th, and even the 7th overtone are available. The 5th- and 7th-overtone types, especially the latter, require special holders and circuits for satisfactory operation, but the 3rd-overtone type needs little more consideration than a regular fundamental type. It is possible in some circuits to operate a crystal on the fundamental and 3rd overtone simultaneously and produce an audio beat between the third harmonic and the third overtone. Unless specifically desired, this operation is to be avoided in conventional circuits.

The overtone frequency for which the crystal is designed is the working frequency, which is not the fundamental, since the crystal actually oscillates on this working frequency when it is functioning in the proper manner. The Q of an overtone crystal, moreover, is much higher than that of a fundamental crystal of the same frequency. As a result, overtone crystals are less prone to frequency change brought about by changes of oscillator input capacitance. Many frequency-standard crystals in the h-f range, therefore, are overtone types.

QUARTZ CRYSTAL HOLDERS

Holder Type	Pin Spacing	Pin Diam.	Size H	Size W	Size T
HC-5/U	0.812	0.156	2.20	1.82	1.60
HC-6/U	0.486	0.050	0.78	0.76	0.35
HC-10/U	(1)	0.060	1.10	—	0.56D
HC-13/U	0.486	0.050	0.78	0.76	0.35
HC-17/U	0.486	0.093	0.78	0.76	0.35
HC-18/U	(2)	——	0.53	0.40	0.15
HC-25/U	0.192	0.040	1.53	0.76	0.35
FT-243	0.500	0.093	1.10	0.90	0.40

(1)—Barrel Mount
(2)—Wire Leads 0.018 Diam.

QUARTZ CRYSTAL TYPES

Mil. Type	Holder Used	Type	Resonance
CR-15B/U	HC-5/U	Fund.	Parallel
CR-16B/U	HC-5/U	Fund.	Series
CR-17/U	HC-10/U	Overtone	Series
CR-18A/U	HC-6/U	Fund.	Parallel
CR-19A/U	HC-6/U	Fund.	Series
CR-23/U	HC-6/U	Overtone	Series
CR-24/U	HC-10/U	Overtone	Series
CR-27/U	HC-6/U	Fund.	Parallel
CR-28A/U	HC-6/U	Fund.	Series
CR-32A/U	HC-6/U	Overtone	Series
CR-52A/U	HC-6/U	Overtone	Series
CR-53A/U	HC-6/U	Overtone	Series

Figure 10
CRYSTAL HOLDERS AND TYPES

Crystal Drive Level Crystal dissipation is a function of the drive level. Excessive crystal current may lead to frequency drift and eventual fracture of the blank. The crystal oscillator should be run at as low a power level as possible to reduce crystal heating. Drive levels of 5 milliwatts or less are recommended for fundamental AT blanks in HC-6/U style holders, and a level of 1 milliwatt maximum is recommended for overtone crystals or fundamental crystals above 10 MHz in HC-6/U holders. The older FT-243 style crystal is capable of somewhat greater drive levels by virtue of the larger blank size.

Series and Parallel Resonance The shunt capacitance of the electrodes and associated wiring is considerably greater than the capacitive component of an equivalent series LC circuit, and unless the shunt capacitance is balanced out, the crystal will exhibit both series- and parallel-resonance frequencies, the latter being somewhat higher than the former. The series-resonant condition is employed in filter circuits and in oscillator circuits wherein the crystal is used in such a manner that the phase shift of the feedback voltage is at the series-resonant frequency.

The only difference between crystals designed for series-resonance and those for parallel-resonance operation is the oscillator input reactance (capacitance) for which they are calibrated. A crystal calibrated for parallel resonance will operate at its calibrated frequency in a series-resonant circuit with the addition of an appropriate value of series capacitance. Thus, a crystal cannot be specified in frequency without stating the reactance with which it is to be calibrated. The older FT-243 fundamental crystals were usually calibrated with a parallel capacitance of 35pf, while many of the new hermetic sealed crystals are calibrated with a capacitance of 32 pf.

11-3 Crystal-Oscillator Circuits

A crystal may replace the conventional tuned circuit in a self-excited oscillator, the crystal oscillating at its series- or parallel-resonant frequency. Basic oscillator circuits are shown in figure 11. Series mode operation of the crystal is used in these circuits. In the solid-state circuits, the holder capacitance of the crystal package is parallel-resonated by the shunt-connected r-f choke, assuring that the crystal oscillates at the correct overtone, as marked on the holder.

Bipolar transistors have a much lower input impedance than the grid of a vacuum tube and this makes the use of the transistor impractical in circuits that use parallel-resonant crystals, such as the Pierce oscillator. Other possible oscillator circuits are suggested in figure 12.

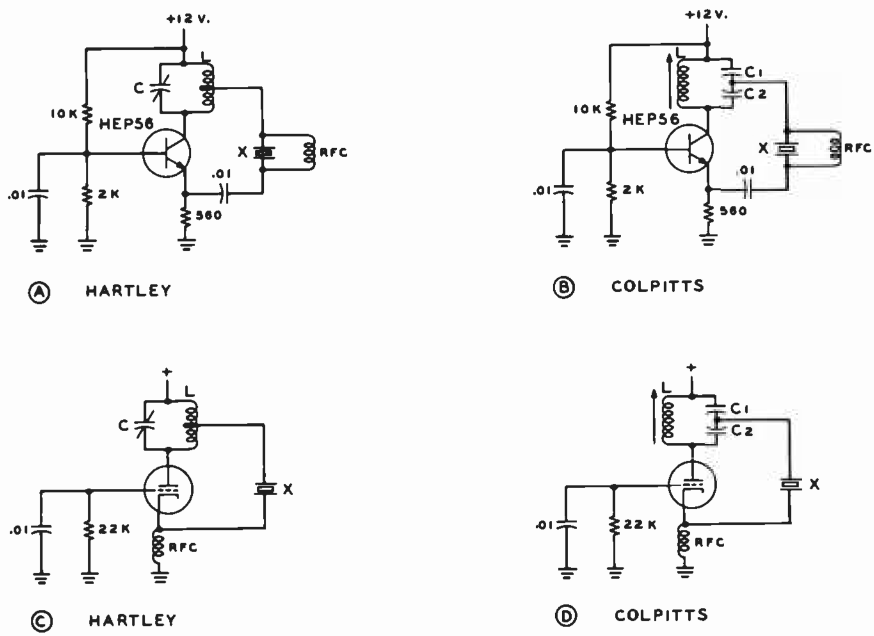


Figure 11

BASIC CRYSTAL OSCILLATOR CIRCUITS

Series-mode operation of the crystal is used in these circuits. Bipolar transistors have a low input impedance that makes parallel-resonant crystal circuits impractical. These circuits are versions of the basic oscillator configurations shown in figure 2.

Tuned-Plate Crystal Oscillator The *Miller,* or tuned-plate crystal oscillator is shown in figure 13A. The plate tank is tuned on the low capacitance side of resonance and oscillation occurs near the parallel-resonant frequency of the crystal.

The diagram shown in figure 13A is the basic circuit. The most popular version of the tuned-plate oscillator employs a pentode or beam tetrode with cathode bias to prevent excessive plate dissipation when the circuit is not oscillating. The cathode resistor is optional. Its omission will reduce both crystal current and oscillator efficiency, resulting in somewhat more output for a given crystal current. The tube usually is an audio or video beam pentode or tetrode, the plate-grid capacitance of such tubes being sufficient to ensure stable oscillation but not so high as to offer excessive feedback with resulting high crystal current. The 6CL6 makes an excellent all-around tube for this type circuit.

Pentode Harmonic Crystal Oscillator Circuits The usual type of crystal-controlled h-f transmitter operates, at least part of the time, on a frequency which is an approximate multiple of the operating frequency of the controlling crystal. Hence, oscillator circuits which are capable of providing output on the crystal frequency if desired, but which also can deliver output energy on harmonics of the crystal frequency have come into wide use. Four such circuits which have found wide application are illustrated in figures 13C through 13F.

The circuit shown in figure 13C is recommended for use with overtone-cut crystals when output is desired on a multiple of the oscillating frequency of the crystal. As an example, a 25-MHz overtone-cut crystal may be used in this circuit to obtain output on 50-MHz or a 48-MHz overtone-cut crystal may be used to obtain output on the 144-MHz amateur band. The circuit is not

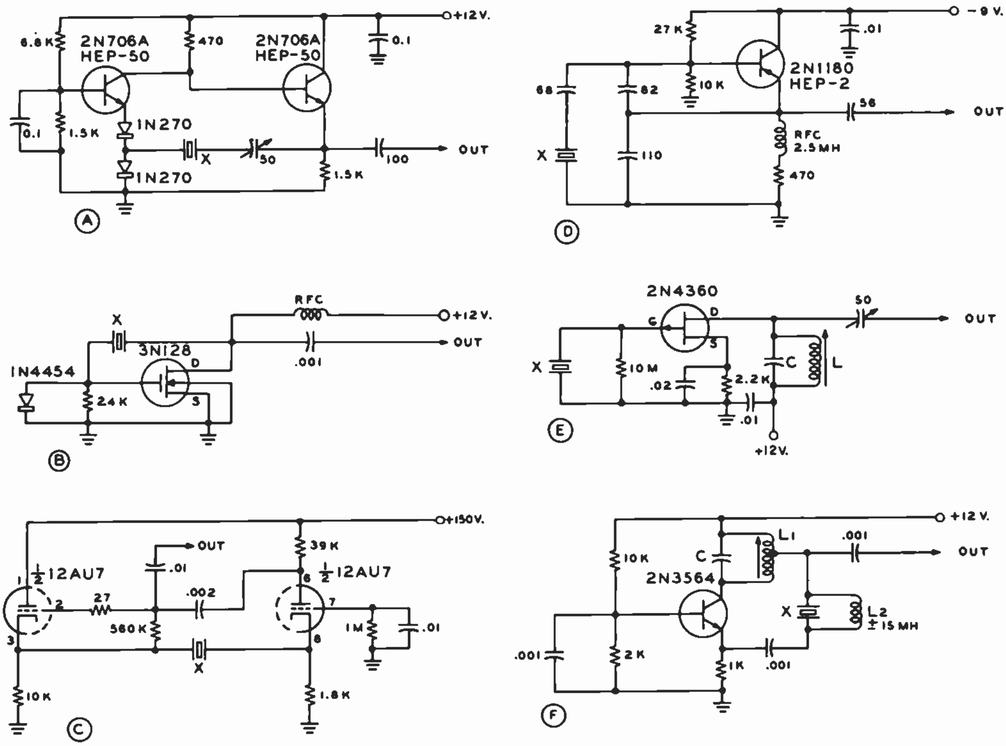


Figure 12

HIGH-FREQUENCY CRYSTAL-OSCILLATOR CIRCUITS

A—Transistorized Butler oscillator with amplitude-limiting diodes. The crystal is adjusted to frequency by series capacitor. Circuit is usable over range of 1 to 25 MHz. B—Pierce oscillator using FET with crystal in series-resonant mode. Drive voltage is clamped by diode. C—Vacuum-tube version of Butler oscillator with second triode section serving as a phase inverter. Circuit is designed for low-frequency operation (80 to 1000 kHz). D—General purpose h-f crystal oscillator for 2- to 30-MHz range. E—FET crystal oscillator for h-f range. Tuned circuit may be adinusted to overtone frequency of crystal. F—Overtone oscilllator. Coil L_2 resonates to crystal frequency with capacitance of crystal holder.

recommended for use with the normal type of fundamental-frequency crystal since more output with fewer variable elements can be obtained with the circuits of 13D and 13F.

The *Pierce harmonic* circuit shown in figure 13D is satisfactory for many applications which require very low crystal current, but has the disadvantage that both sides of the crystal are above ground potential. The *Tri-tet* circuit of figure 13E is widely used and can give excellent output with low crystal current. However, the circuit has the disadvantages of requiring a cathode coil, of requiring careful setting of the variable cathode capacitor to avoid excessive crystal current when changing frequency ranges, and of having both sides of the crystal above ground potential.

The *Colpitts harmonic oscillator* of figure 13F is recommended as being the most generally satisfactory harmonic crystal oscillator circuit since it has the following advantages: (1) the circuit will oscillate with crystals over a very wide frequency range with no change other than plugging in or switching in the desired crystal; (2) crystal current is extremely low; (3) one side of the crystal is grounded, which facilitates crystal-switching circuits; (4) the circuit will operate straight through without frequency pulling,

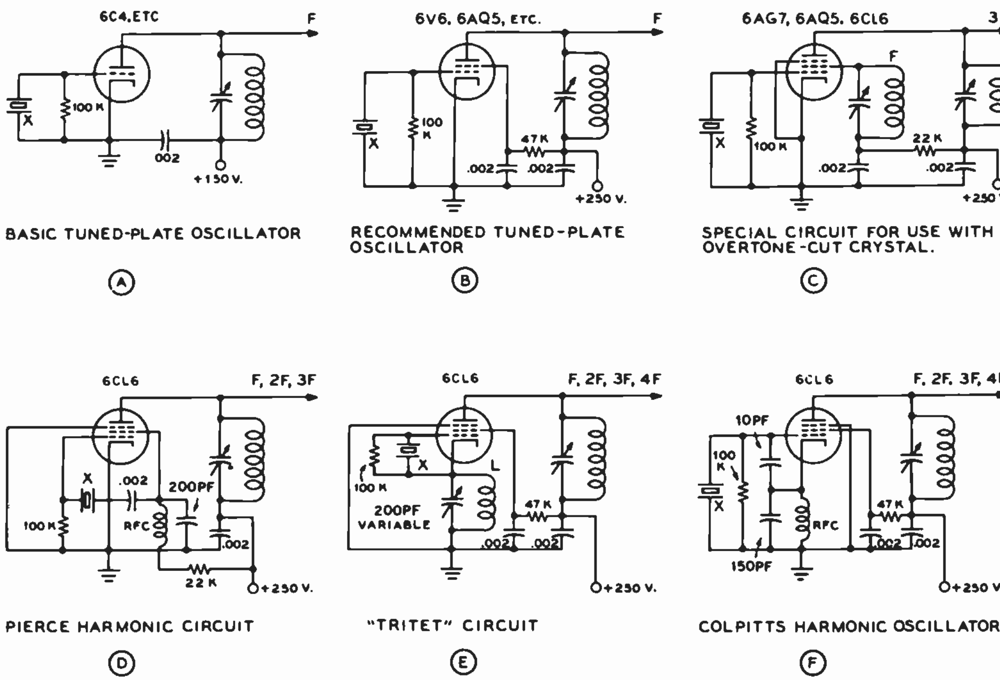


Figure 13

COMMONLY USED CRYSTAL OSCILLATOR CIRCUITS

Shown at A is the basic tuned-plate crystal oscillator with a triode oscillator tube. The plate tank must be tuned on the low-capacitance side of resonance to sustain oscillation. B shows the tuned-plate oscillator as it is normally used, with an a-f power pentode to permit high output with relatively low crystal current. Schematics C, D, E, and F illustrate crystal oscillator circuits which can deliver moderate output energy on harmonics of the oscillating frequency of the crystal. C shows a special circuit which will permit use of an overtone-cut crystal to obtain output energy well into the vhf range. D is valuable when extremely low crystal current is a requirement, but delivers relatively low output. E is commonly used, but is subject to crystal damage if the cathode circuit is mistuned. F is recommended as the most generally satisfactory from the standpoints of: low crystal current regardless of misadjustment, good output on harmonic frequencies, one side of crystal is grounded, will oscillate with crystals from 1.5 to 10 MHz without adjustment, output tank may be tuned to the crystal frequency for fundamental output without stopping oscillation or changing frequency.

or it may be operated with output on the second, third, or fourth harmonic of the crystal frequency.

Crystal Oscillator Tuning The tunable circuits of all oscillators illustrated should be tuned for maximum output as indicated by maximum excitation to the following stage, except that the oscillator tank of tuned-plate oscillators (figure 13A and figure 13B) should be backed off slightly toward the low capacitance side from maximum output, as the oscillator then is in a more stable condition and sure to start immediately when power is applied. This is especially important when the oscillator is keyed, as for break-in c-w operation.

Crystal Switching It is desirable to keep stray shunt capacitances in the crystal circuit as low as possible, regardless of the oscillator circuit. If a selector switch is used, this means that both switch and crystal sockets must be placed close to the oscillator-tube socket. This is especially true of overtone-cut crystals operating on a comparatively high frequency. In fact, on the highest frequency crystals it is preferable to use a turret arrangement for switching, since the stray capacitances can be kept lower.

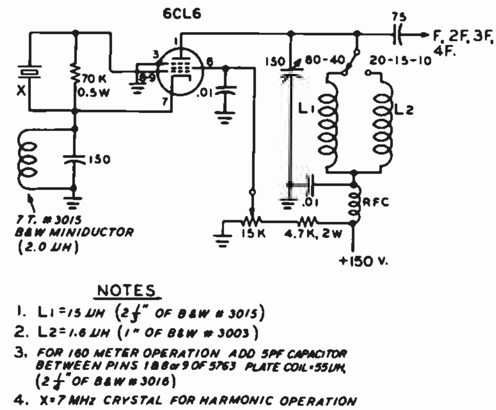


Figure 14

ALL-BAND CRYSTAL OSCILLATOR CAPABLE OF DRIVING BEAM-TETRODE TUBE. 6CL6 OR 5763 MAY BE USED

A Versatile 6CL6 Crystal Oscillator The 6CL6 tube may be used in a modified Tri-tet crystal oscillator, capable of delivering sufficient power on all bands from 160 meters through 10 meters to fully drive a pentode tube such as the 6146. Such an oscillator is extremely useful for portable or mobile work, since it combines all essential exciter functions in one tube. The circuit of this oscillator is shown in figure 14. For 160-, 80-, and 40-meter operation the 6CL6 functions as a tuned-plate oscillator. Fundamental-frequency crystals are used on these three bands. For 20-, 15-, and 10-meter operation the 6CL6 functions as a Tri-tet oscillator with a fixed-tuned cathode circuit. The impedance of this cathode circuit does not affect operation of the 6CL6 on the lower frequency bands so it is left in the circuit at all times. A 7-MHz crystal is used for fundamental output on 40 meters and for harmonic output on 20, 15, and 10 meters. Crystal current is extremely low regardless of the output frequency of the oscillator. The plate circuit of the 6CL6 is capable of tuning a frequency range of 2:1, requiring only two output coils: one for 80- 40-meter operation, and one for 20-, 15-, and 10-meter operation. In some cases it may be necessary to add 5 picofarads of external feedback capacity between the plate and control grid of the 6CL6 tube to sustain oscillation with sluggish 160-meter crystals.

Triode Overtone Oscillators The recent development of reliable overtone crystals capable of operation on the third, fifth, and seventh (or higher) overtones has made possible vhf output from a low-frequency crystal by the use of a double triode regenerative oscillator circuit. Some of the twin triode tubes such as the 12AU7, 12AV7, and 6J6 are especially satisfactory when used in this type of circuit. Crystals that are ground for overtone service may be made to oscillate on odd-overtone frequencies other than the one marked on the crystal holder. A 24-MHz overtone crystal, for example, is a specially ground 8-MHz crystal operating on its third overtone. In the proper circuit it may be made to oscillate on 40 MHz (fifth overtone), 56 MHz (seventh overtone), or 72 MHz (ninth overtone). Even the ordinary 8-MHz crystals not designed for overtone operation may be made to oscillate readily on 24 MHz (third overtone) in these circuits.

A variety of overtone oscillator circuits is shown in figure 15. The circuits of illustrations A and B employ crystal regeneration, with oscillator output on either the third or fifth overtone of the crystal. The regenerative loop consists of a capacitor bridge (C_1, C_2) with the ratio of C_2/C_1 determining the degree of feedback. The second triode section may be tuned to a harmonic of the overtone frequency.

The circuits of illustrations C and D illustrate the use of inductive feedback, the degree determined by the ratio of feedback turns to tank circuit turns in the oscillator tank circuit.

A variation of the Butler circuit is shown in E. The cathode coil (L_1) is chosen so as to resonate the crystal holder capacitance at or near the overtone frequency of the crystal. The cathode chokes may be replaced with resistors, in some instances.

The use of a 144-MHz overtone crystal is illustrated in figure 15F. A 6AB4 or one-half of a 12AT7 tube may be used, with output directly in the 2-meter amateur band. A slight amount of regeneration is provided by the one turn link, (L_2) which is loosely coupled to the 144-MHz tuned tank circuit (L_1) in the plate circuit of the oscillator

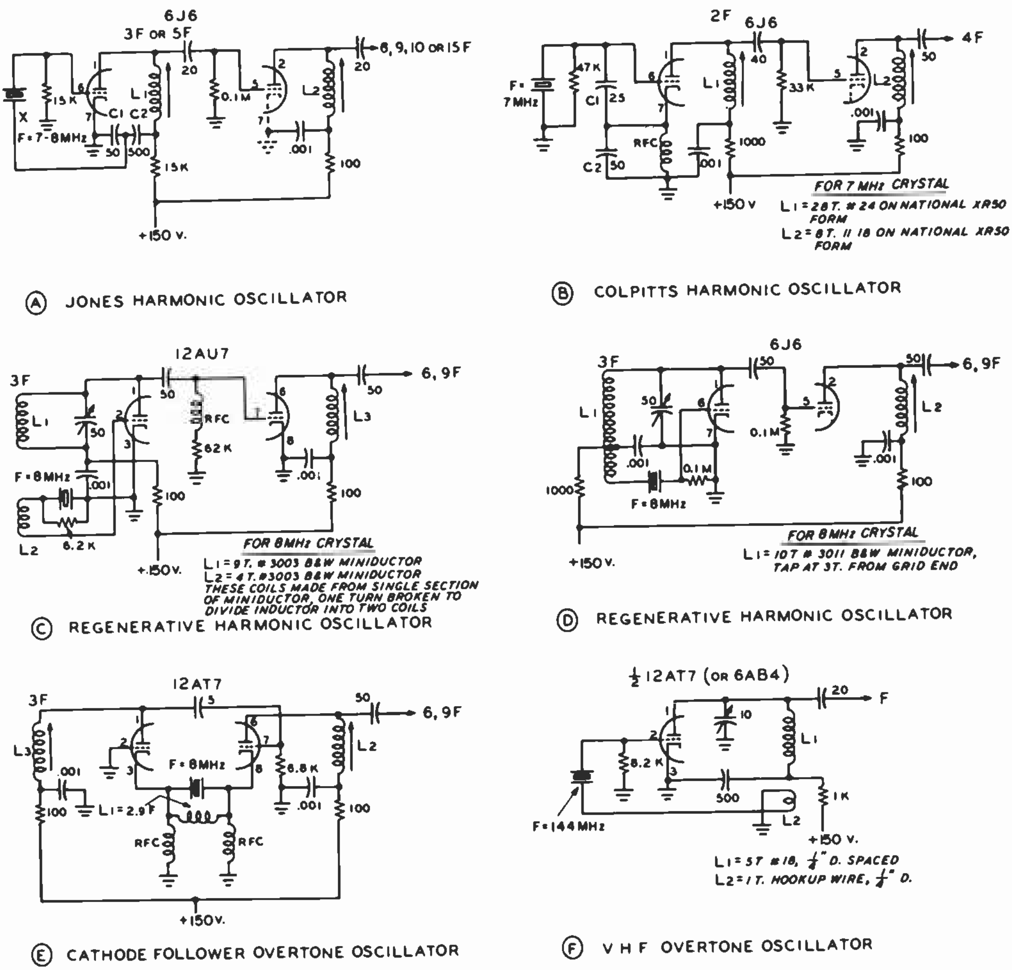


Figure 15

VARIOUS TYPES OF OVERTONE OSCILLATORS USING MINIATURE DOUBLE-TRIODE VACUUM TUBES

tube. If a 12AT7 tube and a 110-MHz crystal are employed, direct output in the 220-MHz amateur band may be obtained from the second half of the 12AT7.

Inductive Loading of Crystals A relatively wide frequency range of operation of a crystal oscillator may be achieved by operating the crystal below its resonant frequency and loading it with an inductance. Frequency stability is reduced by a factor of about 10, but bandwidth operation up to 2 percent of the crystal frequency may be achieved (figure 16). A series combination of a fixed inductor and a variable capacitor will permit oscillation from slightly above to about 2 percent below the parallel-resonant frequency of the crystal.

11-4 Frequency Synthesis

The combination of a master signal with a secondary signal in a suitable mixer provides the choice of a number of controlled frequencies (figure 17). If a stable variable-

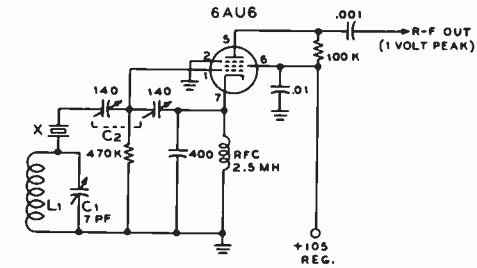


Figure 16

VARIABLE-FREQUENCY CRYSTAL OSCILLATOR

Inductive loading of crystal permits frequency change as great as two percent of nominal crystal frequency. Range covers from slightly above to nearly two percent below resonant frequency. Coil L_1 is 30 μH. Maximum range is adjusted by capacitor C_1. Circuit is tuned by capacitor C_2.

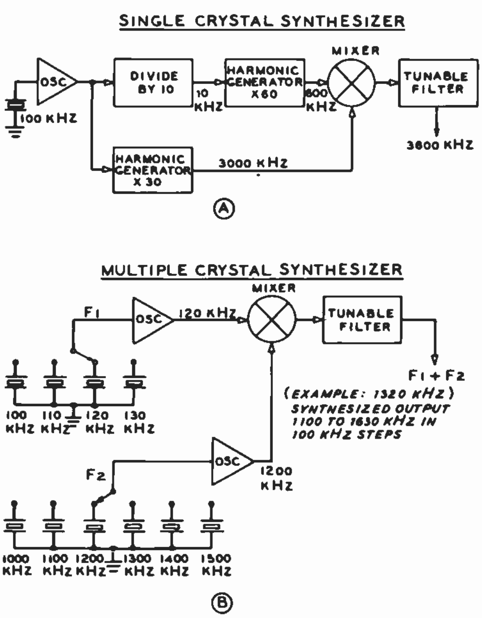


Figure 17

REPRESENTATIVE CRYSTAL SYNTHESIZER CIRCUITS

A—Single crystal frequency of 100 kHz is divided down to 10 kHz, then multiplied by 60 to provide spectrum of signals at 600 kHz. Harmonic generator also provides 3000 kHz signal which is combined in mixer with 600 kHz spectrum. Desired signal is filtered out by tunable filter. B—Multiple crystal synthesizer provides spectrum output from separate crystals, mixed, and is passed through a tunable filter.

frequency oscillator is substituted for one of the crystal oscillators in a digital frequency synthesis technique a virtually unlimited number of discrete frequencies directly related to the frequency of the master oscillator are available. A block diagram of such a device is shown in figure 18. The basic element of frequency synthesis is the *phase-lock loop* circuit in which the output of a *voltage-controlled oscillator* (VCO) is constantly compared with the frequency of the master crystal oscillator. Any unwanted change or drift in frequency of the variable oscillator with respect to the master oscillator is detected by the *phase comparator.*

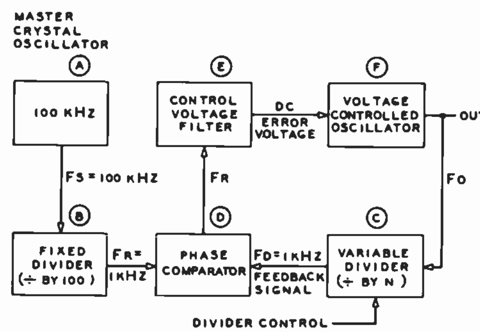


Figure 18

PHASE-LOCK LOOP CIRCUIT

Output of a voltage-controlled oscillator (F) is compared with frequency of master oscillator (A). Any drift of VCO is detected by phase comparator (D) and error voltage (E) returns VCO to correct frequency. Long-term stability of phase-lock loop is that of the master oscillator.

When a phase difference exists, a phase detector generates a control (error) voltage which returns the VCO to the correct frequency. If, for example, the phase difference changes 360° a second, the difference in frequency between the oscillators is one Hz. The long term stability of the phase-lock loop output is thus exactly that of the master oscillator.

A typical voltage-controlled oscillator circuit is shown in figure 19. A common-base Hartley circuit is used. Oscillator frequency is determined by tank circuit (L-C), across a portion of which is placed a variable capacitance (*varactor*) diode (D_1). This reverse-biased diode acts as a capacitance whose

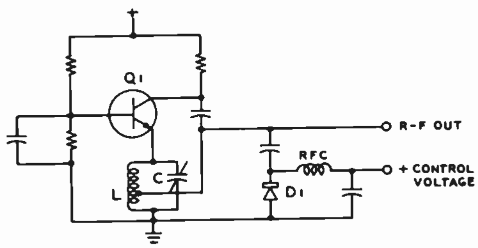


Figure 19

VOLTAGE-CONTROLLED OSCILLATOR

Varactor diode (D_1) acts as capacitance whose value varies in proportion to d-c control voltage across it. Diode is placed across portion of tank circuit of oscillator.

value varies in proportion to the d-c voltage across it. By properly controlling this voltage, the resonant frequency of the oscillator can be varied. Diode control voltage is brought in through an r-f filter for circuit isolation.

A portion of the output signal of the master crystal oscillator (A) is applied to a frequency divider (B) that divides the source signal by 100. Two "divide-by-ten" flip-flop multivibrators (see Chapter 8, Section 3) are commonly used, which provide a square-wave output reference signal (F_R) at 1/100 the frequency of the master oscillator, or 1 kHz.

In a like manner, the output signal of the voltage-controlled oscillator (F_O) is divided down by a variable divider(C). If, for example the output frequency of the VCO is 3500 kHz, and divider C is set to divide down from 3500 kHz, the output frequency (F_D) of the divider is 1 kHz. This signal is compared with the 1-kHz signal from divider B in *phase comparator* D. Any phase difference is detected, and a resultant pulse train is applied through an integrating circuit (Chapter 3, Section 3) and a low-pass filter (E) to the voltage-controlled oscillator. When signals F_D and F_R are equal and in phase, the control loop is "locked."

The phase comparator (D) may consist of a flip-flop multivibrator providing a square-output signal, the pulse width of which is proportional to the phase difference between the two input signals. The output signal, after being processed and filtered provides a d-c control voltage that is highest when the phase difference is greatest. If this voltage is correctly applied to the VCO, it will keep it on the desired frequency.

The output frequency of the synthesizer can be changed by varying the divide ratio of the variable divider (C). This is commonly done by decade switches on the control panel of the synthesizer. When the frequency is changed within the *capture range* of the phase-lock loop, the control voltage will change to bring the frequency of the VCO to the new value demanded by the setting of the variable divider. If the new frequency is outside the capture range of the circuit, the VCO frequency may be manually set by another panel control to within the newly established capture range of the device, or a control signal could be energized that would sweep the VCO through its entire operating range. As the VCO frequency enters the new capture range of the phase-lock loop, the loop will take over frequency control.

11-5 Spurious Frequencies

Spurious frequencies (*spurs*) are generated during every frequency conversion in a receiver or transmitter. These unwanted frequencies mix with the harmonics generated by the mixing oscillators to produce undesired signals that either interfere with reception of the wanted signal or can be radiated along with the desired signal from the transmitter. If the spurs are known, this information can help to determine the required r-f and i-f selectivity characteristics, the number of conversions, the allowable harmonic content of the oscillators, and the optimum intermediate frequencies.

The severity of interference from a given spur depends upon its proximity to the desired signal frequency, rather than the absolute frequency difference. For example, a simple tuned circuit has sufficient selectivity to reject a spur 4 MHz away from a 1-MHz frequency, while much more complicated means are needed to reject a spur that is 4 MHz away from a frequency of 100 MHz. Spur interference is dependent on the ratio

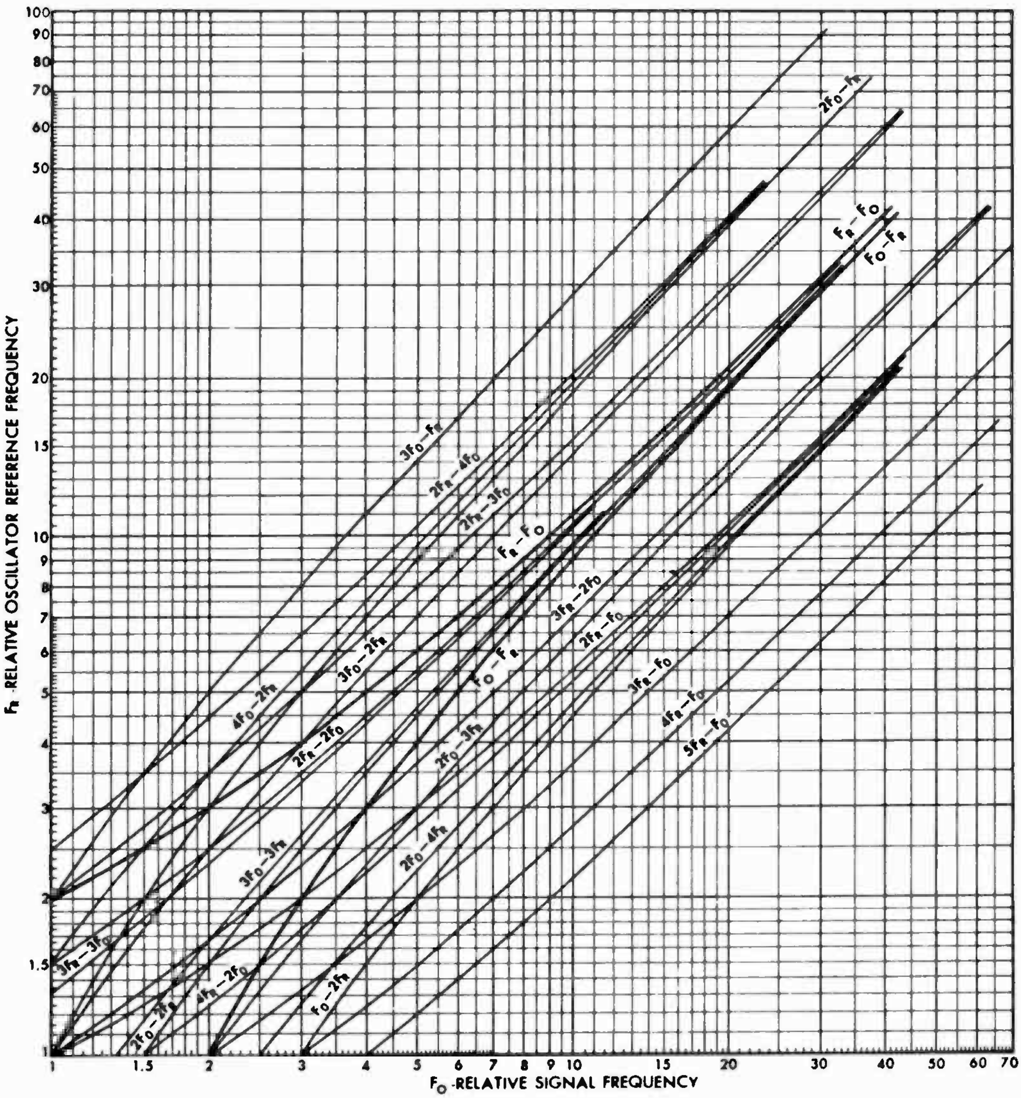


Figure 20

SPUR CHART

Curves cover all spurious mixer products that fall within an octave of the signal frequency.

of the spur frequency to the tuned frequency, and the lower the ratio, the more serious the problem.

Another indication of the importance of a particular spur is contained in the order of response. This order may be defined as the sum of the signal and oscillator harmonics that produce the spur. For example, a spur produced by the second harmonic of the signal and the third harmonic of the oscillator is known as a *fifth-order* spur. Lower-order spurs are more serious because higher harmonics of both input signals are easier to reject by circuit design techniques.

A Spur Chart Graphical relationships between the frequencies of the various spurious signals and the desired signal are presented by the spur chart of figure 20. A given ratio of spur to desired frequency is represented by a constant horizontal distance on the chart.

The local-oscillator frequency is represented by F_R and the relative signal frequency by F_O. The curves cover all spurious products up to the sixth order for spur-signal frequencies that fall within an octave of the signal frequency. Each line on the chart represents a normalized frequency difference of 1 for mF_O+nF_R where m and n may be positive or negative integers. The heavy, central lines labeled F_R-F_O and F_O-F_R are plots of the desired frequency conversion when the oscillator frequency is either higher or lower than the signal frequency. Whichever line represents the desired signal, the other line represents the image spur.

To determine the spurious environment for a given conversion, first normalize the desired signal and oscillator frequencies by dividing both frequencies by the mixing output frequency. Then locate the desired point on one of the heavy lines representing either F_O-F_R *or* F_R-F_O. Since the oscillator frequency does not change for spurs, simply trace horizontally in either direction to determine the relative frequency of the spurs.

Example: Desired signal frequency is 10 MHz.
Mixing output frequency is 2 MHz.
Oscillator frequency is 12 MHz.
Then, relative signal frequency F_O is 10 MHz/2 MHz = 5.

And, relative oscillator frequency F_R is 12 MHz/2 MHz = 6.
Since oscillator frequency is higher, we use the F_R-F_O curve.

Locate the $F_O=5$, $F_R=6$, point on the curve. Tracing horizontally to the left, the spur lines intercepted on the F_O scale are:
$3F_O-2F_R$ at F_O of 4.35, or signal frequency that causes the spur is 8.70 MHz (2 MHz × 4.35).
$2F_R-3F_O$ at F_O of 3.70, equivalent to a signal frequency of 7.40 MHz.
$2F_O-F_R$ at F_O of 3.50, equivalent to a signal frequency of 7.00 MHz.
Tracing right, nearest spur lines are:
$2F_R-2F_O$ at F_O of 5.50, equivalent to a signal frequency of 11.0 MHz.
$3F_R-3F_O$ at F_O of 5.70, equivalent to a signal frequency of 11.4 MHz.
$3F_O-3F_R$ at F_O of 6.36, equivalent to a signal frequency of 12.7 MHz.
And the image frequency, F_O-F_R, occurs at 7.00 or 14.0 MHz.

11-6 Radio-Frequency Amplifiers

The output of the oscillator stage in a transmitter must be held down to a comparatively low level to maintain stability and to retain a factor of safety from fracture of the crystal when one is used. The low-level signal is brought up to the desired power level by means of radio-frequency amplifiers. The three classes of r-f amplifiers that find widest application in modern radio transmitters are the class AB_1, class-B, and class-C types.

The angle of plate-current conduction determines the class of operation. Class B is a 180-degree conduction angle and class C is less than 180 degrees. Class AB is the region between 180 degrees and 360 degrees of conduction. The subscript "1" indicates that no grid current flows, and the subscript "2" means that grid current is present. The class of operation has nothing to do with whether the amplifier is grid driven or cathode driven (grounded grid). A cathode-driven amplifier, for example, can be operated in any desired class, within limitations imposed by the tube.

The Classes of Amplifiers The class-AB amplifier can be operated with very low intermodulation distortion in linear amplifier service. Typical plate efficiency is about 60 percent, and stage gain is about 20 to 25 decibels. The class-B amplifier will generate more intermodulation distortion than the class-AB circuit but the distortion level is acceptable in many applications. Typical plate efficiency is about 66 percent and power gains of 15 to 20 decibels are readily achieved. The class-C amplifier is used where large amounts of r-f power are to be amplified with high efficiency. Class-C amplifiers operate with considerably more than cutoff bias, much like a limiter; therefore, this configuration cannot amplify a modulated signal without serious distortion. Class-C amplifiers are used for high-level amplitude modulation wherein the plate voltage (or plate and screen voltages for tetrodes) is modulated at an audio rate. The output power of a class-C amplifier, adjusted for plate modulation, varies with the square of the plate voltage. That is the same condition that would take place if a resistor equal to the voltage on the amplifier, divided by the plate current, were substituted for the amplifier. Therefore, the stage presents a resistive load to the plate modulator. Typical plate efficiency is 70 percent and stage gain is 8 to 10 decibels.

Grid-Modulated Class C If the grid current to a class-C amplifier is reduced to a low value, and the plate loading is increased to the point where the plate dissipation approaches the rated value, the amplifier may be grid modulated for radiotelephony. If the plate voltage is high, efficiency up to 40 percent is possible.

Grid Excitation Adequate grid excitation must be available for class-B or class-C service. The excitation for a plate-modulated class-C stage must be sufficient to produce a normal value of d-c grid current with rated bias voltage. The bias voltage preferably should be obtained from a combination of grid-resistor and fixed grid-bias supply.

Cutoff bias can be calculated by dividing the amplification factor of the tube into the d-c plate voltage. This is the value normally used for class-B amplifiers (fixed bias, no grid resistor). Class-C amplifiers use from 1.5 to 5 times this value, depending on the available grid drive, or excitation, and the desired plate efficiency. Less grid excitation is needed for c-w operation, and the values of fixed bias (if greater than cutoff) may be reduced, or the value of the grid-bias resistor can be lowered until normal rated d-c grid current flows.

The values of grid excitation listed for each type of tube may be reduced by as much as 50 percent if only moderate power output and plate efficiency are desired. When consulting the tube tables, it is well to remember that the power lost in the tuned circuits must be taken into consideration when calculating the available grid drive. At very-high frequencies, the r-f circuit losses may even exceed the power required for actual grid excitation.

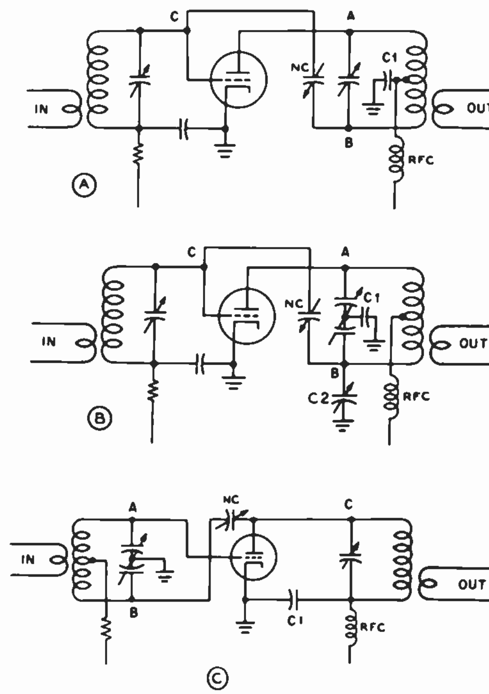


Figure 21

COMMON NEUTRALIZING CIRCUITS FOR SINGLE-ENDED AMPLIFIERS

Excessive grid current damages tubes by overheating the grid structure; beyond a certain point of grid drive, no increase in power output can be obtained for a given plate voltage.

11-7 Neutralization of R-F Amplifiers

The plate-to-grid feedback capacitance of triodes makes it necessary that they be neutralized for operation as r-f amplifiers at frequencies above about 500 kHz. Those screen-grid tubes, pentodes, and beam tetrodes which have a plate-to-grid capacitance of 0.1 pf or less may be operated as an amplifier without neutralization in a well-designed amplifier up to 30 MHz provided the stage gain is less than the over-all feedback gain from output to input circuit.

Neutralizing Circuits The object of *neutralization* is to cancel or neutralize the capacitive feedback of energy from plate to grid. There are two general methods by which this energy feedback may be eliminated: the first, and the most common method, is through the use of a capacitance bridge, and the second method is through the use of a parallel reactance of equal and opposite polarity to the grid-to-plate capacitance, to nullify the effect of this capacitance.

Examples of the first method are shown in figure 21. Figure 21A shows a capacitance-neutralized stage employing a balanced tank circuit. Phase reversal in the tank circuit is obtained by grounding the center of the tank coil to radio-frequency energy by capacitor C_1. Points A and B are 180 degrees out of phase with each other, and the correct amount of out-of-phase energy is coupled through the neutralizing capacitor (NC) to the grid circuit of the tube. The equivalent bridge circuit of this is shown in figure 22A. It is seen that the bridge is not in balance, since the plate-filament capacitance of the tube forms one leg of the bridge, and there is no corresponding capacitance from the neutralizing capacitor (point B) to ground to obtain a complete balance. In addition, it is mechanically difficult to obtain a perfect electrical balance in the tank coil, and the potential between point A and ground and point B and ground, in most cases, is unequal. This circuit, therefore, holds neutralization over a very small operating range and unless tubes of low interelectrode capacitance are used the inherent unbalance of the circuit will permit only approximate neutralization.

Split-Stator Plate Neutralization Figure 21B shows the neutralization circuit which is widely used in single-ended r-f stages. The use of a split-stator plate capacitor makes the electrical balance of the circuit substantially independent of the mutual coupling within the coil and also makes the balance independent of the place where the coil is tapped. With conventional tubes this circuit will allow one neutralization adjustment to be made on, for example, 28 MHz, and this adjustment usually will hold sufficiently close for operation on all lower-frequency bands.

Capacitor C_2 is used to balance out the plate-filament capacity of the tube to allow a perfect neutralizing balance at all frequencies. The equivalent bridge circuit is shown in figure 22B. If the plate-filament capacitance of the tube is extremely low, capacitor C_2 may be omitted, or may merely consist of the residual capacity of NC to ground.

Grid Neutralization A split grid-tank circuit may also be used for neutralization of a triode tube as shown in figure 21C. Out of phase voltage is developed across a balanced grid circuit, and coupled through NC to the single-ended plate circuit of the tube. The equivalent

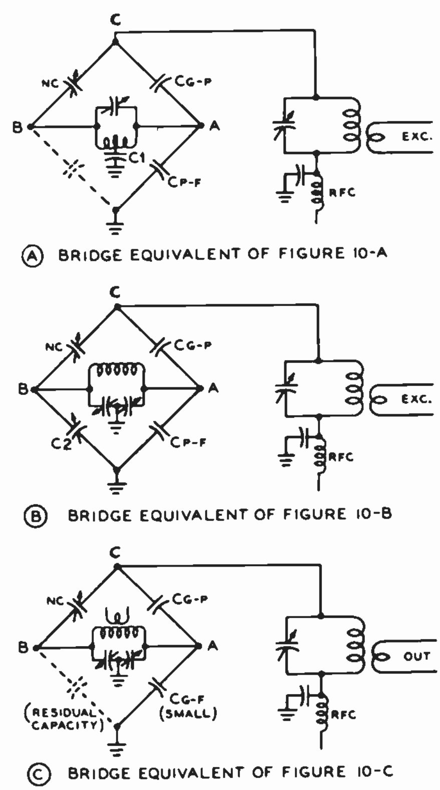


Figure 22

EQUIVALENT NEUTRALIZING CIRCUITS

bridge circuit is shown in figure 22C. This circuit is in balance until the stage is in operation when the loading effect of the tube upon one-half of the grid circuit throws the bridge circuit out of balance. The amount of unbalance depends on the grid-plate capacitance of the tube, and the amount of mutual inductance between the two halves of the grid coil. If an r-f voltmeter is placed between point A and ground, and a second voltmeter placed between point B and ground, the loading effect of the tube will be noticeable. When the tube is supplied excitation with no plate voltage, NC may be adjusted until the circuit is in balance. When plate voltage is applied to the stage, the voltage from point A to ground will decrease, and the voltage from point B to ground will increase, both in direct proportion to the amount of circuit unbalance. The use of this circuit is not recommended above 7 MHz, and it should be used below that frequency only with low internal capacitance tubes.

Push-Pull Neutralization Two tubes of the same type can be connected for *push-pull* operation so as to obtain twice as much output as that of a single tube. A push-pull amplifier, such as that shown in figure 23 also has an advantage in that the circuit can more easily be balanced than a single-tube r-f amplifier. The various interelectrode capacitances and the neutralizing capacitors are connected in such a manner that the reactances on one side of the tuned circuits are exactly equal to those on the opposite side. For this reason, push-pull r-f amplifiers can be more easily neutralized in vhf transmitters; also, they usually remain in perfect neutralization when tuning the amplifier to different bands.

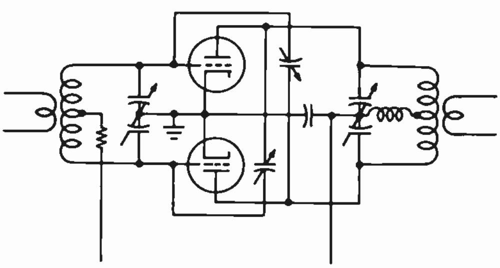


Figure 23

STANDARD CROSS-NEUTRALIZED PUSH-PULL TRIODE AMPLIFIER

Shunt or Coil Neutralization The feedback of energy from grid to plate in an unneutralized r-f amplifier is a result of the grid-to-plate capacitance of the amplifier tube. A neutralization circuit is merely an electrcial arrangement for nullifying the effect of this capacitance. All the previous neutralization circuits have made use of a bridge circuit for balancing out the grid-to-plate energy feedback by feeding back an equal amount of energy of opposite phase.

Another method of eliminating the feedback effect of this capacitance, and hence of neutralizing the amplifier stage, is shown in figure 24. The grid-to-plate capacitance in the triode amplifier tube acts as a capacitive reactance, coupling energy back from the plate to the grid circuit. If this capacitance is paralleled with an inductance having the same value of reactance of opposite sign, the reactance of one will cancel the reactance of the other and a high-impedance tuned circuit from grid to plate will result.

This neutralization circuit can be used on ultra high frequencies where other neutralization circuits are unsatisfactory. This is true because the lead length in the neutrali-

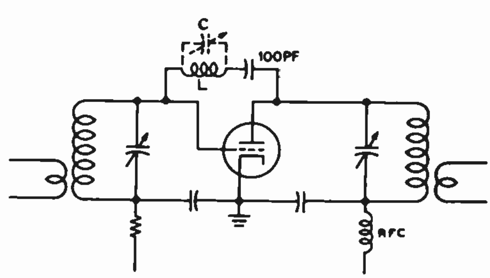


Figure 24

COIL-NEUTRALIZED AMPLIFIER

This neutralization circuit is very effective with triode tubes on any frequency, but is particularly effective in the vhf range. Coil L is adjusted so that it resonates at the operating frequency with the grid-to-plate capacitance of the tube. Capacitor C may be a very small unit of the low-capacitance neutralizing type and is used to trim the circuit to resonance at the operating frequency. If some means of varying the inductance of the coil a small amount is available, the trimmer capacitor is not needed.

zation circuit is practically negligible. The circuit can also be used with push-pull r-f amplifiers. In this case, each tube will have its own neutralizing inductor connected from grid to plate.

The main advantage of this arrangement is that it allows the use of single-ended tank circuits with a single-ended amplifier.

The chief disadvantage of the shunt neutralized arrangement is that the stage must be neutralized each time the stage is returned to a new frequency sufficiently removed that the grid and plate tank circuits must be retuned to resonance. However, by the use of plug-in coils it is possible to change to a different band of operation by changing the neutralizing coil at the same time that the grid and plate coils are changed.

The 100-pF capacitor in series with the neutralizing coil is merely a blocking capacitor to isolate the plate voltage from the grid circuit.

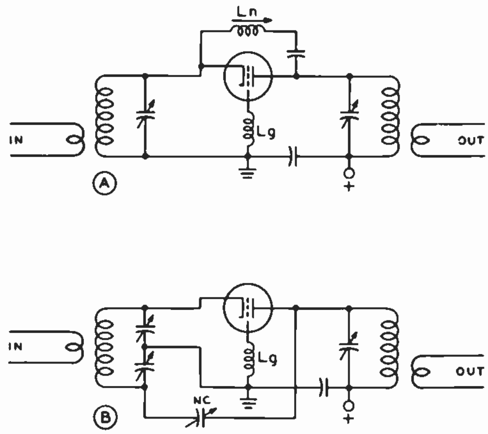


Figure 25

NEUTRALIZATION OF CATHODE DRIVEN AMPLIFIER

A—Cathode-to-plate feedback path may be neutralized by making it part of a parallel-tuned circuit by addition of neutralizing coil L_n. Series capacitor removes plate voltage from neutralizing coil. Adjustments tend to be frequency sensitive.

B—Cathode-to-plate feedback path is neutralized by introducing out-of-phase voltage from drive circuit into plate circuit by means of capacitor NC. Inductor L_g represents grid-lead inductance of vacuum tube, whose effects are not cancelled by either neutralizing circuit.

Neutralization of Cathode-Driven Amplifiers Stable operation of the cathode-driven (grounded-grid) amplifier often requires neutralization, particularly above 25 MHz or so. Complete circuit stability requires neutralization of two feedback paths, as shown in figure 25.

The first path involves the cathode-to-plate capacitance and proper neutralization may be accomplished by a shunt inductance or by a balanced-bridge technique. The bridge technique is less critical of adjustment than the shunt-inductance circuit, and a reasonable bridge balance over a wide frequency range may be achieved with a single setting of the neutralizing capacitance.

The second feedback path includes the grid-to-plate capacitance, the cathode-to-grid capacitance and the series inductance of the grid-to-ground path (figure 26). If this path is not neutralized, a voltage appears on the grid of the tube which either increases or decreases the driving voltage, depending on the values of grid inductance and internal capacitances of the tube. A certain frequency exists at which these two feedback paths nullify each other and this self-neutralizing frequency may be moved about by adding either positive or negative reactance in the grid circuit, as shown in the illustration. If the operating frequency is above the self-neutralizing frequency a series capacitance is used to reduce the grid inductance. If the operating frequency is below the self-neutralizing frequency, the series grid inductance should be increased. For most tubes of the amateur power class, the self-neutralizing frequency lies between 50 and 150 MHz.

11-8 Neutralizing Procedures

Voltage feedback from output to input through the distributed constants of the vacuum tube has a deleterious effect on amplifier performance. The magnitude, phase and rate of change with respect to frequency of this feedback voltage determine the stability of the amplifier. Control of feedback is termed *neutralization*. The purpose of neutralization of an amplifier is to make the input and output circuits independent of each other with respect to voltage feedback. *Proper neutralization may be defined as the state in which, when output and input tank circuits are resonant, maximum drive voltage, minimum plate current, and maximum power output occur simultaneously.*

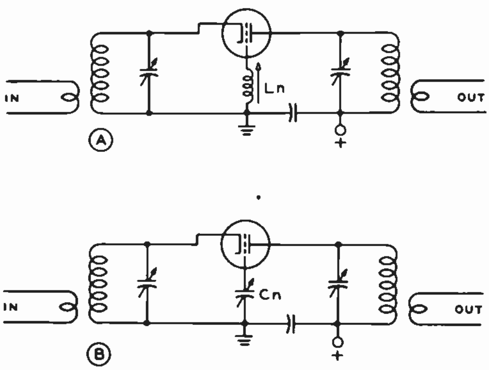


Figure 26

NEUTRALIZATION OF GRID LEAD INDUCTANCE

A—When amplifier is operated below self-neutralizing frequency of tube (h-f range, for example) additional inductance (L_n) in grid-return lead may be required to achieve complete neutralization of amplifier.

B—When amplifier is operated above self-neutralizing frequency of tube (the vhf range, for example) grid inductance is compensated by addition of series capacitance which is adjusted to minimize interaction between input and output circuits of amplifier.

The state of correct neutralization, therefore, may be judged by observing these operating parameters or by observing the degree of feedback present in the amplifier. The amplifier may be neutralized in the active or passive state provided proper instrumentation is used.

Passive Neutralization An amplifier may be neutralized in the passive state with the aid of a signal generator, an r-f voltmeter, and a grid-dip oscillator. The input and output circuits of the amplifier are resonated to the operating frequency and a small signal from the generator is applied to the input circuit of the amplifier. An r-f voltmeter (or well-shielded receiver) is connected to the output circuit of the amplifier. Neutralizing adjustments are now made to reduce to a minimum the feed-through voltage reaching the receiver from the signal generator. Adjustments may be made with no filament or plate voltage applied to the amplifier. Once a null adjustment has been achieved, the amplifier may be activated and the neutralization adjustment touched up at full power level.

Passive neutralization is a highy recommended technique since no voltages are applied to the equipment, and adjustments and circuit modifications may be made without danger to the operator of accidental shock.

Active Neutralization An amplifier may be neutralized by the application of excitation with plate and screen voltage removed. A neutralizing indicator is coupled to the plate circuit and the neutralizing adjustment manipulated for an indication of minimum r-f voltage in the plate tank circuit. This adjustment is normally accomplished with input and output circuits resonated and with a suitable dummy load connected to the amplifier.

Plate (and screen) voltage should be completely removed by actually opening the d-c plate-current return. If a direct-current return circuit exists, a small amount of plate current will flow when grid excitation is applied, even though no high voltage exists on the amplifier stage. Once neutralization has been established, a more accurate check may be made by applying high voltage, and tuning and loading the amplifier while noting if maximum grid and screen current occur at the same point of tuning on the plate circuit tuning capacitor as minimum plate current. As the plate tuning capacitor is detuned slightly from resonance on either side, plate current should rise, and the grid (or screen) current on the stage should decrease smoothly without any sudden jumps on either side of the resonance point. This technique will be found to be a very precise indication of accurate neutralization so long as the amplifier stage is coupled to a load which presents a resistive impedance at the operating frequency.

Neutralization of Screen-Grid R-F Amplifiers Radio-frequency amplifiers using screen-grid tubes can be operated without any additional provision for neutralization at frequencies up to about 15 MHz, provided adequate shielding has been provided between the input and output circuits. Special vhf screen-grid and beam tetrode tubes such as the 2E26, 6146, and 5516 can frequently be operated at frequencies as high as 50 MHz without any additional provision for neutralization.

None of these tubes, however, has perfect shielding between the grid and the plate, a condition brought about by the inherent inductance of the screen leads within the tube itself. In addition, unless "watertight" shielding is used between the grid and plate circuits of the tube a certain amount of external leakage between the two circuits is present. These difficulties may not be serious enough to require neutralization of the stage to prevent oscillation, but in many instances they show up in terms of key-clicks when the stage in question is keyed, or as parasitics when the stage is modulated. Unless the designer of the equipment can carefully check the tetrode stage for miscellaneous feedback between the grid and plate circuits, and make the necessary circuit revisions to reduce this feedback to an absolute minimum, it is wise to neutralize the tetrode just as if it were a triode tube.

In most push-pull tetrode amplifiers the

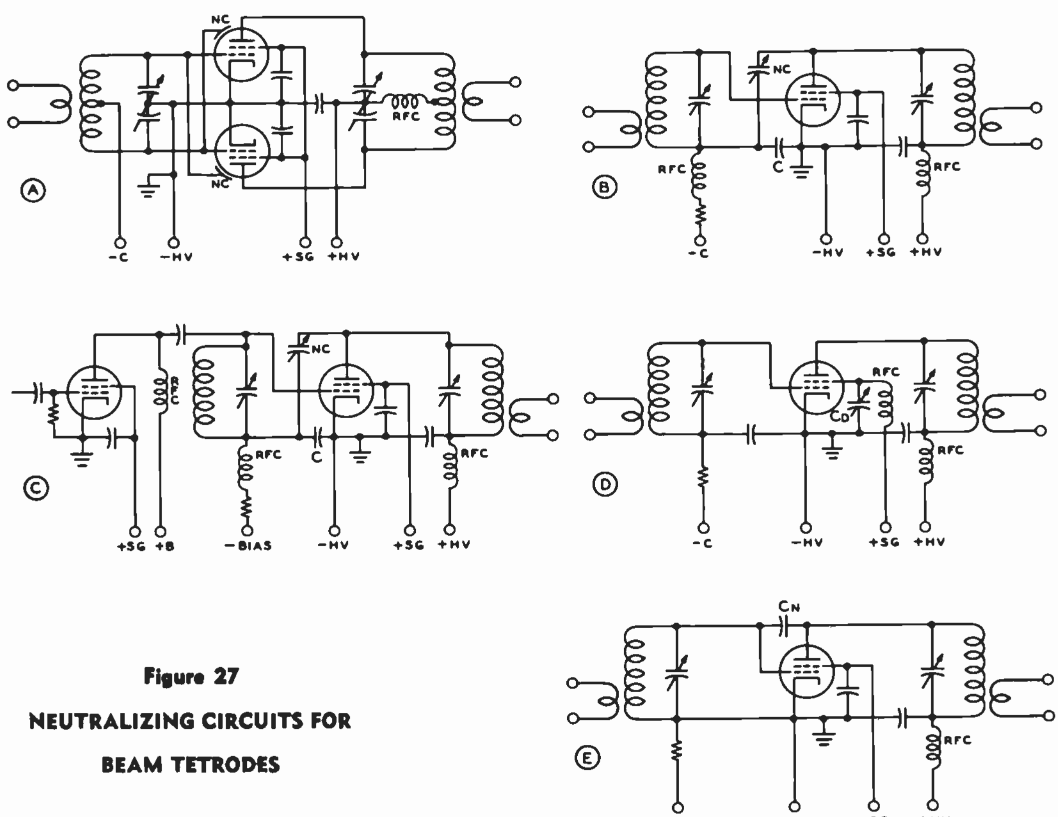


Figure 27

NEUTRALIZING CIRCUITS FOR BEAM TETRODES

A conventional cross-neutralized circuit for use with push-pull beam tetrodes is shown at A. The neutralizing capacitors (NC) usually consist of small plates or rods mounted alongside the plate elements of the tubes. B and C show grid-neutralized circuits for use with a single-ended tetrode having either link coupling or capacitive coupling into the grid tank. D shows a method of tuning the screen-lead inductance to accomplish neutralization in a single-ended vhf tetrode amplifier, while E shows a method of neutralization by increasing the grid-to-plate capacitance on a tetrode when the operating frequency is higher than that frequency where the tetrode is "self-neutralized" as a result of series resonance in the screen lead. Methods D and E normally are not practicable at frequencies below about 50 MHz with the usual types of beam tetrode tubes.

simplest method of accomplishing neutralization is to use the cross-neutralized capacitance bridge arrangement as normally employed with triode tubes. The neutralizing capacitances, however, must be very much smaller than used with triode tubes, values of the order of 0.2 pf normally being required with beam tetrode tubes. This order of capacitance is far less than can be obtained with a conventional neutralizing capacitor at minimum setting, so the neutralizing arrangement is most commonly made especially for the case at hand. Most common procedure is to bring a conductor (connected to the opposite grid) in the vicinity of the plate itself or of the plate tuning capacitor of one of the tubes. Either one or two such capacitors may be used, two being normally used on a higher-frequency amplifier in order to maintain balance within the stage.

An example of this is shown in figure 27A.

Neutralizing Single-Ended Tetrode Stages A single-ended tetrode r-f amplifier stage may be neutralized in the same manner as illustrated for a push-pull stage in figure 27A, provided a split-stator tank capacitor is in use in the plate circuit.

The circuit shown in figure 27B is not a true neutralizing circuit, in that the plate-to-grid capacitance is not balanced out. However, the circuit can afford the equivalent effect by isolating the high resonant impedance of the grid-tank circuit from the energy fed back from plate to grid. When *NC* and *C* are adjusted to bear the following ratio to the grid-to-plate capacitance and the total capacitance from grid-to-ground in the output tube,

$$\frac{NC}{C} = \frac{C_{gp}}{C_{gk}}$$

both ends of the grid tank circuit will be at the same voltage with respect to ground as a result of r-f energy fed back to the grid circuit. This means that the impedance from grid to ground will be effectively equal to the reactance of the grid-to-cathode capacitance in parallel with the stray grid-to-ground capacitance, since the high resonant impedance of the tuned circuit in the grid has been effectively isolated from the feedback path. It is important to note that the effective grid-to-ground capacitance of the tube being neutralized includes the rated grid-to-cathode or input capacitance of the tube, the capacitance of the socket, wiring capacitances and other strays, but it does *not* include the capacitances associated with the grid-tuning capacitor. Also, if the tube is being excited by capacitive coupling from a preceding stage (as in figure 27C), the effective grid-to-ground capacitance includes the output capacitance of the preceding stage and its associated socket and wiring capacitances.

Cancellation of Screen-Lead Inductance The provisions discussed in the previous paragraphs are for neutralization of the small (though still important at the higher frequencies) grid-to-plate capacitance of beam-tetrode tubes. However, in the vicinity of the upper frequency limit of each tube type the inductance of the screen lead of the tube becomes of considerable importance. With a tube operating at a frequency where the inductance of the screen lead is appreciable, the screen will allow a considerable amount of energy leak-through from plate to grid even though the socket terminal on the tube is carefully bypassed to ground. This condition takes place even though the socket pin is bypassed since the reactance of the screen lead will allow a moderate amount of r-f potential to appear on the screen itself inside the electrode assembly in the tube.

The effect of screen-lead inductance on the stability of a stage can be eliminated at any particular frequency by one of two methods. These methods are: (1) Tuning out the screen-lead inductance by series-resonating the screen-lead inductance with a capacitor to ground. This method is illustrated in figure 27D and is commonly employed in commercially built equipment for operation on a narrow frequency band in the range above about 75 MHz. The other method (2) is illustrated in figure 27E and consists in feeding back additional energy from plate to grid by means of a small capacitor connected between these two elements. Note that this capacitor is connected in such a manner as to *increase* the effective grid-to-plate capacitance of the tube.

Note that both these methods of stabilizing a beam-tetrode vhf amplifier stage by cancellation of screen-lead inductance are suitable only for operation over a relatively narrow band of frequencies in the vhf range. At lower frequencies both these expedients for reducing the effects of screen-lead inductance will tend to increase the tendency toward oscillation of the amplifier stage.

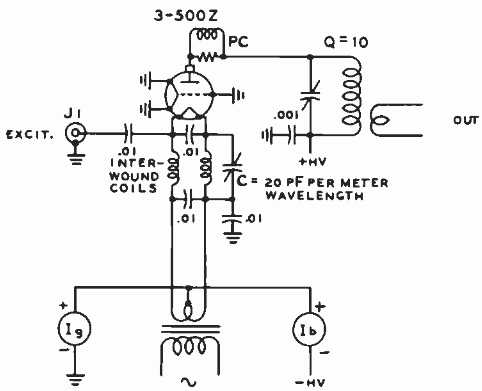


Figure 28

GROUNDED-GRID AMPLIFIER

This type of triode amplifier requires no neutralization, and is very popular as a zero-bias class-B linear stage for SSB service.

11-9 Grounded-Grid Amplifiers

Certain triodes such as the 3-500Z have a grid structure and lead arrangement which result in a very low plate-to-filament capacitance when the grid is grounded, the grid acting as an effective shield much in the manner of the screen of a tetrode tube. By connecting such a triode in the typical circuit of figure 28 taking the usual precautions against stray capacitive and inductive coupling between input and output circuits, a stable power amplifier is realized which requires no neutralization in the hf region. A high-μ triode may not require grid bias to operate in the class-B mode, however, some amount of grid bias may be added to achieve class-C operation.

The *grounded-grid* (cathode-driven) amplifier requires considerably more excitation than if the same tube were employed in a conventional grounded-cathode circuit. The additional drive power required to drive a tube in a grounded-grid circuit is not lost, however, as it shows up in the output circuit and adds to the power delivered to the load. Nevertheless it means that a larger driver stage is required for an amplifier of given output power as a portion of the drive power is delivered to the load (*feedthrough* power). Stage gains of 10 to 12 decibels are common in grounded-grid circuits.

Some tetrodes may be strapped as triodes (screen and grid grounded) and operated as class-B grounded-grid tubes. Data on this class of operation may often be obtained from the tube manufacturer.

11-10 Frequency Multipliers

Quartz crystals and variable-frequency oscillators are not ordinarily used for direct control of the output of high-frequency or vhf transmitters. *Frequency multipliers* are often used to multiply the crystal frequency to the desired value. The multiplier operates on exact multiples of the excitation frequency and when used at twice frequency is termed a *doubler*. The plate circuit of the multiplier is tuned to the desired harmonic, otherwise the circuitry remains the same as that of a simple class-C amplifier (see Chapter 7, figure 1).

The angle of plate current flow is quite small for a frequency multiplier stage to increase the harmonic content of the output wave. Hence higher grid bias is required so that grid excitation voltage will exceed the cutoff value for a shorter portion of the exciting voltage cycle. For a good order of efficiency, frequency doublers have an angle of plate current flow of 90 degrees or less, triplers 60 degrees or less, and quadruplers 45 degrees or less. Under these conditions the efficiency will be of the same order as the reciprocal of the harmonic on which the stage operates. Thus the efficiency of a doubler will be about 50 percent, the efficiency of a tripler about 33 percent, and so on.

The pulses ABC, EFG, and JKL in figure 29 illustrate 180-degree excitation pulses under class-B operation, the solid straight line indicating cutoff bias. If the bias is increased by *N* times, to the value indicated by the dotted straight line, and the excitation increased until the peak r-f voltage with respect to ground is the same as before, then the excitation frequency can be cut in half and the effective excitation pulses will have almost the same shape as before. The only difference is that every other pulse is missing; MNO simply shows where the missing pulse would go. However, if the Q of the plate tank circuit is high, it will have sufficient *flywheel effect* to carry over through the missing pulse, and the only effect will be that the plate input and r-f output at optimum loading drop to approximately half. As the input frequency is half the output frequency, an efficient frequency doubler is the result.

By the same token, a tripler or quadrupler can be analyzed, the tripler skipping two excitation pulses and the quadrupler three. In each case the excitation pulse ideally should be short enough that it does not exceed 180 degrees at the output frequency; otherwise the excitation actually is *bucking* the output over a portion of the cycle.

Push-Push Multipliers Two tubes can be connected in parallel to give twice the output of a single-tube doubler. If the grids are driven *out* of phase instead of *in* phase, the tubes then no longer work simultaneously, but rather one at a time. The effect is to fill in the missing pulses (figure 30). Not only is the output doubled, but several advantages accrue which cannot be obtained by straight parallel operation.

Chief among these is the effective neutralization of the fundamental and all *odd* harmonics, an advantage when spurious emissions must be minimized. Another advantage is that when the available excitation is low and excitation pulses exceed 90 degrees, the output and efficiency will be greater than for the same tubes connected in parallel.

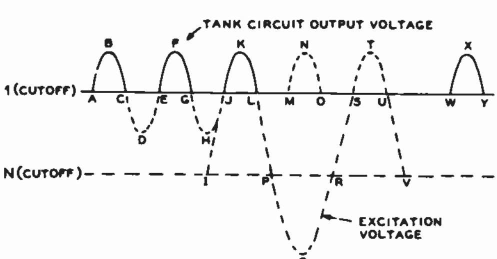


Figure 29

ILLUSTRATING THE ACTION OF A FREQUENCY DOUBLER

Push-Pull Frequency Triplers It is frequently desirable in the case of uhf and vhf transmitters that frequency multiplication stages be balanced with respect to ground. Further it is just as easy in most cases to multiply the crystal or vfo frequency by powers of three rather than multiplying by powers of two as is frequently done in lower-frequency transmitters. Hence the use of push-pull triplers has become quite prevalent in both commercial and amateur vhf and uhf transmitter designs. Such stages are balanced with respect to ground and appear in construction and on paper essentially the same as a push-pull r-f amplifier stage with the exception that the output tank circuit is tuned to three times the frequency of the grid-tank circuit. A circuit for a push-pull tripler stage is shown in figure 31.

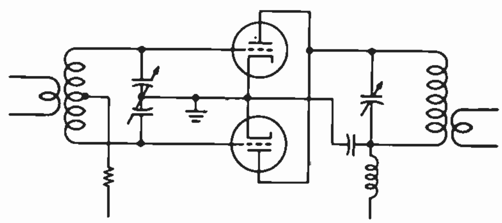


Figure 30

PUSH-PUSH FREQUENCY DOUBLER

The output of a doubler stage may be materially increased through the use of a push-push circuit such as illustrated above.

A push-pull tripler stage has the further advantage in amateur work that it can also be used as a conventional push-pull r-f amplifier merely by changing the grid and plate coils so that they tune to the same frequency.

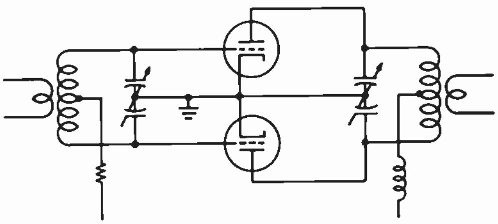


Figure 31

PUSH-PULL FREQUENCY TRIPLER

The push-pull tripler is advantageous in the vhf range since circuit balance is maintained both in the input and output circuits. If the circuit is neutralized it may be used either as a straight amplifier or as a tripler. Either triodes or tetrodes may be used; dual-unit tetrodes such as the 6360, 832A, and 829B are particularly effective in the vhf range.

11-11 Tank-Circuit Design

It is necessary that the proper value of Q be used in the plate tank circuit of any r-f amplifier. The following section has been devoted to a treatment of the subject, and charts are given to assist the reader in the determination of the proper LC ratio to be used in a radio-frequency amplifier stage.

A class-C amplifier draws plate current in the form of very distorted pulses of short duration. Such an amplifier is always operated into a tuned inductance-capacitance or tank circuit which tends to smooth out these pulses, by its storage or tank action, into a sine wave of radio-frequency output. Any waveform distortion of the carrier frequency results in harmonic interference in higher-frequency channels.

A class-A r-f amplifier would produce a sine wave of radio-frequency output if its exciting waveform were also a sine wave. However, a class-A amplifier stage converts its d-c input to r-f output by acting as a variable resistance, and therefore heats considerably. A class-B or-C amplifier driven hard with short pulses at the peak of the exciting waveform acts more as an electronic switch, and therefore can convert its d-c input to r-f output with relatively good efficiency.

Tank Circuit Q As stated before, the tank circuit of a class-C amplifier receives energy in the form of short pulses of plate current which flow in the amplifier tube. But the tank circuit must be able to store enough energy so that it can deliver a current essentially sine wave in form to the load. The ability of a tank to store energy in this manner may be designated as the effective Q of the tank circuit. The effective circuit Q may be stated in any of several ways, but essentially the Q of a tank circuit is the *ratio of the energy stored to 2π times the energy lost per cycle.* Further, the *energy lost per cycle* must, by definition, be equal to *the energy delivered* to the tank circuit by the class-B or-C amplifier tube or tubes.

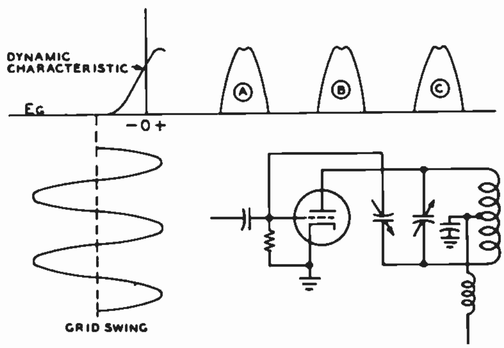


Figure 32

CLASS-C AMPLIFIER OPERATION

Plate current pulses are shown at A, B, and C. The dip in the top of the plate current waveform will occur when the excitation voltage is such that the minimum plate voltage dips below the maximum grid voltage. A detailed discussion of the operation of class-C amplifiers is given in Chapter Seven.

The Q of a tank circuit at resonance is equal to its parallel-resonant impedance (the resonant impedance is resistive at resonance) divided by the reactance of either the capacitor or the inductor which go to make up the tank. The inductive reactance is equal to the capacitive reactance, by definition, at resonance. Hence we may state:

$$Q = \frac{R_L}{X_C} = \frac{R_L}{X_L}$$

where,

R_L is the resonant impedance of the tank,
X_C is the reactance of the tank capacitor,
X_L is the reactance of the tank coil.

This value of resonant impedance (R_L) is the r-f load which is presented to the class-C amplifier tube in a single-ended circuit such as shown in figure 32.

The value of r-f load impedance (R_L) which the class-B/C amplifier tube sees may be obtained, looking in the other direction from the tank coil, from a knowledge of the operating conditions on the class-B/C tube. This load impedance may be obtained from the following expression, which is true in the general case of any class-B/C amplifier:

$$R_L = \frac{(e_{p\ max})^2}{1.8 \times N_p \times I_b \times E_b}$$

where the values in the equation have the characteristics listed in the beginning of Chapter 7.

The expression is academic, since the peak value of the fundamental component of plate voltage swing ($e_{p\ max}$) is not ordinarily known unless a high-voltage peak a-c voltmeter is available for checking. Also, the decimal value of plate-circuit efficiency is not ordinarily known with any degree of accuracy. However, in a *normally operated* class-B/C amplifier the plate voltage swing will be approximately equal to 0.85 to 0.9 times the d-c plate voltage on the stage, and the plate-circuit efficiency will be from 70 to 80 percent (N_p of 0.7 to 0.8), the higher values of efficiency normally being associated with the higher values of plate voltage swing. With these two assumptions as to the normal class-B/C amplifier, the expression for the plate r-f load impedance can be greatly simplified to the following approximate expression, which also applies to class-AB_1 stages:

$$R_L \cong \frac{R_{d.c.}}{1.8}$$

which means simply that the resistance presented by the tank circuit to the class-B/C tube is *approximately equal to one-half the d-c load resistance* which the class-C stage presents to the power supply (and also to the modulator in case high-level modulation of the stage is to be used).

Combining the above simplified expression for the r-f impedance presented by the tank to the tube, with the expression for tank Q given in a previous paragraph we have the following expression which relates the reactance of the tank capacitor or coil to the d-c input to the class-B/C stage.

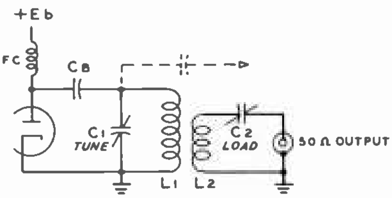
$$X_C = X_L \cong \frac{R_{d.c.}}{2\,Q}$$

The foregoing expression is the basis of the usual charts giving tank capacitance for the various bands in terms of the d-c plate voltage and current to the class-B/C stage, including the chart of figure 33.

Harmonic Radiation versus Q The problem of harmonic radiation from transmitters has long been present, but it has become critical during the past decades along with the extensive occupation of the vhf range. Television signals are particularly susceptible to interference from other signals falling within the passband of the receiver, so that the TVI problem has received the major emphasis of all the services in the vhf range which are susceptible to interference from harmonics of signals in the hf or lower-vhf range.

Inspection of figure 34 will show quickly that the tank circuit of an r-f amplifier should have an operating Q of 10 or greater to afford satisfactory rejection of second-harmonic energy. The curve begins to straighten out above a Q of about 15, so that a considerable increase in Q must be made before an appreciable reduction in second-harmonic energy is obtained. Above a circuit Q of about 10 any increase will not afford appreciable reduction in the third-harmonic energy, so that additional harmonic filtering circuits external to the amplifier proper must be used if increased attenuation of higher-order harmonics is desired. The curves also show that push-pull amplifiers may be operated at Q values of 6 or so, since the second harmonic is cancelled to a large extent if there is no unbalanced coupling between the output tank circuit and the antenna system.

Plate Tank Circuit Design Chart The chart of figure 33 shows circuit capacitance (C) required for a circuit Q of 10, generally considered to be a good compromise value for class AB, B, and C amplifier stages. The capacitance value includes the output capacitance of the tube



PARALLEL RESONANT TANK CIRCUIT Q = 10

COMPONENT		R-F PLATE LOAD RESISTANCE (OHMS) = $\frac{Eb}{2Ib}$ Q = 10 Ib = AMPERES							
	F (MHz)	1000	2000	3000	4000	5000	6000	7000	8000
C1 (pF)	1.8	900	450	300	225	160	150	130	112
	3.5	450	225	150	112	90	75	65	56
	4.0	390	195	130	100	80	65	57	49
	7.0	225	112	75	66	45	38	33	28
	14.0	112	56	38	33	23	19	16	14
	21.0	75	37	25	22	15	12	11	9
	28.0	56	23	19	17	12	9	8	7
	50.0	31	13	11	9	7	5	4	3
L1 (μH)	1.8	9.0	18.0	27.0	36.0	45.0	54.0	63.0	72.0
	3.5	4.5	9.0	13.5	18.0	22.5	27.0	31.5	36.0
	4.0	4.0	8.0	12.0	16.0	20.0	24.0	28.0	32.0
	7.0	2.2	4.5	6.8	9.0	11.0	12.0	15.4	18.0
	14.0	1.1	2.2	3.3	4.5	5.5	6.0	7.7	9.0
	21.0	0.7	1.5	2.1	3.0	3.5	4.0	4.9	6.0
	28.0	0.5	1.1	1.5	2.2	2.7	3.0	3.5	4.5
	50.0	0.3	0.6	0.9	1.2	1.5	1.8	2.1	2.4

COMPONENT	F (MHz)	C2 (pF)	COMPONENT	L2 (μH)
C2 (pF)	1.8	900	L2 (μH)	9.0
	3.5	450		4.5
	4.0	390		4.0
	7.0	225		2.2
	14.0	112		1.1
	21.0	75		0.7
	28.0	56		0.5
	50.0	31		0.3

NOTE:
WHEN SPLIT-STATOR CAPACITOR IS USED IN BALANCED TANK CIRCUIT, CAPACITANCE OF EACH SECTION IS DIVIDED BY TWO AND INDUCTANCE (L1) SHOULD BE MULTIPLIED BY FOUR.

Figure 33

PARALLEL-TUNED-CIRCUIT CHART

Component values listed are for a Q of 10. For other values of Q, use $Q_A/Q_B = C_A/C_B$ and $Q_A/Q_B = L_B/L_A$. Capacitance values shown are divided by four for balanced tank circuit (figure 35C) and inductance is multiplied by four. See figure 35B and D for split-stator circuitry.

and stray circuit capacitances. Total stray capacitance may run from perhaps 5 pF for a low-power vhf stage to as high as 50 pF for a high-power, h-f stage. Also included in the chart are appropriate values for the tank inductance (L_1).

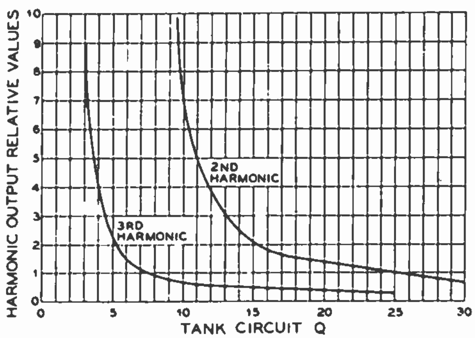


Figure 34

RELATIVE HARMONIC OUTPUT PLOTTED AGAINST TANK CIRCUIT Q

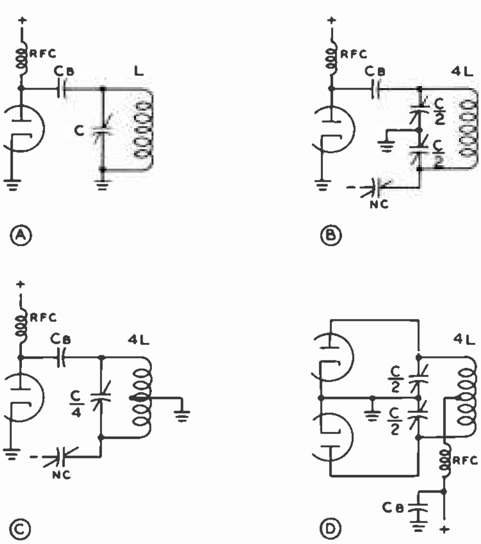


Figure 35

PARALLEL-TUNED TANK CIRCUITS

A—Single ended, use chart of figure 33 for values of L and C. B—Single-ended, split tank. Multiply values of L by four. Each section of split-stator capacitor is 1/2 value listed in figure 33. C—Split tank with single-section capacitor. Capacitor value is 1/4 value listed in figure 33. D—Push-pull circuit with split-stator capacitor. Each section of capacitor is 1/2 value indicated in figure 33.

While tank circuit constants are determined by the r-f load resistance, as discussed earlier, this chart has been modified to read in terms of the d-c load resistance, as determined by the ratio of d-c plate voltage to twice the value of the maximum (peak) d-c plate current in amperes. For linear amplifier service, the maximum plate current may be taken as that noted for proper loading at resonance with full carrier injection.

If a different value of circuit Q is desired, a new Q value may be established by a simple ratio. For example, with a given value of plate voltage to plate current ratio, revised values of constants for a Q of 12 may be found by multiplying the capacitance by 12/10 and the inductance by 10/12. When a split tank circuit is used (figure 35B, D), the capacitance value may be reduced as shown and the inductance raised, while still maintaining a constant value of circuit Q.

At the higher frequencies, stray circuit capacitance may be larger than the value determined for a Q of 10. In this case, the Q must be raised to a higher value. Circuit Q values of 15 to 50 are often unavoidable and commonly used in the vhf range because of high stray circuit capacitance.

USUAL BREAKDOWN RATINGS OF COMMON PLATE SPACINGS

Air-gap in inches	Peak voltage breakdown
.030	1000
.050	2000
.070	3000
.100	4000
.125	4500
.150	5200
.170	6000
.200	7500
.250	9000
.350	11,000
.500	15,000
.700	20,000

Recommended air-gap for use when no d-c voltage appears across plate tank capacitor (when plate circuit is shunt fed, or when the plate tank capacitor is insulated from ground).

D-C plate voltage	C-W	Plate mod.
400	.030	.050
600	.050	.070
750	.050	.084
1000	.070	.100
1250	.070	.144
1500	.078	.200
2000	.100	.250
2500	.175	.375
3000	.200	.500
3500	.250	.600

Figure 36

Spacings should be multiplied by 1.5 for same safety factor when d-c voltage appears across plate tank capacitor.

At the lower frequencies, on the other hand, circuit Q may be decreased to as low as 3 to reduce the cost of the tank tuning capacitor and to reduce circuit selectivity to eliminate sideband clipping. The increased harmonic content of the output waveform, in this instance, is reduced by placing a suitable harmonic filter in the transmission line from amplifier to antenna.

The tank circuit operates in the same manner whether the tube driving it is a pentode, triode, or tetrode; whether the circuit is single-ended or push-pull; or whether it is shunt-fed or series-fed. The prime factor in establishing the operating Q of the tank circuit is the ratio of the loaded resonant impedance across its terminals to the reactance of the coil and capacitor which make up the circuit.

Effect of Loading on Q The Q of a circuit depends on the resistance in series with the capacitance and inductance. This series resistance is very low for a low-loss coil not loaded by an antenna circuit. The value of Q may be from 100 to 400 under these conditions. Coupling an antenna circuit has the effect of increasing the series resistance, though in this case the power is consumed as useful radiation by the antenna. Mathematically, the antenna increases the value of R in the expression $Q = \omega L/R$ where L is the coil inductance in microhenrys and ω is the term $2\pi f$ (f being in MHz).

The coupling from the final tank circuit to the antenna or antenna transmission line can be varied to obtain values of Q from perhaps 3 at maximum coupling to a value of Q equal to the unloaded Q of the circuit at zero antenna coupling. This value of unloaded Q can be as high as 400, as mentioned in the preceding paragraph. However, the value of $Q = 10$ will not be obtained at values of normal d-c plate current in the class-C amplifier stage unless the C-to-L ratio in the tank circuit is correct for that frequency of operation.

Tuning Capacitor Air Gap To determine the required tuning-capacitor air gap for a particular amplifier circuit it is first necessary to estimate the peak r-f voltage which will appear between the plates of the tuning capacitor. Then, using figure 36, it is possible to estimate the plate spacing which will be required.

The instantaneous r-f voltage in the plate circuit of a class-C amplifier tube varies from nearly zero to nearly twice the d-c plate voltage. If the d-c voltage is being 100 percent modulated by an audio voltage, the r-f peaks will reach nearly four times the d-c voltage.

These rules apply to a loaded amplifier or buffer stage. If either is operated without an r-f load, the peak voltages will be greater and can *exceed* the d-c plate supply voltage. For this reason no amplifier should be operated without load when anywhere near normal d-c plate voltage is applied.

If a plate blocking capacitor is used, it must be rated to withstand the d-c plate voltage plus any audio voltage. This capacitor should be rated at a d-c working voltage of at least *twice the d-c plate supply in a plate-modulated amplifier*, and at least *equal to the d-c supply* in any other type of r-f amplifier.

Inductive Coupling to a Coaxial Line The chart of figure 33 provides data for coupling the resonant tank circuit to a low-impedance coaxial transmission line. To achieve proper coupling the coupling coil should be series-resonated to the tank frequency. The inductance of the link coil is such that its reactance at the operating frequency is equal to the characteristic impedance of the transmission line. The circuit Q of the link-capacitor combination may be as low as 2. In such a case, the value of series capacitance is quite large and the value may be reduced to a more practical amount by placing an auxiliary inductance (L) in series with the link coil as shown in figure 37.

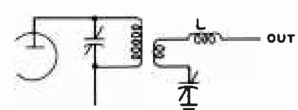
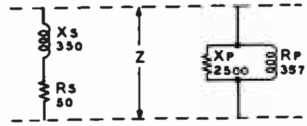


Figure 37

AUXILIARY LOADING COIL (L) USED IN SERIES-TUNED ANTENNA CIRCUIT TO ACHIEVE MAXIMUM COUPLING

11-12 L, Pi, and Pi-L Matching Networks

Various types of networks are used to transform one impedance to another and network types known as *L, pi,* and *pi-L* are commonly used in transmitter circuitry for this purpose. The reason these networks are able to complete a transformation is that, for any series circuit consisting of a series reactance and resistance, there can be found an equivalent parallel network which possesses the same impedance characteristics (figure 38). Such networks are used to accomplish a match between the tube or device of an amplifier and a transmission line.



(1) $Z_S = \sqrt{R_S^2 + X_S^2}$

(2) $Q = \frac{X_S}{R_S}$

(3) $Z_P = \frac{R_P X_P}{\sqrt{R_P^2 + X_P^2}}$

(4) $Q = \frac{R_P}{X_P}$

AND (5)

$\frac{R_P}{R_S} = Q^2 + 1$

Figure 38

SERIES TO PARALLEL IMPEDANCE CONVERSION

The L-Network The *L-network* is the simplest of the matching networks and may take either of the two forms of figure 39. The two configurations are equivalent, and the choice is usually made on the basis of other component and circuit considerations apart from the impedance matching characteristics. The circuit shown in illustration (B) is generally preferred because the shunt capacitor (C) provides a low impedance path to ground for the higher harmonic frequencies.

The L-network is of limited utility in impedance matching since its ratio of impedance transformation is fixed at a value equal to (Q^2+1). The operating Q may be relatively low (perhaps 3 to 6) in a matching network between the plate *tank circuit* of an amplifier and a transmission line; hence impedance transformation ratios of 10 to 1 and even lower may be attained. But when the network also acts as the plate tank circuit of the amplifier stage, as in figure 40, the operating Q should be at least 10 and preferably 15. An operating Q of 15 represents an impedance transformation of 225; this value normally will be too high even for transforming from the 2000- to 10,000-ohm plate impedance of a class B/C amplifier stage down to a 50-ohm transmission line.

However, the L-network is interesting since it forms the basis of design for the pi-network. Inspection of figure 40 will show that the L-network in reality must be con-

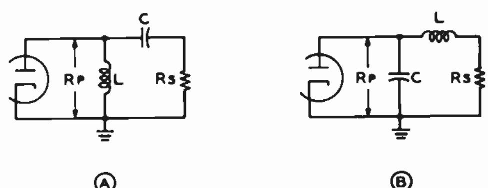
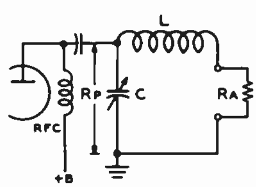


Figure 39

TWO EQUIVALENT L-NETWORKS

A—Inductance in parallel leg, capacitance in series leg. B—Capacitance in parallel leg, inductance in series leg. Impedance values for both circuits are given in figure 40.



$R_P = R_A(Q^2+1)$ (EXACT)

$R_P = Q^2 R_A$ (APPROX.)

$Q = \frac{X_L}{R_A} = \frac{X_C}{R_A} = \frac{R_P}{X_C} = \frac{R_P}{X_L}$

$X_L = X_C$

FOR OPERATING CIRCUIT Q OF 15:

$R_P =$ APPROX. $\frac{\text{PLATE VOLTAGE}}{\text{PLATE CURRENT}}$

$R_P = 225\ R_A$

$X_C = \frac{R_P}{15}$

$X_L = \frac{R_P}{15}$

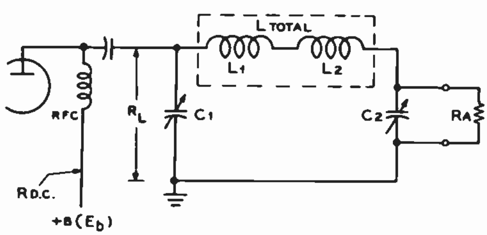
Figure 40

THE L-NETWORK IMPEDANCE TRANSFORMER

The L-network is useful with a moderate operating Q for high values of impedance transformation, and it may be used for applications other than in the plate circuit of a tube with relatively low values of operating Q for moderate impedance transformations. Exact and approximate design equations are given.

sidered as a parallel-resonant tank circuit in which R_A represents the coupled-in load resistance; only in this case the load resistance is directly coupled into the tank circuit rather than being inductively coupled as in the conventional arrangement where the load circuit is coupled to the tank circuit by means of a link. When R_A is shorted, L and C comprise a conventional parallel-resonant tank circuit, since for proper operation L and C must be resonant in order for the network to present a resistive load to the class-C amplifier.

The Pi-Network The *pi-network* can be considered as two back-to-back L-networks as shown in figure 41. This network is much more general in its application than the L network since it offers greater harmonic attenuation, and since it can be used to match a relatively wide range of impedances, while still maintaining any desired operating Q. The values of C_1 and L_1 in the pi-network of figure 41 can be thought of as having the same values of the L network in figure 40 for the same operating Q, but, what is more important from the comparison standpoint these values will be about the *same as in a conventional tank circuit.*



$$R_{D.C.} = \frac{E_b}{I_b}$$

$$R_P \approx \frac{R_{D.C.}}{2}$$

$$X_{C1} = \frac{R_L}{Q}$$

$$X_{L1} = \frac{R_L}{Q}$$

$$X_{C2} = -R_A \sqrt{\frac{R_L}{R_A(Q^2+1)-R_L}}$$

$$X_{L2} = -\frac{R_A^2 X_{C2}}{R_A^2 + X_{C2}^2}$$

$$X_{L\,TOT.} = X_{L1} + X_{L2}$$

Figure 41

THE PI-NETWORK

The pi-network is valuable for use as an impedance transformer over a wide ratio of transformation values. The operating Q should be at least 12 when the circuit is to be used in the plate circuit of a class-C amplifier. Design equations are given above. Inductor L_{Tot} represents a single inductance, usually variable, with a value equal to the sum of L_1 and L_2.

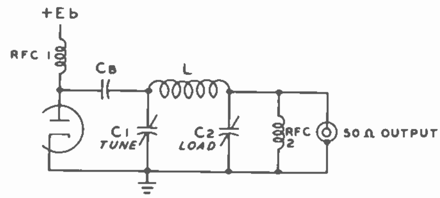
The value of the capacitance may be determined by calculation with the operating Q and the load impedance which should be reflected to the plate of the class-C amplifier as the two known quantities—or the actual values of the capacitance may be obtained for an operating Q of 10 by reference to the chart of figure 42.

The inductive arm in the pi-network can be thought of as consisting of two inductances in series, as illustrated in figure 41.

The first portion of this inductance (L_1) is that value of inductance which would resonate with C_1 at the operating frequency —the same as in a conventional tank circuit. However, the actual value of inductance in this arm of the pi-network, L_{Tot} will be greater than L_1 for normal values of impedance transformation. For high transformation ratios L_{Tot} will be only slightly greater than L_1; for a transformation ratio of 1.0, L_{Tot} will be twice as great as L_1. The amount of inductance which must be added to L_1 to restore resonance and maintain circuit Q is obtained through use of the expression for X_{L1} and X_{L2} in figure 41.

The peak voltage rating of the main tuning capacitor (C_1) should be the normal value for a class-C amplifier operating at the plate voltage to be employed. The inductor (L_{Tot}) may be a plug-in coil which is changed for each band of operation, or some sort of variable inductor may be used. A continuously variable slider-type variable inductor may be used to good advantage if available, or a tapped inductor may be employed. However, to maintain good circuit Q on the higher frequencies when a variable or tapped coil is used on the lower frequencies, the tapped or variable coil should be removed from the circuit and replaced by a smaller coil which has been especially designed for the higher frequency ranges.

The peak voltage rating of the output or loading capacitor (C_2) is determined by the power level and the impedance to be fed. If a 50-ohm coaxial line is to be fed from the pi-network, receiving-type capacitors will be satisfactory even up to the power level of a plate-modulated kilowatt amplifier. In any



PI-NETWORK Q=10

PI-NETWORK COMPONENT	R-F PLATE LOAD RESISTANCE (OHMS) = $\frac{Eb}{2Ib}$ Q=10 Ib=AMPERES								
	F (MHz)	1000	1500	2000	2500	3000	3500	4000	4500
C1 (PF)	1.8	900	600	450	360	300	255	225	200
	3.5	450	300	225	180	150	128	112	100
	4.0	395	280	200	180	130	115	100	88
	7.0	225	150	112	90	75	64	58	50
	14.0	112	75	56	45	37	32	28	25
	21.0	75	50	38	30	25	21	19	17
	28.0	56	37	28	23	18	18	14	13
	50.0	32	22	18	13	11	9	8	7
C2 (PF)	1.8	3700	2840	2200	1800	1420	1180	940	880
	3.5	1850	1420	1100	900	710	580	470	340
	4.0	1850	1250	980	780	620	510	410	270
	7.0	925	710	550	450	355	290	235	170
	14.0	480	355	275	225	175	145	117	85
	21.0	310	238	183	150	120	97	78	57
	28.0	230	180	138	112	87	72	58	42
	50.0	130	100	77	63	49	40	33	24
L (µH)	1.8	10.8	15.8	20.2	25.0	29.4	34.0	38.0	42.0
	3.5	5.4	7.9	10.1	12.5	14.7	17.0	19.0	21.0
	4.0	4.7	8.8	9.0	11.0	13.0	14.5	18.5	18.0
	7.0	2.7	4.0	5.0	8.3	7.3	8.5	9.5	10.5
	14.0	1.4	2.0	2.5	3.1	3.6	4.2	4.7	5.3
	21.0	0.9	1.3	1.7	2.1	2.4	2.8	3.2	3.5
	28.0	0.7	1.0	1.3	1.5	1.8	2.1	2.3	2.6
	50.0	0.4	0.5	0.7	0.9	1.0	1.2	1.3	1.4

Figure 42

PI-NETWORK CHART

Component values listed are for class-AB/B service for a Q of 10. For other values of Q, use $Q_A/Q_B = C_A/C_B$ and $Q_A/Q_B = L_B/L_A$. When plate load resistance is higher than 3000 ohms, or for class-C service, it is recommended that components be selected for a circuit Q between 12 and 15. For 70-ohm termination, multiply values of capacitor C_2 by 0.72.

event, the peak voltage which will be impressed across the output capacitor is expressed by:

$$e_p = \sqrt{2 \times R_a \times P_o}$$

where,

e_p is the peak voltage across the capacitor,
R_a is the value of resistive load which the network is feeding,
P_o is the maximum value of the average power output of the stage.

The harmonic attenuation of the pi-network is greater than that of the simple L-network but is not considered great enough to meet the FCC transmitter requirements for harmonic attenuation. The attenuation to second harmonic energy is approximately −35 decibels for the pi-network for a transformation ratio of 40, and increases to −40 db when the operating Q is raised from 10 to 15.

The Pi-L Network The *pi-L network* is made up of three L-networks and provides a greater transformation ratio and higher harmonic suppression than do either of the simpler networks (figure 43). Because the loading capacitor is placed at the *image impedance* level (R_1), which is usually of the order of 300 to 700 ohms, the peak voltage across the capacitor ($C_2A + C_2B$) will be higher than that across the output capacitor of an equivalent pi-network, and the value of the pi-L capacitor will be appreciably less than that of the equivalent pi-network loading capacitor. A formal calculation of the pi-L circuit parameters is given in the article *"The Pi-L Plate Circuit in Kilowatt Amplifiers,"* QST, July, 1962. A free reprint of this article may be obtained by writing to: Amateur Service Department, EIMAC division of Varian, San Carlos, Calif. Typical components for pi-L network design for the various h-f amateur bands is given in the chart of figure 44.

For a transformation ratio of 40 the attenuation to second harmonic energy is about −52 decibels for a pi-L network having a Q of 10 and and image impedance of 300 ohms, rising to −55 decibels for a Q of 15 (figure 45).

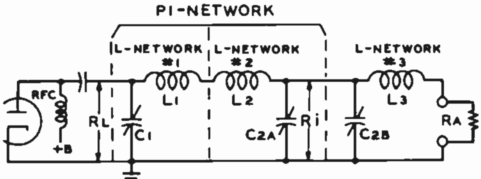


Figure 43

PI-L NETWORK IS MADE UP OF THREE L-NETWORKS IN SERIES

Pi-L network provides greater transformation ratio and higher harmonic suppression than do either the L- or the pi-networks. Loading capacitor (C_2) is common to networks 2 and 3 and is placed at image impedance level (R_i) which is usually of the order of 300 to 700 ohms.

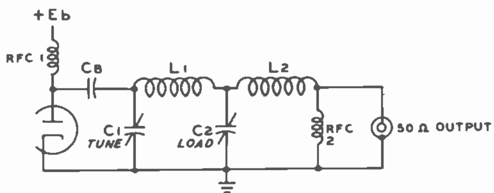
11-13 Toroidal-Wound Tank Coils

Ferrite-core toroids are often used in tank circuits up to the multikilowatt power level (figure 46). Because the toroid coil is self-shielding, its use permits much greater component density than does an equivalent air coil having a large field about it. The toroid, moreover, may be mounted directly against a metal surface without a significant change of circuit Q, and only a small change in overall distributed capacitance of the coil.

When used at h-f, the core and winding losses of a ferrite-core inductor cannot be reduced by traditional design as in the case of lower-frequency transformers. In order to keep the leakage inductance small, the toroid winding must have the minimum possible number of turns, which means the core material is very heavily loaded. The ultimate power rating of the ferrite-core inductor therefore depends upon the effectiveness of core cooling. As the thermal conductivity of ferrite material is quite low, this means that core temperature can become quite high.

The working temperature of a ferrite core is limited to a medium value, and, as the temperature rises, the core loss increases rapidly and core permeability drops. There exists a temperature known as the *runaway temperature*, above which any increase in cooling is more than offset by the increase in losses. The working temperature must be held well below this point.

As the operating temperature of the ferrite core rises, both permeability and Q drop,



PI-L NETWORK Q = 10

IMAGE RESISTANCE = 350 Ω

PI-L NETWORK COMPONENT		R-F PLATE LOAD RESISTANCE (OHMS) = $\frac{Eb}{2 Ib}$ Q = 10 Ib = AMPERES							
	F (MHz)	1000	1500	2000	2500	3000	3500	4000	4500
C1 (pF)	1.8	900	600	450	360	300	256	224	200
	3.5	450	300	225	180	150	128	112	100
	4.0	395	260	200	160	130	115	100	88
	7.0	225	150	112	90	75	84	56	50
	14.0	112	75	58	45	37	32	28	25
	21.0	76	50	38	30	25	21	19	17
	28.0	56	37	28	23	18	16	14	13
	50.0	32	22	16	13	11	9	8	7
C2 (pF)	1.8	2160	1880	1690	1600	1500	1440	1380	1320
	3.5	1080	940	845	800	750	720	690	660
	4.0	940	820	740	690	650	620	600	570
	7.0	540	470	422	400	375	360	345	330
	14.0	270	235	211	200	187	180	175	165
	21.0	180	155	140	130	125	120	117	110
	28.0	135	117	105	100	93	90	87	82
	50.0	75	66	59	56	53	50	48	46
L1 (μH)	1.8	14.2	20.0	25.2	31.0	36.0	41.0	46.0	51.0
	3.5	7.1	10.0	12.6	15.5	18.0	20 5	23.3	25.5
	4.0	6.3	8.8	11.0	13.5	15.5	18.0	20.0	22.5
	7.0	3.6	5.0	6.3	7.6	9.0	10.3	11.5	12.7
	14.0	1.8	2.5	3.2	3.8	4.5	5.2	5.7	6.3
	21.0	1.2	1.6	2.1	2.5	3.0	3.4	3.8	4.2
	28.0	0.9	1.3	1.8	1.8	2.2	2.6	2.8	3.1
	50.0	0.5	0.7	0.9	1.0	1.2	1.4	1.6	1.6

F (MHz)	1.8	3.5	4.0	7.0	14.0	21.0	28.0	50.0
L2 (μH)	11.2	5.6	4.8	2.7	1.4	0.9	0.7	0.4

Figure 44

PI-L NETWORK CHART

Component values are listed for class AB/B/C service for a Q of 10. For other values of Q, use conversion transformations listed in figures 33 and 42. Image impedance of 350 ohms is used for calculations.

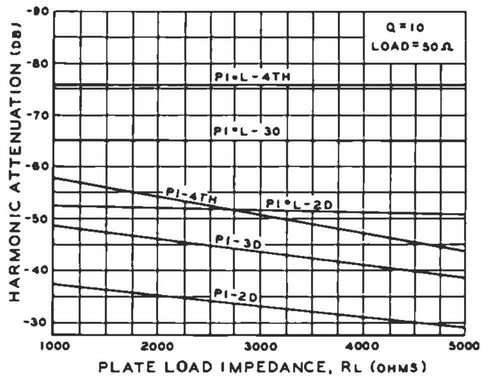


Figure 45

HARMONIC ATTENUATION OF PI- AND PI-L NETWORKS

Second, third and fourth harmonic levels are shown relative to fundamental signal. Pi-L configuration provides improved attenuation to all higher harmonics as compared to pi-network.

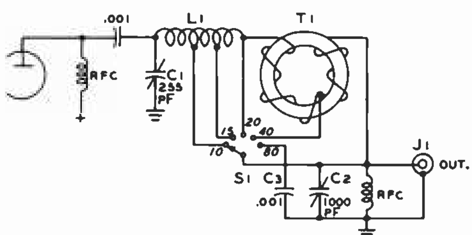


Figure 46

FERRITE TOROID TANK CIRCUIT

L_1—9 turns of 3⁄16" copper tubing, 1¼" inside diameter, 2½" long. 10 meter tap is 3.5 turns from plate, 15 meter tap 5.5 turns. Toroid inductor is 16 turns #10 wire tapped 5 turns from output end.

causing a decrease in circuit efficiency and a detuning action both of which will contribute to higher core temperature. In addition, the nonlinearity of the ferrite material under a varying r-f flux can cause intermodulation distortion when the ferrite is being driven by a complex signal such as encountered in SSB service.

Each class of ferrite material achieves maximum circuit Q at a specific frequency and the ferrite is graded for initial permeability at a stated operating frequency.

11-14 Grid Bias

Radio-frequency amplifiers require some form of *grid bias* for proper operation. Practically all r-f amplifiers operate in such a manner that plate current flows in the form of short pulses which have a duration of only a fraction of an r-f cycle. To accomplish this with a sinusoidal excitation voltage, the operating grid bias must be at least sufficient to cut off the plate current. In a high efficiency class-C amplifier the operating bias may be several times the cutoff value. Cutoff bias, it will be recalled, is that value of grid voltage which will reduce the plate current to zero at the plate voltage employed. The method for calculating it has been indicated previously. This theoretical value of cutoff will not reduce the plate current completely to zero, due to the variable-μ tendency or "knee" which is characteristic of all tubes as the cutoff point is approached.

Class-C Bias Amplitude-modulated class-C amplifiers should be operated with the grid bias adjusted to a value greater than twice cutoff at the operating plate voltage. This procedure will ensure that the tube is operating at a bias greater than cutoff when the plate voltage is doubled on positive modulation peaks. C-w, RTTY, and f-m transmitters can be operated with bias as low as cutoff, if only limited excitation is available and moderate plate efficiency is satisfactory. In a c-w transmitter, the bias supply or resistor should be adjusted to the point which will allow normal grid current to flow for the particular amount of grid driving r-f power available.

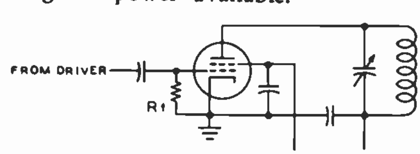


Figure 47

SELF BIAS

The grid resistor on an amplifier or multiplier stage may also be used as the shunt feed impedance to the grid of the tube when a high value of resistor (greater than perhaps 20,000 ohms) is used. When a lower value of grid resistor is to be employed, an r-f choke should be used between the grid of the tube and the grid resistor to reduce r-f losses in the grid resistance.

Self Bias A resistor can be connected in the grid circuit of a class-C amplifier to provide *self-bias.* This resistor (R_1 in figure 47), is part of the d-c path in the grid circuit.

The r-f excitation applied to the grid circuit of the tube causes a pulsating direct current to flow through the bias supply lead, due to the rectifying action of the grid, and any current flowing through R_1 produces a voltage drop across that resistor. The grid of the tube is positive for a short duration of each r-f cycle, and draws electrons from the filament or cathode of the tube during that time. These electrons complete the circuit through the d-c *grid return.* The voltage drop across the resistance in the grid return provides a *negative bias* for the grid.

Self bias automatically adjusts itself over fairly wide variations of r-f excitation. The value of grid resistance should be such that normal values of grid current will flow at the maximum available amount of r-f excitation. Self bias cannot be used for grid-modulated or linear amplifiers in which the average d-c current is constantly varying with modulation.

Safety Bias Self bias alone provides no protection against excessive plate current in case of failure of the source of r-f grid excitation. A well-regulated low-voltage bias supply can be connected in series with the grid resistor as shown in figure 48. This fixed protective bias will protect the tube in the event of failure of grid excitation. "Zero-bias" tubes do not require this bias source, since their plate current will drop to a safe value when the excitation is removed.

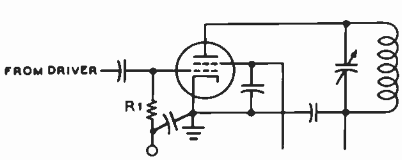


Figure 48

COMBINATION SELF- AND FIXED BIAS

Self-bias often is used in conjunction with a fixed minimum value of power supply bias. This arrangement permits the operating bias to be established by the excitation energy, but in the absence of excitation the electrode currents to the tube will be held to safe values by the fixed-minimum power supply bias. If a relatively low value of grid resistor is to be used, an r-f choke should be connected between the grid of the tube and the resistor as discussed in figure 47.

Cathode Bias A resistor can be connected in series with the cathode or center-tapped filament lead of an amplifier to secure *automatic bias.* The plate current flows through this resistor, then back to the cathode or filament, and the voltage drop across the resistor can be applied to the grid circuit by connecting the grid bias lead to the grounded or power-supply end of resistor R, as shown in figure 49.

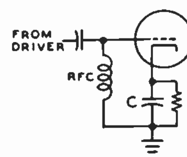


Figure 49

R-F STAGE WITH CATHODE BIAS

Cathode bias sometimes is advantageous for use in an r-f stage that operats with a relatively small amount of r-f excitation.

The grounded (B-minus) end of the cathode resistor is negative relative to the cathode by an amount equal to the voltage drop across the resistor. The value of resistance must be so chosen that the sum of the desired grid and plate current flowing through the resistor will bias the tube for proper operation.

This type of bias is used more extensively in audio-frequency than in radio-frequency amplifiers. The voltage drop across the resistor must be subtracted from the total plate supply voltage when calculating the power input to the amplifier, and this loss of plate voltage in an r-f amplifier may be excessive. A class-A audio amplifier is biased only to approximately one-half cutoff, whereas an r-f amplifier may be biased to twice cutoff, or more, and thus the plate supply voltage loss may be a large percentage of the total available voltage when using low- or medium-μ tubes.

Often just enough cathode bias is employed in an r-f amplifier to act as safety bias to protect the tubes in case of excitation failure, with the rest of the bias coming from a grid resistor.

Separate Bias Supply An external supply often is used for grid bias, as shown in figure 50. The bleeder resistance across the output of the filter can be made sufficiently low in value that the grid current of the amplifier will not appreciably change the amount of negative grid-bias voltage. Alternately, a voltage-regulated grid-bias supply can be used. This type of bias supply is used in class-B audio and class-

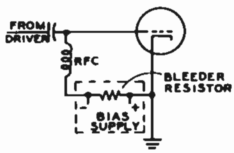


Figure 50

SEPARATE BIAS SUPPLY

A separate bias supply may be used for triodes or tetrodes. Bias is applied across a low-resistance bleeder. Grid current (if any) flowing through bleeder will boost bias voltage over nominal value of supply. Bias supply for AB_1 linear amplifier, even though no grid curernt is encountered, must still have low-resistance bleeder to help overcome rise in bias due to collection of primary electrons on grid of tube.

B r-f linear amplifier service where the voltage regulation in the C-bias supply is important. For a class-C amplifier, regulation is not so important, and an economical design of components in the power supply, therefore, can be utilized. In this case, the bias voltage must be adjusted with normal grid current flowing, as the grid current will raise the bias considerably when it is flowing through the bias-supply bleeder resistance.

Zener Bias A few volts of bias may be needed to reduce the zero-signal plate current of a "zero-bias" triode, particularly if the equipment is power-supply limited. A low-impedance bias source is required and the simplest way of obtaining well-regulated bias voltage is to place a zener diode in the filament or cathode return circuit (figure 51). The 1N4551, for example, has a nominal voltage drop of 4.7 volts and an impedance of 0.1 ohm, making it ideal for this service. At this value of bias, the zero-signal plate current of a 3-500Z at a plate potential of 3250 volts is reduced from 160 to approximately 90 milliamperes.

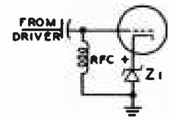


Figure 51

ZENER CATHODE BIAS

Zener diode may be used to obtain a few volts of well-regulated cathode bias. This circuit may be used to reduce zero-signal plate current of high-μ triodes in cathode-driven (grounded-grid) service.

The 1N4551 diode may be bolted directly to the chassis which will act as a heat sink.

11-15 Protective Circuits for Transmitting Tubes

The tetrode transmitting tube requires three operating voltages: grid bias, screen voltage, and plate voltage. The current requirements of these three operating voltages are somewhat interdependent, and a change in potential of one voltage will affect the current drain of the tetrode in respect to the other two voltages. In particular, if the grid excitation voltage is interrupted as by keying action, or if the plate supply is momentarily interrupted, the resulting voltage or current surges in the screen circuit are apt to permanently damage the tube.

The Series Screen Supply A simple method of obtaining screen voltage is by means of a dropping resistor from the high-voltage plate supply, as shown in figure 52. This circuit is recommended for use with low power tetrodes (6146, 5763, etc.) in class-C service. Because of poor regulation with varying screen current it should not be used in a linear amplifier stage. Since the current drawn by the screen is a function of the exciting

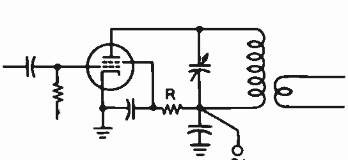


Figure 52

DROPPING-RESISTOR SCREEN SUPPLY

voltage applied to the tetrode, the screen voltage will rise to equal the plate voltage under conditions of no exciting voltage. If the control grid is overdriven, on the other hand, the screen current may become excessive. In either case, damage to the screen and its associated components may result. In addition, fluctuations in the plate loading of

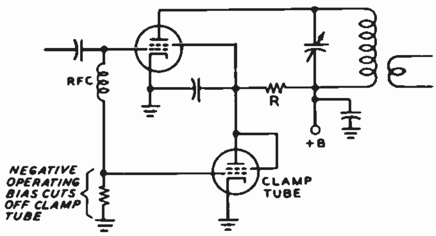


Figure 53

CLAMP-TUBE SCREEN SUPPLY

the tetrode stage will cause changes in the screen current of the tube. This will result in screen voltage fluctuations due to the inherently poor voltage regulation of the screen series dropping resistor. These effects become dangerous to tube life if the plate voltage is greater than the screen voltage by a factor of 2 or so.

The Clamp Tube A clamp tube may be added to the series screen supply, as shown in figure 53. The clamp tube is normally cut off by virtue of the d-c grid bias drop developed across the grid resistor of the tetrode tube. When excitation is removed from the tetrode, no bias appears across the grid resistor, and the clamp tube conducts heavily, dropping the screen voltage to a safe value. When excitation is applied to the tetrode the clamp tube is inoperative, and fluctuations of the plate loading of the tetrode tube could allow the screen voltage to rise to a damaging value. Because of this factor, the clamp tube does not offer complete protection to the tetrode.

The Separate Screen Supply A low-voltage screen supply may be used instead of the series screen-dropping resistor. This will protect the screen circuit from excessive voltages when the other tetrode operating parameters shift. However, the screen can be easily damaged if plate or bias voltage is removed from the tetrode, as the screen current will reach high values and the screen dissipation will be exceeded. If the screen supply is capable of providing slightly more screen voltage than the tetrode requires for proper operation, a series wattage-limiting resistor may be added to the circuit as shown in figure 54. With this resistor in the circuit it is possible to apply excitation to the tetrode tube with screen voltage present (but in the absence of plate voltage) and still not damage the screen of the tube. The value of the resistor should be chosen so that the product of the voltage applied to the screen of the tetrode times the screen current never exceeds the maximum rated screen dissipation of the tube.

This circuit is not suited for linear amplifiers since the screen voltage regulation is poor.

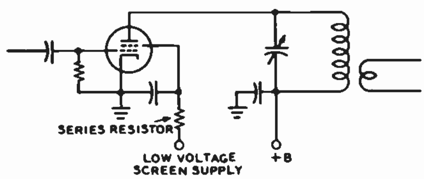


Figure 54

A PROTECTIVE WATTAGE-LIMITING RESISTOR FOR USE WITH LOW-VOLTAGE SCREEN SUPPLY

Screen Protection In designing equipment using high-power tetrodes, consideration must be given to control of secondary emission from the screen element of the tube. The screen is normally operated at a relatively low potential to accelerate the electrons emitted from the cathode. Not all of the electrons pass through the screen grid on the way to the plate, some of them being intercepted by the grid. In the process of striking the screen grid, other electrons are emitted, some of which may be attracted by the higher potential of the plate. The result is a flow of electrons from the screen to the plate. It is possible that more electrons will leave the screen than will arrive and a screen meter will indicate a reverse electron flow, or negative screen current, under this condition. A low-impedance path to ground must be provided for this flow, otherwise the screen voltage will attempt to rise to the value of the plate voltage, by virtue of the IR drop created by the negative screen current flowing across the high-impedance screen circuit. As the screen voltage rises, the plate current of the tetrode increases and the tube is in a runaway condition. The addition of a resistor from screen to ground will compensate for the effect of negative screen current. The value of this resistor will be such that the bled current will run from 20 ma to as high as

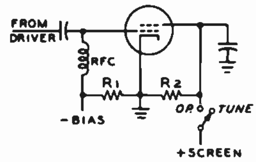


Figure 55

SCREEN CONTROL CIRCUIT

The d-c return path to ground for screen of a tetrode should not be broken. Resistor R_1 completes the circuit and screen high-voltage lead may be open to reduce stage gain for tuneup purposes.

70 ma, depending on the tube type. Tube data sheets normally state the amount of bled current required to counteract the emission current.

A correct circuit for the screen supply of a linear amplifier, including a "tune-operate" switch is shown in figure 55. In the "tune" position, screen voltage is removed, permitting adjustments to be made to the circuit at a very low power level for tuneup purposes.

Grid Protection The impedance of the grid circuit must be considered, particularly in class AB_1 amplifiers wherein a regulated bias source is required. Primary grid emission can cause trouble if the impedance of the grid circuit is too high. The d-c resistance to ground of the bias supply should be sufficiently low (below 1000 ohms or so) to prevent appreciable reverse bias from being developed by the flow of emission current through the internal resistance of the bias supply. The reverse bias produced by this effect tends to subtract from the grid bias, causing a runaway condition if not controlled.

Arc Protection Modern transmitting tubes have a very close internal spacing between elements to achieve high power gain and good performance at very high operating frequencies. Components, too, tend toward more compact sizes to allow high-density construction in modern equipment. Under these conditions, flashovers or arcing between high- and low-potential points in the circuit or tube may possibly occur. The impedance of an arc is very low, of the order of an ohm or so, and extremely high values of *fault current* flow during the flashover. Fault current flowing through a small resistance or impedance creates a high voltage drop in unexpected places and may result in damaged equipment. A flashover in a d-c plate circuit, for example, can discharge the power-supply filter capacitor in a fraction of a second and allow thousands of amperes of current to pass through the arc and any components in series with the discharge path.

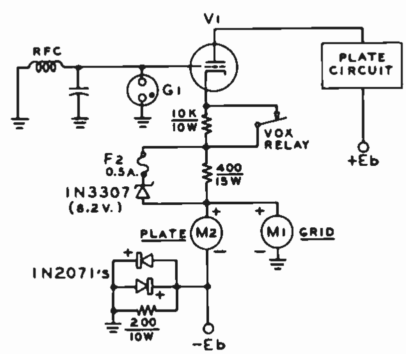


Figure 56

FLASHOVER PROTECTION

Equipment can be protected from flashover and high flashover currents by placing spark gap (G_1) from grid to ground, zener diode fuse in cathode lead and reverse-connected diodes placed across metering circuit. Spark gap arcs over at a predetermined voltage to provide low-impedance path from grid to ground, thus protecting r-f choke and grid bypass capacitor. Cathode fuse opens under heavy arc current, protecting zener diode, while shunt resistor provides path to ground for fault current. Reverse-connected diodes across plate and grid meters provide low-impedance shunt when voltage across meters reaches level of forward voltage drop across the diodes (about 0.4 to 0.8 volt, depending on diode temperature). Filter capacitors in power supply may also be series-connected with a high-voltage quick-action fuse to prevent discharge through fault circuit in equipment.

A sparking gap (G_1) may be placed at a critical point, as shown in figure 56 to protect tube and components against transient arc voltages and a high-voltage, quick-action fuse can be placed in series with high capacity filter circuits to prevent damaging fault currents from flowing through delicate metering circuits or zener diodes. Meters may be protected from overload by placing reverse-connected silicon diodes, across them to carry the fault current, as shown in the illustration.

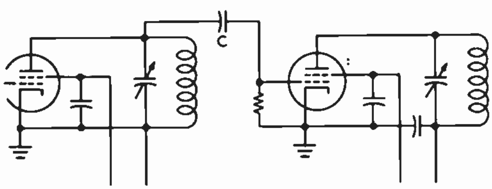


Figure 57

CAPACITIVE INTERSTAGE COUPLING

11-16 Interstage Coupling

Energy is usually coupled from one circuit of a transmitter into another either by *capacitive coupling*, *inductive coupling*, or *link coupling*. The latter is a special form of inductive coupling. The choice of a coupling method depends on the purpose for which it is to be used.

Capacitive Coupling Capacitive coupling between an amplifier or doubler circuit and a preceding driver stage is shown in figure 57. The coupling capacitor (C) isolates the d-c plate supply from the next grid and provides a low-impedance path for the rf energy between the tube being driven and the driver tube. This method of coupling is simple and economical for low-power amplifier or exciter stages, but has certain disadvantages, particularly for high-frequency stages. The grid leads in an amplifier should be as short as possible, but this is difficult to attain in the physical arrangement of a high-power amplifier with respect to a capacitively coupled driver stage.

Disadvantages of Capacitive Coupling One significant disadvantage of capacitive coupling is the difficulty of adjusting the load on the driver stage. Impedance adjustment can be accomplished by tapping the coupling lead a part of the way down on the plate coil of the tuned stage of the driver circuit; but often when this is done a parasitic oscillation will take place in the stage being driven.

One main disadvantage of capacitive coupling lies in the fact that the grid-to-filament capacitance of the driven tube is placed directly across the driver tuned circuit. This condition sometimes makes the r-f amplifier difficult to neutralize, and the increased minimum circuit capacitance makes it difficult to use a reasonable size coil in the vhf range. Difficulties from this source can be partially eliminated by using a center-tapped or split-stator tank circuit in the plate of the driver stage, and coupling capacitively to the opposite end from the plate. This method places the plate-to-filament capacitance of the driver across one-half of the tank and the grid-to-filament capacitance of the following stage across the other half, as shown in figure 58.

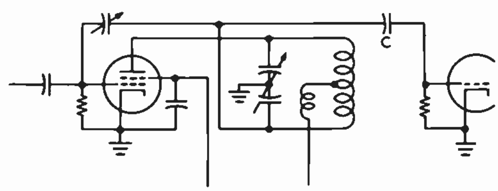


Figure 58

BALANCED CAPACITIVE COUPLING

Balanced capacitive coupling sometimes is useful when it is desirable to use a relatively large inductance in the interstage tank circuit, or where the exciting stage is neutralized as shown above.

Capacitive coupling, generally speaking, does not provide a high degree of attenuation to harmonics of the driving signal and its use (particularly in driver chains for vhf equipment) should be tempered caution.

Inductive Coupling *Inductive coupling* (figure 59) results when two coils are electromagnetically coupled to one another. The degree of coupling is controlled by varying the mutual inductance of the two coils, which is accomplished by changing the spacing or the relationship between the axes of the coils.

Inductive coupling is used extensively for coupling r-f amplifiers in radio receivers and in vhf exciters to attenuate harmonics and subharmonics of the signal frequency.

Unity Coupling If the grid-tuning capacitor of figure 59 is removed and the coupling increased to the maximum practicable value by interwinding the turns of the two coils, the circuit insofar as r.f. is concerned, acts like that of figure 57, in which one tank serves both as plate tank for the driver and grid tank for the driven stage. The interwound grid winding serves simply to isolate the d-c plate voltage of the driver from the grid of the driven stage, and to provide a return for d-c grid current. This type of coupling is known as *unity coupling*. Because of the high mutual inductance, both primary and secondary are resonated by the one tuning capacitor.

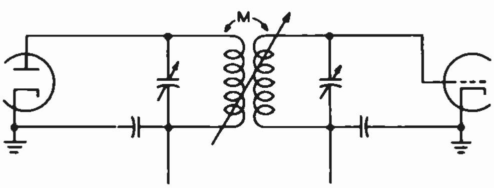


Figure 59

INDUCTIVE INTERSTAGE COUPLING

Link Coupling A modified form of inductive coupling termed *link coupling* is often used in transmitting equipment when two stages are separated by a distance comparable to a fraction of the operational wavelength. A low-impedance r-f transmission line couples the two tuned circuits together. Each end of the line is terminated in one or more turns of wire, or links, wound around the coils which are being coupled together. These links should be coupled to each tuned circuit at the point of zero r-f potential, or *nodal point*. A ground connection to one side of the link usually is used to reduce harmonic coupling, or where capacitive coupling between two circuits must be minimized. Coaxial line is commonly used to transfer energy between the two coupling links (figure 60).

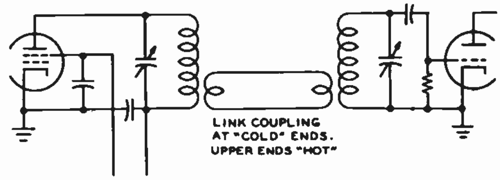


Figure 60

INTERSTAGE COUPLING BY MEANS OF A LINK

Link interstage coupling is very commonly used since the two stages may be separated by a considerable distance, since the amount of a coupling between the two stages may be easily varied, and since the capacitances of the two stages may be isolated to permit use of larger inductances in the vhf range.

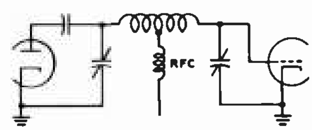


Figure 61

PI-NETWORK INTERSTAGE COUPLING

Network Coupling The L-, pi-, or pi-L network may be used as an interstage coupling device providing a high degree of harmonic attenuation. The pi-network (figure 61) is very effective in harmonic reduction when the output capacitor is connected directly across the input terminals of the amplifier stage, providing a direct capacitive shunt to ground for harmonic voltages. The network may be either a step-up or step-down design depending on the relative impedance levels of the driver output and amplifier input circuits.

Part II—VHF Circuits

The representative circuits discussed in the first part of this chapter apply equally as well to the vhf portion of the spectrum as they do to the lower frequencies. At the very-high frequencies, however, the clear distinction between external lumped circuit parameters and the amplifying device becomes indistinct and different design techniques are required to achieve proper circuit and tube efficiency.

11-17 Vacuum-Tube Limitations

The vacuum tube becomes progressively less efficient as the frequency of operation is raised requiring more drive power for a given power output level. At the same time, the input impedance of the tube drops as does the maximum impedance realizable in the plate circuit. *Lead inductance* of tube and socket create undesirable r-f voltage drops so that the available driving voltage does not appear across the tube elements (figure 1A). In addition, the interelectrode capacitance of the tube approaches a large fraction of the capacitance required to establish circuit resonance with the result that the tank circuit may "disappear" within the tube (figure 1B). The combination of lead inductance and interelectrode capacitance of the tube will cause an internal resonance in the upper vhf region, possibly leading to parasitic oscillation and instability.

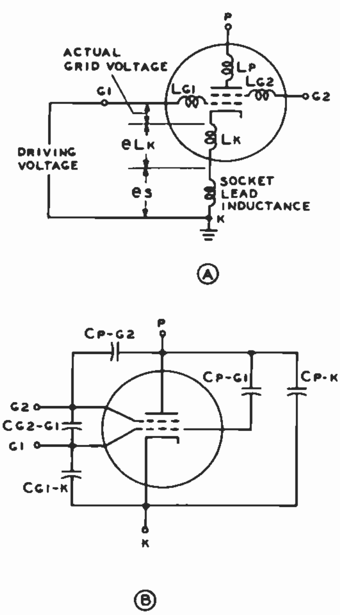


Figure 1

LEAD INDUCTANCE AND INTERNAL CAPACITANCE

A—Interelectrode capacitances of the tube may approach a large fraction of the capacitance required to establish circuit resonance. B—Lead inductance of the tube and socket creates voltage drops so that only a portion of the drive voltage appears between grid and cathode.

Cathode Lead Inductance Tube gain is adversely affected by *cathode lead inductance* which, in conjunction with grid-cathode capacitance, causes a resistive load to appear across the input of the tube. This load results from a voltage drop across the cathode lead inductance which drives the cathode as in a grounded-grid amplifier stage. A portion of the drive signal thus appears in the output circuit (termed *feedthrough* power) which must be supplied by the driving stage. As the frequency of operation is raised, input loading due to cathode lead inductance rises, roughly as the square of the increase in frequency. Thus, input loading is nine times as great at 432 MHz as it is at 144 MHz for a given tube.

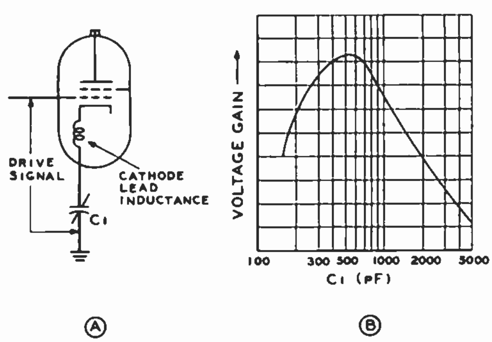


Figure 2

CATHODE LEAD INDUCTANCE

A—Cathode lead inductance is neutralized by series-resonant cathode circuit. B—Voltage gain of the tube may be peaked by adjustment of cathode bypass capacitor.

The cathode lead inductance may be neutralized by choosing a value of cathode bypass capacitance such that the total lead inductance (tube, socket, and stray circuit inductance) is approximately series-resonant at the operating frequency, as shown in figure 2.

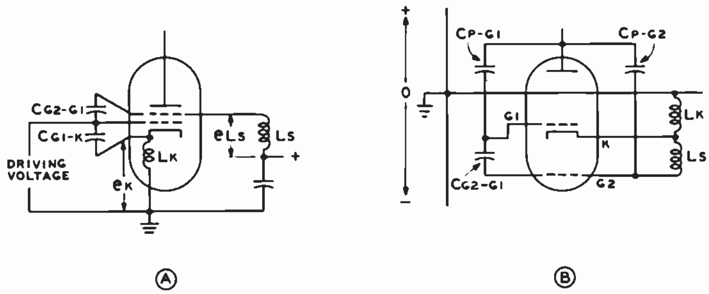


Figure 3

VHF SCREEN NEUTRALIZATION

A—Cathode lead inductance may be neutralized by placing inductance in series with screen-to-ground circuit. B—Cathode and screen lead inductances form bridge with grid-to-screen and grid-to-plate capacitances. Bridge balance places grid and cathode at same voltage level as far as internal feedback is concerned. Bridge is balanced by adjustment of screen inductor L_s.

Cathode lead inductance may also be neutralized by placing an inductance (L_K) in series with the screen-to-ground circuit as shown in figure 3 or by utilizing the grid structure of the tube as a screen and placing the exciting signal on the cathode (figure 4). The cathode lead inductance is now a part of the input tuned circuit and the *grid lead inductance* (while having a voltage drop across it) usually is of much smaller magnitude than cathode lead inductance in a well designed vhf tube.

The grid lead inductance can either cause instability and a loss of drive voltage or it may provide a method of neutralizing the amplifier, as discussed in the previous part of this chapter.

Screen Lead Inductance *Screen lead inductance* may help or hinder the operation of the tube. Below the self-neutralizing frequency of the tube (see Part I, Section 11-8) screen lead inductance is detrimental to amplifier stability as r-f current flowing through the inductance will cause an unwanted r-f voltage to be developed on the screen element. At operating frequencies above the self-neutralizing frequency, a variable screen-bypass capacitor is sometimes added to allow the self-neutralizing frequency to be moved up to the operating frequency.

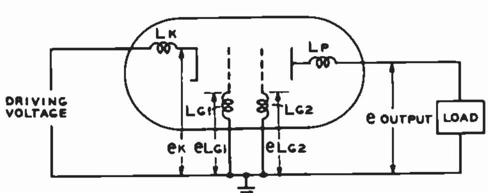
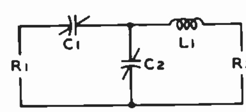


Figure 4

CATHODE DRIVEN VHF AMPLIFIER

Cathode lead inductance is part of the input circuit and a degenerative signal now appears across grid-to-ground inductance. Grid inductance (L_G) may be used for neutralization of the stage when proper shift is present.

Input Capacitance The *input capacitance* of a grid-driven tetrode is the sum of the grid-cathode and grid-



$R_1 > R_2$

$X_{L1} = Q_L R_2$

$X_{C1} = R_1 \sqrt{\dfrac{R_2(Q_L^2+1)}{R_1} - 1}$

$X_{C2} = \dfrac{R_2(Q_L^2+1)}{Q_L} \cdot \dfrac{1}{1-\left(\dfrac{X_{C1}}{Q_L R_1}\right)}$

Q_L = LOADED Q OF NETWORK

Figure 5

T-NETWORK FOR CATHODE-DRIVEN AMPLIFIER

Simple T-network can be used for step-down or step-up transformation between cathode impedance and nominal 50-ohm termination. In this circuit, R_1 is greater than R_2. Network Q of 2 to 5 is commonly used.

screen capacitances. The larger the input capacitance the lower the reactance and the greater the exciting current needed to charge the capacitance. The driving stage must supply the current to charge this capacitance. Stray input capacitance external to the tube must be held to the minimum value, and peak driving voltage should be limited by operating with low bias to reduce the effects of charging current and accompanying waste of drive power. The charging current can cause heating of the tube seals and expansion and detuning of the resonant circuits.

The cathode-driven amplifier has a lower input capacitance for a given tube than the grid-driven equivalent since the input capacitance consists only of the cathode-grid capacitance, and its use is widespread in vhf equipment.

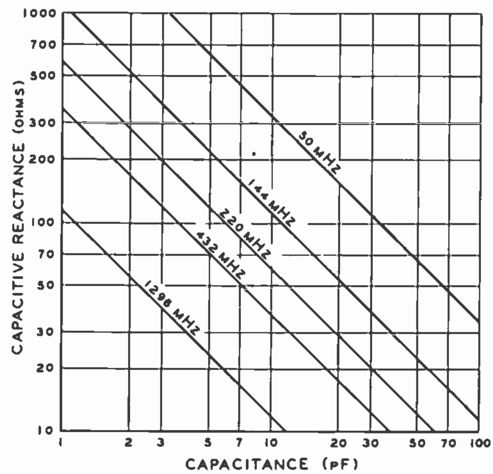


Figure 6

REACTANCE CHART FOR VHF BANDS

Feedback Capacitance The *feedback capacitance* in a grid-driven amplifier is the grid-plate capacitance of the tube, which becomes a larger factor in circuit design as the frequency of operation is raised. The cathode-driven amplifier minimizes feedback capacitance since the cathode-plate capacitance is usually quite small in most vhf tetrode tubes, with the grid (or grids) shielding the output from the input circuit.

Regardless of circuitry, the higher the operating frequency is, the greater are the chances for amplifier instability due to r-f feedback from the output through the feedback capacitance of the tube to the input circuit.

Circuit and Tube Losses The power losses associated with tube and circuit all tend to increase with frequency. In the vhf region all r-f current flows in the surface layers of a conductor because of *skin effect.* Resistance and r-f losses in a conductor increase with the square root of the frequency, since the layer in which the current flows decreases in thickness as the frequency of operation increases. Additional circuit losses will accrue due to *radiation of energy* from wires and components carrying r-f current. The power radiated from a short length of conductor increases as the square of the frequency.

Dielectric loss within insulating supports in the tube and in external circuitry increases directly with frequency and is due to the molecular movements produced within the dielectric by the electric field. Both dielectric and radiation loss contribute to a general reduction tube and circuit efficiency as the frequency of operation is raised.

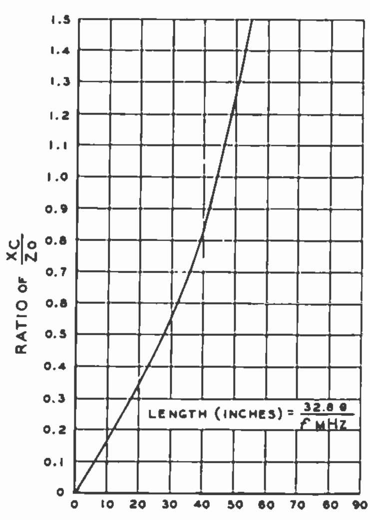


Figure 7

ELECTRICAL LENGTH OF LINE AS FUNCTION OF X_C/Z_O

Transit-Time Effect *Transit time* is the finite time an electron takes in passing from the cathode to the grid of a tube and is a function of the grid-to-cathode spacing and grid-to-cathode voltage, increasing as the frequency of operation is increased. If transit time is an appreciable fraction of one operating cycle, electrons in transit will be "out of step" with instantaneous grid potential, and the resulting plate current pulses are not as sharp and defined as the current pulses liberated from the cathode. This increases the conduction angle of operation and reduces the plate efficiency of the tube.

11-18 Input and Output

Single-ended vhf amplifiers make use of linear versions of parallel-tuned or network circuits in the input and output configurations. A practical and simple input circuit

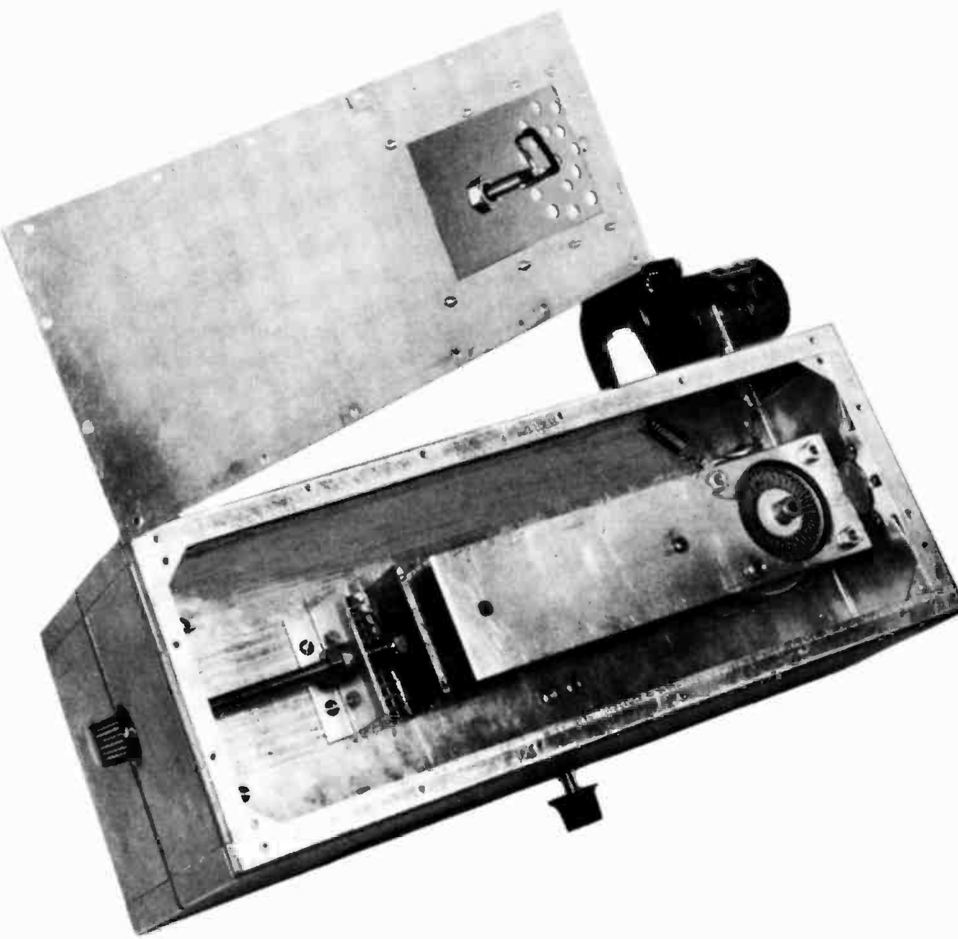


Figure 8

HALF-WAVELENGTH STRIP-LINE PLATE CIRCUIT

Tuning capacitor is placed at the high-impedance end of the line away from the tube. Inductive output coupling loop is placed at a low-impedance on the line, near the center.

for a cathode-driven amplifier is the version of the T-network shown in figure 5. For the lower portion of the vhf region the network can be made up of lumped constants.

The output circuitry, in addition to matching the tube to the transmission line may also be called upon to dissipate the anode heat of the tube. In order to do this, and to prevent rapid detuning of the circuit with rising temperature, the circuit Q should be as low as practicable. The strip-line technique (see chapter 10, part II) is often used since it provides a large thermal capacity and requires a minimum of machine work, as compared to a coaxial cavity.

The strip line (or cavity) can operate in the ¼-, ½- or ¾-wave mode, with increasing loaded Q, increasing impedance, and decreasing bandwith as the electrical length is increased. The impedance of the output circuit is limited by tube and stray output capacitance:

$$X_c = Z_o \times \tan l$$

where,

X_c = tube and stray output capacitance,

Z_o = characteristic impedance of line or cavity,

l = length of line or cavity in electrical degrees.

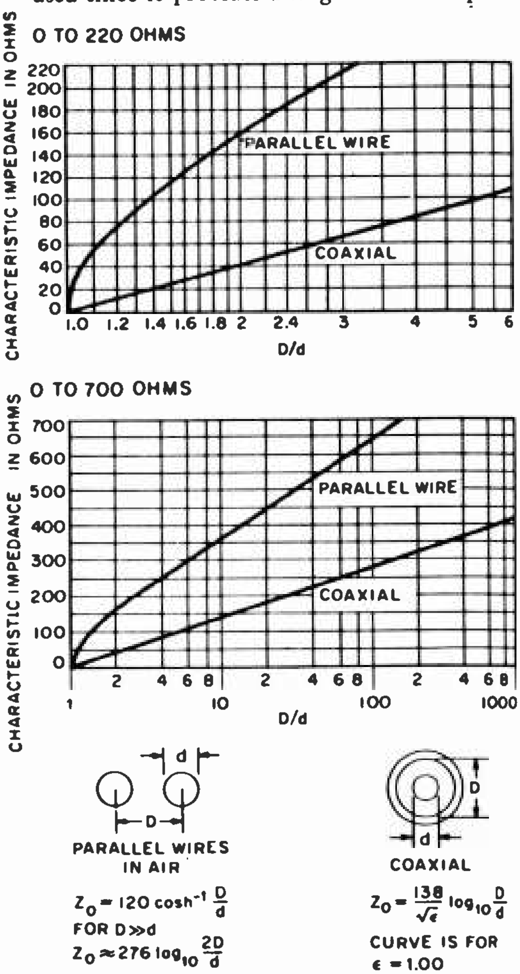


Figure 9

CHARACTERISTIC IMPEDANCE OF PARALLEL AND COAXIAL LINES HAVING AIR DIELECTRIC

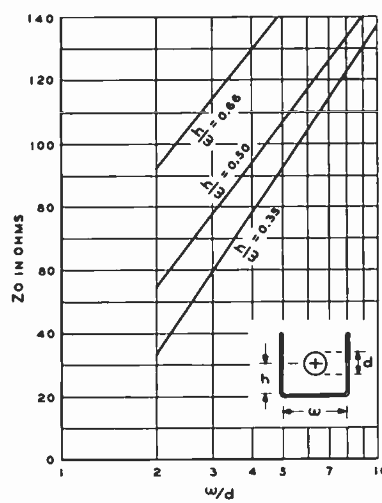
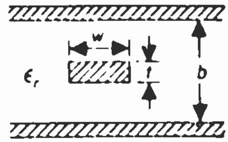


Figure 10

CHARACTERISTIC IMPEDANCE OF OPEN TROUGH LINE FOR VARIOUS HEIGHT TO WIDTH RATIOS

For minimum loaded Q and greatest bandwidth, the ratio X_c/Z_o should approximate 0.5 for a quarter-wave circuit and 0.83 for a half-wave or three quarter-wave circuit.

Strip-line or coaxial circuit design may be aided by the charts of figures 6 and 7. For example, a 3CX1000A7 high-mu triode in grounded-grid configuration has an average output capacitance (plate-to-grid) of 15 pF. Circuit stray and tuning capacitance are estimated to total 15 pF. At 144 MHz, X_c



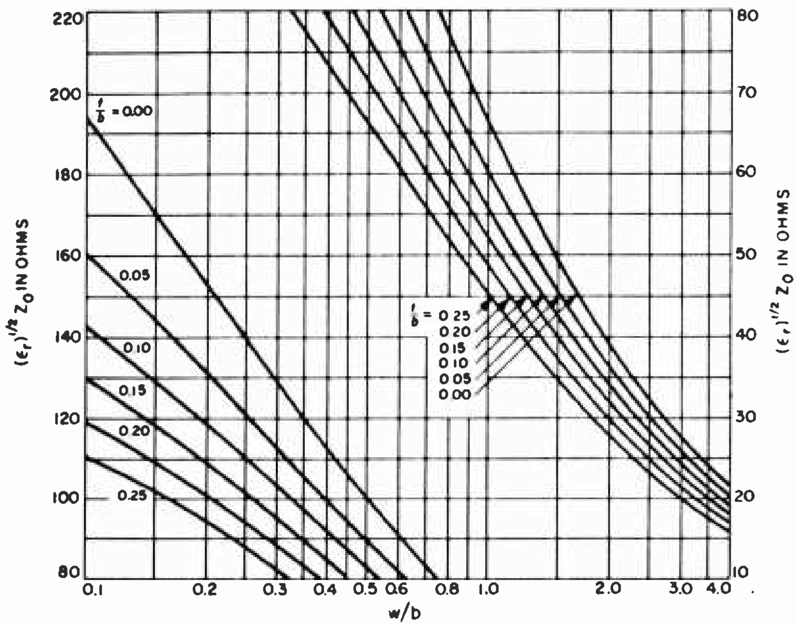


Figure 11

PLOT OF STRIP TRANSMISSION LINE Z_0 VERSUS *w/b* FOR VARIOUS VALUES OF *t/b*.

For lower left family of curves, refer to left-hand ordinate values; for upper right curves, use right-hand scale.

is about 35 ohms for the total value of 30 pF. For an X_c/Z_0 ratio of 0.5 and given the X_c value of 35 ohms the line impedance should be about 70 ohms. From figure 7, the point $X_c/Z_0 = 35$ is found and the line length noted to be 27 electrical degrees, or about 6⅛ inches. This is the total physical length of the strip line and includes the path through the tube anode cooler and tuning capacitor. If this short a line poses coupling problems, the experimenter may go to a longer half wavelength line, with the attendant problems of increased circuit Q for the longer length.

The line, in any event, resonates with a fixed value of capacitance and decreasing line impedance increases the electrical length, whereas increasing line impedance decreases the electrical length.

The Half-Wavelength Line The half-wavelength line or cavity is useful when the capacitance of the tube is appreciable and the use of a quarter-wavelength line placed the low impedance end of the line close to the tube socket terminals. A single ended, half-wave strip-line circuit is shown in figure 8 with the tuning adjustment placed at the high-impedance end of the line at the point of low impedance and minimum r-f voltage. The whole circuit, including the output capacitance of the tube, becomes an electrical half wavelength, capacitively loaded at one end by the tube, and at the other by the tuning capacitor.

Though plate circuitry is shown in these examples, the principle applies equally well to grid circuitry.

Tank Circuit Impedance The characteristic impedance of the transmission line making up the resonant tank circuit must be known in order to determine the physical attributes of the configuration. The characteristic impedance of parallel and coaxial lines having an air dielectric are given in figure 9. The impedance of an open *trough line* having height to width ratios of 0.33, 0.50 and 0.66 may be determined from the graph of figure 10. The characteristic impedance of a strip line having various height to width ratios can be calculated with the aid of the nomograph of figure 11.

CHAPTER TWELVE

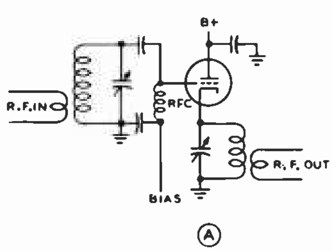
R-F Feedback

Comparatively high gain is required in single-sideband equipment because the signal is usually generated at levels of one watt or less. To get from this level to a kilowatt requires about 30 db of gain. High gain tetrodes may be used to obtain this increase with a minimum number of stages and circuits. Each stage contributes some distortion; therefore, it is good practice to keep the number of stages to a minimum. It is generally considered good practice to operate the low-level amplifiers below their maximum power capability in order to confine most of the distortion to the last two amplifier stages. *R-f feedback* can then be utilized to reduce the distortion in the last two stages. This type of feedback is no different from the common audio feedback used in high-fidelity sound systems. A sample of the output waveform is applied to the amplifier input to correct the distortion developed in the amplifier. The same advantages can be obtained at radio frequencies that are obtained at audio frequencies when feedback is used.

12-1 R-F Feedback Circuits

R-f feedback circuits have been developed by the *Collins Radio Co.* for use with linear amplifiers. Tests with large receiving and small transmitting tubes showed that amplifiers using these tubes without feedback developed signal-to-distortion ratios no better than 30 db or so. Tests were run employing cathode-follower circuits, such as shown in figure 1A. Lower distortion was achieved, but at the cost of low gain per stage. Since the voltage gain through the tube is less than unity, all gain has to be achieved by voltage step-up in the tank circuits. This gain is limited by the dissipation of the tank coils, since the circuit capacitance across the coils in a typical transmitter is quite high. In addition, the tuning of such a stage is sharp because of the high-Q circuits.

The cathode-follower performance of the tube can be retained by moving the r-f ground



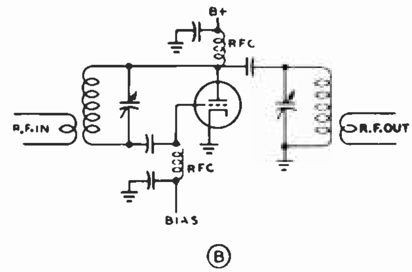


Figure 1
SIMILAR CATHODE FOLLOWER CIRCUITS HAVING DIFFERENT R-F GROUND POINTS

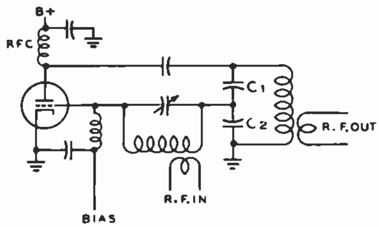


Figure 2
SINGLE STAGE AMPLIFIER WITH R-F FEEDBACK CIRCUIT

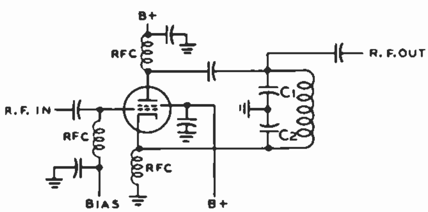


Figure 3
SINGLE STAGE FEEDBACK AMPLIFIER WITH GROUND RETURN POINT MODIFIED FOR UNBALANCED INPUT AND OUTPUT CONNECTIONS

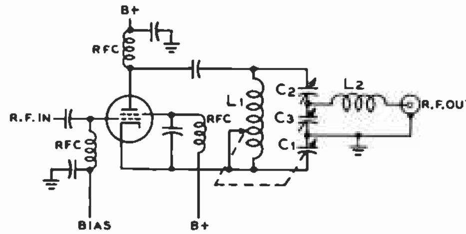


Figure 4
R-F AMPLIFIER WITH FEEDBACK AND IMPEDANCE MATCHING OUTPUT NETWORK

Tuning and loading are accomplished by C_2 and C_3. C_1 and L_1 are tuned in unison to establish the correct degree of feedback.

point of the circuit from the plate to the cathode as shown in figure 1B. Both ends of the input circuit are at high r-f potential so inductive coupling to this type of amplifier is necessary.

Inspection of figure 1B shows that by moving the top end of the input tank down on a voltage-divider tap across the plate tank circuit, the feedback can be reduced from 100%, as in the case of the cathode-follower circuit, down to any desired value. A typical feedback circuit is illustrated in figure 2. This circuit is more practical than those of figure 1, since the losses in the input tank are greatly reduced. A feedback level of 12 db may be achieved as a good compromise between distortion and stage gain. The voltage developed across C_2 will be three times the grid-cathode voltage. Inductive coupling is required for this circuit, as shown in the illustration.

The circuit of figure 3 eliminates the need for inductive coupling by moving the r-f ground to the point common to both tank circuits. The advantages of direct coupling between stages far outweigh the disadvantages of having the r-f feedback voltage appear on the cathode of the amplifier tube.

In order to match the amplifier to a load, the circuit of figure 4 may be used. The ratio of X_{L1} to X_{C1} determines the degree of feedback, so it is necessary to tune them in unison when the frequency of operation is changed. Tuning and loading functions are accomplished by varying C_2 and C_3. L_2 may also be varied to adjust the loading.

Feedback Around a Two-Stage Amplifier The maximum phase shift obtainable over two simple tuned circuits does not exceed 180 degrees, and feedback around a two-stage amplifier is possible. The basic circuit of a two stage feedback amplifier is shown in figure 5. This circuit is a conventional two-stage tetrode amplifier except that r.f. is fed back from the plate circuit of the PA tube to the cathode of the driver tube. This will reduce the distortion

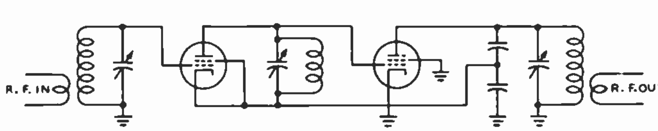


Figure 5
BASIC CIRCUIT OF TWO-STAGE AMPLIFIER WITH R-F FEEDBACK

Feedback voltage is obtained from a voltage divider across the output circuit and applied directly to the cathode of the first tube. The input tank circuit is thus outside the feedback loop.

of both tubes as effectively as using individual feedback loops around each stage, yet will allow a higher level of over-all gain. With only two tuned circuits in the feedback loop, it is possible to use 12 to 15 db of feedback and still leave a wide margin for stability. It is possible to reduce the distortion by nearly as many db as are used in feedback. This circuit has two advantages that are lacking in the single-stage feedback amplifier. First, the filament of the output stage can now be operated at r-f ground potential. Second, any conventional pi output network may be used.

R-f feedback will correct several types of distortion. It will help correct distortion caused by poor power supply regulation, too low grid bias, and limiting on peaks when the plate voltage swing becomes too high.

Neutralization and R-F Feedback The purpose of neutralization of an r-f amplifier stage is to balance out effects of the grid-plate capacitance coupling in the amplifier. In a conventional amplifier using a tetrode tube, the effective input capacity is given by:

$$\text{Input capacitance} = C_{in} + C_{gp}\,(1 + A\cos\theta)$$

where,

C_{in} equals tube input capacitance,
C_{gp} equals grid-plate capacitance,
A equals grid-to-plate voltage amplification,
θ equals angle of load.

In a typical unneutralized tetrode amplifier having a stage gain of 33, the input capacitance of the tube with the plate circuit in resonance is increased by 8 pf due to the unneutralized grid-plate capacitance. This is unimportant in amplifiers where the gain (A) remains constant but if the tube gain varies, serious detuning and r-f phase shift may result. A grid or screen modulated r-f amplifier is an example of the case where the stage gain varies from a maximum down to zero. The gain of a tetrode r-f amplifier operating below plate current saturation varies with loading so that if it drives a following stage into grid current the loading increases and the gain falls off.

The input of the grid circuit is also affected by the grid-plate capacitance, as shown in this equation:

$$\text{Input resistance} = \frac{1}{2\pi f \times C_{gp}\,(A\sin\theta)}$$

This resistance is in shunt with the grid current loading, grid tank circuit losses, and driving source impedance. When the plate circuit is inductive there is energy transferred from the plate to the grid circuit (positive feedback) which will introduce negative resistance in the grid circuit. When this shunt negative resistance across the grid circuit is lower than the equivalent positive resistance of the grid loading, circuit losses, and driving source impedance, the amplifier will oscillate.

When the plate circuit is in resonance (phase angle equal to zero) the input resistance due to the grid-plate capacitance becomes infinite. As the plate circuit is tuned to the capacitive side of resonance, the input resistance becomes positive and power is actually transferred from the grid to the plate circuit. This is the reason that the grid current in an unneutralized tetrode r-f amplifier varies from a low value with the plate circuit tuned on the low-frequency side of resonance to a high value on the high-frequency side of resonance The grid current is proportional to the r-f voltage on the grid which is varying under these conditions. In a tetrode class-AB_1 amplifier, the effect of grid-plate feedback can be observed by placing a r-f voltmeter across the grid circuit and observing the voltage change as the plate circuit is tuned through resonance.

If the amplifier is over-neutralized, the effects reverse so that with the plate circuit tuned to the low-frequency side of resonance, the grid voltage is high, and on the high-frequency side of resonance, it is low.

Amplifier Neutralization Check A useful "rule of thumb" method of checking neutralization of an amplifier stage (assuming that it is nearly correct to start with) is to tune both grid and plate circuits to resonance. Then, observing the r-f grid current, tune the plate circuit to the high-frequency side of resonance. If the grid current rises, more neutralization capacitance is required. Conversely, if the grid current decreases, less capacitance is needed. This indication is very sensitive in a neutralized triode amplifier, and correct neutralization exists when the grid current peaks at the point of plate current dip. In tetrode power amplifiers this indication is less pronounced. Sometimes in a supposedly neutralized tetrode amplifier, there is practically no change in grid voltage as the plate circuit is tuned through resonance, and in some amplifiers it is unchanged on one side of resonance and drops slightly on the other side. Another observation sometimes made is a small dip in the center of a broad peak of grid current. These various effects are probably caused by

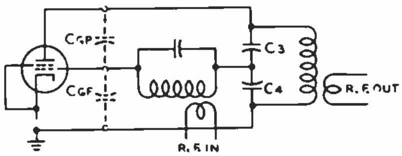


Figure 6
SINGLE STAGE R-F AMPLIFIER WITH FEEDBACK RATIO OF C_3/C_4 to C_{gp}/C_{gf} DETERMINES STAGE NEUTRALIZATION

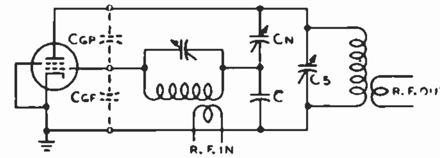


Figure 7
NEUTRALIZED AMPLIFIER AND INHERENT FEEDBACK CIRCUIT
Neutralization is achieved by varying the capacity of C_N.

coupling from the plate to the grid circuit through other paths which are not balanced out by the particular neutralizing circuit used.

Feedback and Neutralization of a One-Stage R-F Amplifier Figure 6 shows an r-f amplifier with negative feedback. The voltage developed across C_4 due to the divider action of C_3 and C_4 is introduced in series with the voltage developed across the grid tank circuit and is in phase-opposition to it. The feedback can be made any value from zero to 100% by properly choosing the values of C_3 and C_4.

For reasons stated previously, it is necessary to neutralize this amplifier, and the relationship for neutralization is:

$$\frac{C_3}{C_4} = \frac{C_{gp}}{C_{gf}}$$

It is often necessary to add capacitance from plate to grid to satisfy this relationship

Figure 7 is identical to figure 6 except that it is redrawn to show the feedback inherent in this neutralization circuit more clearly. C_N and C replace C_3 and C_4, and the main plate tank tuning capacitance is C_5. The circuit of figure 7 presents a problem in coupling to the grid circuit. Inductive coupling is ideal, but the extra tank circuits complicate the tuning of a transmitter which uses several cascaded amplifiers with feedback around each one. The grid could be coupled to a high source impedance such as a tetrode plate, but the driver then cannot use feedback because this would cause the source impedance to be low. A possible solution is to move the circuit ground point from the cathode to the bottom end of the grid tank circuit. The feedback voltage then appears between the cathode and ground (figure 8). The input can be capacitively coupled, and the plate of the amplifier can be capacitively coupled to the next stage. Also, cathode type transmitting tubes are available that allow the heater to remain at ground potential when r.f. is impressed upon the cathode. The output voltage available with capacity coupling, of course, is less than the plate-cathode r-f voltage developed by the amount of feedback voltage across C_4.

12-2 Feedback and Neutralization of a Two-Stage R-F Amplifier

Feedback around two r-f stages has the advantage that more of the tube gain can be realized and nearly as much distortion reduction can be obtained using 12 db around two stages as is realized using 12 db around each of two stages separately. Figure 9 shows a basic circuit of a two-stage feedback amplifier. Inductive output coupling is used, although a pi-network configuration will also work well. The small feedback voltage required is obtained from the voltage divider (C_1-C_2) and is applied to the cathode of the driver tube. C_1 is only a few pf, so this feedback voltage divider may be left fixed for a wide frequency range. If the combined tube gain is 160, and 12 db of feedback is desired, the ratio of C_2 to C_1 is about 40 to 1. This ratio in practice may be 100 pf to 2.5 pf, for example.

A complication is introduced into this simplified circuit by the cathode-grid capacitance

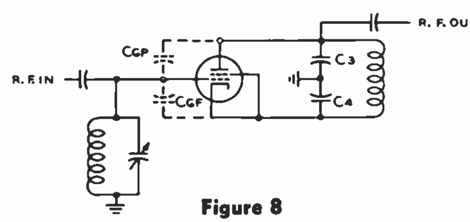


Figure 8

UNBALANCED INPUT AND OUTPUT CIRCUITS FOR SINGLE-STAGE R-F AMPLIFIER WITH FEEDBACK

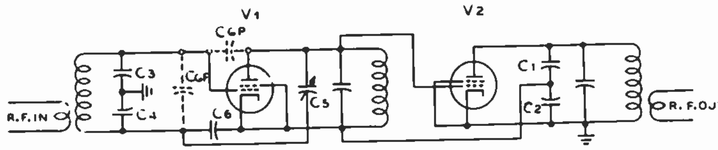


Figure 9

TWO-STAGE AMPLIFIER WITH FEEDBACK

Included is a capacitor (C_4) for neutralizing the cathode-grid capacity of the first tube. V_1 is neutralized by capacitor C_5, and V_2 is neutralized by the correct ratio of C_1/C_2.

of the first tube which causes an. undesired coupling to the input grid circuit. It is necessary to neutralize out this capacitance coupling, as illustrated in figure 9. The relationship for neutralization is:

$$\frac{C_3}{C_4} = \frac{C_{gf}}{C_6}$$

The input circuit may be made unbalanced by making C_4 five times the capacity of C_3. This will tend to reduce the voltage across the coil and to minimize the power dissipated by the coil. For proper balance in this case, C_6 must be five times the grid-filament capacitance of the tube.

Except for tubes having extremely small grid-plate capacitance, it is still necessary to properly neutralize both tubes. If the ratio of C_1 to C_2 is chosen to be equal to the ratio of the grid-plate capacitance to the grid-filament capacitance in the second tube (V_2), this tube will be neutralized. Tubes such as a 4X-150A have very low grid-plate capacitance and probably will not need to be neutralized when used in the first (V_1) stage. If neutralization is necessary, capacitor C_5 is added for this purpose and the proper value is given by the following relationship:

$$\frac{C_{gp}}{C_5} = \frac{C_{gf}}{C_6} = \frac{C_3}{C_4}$$

If neither tube requires neutralization, the bottom end of the interstage tank circuit may be returned to r-f ground. The screen and suppressor of the first tube should then be grounded to keep the tank output capacitance directly across this interstage circuit and to avoid common coupling between the feedback on the cathode and the interstage circuit. A slight amount of degeneration occurs in the first stage since the tube also acts as a grounded grid amplifier with the screen as the grounded grid. The μ of the screen is much lower than that of the control grid so that this effect may be unnoticed and would only require slightly more feedback from the output stage to overcome.

Tests For Neutralization Neutralizing the circuit of figure 9 balances out coupling between the input tank circuit and the output tank circuit, but it does not remove all coupling from the plate circuit to the grid-cathode tube input. This latter coupling is degenerative, so applying a signal to the plate circuit will cause a signal to appear between grid and cathode, even though the stage is neutralized. A bench test for neutralization is to apply a signal to the plate of the tube and detect the presence of a signal in the grid coil by inductive coupling to it. No signal will be present when the stage is neutralized. Of course, a signal could be inductively coupled to the input and neutralization accomplished by adjusting one branch of the neutralizing circuit bridge (C_5 for example) for minimum signal on the plate circuit.

Neutralizing the cathode-grid capacitance of the first stage of figure 9 may be accomplished by applying a signal to the cathode of the tube and adjusting the bridge balance for minimum signal on a detector inductively coupled to the input coil.

Tuning a Two-Stage Feedback Amplifier Tuning the two-stage feedback amplifier of figure 9 is accomplished in an unconventional way because the output circuit cannot be tuned for maximum output signal. This is because the output circuit must be tuned so the feedback voltage applied to the cathode is in-phase with the input signal applied to the first grid. When the feedback voltage is not in-phase, the resultant grid-cathode voltage increases as shown in figure 10. When the output circuit is properly tuned, the resultant grid-cathode voltage on the first tube will be at a minimum, and the voltage on the interstage tuned circuit will also be at a minimum.

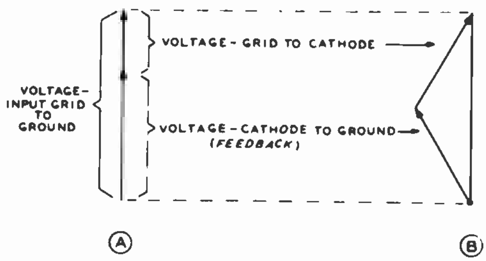


Figure 10
VECTOR RELATIONSHIP OF FEEDBACK VOLTAGE
A = Output Circuit Properly Tuned
B = Output Circuit Mis-Tuned

The two-stage amplifier may be tuned by placing a r-f voltmeter across the interstage tank circuit ("hot" side to ground) and tuning the input and interstage circuits for maximum meter reading, and tuning the output circuit for minimum meter reading. If the second tube is driven into the grid current region, the grid current meter may be used in place of the r-f voltmeter. On high powered stages where operation is well into the class-AB region, the plate current dip of the output tube indicates correct output circuit tuning, as in the usual amplifier.

Parasitic Oscillations in the Feedback Amplifier Quite often low frequency parasitics may be found in the interstage circuit of the two-stage feedback amplifier. Oscillation occurs in the first stage due to low frequency feedback in the cathode circuit. R-f chokes, coupling capacitors, and bypass capacitors provide the low frequency tank circuits. When the feedback and second stage neutralizing circuits are combined, it is necessary to use the configuration of figure 11. This circuit has the advantage that only one capacitor (C_3) is required from the plate of the output tube, thus keeping the added capacitance across the output tank at a minimum.

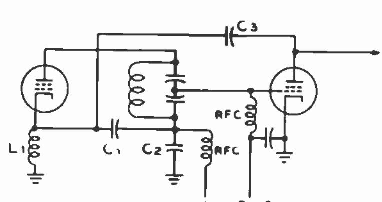


Figure 11
INTERSTAGE CIRCUIT COMBINING NEUTRALIZATION AND FEEDBACK NETWORKS

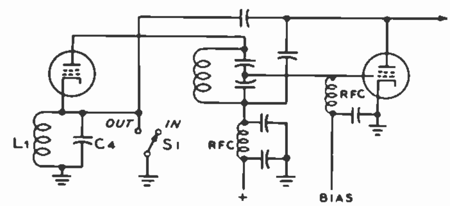


Figure 12
INTERSTAGE CIRCUIT WITH SEPARATE NEUTRALIZING AND FEEDBACK CIRCUITS

It is convenient, however, to separate these circuits so neutralization and feedback can be adjusted independently. Also, it may be desirable to be able to switch the feedback out of the circuit. For these reasons, the circuit shown in figure 12 is often used. Switch S_1 removes the feedback loop when it is closed.

A slight tendency for low-frequency parasitic oscillations still exists with this circuit. L_1 should have as little inductance as possible without upsetting the feedback. If the value of L_1 is too low, it cancels out part of the reactance of feedback capacitor C_1 and causes the feedback to increase at low values of radio frequency. In some cases, a swamping resistor may be necessary across L_1. The value of this resistor should be high compared to the reactance of C_1 to avoid phase-shift of the r-f feedback.

12-3 Neutralization Procedure in Feedback-Type Amplifiers

Experience with feedback amplifiers has brought out several different methods of neutralizing. An important observation is that when all three neutralizing adjustments are correctly made the peaks and dips of various tuning meters all coincide at the point of circuit resonance. For example, the coincident indications when the various tank circuits are tuned through resonance with feedback operating are:

A—When the *PA plate circuit* is tuned through resonance:
1—PA plate current dip
2—Power output peak
3—PA r-f grid voltage dip
4—PA grid current dip
(Note: The PA grid current peaks when feedback circuit is disabled and the tube is heavily driven)

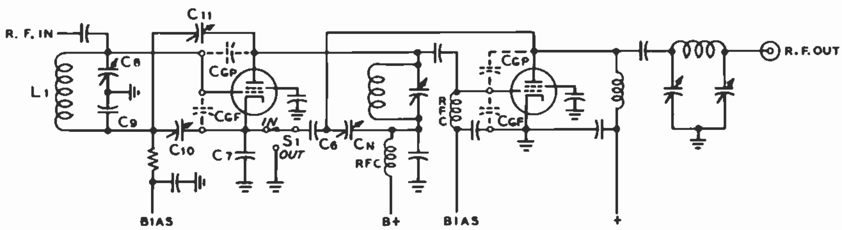


Figure 13
TWO-STAGE AMPLIFIER WITH FEEDBACK CIRCUIT

B—When the *PA grid circuit* is tuned through resonance:
1—Driver plate current dip
2—PA r-f grid voltage peak
3—PA grid current peak
4—PA power output peak

C—When the *driver grid circuit* is tuned through resonance:
1—Driver r-f grid voltage peak
2—Driver plate current peak
3—PA r-f grid current peak
4—PA plate current peak
5—PA power output peak

Four meters may be employed to measure the most important of these parameters. The meters should be arranged so that the following *pairs of readings* are displayed on meters located close together for ease of observation of coincident peaks and dips:

1—PA plate current and power output
2—PA r-f grid current and PA plate current
3—PA r-f grid voltage and power output
4—Driver plate current and PA r-f grid voltage

The third pair listed above may not be necessary if the PA plate current dip is pronounced. When this instrumentation is provided, the neutralizing procedure is as follows:

1—Remove the r-f feedback
2—Neutralize the grid-plate capacitance of the driver stage
3—Neutralize the grid-plate capacitance of the power amplifier (PA) stage
4—Apply r-f feedback
5—Neutralize driver grid-cathode capacitance

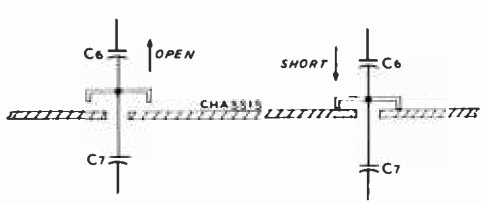


Figure 14
FEEDBACK SHORTING DEVICE

These steps will be explained in more detail in the following paragraphs:

Step 1. The removal of r-f feedback through the feedback circuit must be complete. The switch (S_1) shown in the feedback circuit (figure 13) is one satisfactory method. Since C_6 is effectively across the PA plate tank circuit it is desirable to keep it across the circuit when feedback is removed to avoid appreciable detuning of the plate tank circuit. Another method that can be used if properly done is to ground the junction of C_6 and C_7. Grounding this common point through a switch or relay is not good enough because of common coupling through the length of the grounding lead. The grounding method shown in figure 14 is satisfactory.

Step 2. Plate power and excitation are applied. The driver grid tank is resonated by tuning for a peak in driver r-f grid voltage or driver plate current. The power amplifier grid tank circuit is then resonated and adjusted for a dip in driver plate current. Driver neutralization is now adjusted until the PA r-f grid voltage (or PA grid current) peaks at exactly the point of driver plate current dip. A handy rule for adjusting grid-plate neutralization of a tube without feedback: with all circuits in resonance, detune the plate circuit to the high frequency side of resonance: If grid current to next stage (or power output of the stage under test) increases, more neutralizing capacitance is required and vice versa.

If the driver tube operates class A so that a plate current dip cannot be observed, a dif-

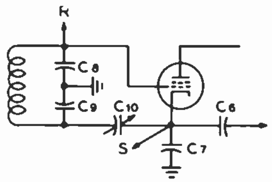


Figure 15
FEEDBACK NEUTRALIZING CIRCUIT USING AUXILIARY RECEIVER

ferent neutralizing procedure is necessary This will be discussed in a subsequent section.

Step 3. This is the same as step 2 except it is applied to the power amplifier stage. Adjust the neutralization of this stage for a peak in power output at the plate current dip.

Step 4. Reverse step 1 and apply the r-f feedback.

Step 5. Apply plate power and an exciting signal to drive the amplifier to nearly full output. Adjust the feedback neutralization for a peak in amplifier power output at the exact point of minimum amplifier plate current. Decrease the feedback neutralization capacitance if the power output rises when the tank circuit is tuned to the high frequency side of resonance.

The above sequence applies when the neutralizing adjustments are approximately correct to start with. If they are far off, some "cut-and-try" adjustment may be necessary. Also, the driver stage may break into oscillation if the feedback neutralizing capacitance is not near the correct setting.

It is assumed that a single-tone test signal is used for amplifier excitation during the above steps, and that all tank circuits are at resonance except the one being detuned to make the observation. There is some interaction between the driver neutralization and the feedback neutralization so if an appreciable change is made in any adjustment the others should be rechecked. It is important that the grid-plate neutralization be accomplished first when using the above procedure, otherwise the feedback neutralization will be off a little, since it partially compensates for that error.

Neutralization Techniques The method of neutralization employing a sensitive r-f detector inductively coupled to a tank coil is difficult to apply in some cases because of mechanical construction of the equipment, or because of undesired coupling. Another method for observing neutralization can be used, which appears to be more accurate in actual practice. A sensitive r-f detector such as a receiver is loosely coupled to the grid of the stage being neutralized, as shown in figure 15. The coupling capacitance is of the order of one or two pf. It must be small enough to avoid upsetting the neutralization when it is removed because the total grid-ground capacitance is one leg of the neutralizing bridge. A signal generator is connected at point S and the receiver at point R. If C_{10} is not properly adjusted the S-meter on the receiver will either kick up or down as the grid tank circuit is tuned through resonance. C_{10} may be adjusted for minimum deflection of the S-meter as the grid circuit is tuned through resonance.

The grid-plate capacitance of the tube is then neutralized by connecting the signal generator to the plate of the tube and adjusting C_{11} of figure 13 for minimum deflection again as the grid tank is tuned through resonance. The power amplifier stage is neutralized in the same manner by connecting a receiver loosely to the grid circuit, and attaching a signal generator to the plate of the tube. The r-f signal can be fed into the amplifier output terminal if desired.

Some precautions are necessary when using this neutralization method. First, some driver tubes (the 6CL6, for example) have appreciably more effective input capacitance when in operation and conducting plate current than when in standby condition. This increase in input capacitance may be as great as three or four pf, and since this is part of the neutralizing bridge circuit it must be taken into consideration. The result of this change in input capacitance is that the neutralizing adjustment of such tubes must be made when they are conducting normal plate current. Stray coupling must be avoided, and it may prove helpful to remove filament power from the preceding stage or disable its input circuit in some manner.

It should be noted that in each of the above adjustments that minimum reaction on the grid is desired, not minimum voltage. Some residual voltage is inherent on the grid when this neutralizing circuit is used.

CHAPTER THIRTEEN

Frequency Modulation and Repeaters

Exciter systems for f-m and single-sideband transmission are basically similar in that modulation of the signal in accordance with the intelligence to be transmitted is normally accomplished at a relatively low level. Then the intelligence-bearing signal is amplified to the desired power level for ultimate transmission. True, amplifiers for the two types of signals are basically different; linear amplifiers of the class-A or class-B type being used for SSB signals, while class-C or nonlinear class-B amplifiers may be used for f-m amplification. But the principle of low-level modulation and subsequent amplification is standard for both types of transmission.

13-1 Frequency Modulation

Early frequency-modulation experiments were conducted by Major Edwin H. Armstrong of Columbia University based on the belief that noise and static were amplitude variations that had no orderly variations in frequency. In 1934 Armstrong conducted his classic f-m transmissions in the old 2½ meter amateur band in conjunction with W2AG in Yonkers, N.Y. Subsequent amateur experiments in 1936 showed that f-m promised excellent prospects for static-free, reliable, mobile communication in the vhf bands.

Postwar vhf development centered around amplitude modulation in the amateur bands for over two decades, aided by the flood of surplus military vhf gear, and it was not until the "mid-sixties" that amateur interest in f-m was stimulated by a quantity of obsolete commercial mobile f-m gear available on the surplus market at modest prices.

Vhf commercial two-way mobile radio is now standardized on channelized frequency-modulation techniques which provide superior rejection to random noise, interference, and fading as compared to conventional a-m systems. When the amplitude of the r-f signal is held constant (limited) and the intelligence transmitted by varying the frequency or phase of the signal, some of the disruptive effects of noise can be eliminated. In addition, audio squelch circuits silence noise peaks and background effects in the receiver until an intelligible signal appears on the frequency. The combination of noise rejection and squelch control provides superior range for a given primary power, as compared to an equivalent a-m power allocation.

Amateur vhf f-m techniques are based on the channel concept. Transmitters and receivers are mainly crystal controlled on a given frequency and random tuning techniques common to the lower frequency amateur bands are absent. F-m channels on the 10-meter band are standardized by common agreement at 20-kHz separation, starting at 29.02 MHz; 20-kHz spearation, starting at 52.54 MHz on the 6-meter band; 30-kHz separation, starting at 146.01 MHz on the 2-meter band; 40-kHz separation, starting at 220.02 MHz on the 220-225 MHz band; and 50-kHz separation, starting at 440.00 MHz on the 420-450 MHz band. National calling frequencies are established on all the vhf bands for emergency communications.

In this chapter various points of difference between frequency-modulation and amplitude-modulation transmission and reception will be discussed and the advantages of frequency modulation for certain types of communication pointed out. Since the distinguishing features of the two types of transmission lie entirely in the modulating circuits at the transmitter and in the detector and limiter circuits in the receiver, these parts of the communication system will receive the major portion of attention.

Modulation *Modulation* is the process of altering a radio wave in accord with the intelligence to be transmitted. The nature of the intelligence is of little importance as far as the process of modulation is concerned; it is the *method*, by which this intelligence is made to give a distinguishing characteristic to the radio wave which will enable the receiver to convert it back into intelligence, that determines the type of modulation being used.

Figure 1 is a drawing of an r-f carrier amplitude-modulated by a sine-wave audio voltage. After modulation the resultant modulated r-f wave is seen still to vary about the zero axis at a constant rate, but the strength of the individual r-f waves is proportional to the amplitude of the modulation voltage.

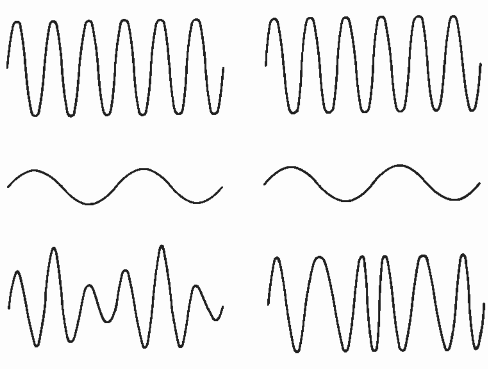


Figure 1 **Figure 2**

A-M AND F-M WAVES

Figure 1 shows a sketch of the scope pattern of an amplitude-modulated wave at the bottom. The center sketch shows the modulating wave and the upper sketch shows the carrier wave.
Figure 2 shows at the bottom a sketch of a frequency-modulated wave. In this case the center sketch also shows the modulating wave and the upper sketch shows the carrier wave. Note that the carrier wave and the modulating wave are the same in either case, but that the waveform of the modulated wave is quite different in the two cases.

In figure 2, the carrier of figure 1 is shown frequency-modulated by the same modulating voltage. Here it may be seen that modulation voltage of one polarity causes the carrier frequency to decrease, as shown by the fact that the individual r-f waves of the carrier are spaced farther apart. A modulating voltage of the opposite polarity causes the frequency to increase, and this is shown by the r-f waves being compressed together to allow more of them to be completed in a given time interval.

Figures 1 and 2 reveal two very important characteristics about amplitude- and frequency-modulated waves. First, it is seen that while the amplitude (power) of the signal is varied in a-m transmission, no such variation takes place in frequency modulation. In many cases this advantage of frequency modulation is probably of equal or greater importance than the widely publicized noise-reduction capabilities of the system. When 100 percent amplitude modulation is obtained, the average power output of the transmitter must be increased by 50 percent. This additional output must be supplied either by the modulator itself, in the high-level system, or by operating one or more of the transmitter stages at such a low output level that they are capable of producing the additional output without distortion in the low-level system. On the other hand, a frequency-modulated transmitter requires an insignificant amount of power from the modulator and needs no provision for increased power output on modulation peaks. All of the stages between the oscillator and the antenna may be operated as high-efficiency class-B or class-C amplifiers or frequency multipliers.

Carrier-Wave Distortion The second characteristic of f-m and a-m waves revealed by figures 1 and 2 is that both types of modulation result in distortion of the r-f carrier. That is, after modulation, the r-f waves are no longer sine waves, as they would be if no frequencies other than the fundamental carrier frequency were present.

It may be shown in the amplitude-modulation case illustrated, that there are only two additional frequencies present, and these are

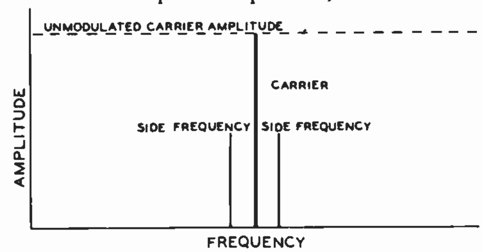


Figure 3

A-M SIDE FREQUENCIES

For each a-m modulating frequency, a pair of side frequencies is produced. The side frequencies are spaced away from the carrier by an amount equal to the modulation frequency, and their amplitude is directly proportional to the amplitude of the modulation. The amplitude of the carrier does not change under modulation.

the familiar *side frequencies*, one located on each side of the carrier, and each spaced from the carrier by a frequency interval equal to the modulation frequency. In regard to frequency and amplitude, the situation is as shown in figure 3. The strength of the carrier itself does not vary during modulation, but the strength of the side frequencies depends on the percentage of modulation. At 100 percent modulation the power in the side frequencies is equal to one-half that of the carrier.

Under frequency modulation, the carrier wave again becomes distorted, as shown in figure 2. But, in this case, many more than two additional frequencies are formed. The first two of these frequencies are spaced from the carrier by the modulation frequency, and the additional side frequencies are located out on each side of the carrier and are also spaced from each other by an amount equal to the modulation frequency. Theoretically, there are an infinite number of side frequencies formed, but, fortunately, the strength of those beyond the frequency *swing* of the transmitter under modulation is relatively low.

One set of side frequencies that might be formed by frequency modulation is shown in figure 4. Unlike amplitude modulation, the strength of the component at the carrier frequency varies widely in frequency modulation and it may even disappear entirely under certain conditions. The variation of strength of the carrier component is useful in measuring the amount of frequency modulation, and will be discussed in detail later in this chapter.

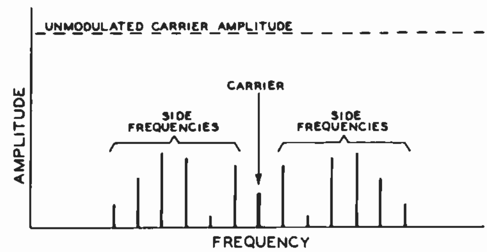


Figure 4

F-M SIDE FREQUENCIES

With frequency modulation, each modulation frequency component causes a large number of side frequencies to be produced. The side frequencies are separated from each other and the carrier by an amount equal to the modulation frequency, but their amplitude varies greatly as the amount of modulation is changed. The carrier strength also varies greatly with frequency modulation. The side frequencies shown represent a case where the deviation each side of the "carrier" frequency is equal to five times the modulating frequency. Other amounts of deviation with the same modulation frequency would cause the relative strengths of the various sidebands to change widely.

One of the great advantages of frequency modulation over amplitude modulation is the reduction in noise at the receiver which the system allows. If the receiver is made responsive only to changes in frequency, a considerable increase in signal-to-noise ratio is made possible through the use of frequency modulation, when the signal is of greater strength than the noise. The noise-reducing capabilities of frequency modulation arise from the inability of noise to cause appreciable frequency modulation of the noise-plus-signal voltage which is applied to the detector in the receiver.

F-M Terms Unlike amplitude modulation, the term *percentage modulation* means little in f-m practice, unless the receiver characteristics are specified. There are, however, three terms, *deviation*, *modulation index*, and *deviation ratio*, which convey considerable information concerning the character of the f-m wave.

Deviation is the amount of frequency

shift each side of the unmodulated carrier frequency which occurs when the transmitter is modulated. Deviation is ordinarily measured in kilohertz, and in a properly operating f-m transmitter it will be directly proportional to the amplitude of the modulating signal. When a symmetrical modulating signal is applied to the transmitter, equal deviation each side of the resting frequency is obtained during each cycle of the modulating signal, and the total frequency range covered by the f-m transmitter is sometimes known as the *swing*. If, for instance, a transmitter operating on 1000 kHz has its frequency shifted from 1000 kHz to 1010 kHz, back to 1000 kHz, then to 990 kHz, and again back to 1000 kHz during one cycle of the modulating wave, the *deviation* would be 10 kHz and the *swing* 20 kHz.

The *modulation index* of an f-m signal is the ratio of the deviation to the audio modulating frequency, when both are expressed in the same units. Thus, in the example above if the signal is varied from 1000 kHz to 1010 kHz to 990 kHz, and back to 1000 kHz at a rate (frequency) of 2000 times a second, the modulation index would be 5, since the deviation (10 kHz) is 5 times the modulating frequency (2 kHz).

The *deviation ratio* is similar to the modulation index in that it involves the ratio between a modulating frequency and deviation. In this case, however, the deviation in question is the peak frequency shift obtained under full modulation, and the audio frequency to be considered is the maximum audio frequency to be transmitted. When the maximum audio frequency to be transmitted is 5000 Hz, for example, a deviation ratio of 3 would call for a peak deviation of 3 × 5000, or 15 kHz at full modulation. The noise-suppression capabilities of frequency modulation are directly related to the deviation ratio. As the deviation ratio is increased, the noise suppression becomes better *if* the signal is somewhat stronger than the noise. Where the noise approaches the signal in strength, however, low deviation ratios allow communication to be maintained in many cases where high-deviation-ratio frequency modulation and conventional amplitude modulation are incapable of giving service. This assumes that a narrow-band f-m receiver is in use. For each value of r-f signal-to-noise ratio at the receiver, there is a maximum deviation ratio which may be used, beyond which the output audio signal-to-noise ratio decreases. Up to this critical deviation ratio, however, the noise suppression becomes progressively better as the deviation ratio is increased.

For high-fidelity f-m broadcasting purposes, a deviation ratio of 5 is ordinarily used, the maximum audio frequency being 15,000 Hz, and the peak deviation at full modulation being 75 kHz. Since a swing of 150 kHz is covered by the transmitter, it is obvious that wide-band f-m transmission must necessarily be confined to the vhf range or higher, where room for the signals is available.

In the case of television sound, the deviation ratio is 1.67; the maximum modulation frequency is 15,000 Hz, and the transmitter deviation for full modulation is 25 kHz. The sound carrier frequency in a standard TV signal is located exactly 4.5 MHz higher than the picture carrier frequency. In the *intercarrier* TV sound system, which is widely used, this constant difference between the picture carrier and the sound carrier is employed within the receiver to obtain an f-m subcarrier at 4.5 MHz. This 4.5 MHz subcarrier then is demodulated by the f-m detector to obtain the sound signal which accompanies the picture.

Narrow-Band F-M Transmission Narrow-band f-m transmission has become standardized for use by the mobile services such as police, fire, and taxicab communications, and is also authorized for amateur work in portions of each of the amateur radiotelephone bands. A maximum deviation of 15 kHz has been standardized for the mobile and commercial communication services, while a maximum deviation of 3 kHz is authorized for amateur nbfm h-f communication. For a maximum audio frequency of 3000 Hz, the maximum deviation ratio is 1.0. For vhf f-m, the deviation ranges from 3 kHz to 15 kHz for a deviation ratio of up to 5.0.

The new channelized f-m concept for amateur communication has standardized on 5 kHz deviation on 10 meters and 6 meters, 5 to 10 kHz deviation on 2 meters, and 15 kHz deviation on the higher vhf bands. F.C.C. amateur regulations limit the band-

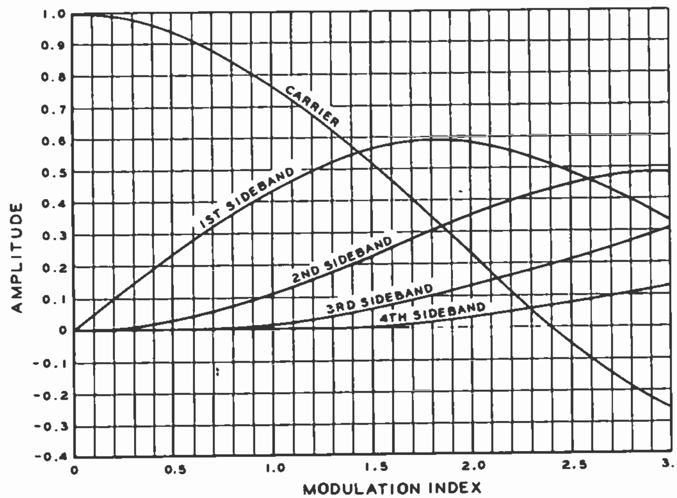


Figure 5

BESSEL CURVES SHOW VARIATION IN CARRIER AND SIDEBAND AMPLITUDE AS MODULATION INDEX IS INCREASED

The carrier and sideband frequency pairs rise and fall with increasing modulation index and pass through zero at certain values. Carrier drops to zero at modulation index of 2.40. The negative amplitude of the carrier above the 2.40 index indicates that the phase is reversed as compared to the phase without modulation.

width of f-m to that of an a-m transmission having the same audio characteristics below 29.0 MHz and in the 50.1 to 52.5 MHz frequency segment. Greater bandwidths are allowed above 29 MHz and above 52.5 MHz.

F-M Sidebands Sidebands are set up when a radio-frequency carrier is frequency modulated. These sidebands differ from those resulting from a-m in that they occur at integral multiples of the modulating frequency; in a-m a single set of sidebands is generated for each modulating frequency. A simple method of determining the amplitude of the various f-m sidebands is the family of *Bessel curves* shown in figure 5. There is one curve for the carrier and one for each pair of sideband frequencies up to the fourth.

The Bessel curves show how the carrier and sideband frequency pairs rise and fall with increasing modulation index, and illustrate the particular values at which they disappear as they pass through zero. If the curves were extended for greater values of modulation index, it would be seen that the carrier amplitude goes through zero at modulation indices of 5.52, 8.65, 11.79, 14.93, etc. The modulation index, therefore, can be measured at each of these points by noting the disappearance of the carrier.

The relative amplitudes of carrier and sideband frequencies for any modulation index can be determined by finding the y-axis amplitude intercept for the particular function. Representative spectrum plots for three different values of modulation index are shown in figure 6. The negative amplitude in the Bessell curves indicate that the phase of the particular function is reversed as compared to the phase without modulation. In f-m, the energy that goes into the sideband frequencies is taken from the carrier; the total power in the over-all composite signal remains the same regardless of the modulation index.

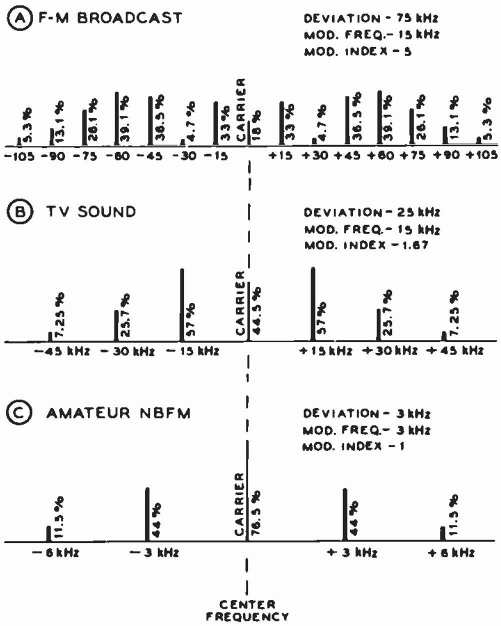


Figure 6

EFFECT OF F-M MODULATION INDEX

Showing the side-frequency amplitude and distribution for the three most common modulation indices used in f-m work. The maximum modulating frequency and maximum deviation are shown in each case.

It might be thought that the large number of side frequencies thus formed might make the frequency spectrum produced by an f-m transmitter prohibitively wide. However, the additional side frequencies are of small amplitude, and, instead of increasing the bandwidth, modulation by a complex wave actually reduces the effective bandwidth of the f-m spectrum. This is especially true when speech modulation is being used, since most of the power of voice sounds is concentrated in the lower audio frequencies.

13-2 Direct F-M Circuits

Frequency modulation may be obtained either by the direct method, in which the frequency of an oscillator is changed directly by the modulating signal, or by the indirect method which makes use of phase modulation. Phase-modulation circuits will be discussed in the following section.

A successful frequency-modulated transmitter must meet two requirements: (1) The frequency deviation must be symmetrical about a fixed frequency, for symmetrical modulation voltage. (2) The deviation must be directly proportional to the amplitude of the modulation, and independent of the modulation frequency. There are several methods of direct frequency modulation which will fullfill these requirements. Some of these methods will be described in the following paragraphs.

Reactance Modulators One of the most practical ways of obtaining direct frequency modulation is through the use of a *reactance modulator.* In this arrangement the modulator output circuit is connected across the oscillator tank circuit, and made to appear as either a capacitive or inductive reactance by exciting the modulator with a voltage which either leads or lags the oscillator tank voltage by 90 degrees. The leading or lagging input voltage causes a corresponding leading or lagging output current, and the output circuit appears as capacitive or inductive reactance across the oscillator tank circuit. When the transconductance of the modulator is varied by varying one of the element voltages, the magnitude of the reactance across the oscillator tank is varied. By applying audio modulating voltage to one of the elements, the transconductance (and hence the frequency) may be varied at an audio rate. When properly designed and operated, the reactance modulator provides linear frequency modulation, and is capable of producing large amounts of deviation.

There are numerous possible configurations of the reactance modulator circuit. The difference in the various arrangements lies principally in the type of phase-shifting circuit used to give a grid voltage which is in phase quadrature with the r-f voltage at the modulator plate (figure 7).

A reactance tube modulator is shown in figure 8. The pentode is plate coupled through a blocking capacitor (C_1) to the "hot" side of the oscillator grid circuit. Another blocking capacitor (C_2) feeds r-f to the phase-shifting network (R-C_3) in the modulator grid circuit. If the resistance of R is made large in comparison with the reactance of C_3 at the oscillator frequency, the current through the R-C_3 combination

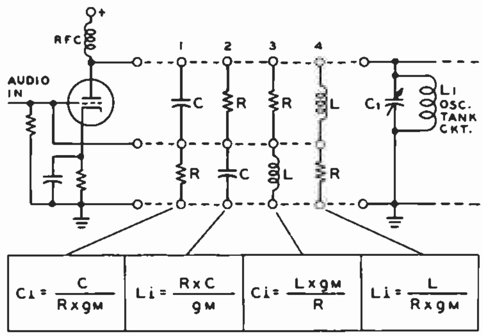


Figure 7

FOUR POSSIBLE LOAD ARRANGEMENTS FOR REACTANCE MODULATOR

will be nearly in phase with the voltage across the tank circuit, and the voltage across C_3 will lag the oscillator tank voltage by almost 90 degrees. The result of the 90-degree lagging voltage on the modulator grid is that its plate current lags the tank voltage by 90 degrees, and the reactance tube appears as an inductance in shunt with the oscillator inductance, thus raising the oscillator frequency.

The phase-shifting capacitor (C_3) can consist of the input capacitance of the modulator tube and stray capacitance between

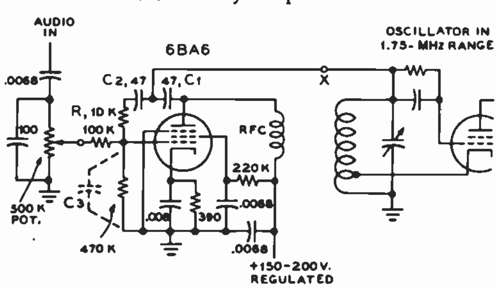


Figure 8

REACTANCE-TUBE MODULATOR

grid and ground. However, better control of the operating conditions of the modulator may be had through the use of a variable capacitor as C_3. Resistance R will usually have a value of between 4700 and 100,000 ohms.

Stabilization Due to the presence of the reactance-tube frequency modulator, the stabilization of an f-m oscillator in regard to voltage changes is considerably more involved than in the case of a simple self-controlled oscillator for transmitter frequency control. If desired, the oscillator itself may be made perfectly stable under voltage changes, but the presence of the frequency modulator destroys the beneficial effect of any such stabilization. It thus becomes desirable to apply the stabilizing arrangement to the modulator as well as the oscillator.

Linearity Test It is almost a necessity to run a static test on the reactance-tube frequency modulator to determine its linearity and effectiveness, since small changes in the values of components, and in stray capacitances will almost certainly alter the modulator characteristics. A frequency-versus-control-voltage curve should be plotted to ascertain that equal increments in control voltage, both in a positive and a negative direction, cause equal changes in frequency. If the curve shows that the modulator has an appreciable amount of non-linearity, changes in bias, electrode voltages, r-f excitation, and resistance values may be made to obtain a straight-line characteristic.

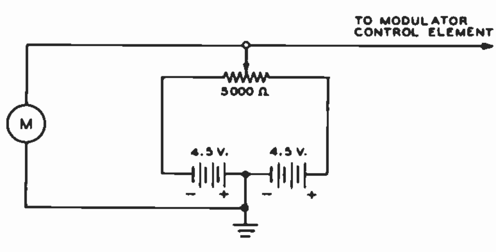


Figure 9

REACTANCE-TUBE LINEARITY CHECKER

Figure 9 shows a method of connecting two 4.5-volt C batteries and a potentiometer to plot the characteristic of the modulator. It will be necessary to use a zero-center voltmeter to measure the grid voltage, or else reverse the voltmeter leads when changing from positive to negative grid voltage. When a straight-line characteristic for the modulator is obtained by the static test method, the capacitances of the various bypass capacitors in the circuit must be kept small to retain this characteristic when an audio voltage is used to vary the frequency in place of the d-c voltage with which the characteristic was plotted.

The Diode Modulator When a resistor and a capacitor are placed in series across an oscillator tank circuit, the current flowing in the series circuit is out of phase with the voltage. If the resistance or capacitance is made variable, the phase difference may be varied. If the variation is controlled at an audio rate, the resultant current can be used to frequency-modulate an oscillator (figure 10). The *diode modulator* may be a vacuum tube acting as a variable resistance or a solid-state voltage-variable capacitor whose capacitance varies inversely as the magnitude of the reverse bias. The variable element is placed in series with a small capacitance across the tank circuit of an oscillator to produce a frequency-modulated signal. The bias voltage applied to the diode should be regulated for best results.

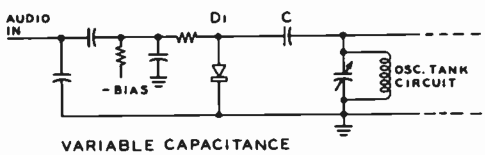


Figure 10

THE DIODE MODULATOR

13-3 Phase Modulation

By means of *phase modulation* (pm) it is possible to dispense with self-controlled oscillators and to obtain directly crystal-controlled frequency modulation. In the final analysis, phase modulation is simply frequency modulation in which the deviation is directly proportional to the modulation frequency. If an audio signal of 1000 Hz causes a deviation of 0.5 kHz, for example, a 2000-Hz modulating signal of the same amplitude will give a deviation of 1 kHz, and so on. To produce an f-m signal, it is necessary to make the deviation independent of the modulation frequency, and proportional only to the modulating signal (figure 11). With phase modulation this is done by including a frequency-correcting network in the transmitter. The audio-correction network must have an attenuation that varies directly with frequency, and this requirement is easily met by a very simple resistance capacitance network.

The only disadvantage of phase modulation, as compared to direct frequency modulation such as is obtained through the use of a reactance-tube modulator, is the fact that very little frequency deviation is produced directly by the phase moduator. The deviation produced by a phase modulator is independent of the actual carrier frequency on which the modulator operates, but is dependent only on the phase deviation which is being produced and on the modulation frequency. Expressed as an equation:

$$F_d = M_p \text{ modulating frequency}$$

where,

F_d is the frequency deviation one way from the mean value of the carrier,

M_p is the phase deviation accompanying modulation expressed in radians (a radian is approximately 57.3°).

Thus, to take an example, if the phase deviation is ½ radian and the modulating frequency is 1000 Hz, the frequency deviation applied to the carrier being passed through the phase modulator will be 500 Hz.

It is easy to see that an enormous amount of multiplication of the carrier frequency is required in order to obtain from a phase modulator the frequency deviation of 75 kHz required for commercial f-m broadcasting. However, for amateur and commercial f-m work only a quite reasonable number of multiplier stages are required to obtain a deviation ratio of approximately one.

Many vhf f-m transmitters employ crystal control with the crystal frequency one twenty-fourth or one thirty-second of the carrier frequency. A deviation of 15 kHz at 144 MHz, for example, is equivalent to a deviation of 0.625 kHz at a crystal frequency of 6 MHz, which is well within the linear capability of a phase modulator. Some high-frequency f-m gear for the 30-MHz region employs crystals in the 200- to 500-kHz region to achieve sufficient frequency

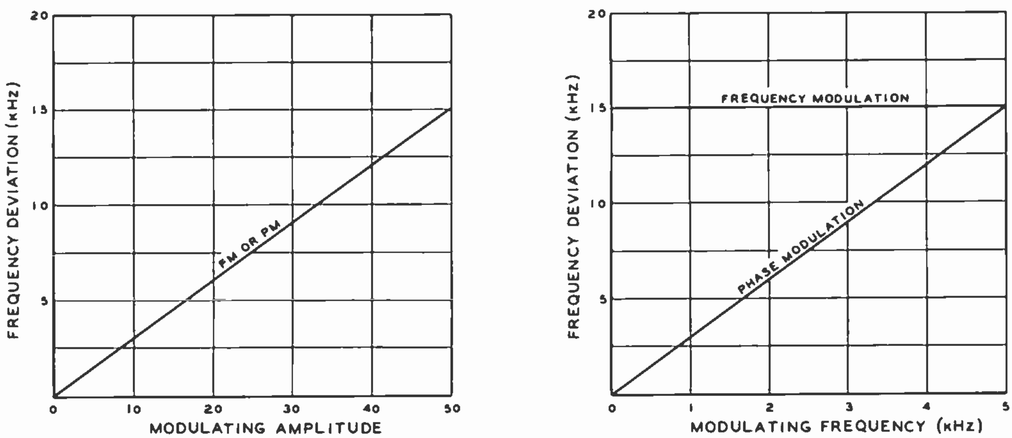


Figure 11

RELATIONSHIP BETWEEN FREQUENCY AND PHASE MODULATION

Frequency deviation is a function of amplitude and frequency of modulating signal for phase modulation (left) and a function of the amplitude only of modulating signal for frequency modulation (right). Most modern f-m transmitters use phase modulation as it may be easily applied to a crystal-controlled circuit.

multiplication for satisfactory phase modulation at the crystal frequency.

Odd-harmonic distortion is produced when frequency-modulation is obtained by the phase-modulation method, and the amount of this distortion that can be tolerated is the limiting factor in determining the amount of phase modulation that can be used. Since the aforementioned frequency-correcting network causes the lowest modulating frequency to have the greatest amplitude, maximum

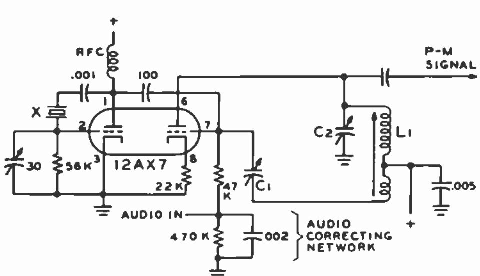


Figure 12

PRACTICAL PHASE MODULATOR

Capacitor C_1 provides adjustment for the phase of the r-f voltage acting between grid and plate of the phase modulator. Capacitor C_2 serves as phase and magnitude control.

phase modulation takes place at the lowest modulating frequency, and the amount of distortion that can be tolerated at this frequency determines the maximum deviation that can be obtained by the p-m method. For high-fidelity broadcasting, the deviation produced by phase modulation is limited to an amount equal to about one-third of the lowest modulating frequency. But for nbfm work the deviation may be as high as 0.6 of the modulating frequency before distortion becomes objectionable on voice modulation. In other terms this means that phase deviations as high as 0.6 radian may be used for amateur and commercial nbfm transmission.

The Phase Modulator A change in the phase of a signal can be produced by passing the signal through a network containing a resistance and a reactance. If the series combination is considered to be the input, and the output voltage is taken from across the resistor, a definite amount of phase shift is introduced, the amount depending on the frequency of the signal and the ratio of the reactance to the resistance. When the resistance is varied with an applied audio signal, the phase angle of the output changes in direct proportion to the audio signal amplitude and produces a phase-modulated signal.

A practical phase modulator is shown in figure 12. The basic RC phase-shift network is replaced by the variable plate resistance of a vacuum tube. The plate resistance of the second section of the double triode changes with grid voltage and therefore serves as the variable resistor element. As the plate resistance of the triode changes with the audio signal applied to the grid circuit, the phase between input and output circuits changes accordingly. A variation of this circuit, in which the transconductance changes with varying input signal, is often used as the basis for a p-m signal.

Capacitor C_1 of the second triode section is not a neutralizing capacitor but is an adjustment for the phase of the r-f voltage acting between grid and plate of the phase modulator. Capacitor C_2 serves as a phase-angle and magnitude control, and both capacitors are adjusted for maximum phase-modulation capability of the circuit. Resonance is established by tuning the slug of coil L_1.

Shown in figure 13 is a simple phase modulator which employs a varactor diode to vary the phase of a tuned circuit. The modulator is installed between the oscillator and the subsequent frequency multiplier stage.

A phase modulator capable of a greater degree of modulation is shown in figure 14. This configuration is often used in vhf crystal-controlled f-m transmitters. In general a FET is used as a crystal oscillator, followed by a second FET as a phase modulator, with the modulating network in the gate circuit. Two inexpensive silicon diodes used as varactors across a phasing coil are driven by the modulating voltage. The r-f output of the 2N5459 is about 30 milliwatts.

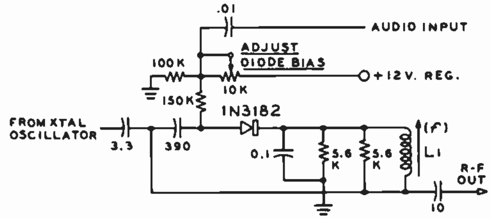


Figure 13

PHASE MODULATOR EMPLOYING VARACTOR DIODE

Audio voltage applied to varactor diode varies the phase of the tuned circuit. Diode bias is adjusted for largest phase shift consistent with linearity.

The F-M Transmitter The various direct and indirect methods of producing f-m involve changing either the frequency or the phase of an r-f carrier in accordance with the modulating signal. The f-m signal is then raised to the operating frequency by passing it through a series of frequency multipliers. When the frequency is multiplied, the frequency deviation is multiplied by a like amount.

Inexpensive and highly stable crystals are available in the 3- to 10-MHz range and many popular f-m transmitters in the vhf region use such crystals, multiplying the crystal frequency by a factor of 12, 18 or 24. Because the amplitude of an f-m signal is constant, the signal may be amplified by nonlinear stages such as doublers and class-C amplifiers without introducing signal distortion. Actually, it is an advantage to pass an f-m signal through nonlinear stages, since any vestige of amplitude modulation generated in the phase modulator may be smoothed out by the inherent limiting action of a class-C amplifier.

Measurement of Deviation When a single-frequency modulating voltage is used with an f-m transmitter the relative amplitudes of the various sidebands and the carrier vary widely as the deviation is varied by increasing or decreasing the amount of modulation. Since the relationship between the amplitudes of the various sidebands and carrier to the audio modulating frequency and the deviation is known, a simple method of measuring the deviation of a frequency-modulated transmitter is possible. In making the measurement, the result is given in the form of the modulation index for a certain amount of audio input.

The measurement is made by applying a sine-wave audio voltage of known frequency to the transmitter, and increasing the modulation until the amplitude of the carrier component of the frequency-modulated wave reaches zero. The modulation index for zero carrier may then be determined from the table below. As may be seen from the table, the first point of zero carrier is obtained when the modulation index has a value of 2.405—in other words, when the deviation is 2.405 times the modulation frequency. For example, if a modulation frequency of 1000 Hz is used, and the modulation is increased until the first carrier null is obtained, the deviation will then be 2.405 times the modulation frequency, or 2.405 kHz. If the modulating frequency happened to be 2000 Hz, the deviation at the first null would be 4.810 kHz. Other carrier nulls will be obtained when the index is 5.52, 8.654, and at increasing values separated approximately by π. The following is a listing of the modulation index at successive carrier nulls up to the tenth:

Zero carrier point no.	Modulation index
1	2.405
2	5.520
3	8.654
4	11.792
5	14.931
6	18.071
7	21.212
8	24.353
9	27.494
10	30.635

The only equipment required for making the measurements is a calibrated audio oscillator of good wave form, and a communication receiver equipped with a narrow passband i-f filter, to exclude sidebands spaced from the carrier by the modulation frequency. The unmodulated carrier is accurately tuned on the receiver. Then modulation from the audio oscillator is applied to the transmitter, and the modulation is increased until the first carrier null is obtained. This carrier null will correspond to a modulation index of 2.405, as previously mentioned. Successive null points will correspond to the indices listed in the table.

A volume indicator in the transmitter audio system may be used to measure the audio level required for different amounts of deviation, and the indicator thus calibrated in terms of frequency deviation. If the measurements are made at the fundamental frequency of the oscillator, it will be necessary to multiply the frequency deviation by the harmonic upon which the transmitter is operating, of course.

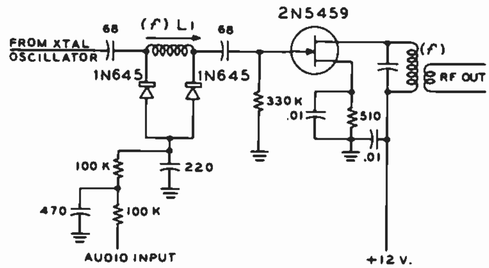


Figure 14

FET PHASE-MODULATED IN GATE CIRCUIT

Two silicon diodes are used as varactors across a phasing coil (L_1). R-f output of 2N5459 is about 30 milliwatts. Circuit permits a small degree of amplitude modulation which is limited out by succeeding stages of f-m exciter.

Modulation Limiting Deviation in an f-m transmitter can be controlled by a circuit that holds the audio level within prescribed limits. Simple audio clipping circuits may be used, as well as more complex deviation control circuits. Diode limiting circuits, such as discussed in Chapter 9 are commonly used, followed by a simple audio filter which removes the harmonics of the clipped audio signal. A representative clipping and filtering circuit is shown in figure 15.

13-4 Reception of F-M Signals

A conventional communications receiver may be used to receive narrow-band f-m transmission, although performance will be much poorer than can be obtained with an nbfm receiver or adapter. However, a receiver specifically designed for f-m reception must be used when it is desired to receive high deviation f-m such as used by f-m broadcast stations, TV sound, and mobile communications.

The f-m receiver must have, first of all, a bandwidth sufficient to pass the range of frequencies generated by the f-m transmit-

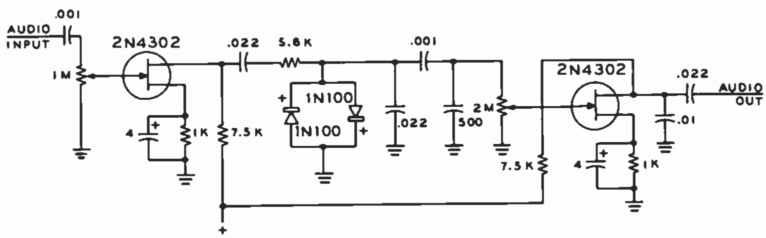


Figure 15

MODULATION LIMITING

Deviation in an f-m transmitter can be controlled by a clipping circuit which holds peak audio level within prescribed limits. Simple audio filter removes higher harmonics of clipped signal.

ter. And since the receiver must be superheterodyne if it is to have good sensitivity at the frequencies to which frequency modulation is restricted, i-f bandwidth is an important factor in its design.

The second requirement of the f-m receiver is that it incorporate some sort of device for converting frequency changes into amplitude changes, in other words, a detector operating on frequency variations rather than amplitude variations. Most f-m equipment operates in the vhf region, and at these frequencies it is not always possible to obtain optimum performance at reasonable cost with a single-conversion superheterodyne receiver. When good adjacent-channel selectivity is necessary, a low i-f channel is desirable; this, however lowers the image rejection ability of the receiver. Similarly, if good image rejection is desired, a high i-f channel should be used, but this is not compatible with good adjacent-channel rejection unless an expensive i-f filter is employed.

These difficulties are compromised by the use of a double-conversion receiver, such as the one shown in the block diagram of figure 16. In many receiver designs, the high i-f channel is chosen so that a harmonic of the mixing oscillator used for the second mixer may be used with the first mixer to reduce the number of crystals in the receiver. In other cases, a frequency synthesizer is used to generate the proper mixing frequencies.

The third requirement, and one which is necessary if the full noise-reducing capabilities of the f-m system of transmission are desired, is a limiting device to eliminate

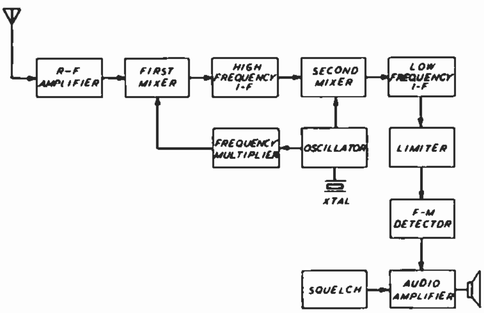


Figure 16

DOUBLE-CONVERSION RECEIVER FOR VHF F-M RECEPTION

amplitude variations before they reach the detector.

The Frequency Detector The simplest device for converting frequency variations to amplitude variations is an "off-tune" resonant circuit, as illustrated in figure 17. With the carrier tuned in at point A, a certain amount of r-f voltage will be developed across the tuned circuit, and, as the frequency is varied either side of this frequency by the modulation, the r-f voltage will increase and decrease to point C and B in accordance with the modulation. If the voltage across the tuned circuit is applied to an ordinary detector, the detector output will vary in accordance with the modulation, the amplitude of the variation being proportional to the deviation of the signal, and the rate being equal to the modulation frequency. It is obvious from figure 17 that only a small portion of the resonance curve is usable for linear conversion of frequency

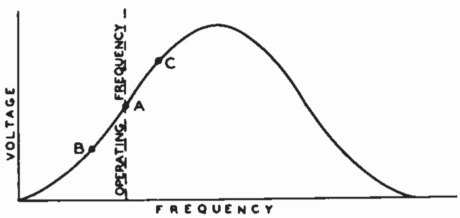


Figure 17

SLOPE DETECTION OF F-M SIGNAL

variations into amplitude variations, since the linear portion of the curve is rather short. Any frequency variation which exceeds the linear portion will cause distortion of the recovered audio. It is also obvious by inspection of figure 17 that an a-m receiver used in this manner is vulnerable to signals on the peak of the resonance curve and also to signals on the other side of the resonance curve. Further, no noise-limiting action is afforded by this type of reception. This system, therefore, is not recommended for f-m reception, although it may be widely used by amateurs for occasional nbfm reception.

Double-Tuned Discriminator A better frequency detector or *discriminator,* is shown in figure 18A. In this arrangement two tuned circuits are used, one tuned on each side of the i-f amplifier frequency, and with their resonant frequencies spaced slightly more than the expected transmitter swing. Their outputs are combined in a differential rectifier so that the voltage across series load resistors R_1 and R_2 is equal to the algebraic sum of the individual output voltages of each rectifier. When a signal at the i-f midfrequency is received, the voltages across the load resistors are equal and opposite, and the sum voltage is zero. As the r-f signal varies from the midfrequency, however, these individual voltages become unequal, and a voltage having the polarity of the larger voltage and equal to the difference between the two voltages appears across the series resistors, and is applied to the audio amplifier. The relationship between frequency and discriminator output voltage is shown in figure 19. The separation of the discriminator peaks and the linearity of the output voltage-versus-frequency curve depend on the discriminator frequency, the Q of the tuned circuits, and the value of the diode load resistors.

Foster-Seeley Discriminator A popular form of discriminator is that shown in figure 18B. This type of discriminator yields an output voltage-versus-frequency characteristic similar to that shown in figure 20. Here, again, the output voltage is equal to the algebraic sum of the voltages developed across the load resistors of the two diodes, the resistors being connected in series to ground. However, this *Foster-Seeley* discriminator requires only two tuned circuits instead of the three used in the previous discriminator. The operation of the circuit results from the phase relationships existing in a transformer having a tuned secondary. In effect, as a close examination of the circuit will reveal, the primary circuit is in series for r.f. with each half of the secondary to ground. When the received. signal is at the resonant frequency of the secondary, the r-f voltage across the secondary is 90 degrees out of phase with that across the primary. Since each diode is connected across one half of the secondary

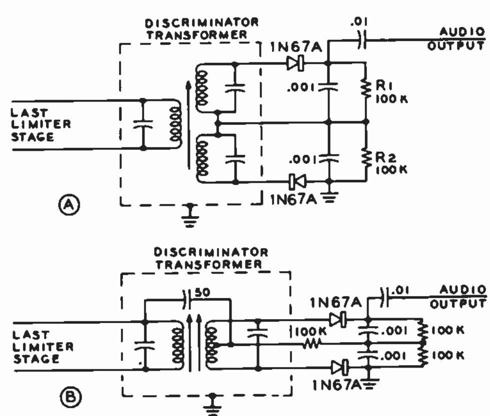
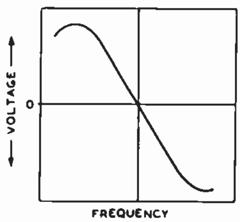


Figure 18

THE F-M DETECTOR

A—The double-tuned discriminator uses two secondary windings on the detector transformer, one tuned on each side of the i-f amplifier center frequency. On either side of center frequency a voltage of polarity and magnitude proportional to direction and magnitude of frequency shift is developed. B—Foster-Seeley discriminator employs a single, tapped secondary winding. Vector diagram of summed output voltages is shown in figure 20.

winding and the primary winding in series, the resultant r-f voltages applied to each are equal, and the voltages developed across each diode load resistor are equal and of opposite polarity. Hence, the net voltage between the top of the load resistors and ground is zero. This is shown vectorially in figure 20A where the resultant voltages R and R′ which are applied to the two diodes are shown to be equal when the phase angle between primary and secondary voltages is 90 degrees. If, however, the signal varies from the resonant frequency, the 90-degree phase relationship no longer exists between primary and secondary.



As its "center" frequency the discriminator produces zero output voltage. On either side of this frequency it gives a voltage of a polarity and magnitude which depend on the direction and amount of frequency shift.

Figure 19

DISCRIMINATOR VOLTAGE-FREQUENCY CURVE

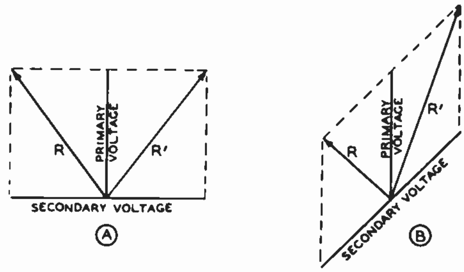


Figure 20

DISCRIMINATOR VECTOR DIAGRAM

A signal at the resonant frequency of the secondary will cause the secondary voltage to be 90 degrees out of phase with the primary voltage, as shown at A, and the resultant voltages R and R′ are equal. If the signal frequency changes, the phase relationship also changes, and the resultant voltages are no longer equal, as shown at B. A differential rectifier is used to give an output voltage proportional to the difference between R and R′.

The result of this effect is shown in figure 20B where the secondary r-f voltage is no longer 90 degrees out of phase with respect to the primary voltage. The resultant voltages applied to the two diodes are now no longer equal, and a d-c voltage proportional to the difference between the r-f voltages applied to the two diodes will exist across the series load resistors. As the signal frequency varies back and forth across the resonant frequency of the discriminator, an a-c voltage of the same frequency as the original modulation, and proportional to the deviation, is developed and passed on to the audio amplifier.

Ratio Detector A third form of f-m detector circuit, called the *ratio detector* is diagrammed in figure 21. The input transformer can be designed so that the parallel input voltage to the diodes can be taken from a tap on the primary of the transformer.

The circuit of the ratio detector appears very similar to that of the more conventional discriminator arrangement. However, it will be noted that the two diodes in the ratio detector are polarized so that their d-c output voltages add, as contrasted to the Foster-Seeley circuit wherein the diodes are polarized so that the d-c output voltages buck each other. At the center frequency to which the discriminator transformer is tuned, the voltage appearing at the top of the 100K resistor will be one-half the d-c voltage appearing at the *avc output* terminal, since the contribution of each diode will be the same. However, as the input frequency varies to one side or the other of the tuned value (while remaining within the passband of the i-f amplifier feeding the detector) the relative contributions of the two diodes will be different. The voltage appearing at the top of the 100K resistor will increase for frequency deviations in one direction and will decrease for frequency deviations in the other direction from the mean or tuned value of the transformer. The audio output voltage is equal to the ratio of the relative contributions of the two diodes, hence the name ratio detector.

The ratio detector offers several advantages over the simple discriminator circuit. The circuit does not require the use of a limiter preceding the detector since the circuit is inherently insensitive to amplitude modulation on an incoming signal. This factor alone means that the r-f and i-f gain ahead of the detector can be much less than the conventional discriminator for the same over-all sensitivity. Further, the circuit provides avc voltage for controlling the gain of the preceding r-f and i-f stages. The ratio detector is, however, susceptible to variations in the amplitude of the incoming signal as in any other detector circuit except the discriminator *with* a limiter preceding it, so that avc should be used on the stage preceding the detector.

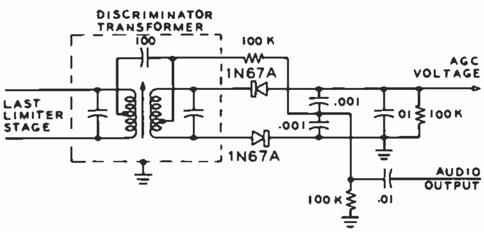


Figure 21

THE RATIO DETECTOR

This detector is inherently insensitive to amplitude modulation and does not require the use of a limiter ahead of it. Automatic volume control voltage is provided for controlling gain of r-f and i-f stages ahead of the detector.

The Pulse-Counting Detector Shown in figure 22A is a compact detector that provides inherent quieting and linear detection over wide frequency ranges. Two ICs (RTL logic) provide the functions of a limiter and discriminator. The first inverter serves as a signal amplifier and the following two stages provide limiting to produce a pulse train at the intermediate frequency. This train is fed to a "divide-by-four" circuit composed of flip-flops FF_1 and FF_2. The low-frequency signal triggers a monostable multivibrator (U_{1D}), whose period is about 0.5 that of the i-f signal. The output pulses of the multivibrator have a repetition rate which varies in direct proportion to the frequency variation of the i-f signal. The pulses are amplified by two inverter stages and converted to an audio signal by the RC de-emphasis network at the output of U_{1F}.

The Phase-Locked Loop Detector The phase-locked loop, discussed in Chapter 11 is now available in a single IC package or in separate building block ICs. The PLL consists of a phase detector, a filter, a d-c amplifier, and a voltage-controlled oscillator which runs at a frequency close to that of an incoming signal. The phase detector produces an error voltage proportional to the difference in frequency between the oscillator and the incoming signal, the error voltage being applied to the voltage-controlled oscillator. Any change in frequency of the incoming signal is sensed, and the resulting error voltage readjusts the oscillator frequency so that it remains locked to the incoming signal. As a result, the error voltage is a replica of the audio variations originally used to shift the frequency of the f-m signal, and the PLL functions directly as an f-m detector. The functional bandwidth of the system is determined by a filter placed on the error voltage line. The *Signetics* NE565 is especially designed for this service (figure 22B).

Limiters The limiter of an f-m receiver using a conventional discriminator serves to remove amplitude modulation and pass on to the discriminator a frequency-modulated signal of constant amplitude; a typical circuit is shown in figure 23. The limiter tube is operated as an i-f stage with very low plate voltage and with grid-resistor bias, so that it overloads quite easily. Up to a certain point the output of the limiter will increase with an increase in signal. Above this point, however, the limiter becomes overloaded, and further large increases in signal will not give any increase in output. To operate successfully, the limiter must be supplied with a large amount of signal, so that the amplitude of its output will not change for rather wide variations in amplitude of the signal. Noise, which causes little frequency modulation but much amplitude modulation of the received signal, is virtually wiped out in the limiter.

The voltage across the grid resistor varies with the amplitude of the received signal. For this reason, conventional amplitude-modulated signals may be received on the f-m receiver by connecting the input of the audio amplifier to the top of this resistor, rather than to the discriminator output. When properly filtered by a simple RC circuit, the voltage across the grid resistor may also be used as avc voltage for the receiver.

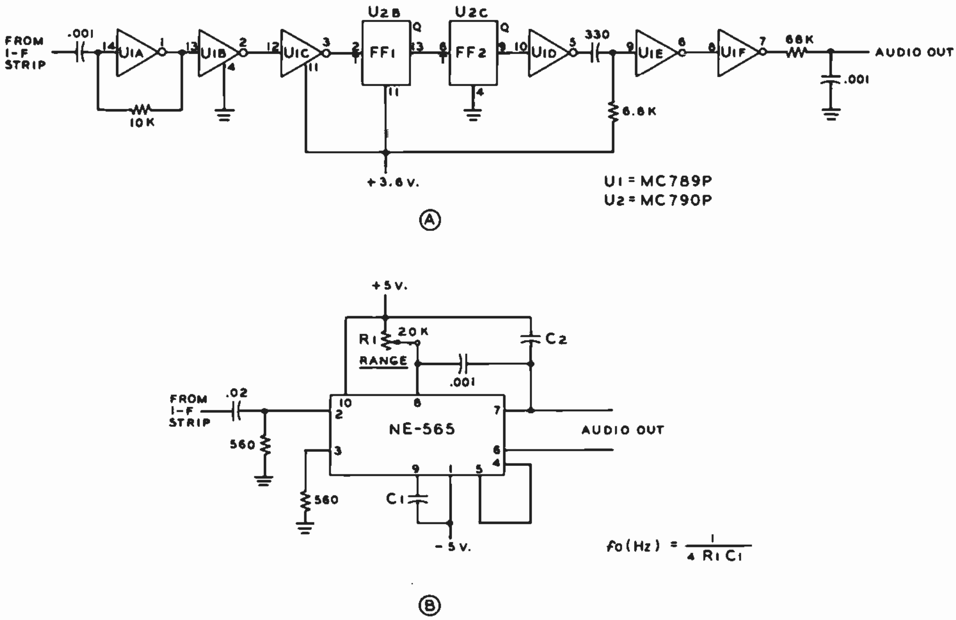


Figure 22

UNUSUAL F-M DETECTORS MAKE USE OF INTEGRATED CIRCUITS

A—Pulse counting detector uses two small ICs and provides quieting and linear detection over wide frequency ranges. First three stages provide limiting and produce a pulse train which is fed to a "divide-by-four" pair of flip-flops. Low-frequency pulses trigger a multivibrator (U_{1D}) whose repetition rate varies in direct proportion to frequency variation of i-f signal. Pulses are converted to audio signal by R-C deemphasis network at output of detector. B—Single IC performs as phase-locked loop detector for f-m. Error voltage proportional to frequency deviation is applied to voltage-controlled oscillator, locking it to incoming signal. Error voltage is replica of frequency shift on incoming signal.

When the limiter is operating properly avc is neither necessary nor desirable, however, for f-m reception alone.

A two-stage solid-state limiter is shown in figure 24.

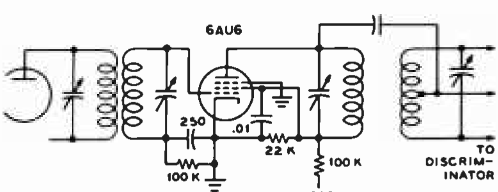


Figure 23

LIMITER CIRCUIT

One, or sometimes two, limiter stages normally precede the discriminator so that a constant signal level will be fed to the f-m detector. This procedure eliminates amplitude variations in the signal fed to the discriminator, so that it will respond only to frequency changes.

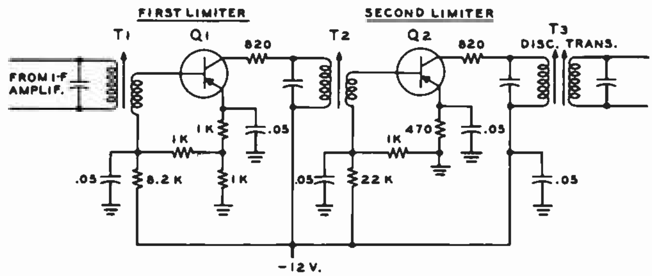
Proper limiting action calls for a signal of considerable strength to insure full clipping, typically several volts for tubes and about one volt for transistors. Limiting action should start with an r-f input of 0.2 μV, or less, at the receiver antenna terminals, consequently a large amount of signal gain is required between antenna and the limiter stages. Typically 100 db to 140 db gain is used in modern f-m receivers, most of this gain being achieved in the i-f amplifier chain. The high gain level amplifies internal and external noise and an annoying blast of noise emits from the speaker of the f-m receiver unless some form of *audio squelch* is provided, as discussed later in this chapter.

Receiver Bandwidth One of the most important factors in the design of an f-m receiver is the frequency swing which it is intended to handle. It will be apparent from figure 19 that if the straight

Figure 24

TWO-STAGE SOLID-STATE F-M LIMITER

F-m limiter circuit serves to remove amplitude variations of incoming f-m signal. Limiter saturates with small signal and further increases in strength of incoming signal will not give any increase in output level. Noise, which causes little f-m but much a-m, is virtually eliminated in effective limiter stages.



portion of the discriminator circuit covers a wider range of frequencies than those generated by the transmitter, the audio output will be reduced from the maximum value of which the receiver is capable.

In this respect, the term *modulation percentage* is more applicable to the f-m receiver than it is to the transmitter, since the modulation capability of the communication system is limited by the receiver bandwidth and the discriminator characteristic; full utilization of the linear portion of the characteristic amounts, in effect, to 100 percent modulation. This means that some sort of standard must be agreed on, for any particular type of communication, to make it unnecessary to vary the transmitter swing to accommodate different receivers.

Two considerations influence the receiver bandwidth necessary for any particular type of communication. These are the maximum audio frequency which the system will handle, and the deviation ratio which will be employed. For voice communication, the maximum audio frequency is more or less fixed at 3000 to 4000 Hz. In the matter of deviation ratio, however, the amount of noise suppression which the f-m system will provide is influenced by the ratio chosen, since the improvement in signal-to-noise ratio which the f-m system shows over amplitude modulation is equivalent to a constant *multiplied by the deviation ratio.* This assumes that the signal is somewhat stronger than the noise at the receiver, however, as the advantages of wideband frequency modulation in regard to noise suppression disappear when the signal-to-noise ratio approaches unity.

As mentioned previously, broadcast f-m practice is to use a deviation ratio of 5. When this ratio is applied to a voice-communication system, the total swing becomes 30 to 40 kHz. With lower deviation ratios, such as are most frequently used for voice work, the swing becomes proportionally less, until at a deviation ratio of 1 the swing is equal to twice the highest audio frequency.

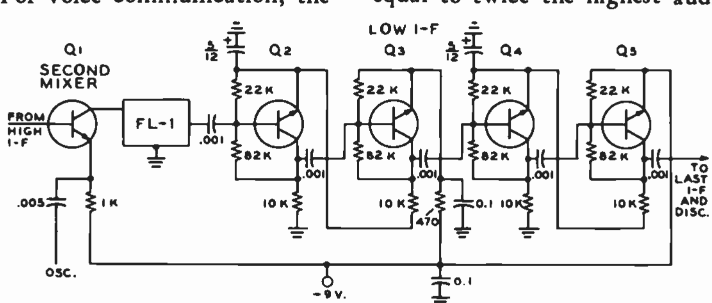


Figure 25

TRANSISTOR I-F STRIP USES CASCODE CIRCUIT

Transistors in pairs (Q_2-Q_3 and Q_4-Q_5) are placed in series in regard to the supply voltage in the manner of a cascode amplifier so that each transistor of a pair has half the d-c voltage across it. A crystal or mechanical filter provides good adjacent-channel selectivity.

Actually, however, the receiver bandwidth must be greater than the expected transmitter swing, since for distortionless reception the receiver must pass the complete band of energy generated by the transmitter, and this band will always cover a range somewhat wider than the transmitter swing.

On the other hand, a low deviation ratio is more satisfactory for strictly communication work, where readability at low signal-to-noise ratios is more important than additional noise suppression when the signal is already appreciably stronger than the noise.

Deviations of 15, 5, and 2.5 kHz are common on the amateur vhf bands and are termed wideband, narrowband, and sliver band, respectively. Bandwidth required in an f-m receiver is about 2.4 times the deviation: 36 kHz for wideband reception and 13 kHz for narrowband reception.

The proper degree of i-f selectivity may be achieved by using a number of overcoupled transformers or by the use of a ceramic or crystal filter. Shown in figure 25 is a transistorized i-f strip using a packaged filter for adjacent channel selectivity and four stages of resistance-coupled amplification to provide adequate gain. The stages are paired in regard to the supply voltage, with the paired transistors placed in series so that each has half the supply voltage. I-f filters for vhf f-m service generally have a center frequency of 455 kHz, 9.0, 10.7, or 21.5 MHz with bandwidths ranging from 12 kHz to 36 kHz.

Pre-Emphasis and De-Emphasis Standards in f-m broadcast and TV sound work call for the pre-emphasis of all audio modulating frequencies above about 2000 Hz, with a rising slope such as would be produced by a 75-microsecond RL network. Thus the f-m receiver should include a compensating de-emphasis RC network with a time constant of 75 microseconds so that the over-all frequency response from microphone to speaker will approach linearity. The use of pre-emphasis and de-emphasis in this manner results in a considerable improvement in the over-all signal-to-noise ratio of an f-m system. Appropriate values for the de-emphasis network, for different values of circuit impedance are given in figure 26.

Squelch Circuits *Squelch circuits* are used to mute the audio of an f-m receiver when no signal is present. In a high-gain receiver, speaker noise can be very annoying to the operator who must monitor a channel for a long period. When the receiver is squelched, no background noise is heard; when an r-f signal comes on, squelch is turned off and the audio system becomes operative. Squelch circuits may be carrier operated or noise operated.

A simple squelch circuit is shown in figure 27. The squelch voltage is derived from the discriminator circuit, a negative voltage being obtained to bias the squelch tube (V_1) to cutoff and to permit the received signal to pass through the audio amplifier (V_2). Without an input signal, the positive bias is greater than the adjustable bias applied to the cathode of the squelch tube, causing the

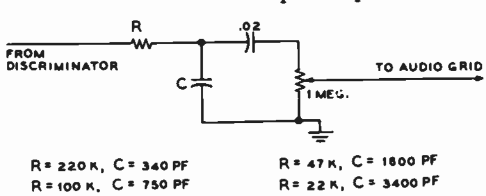


Figure 26

75-MICROSECOND DE-EMPHASIS CIRCUIT

The audio signal transmitted by f-m and TV stations has received high-frequency pre-emphasis, so that a de-emphasis circuit should be included between the output of the f-m detector and the input of the audio system.

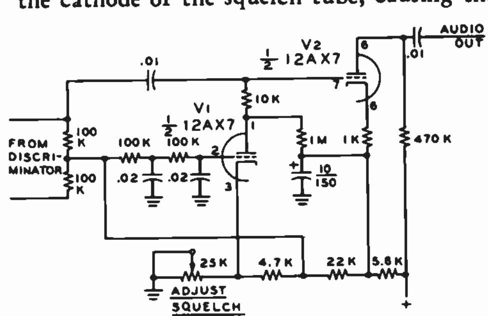


Figure 27

ONE-TUBE AUDIO SQUELCH CIRCUIT

Squelch circuit mutes audio of f-m receiver when no signal is present. Squelch voltage is derived from discriminator and cuts off squelch tube, allowing audio signal to pass through amplifier tube. Squelch level may be adjusted by 25K potentiometer.

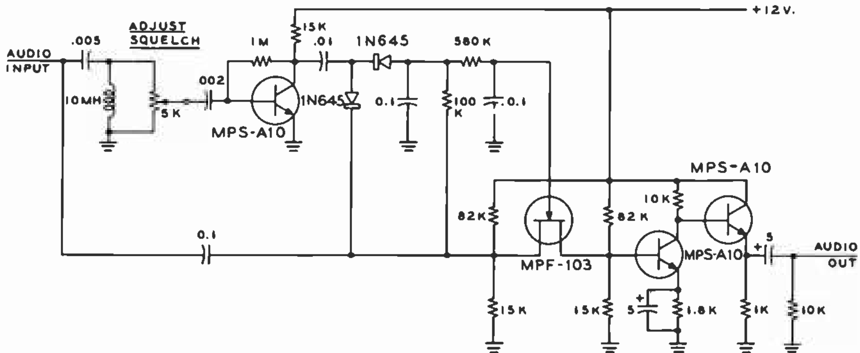


Figure 28

SOLID-STATE SQUELCH CIRCUIT

Audio voltage is amplified, rectified, and applied to the gate element of JFET which acts as a series audio gate. Squelch level is controlled by varying the signal gate voltage of MPF-103 squelch amplifier stage.

tube to conduct, dropping the plate voltage and cutting off the audio tube through the 10K resistor to the control grid.

A solid-state squelch circuit is shown in figure 28. Audio voltage is amplified and rectified and applied to the gate of a JFET which acts as a series audio gate. Squelch level is controlled by varying the signal gate voltage of the MPF-103 device. The output impedance of the MPS-A10 amplifier is quite low and suitable for running into an audio line, if required.

13-5 The F-M Repeater

Since radio transmission in the vhf region is essentially short range, a form of radio relay station termed a *repeater* may be employed to expand the communication range of base or mobile stations over an extended distance. Various types of relays are in use in the United States, their operation depending on the requirements of the communications circuit.

The *relay unit* is a fixed repeating station whose specific purpose is to extend station-to-station communication capability. The user's transmitter is on the input frequency while his receiver is on the output frequency of the relay (figure 29). When desired, direct communication between stations may take place by using a closely spaced frequency domain and a two-frequency transmitter.

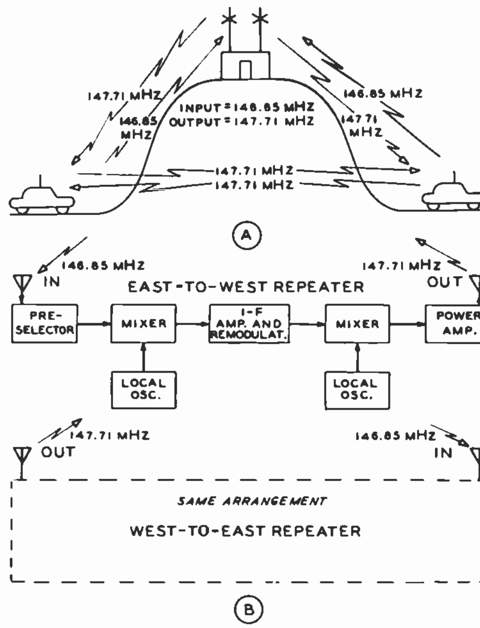


Figure 29

F-M REMOTE REPEATER FOR MOBILE SERVICE

Radio relay station serves as a repeater to extend the range of base or mobile f-m stations. Communication between units may be achieved either directly, or through repeater. The repeater consists of a back-to-back receiver and transmitter having a common i-f and remodulator system. Most repeaters are limited to a single channel, but multiplex operation permits simultaneous transmission of different information forms on the channel.

The *remote base* is a form of relay unit whose location has a height or tactical advantage. Means must be provided to control such an installation which in amateur service most often is working in conjunction with a pair of frequencies—input and output. In so doing, remote bases serve on common frequencies by which individual groups operating their own installation can cross-communicate. Frequencies above 220 MHz or direct-wire lines must be used for remote control.

Simplex communication, on the other hand, refers to communication between individual units operating on a common transmit and receive frequency. Thus simplex operation can be interfaced with relay operation, using either a local or remote base. Remote base operation must take place under FCC license to a responsible controlling authority and each application for such service is judged individually on the merits of the case.

Repeater Types There are two basic categories of repeaters: *open* and *closed*. The open repeater is one which has been installed for the benefit of all who wish to use it for communications; the *closed* repeater is one which is designed to selectively benefit a specific group of users. Both types are in widespread use throughout the United States and many foreign countries. Early repeaters were a-m open types, which later gave way to the f-m open and closed repeaters. The open repeater is virtually always carrier operated, switching to the trnsmitting mode only with an incoming signal. The frequency of 146.34 MHz has been unofficially adopted as the "national repeater input channel" in the United States and Canada, although many repeaters use other

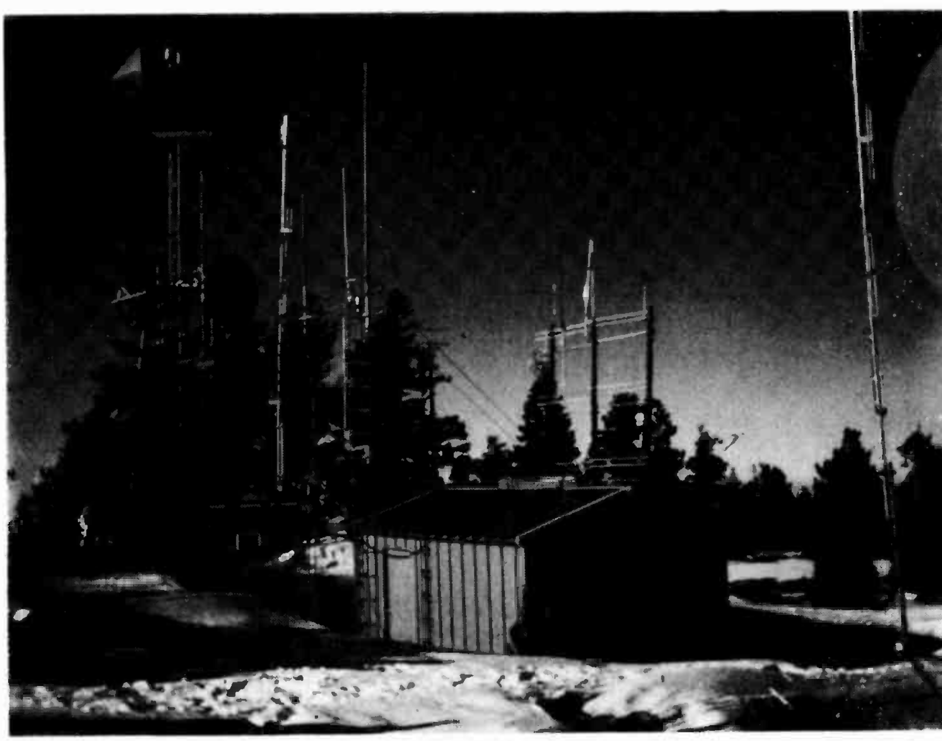


Figure 30

TYPICAL REMOTE REPEATER INSTALLATION

WB6SLR's remote repeater installation at a commercial facility atop 8500-foot Blueridge Summit in California.

inputs. Both 146.76 MHz and 146.94 MHz are popular repeater output channels.

The closed repeater, as the name implies, gives the benefits of repeater coverage to a select group of subscribers or users. Special selective circuits are used on the repeater to reject all signals other than those for which the system was designed. This function is almost universally achieved with a system of *access tones*, whereby a specific tone on the incoming signal is a prerequisite to being automatically relayed to the repeater output. One technique calls for a continuous low-frequency tone (below 120 Hz) to be transmitted. A decoding device is employed at the repeater that responds only to signals bearing this tone. This is termed a *continuous tone squelched private line* (PL) system. A second technique requires that the incoming signal be accompanied by a short high-frequency *tone burst* of a few milliseconds. The decoder at the repeater allows the transmitter to be energized only when the signal bears the proper tone. This access approach is called the *single-tone*, or "whistle-on" system, since it may be activated by an operator with a good ear for tone and a talent for whistling!

Many repeaters make use of a *transmission limiter*, which consists of a timer which disables the repeater when input time exceeds 3 minutes or so. The repeater is reactivated when the input signal is removed. More complicated control techniques exist, too, which make use of channelized tones between 1500 and 1650 Hz.

Control Techniques The basic control element of most amateur repeaters is the *carrier-operated relay* (COR), a squelch-responding circuit that provides a relay closure (K_1) with each signal that occupies the channel (figure 31). When the repeater is at a remote location, functional control may be exerted over a wire (telephone line) or by a uhf radio link. The control scheme is based upon the transmission of specific and precise audio frequency signals which activate turn-on and shut-down systems, frequency selections, and automatic time-out devices. The audio frequencies are generated by a tone generator termed an *encoder* and the responding relay is called a *decoder*. Multiple functions may be achieved through the use of a single decoder by the use of timers, ratchet relays, or stepping switches (figure 32). One of the most promising tone-control techniques makes use of the multitone (*Touchtone*) technique. *Touchtone* command signals are generated with a conventional *Touchtone* telephone dial which has an integral multitone encoder. The system makes use of eight discrete tone frequencies arranged in two groups of four tones each (a high group and a low group). Sixteen digits can then be represented by the combination of one tone from the high group with one tone from the low. The individual frequencies and various combinations are shown in figure 33, which is a schematic of the standard 25A3 10-button *Touchtone telephone pad*. The

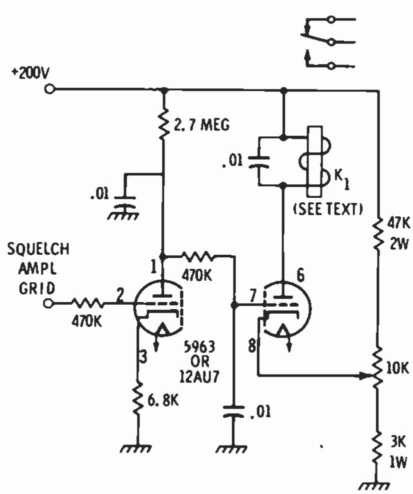


Figure 31

UNIVERSAL CARRIER-OPERATED RELAY

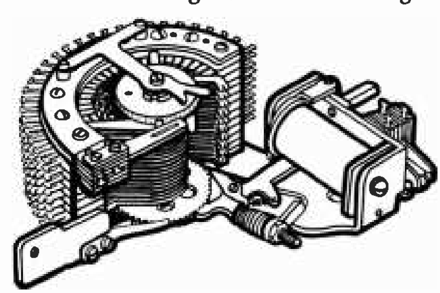


Figure 32

STANDARD ROTARY STEPPER SWITCH

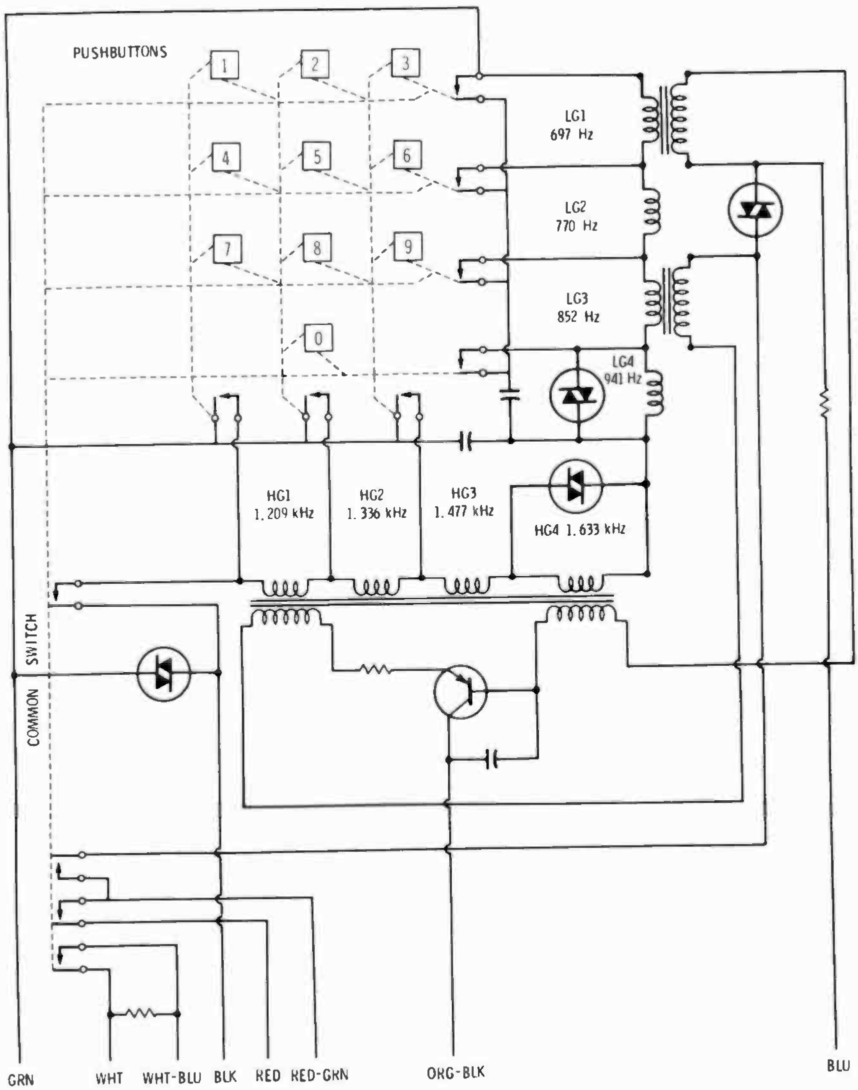


Figure 33

TOUCHTONE PAD

supply voltage is fed to the pad over the same path as the output of the tones.

The *Touchtone* encoder pad can be connected directly into the microphone amplifier of an f-m transmitter for transmission of the tones over the air to the decoder unit at the repeater site.

The *Touchtone* signal can be decoded by separating the two-tone combination via bandpass and band-elimination filters into groups so that each tone can be regulated, limited, and applied to the desired control circuit.

Other tone systems exist, including the dual-tone (*Secode*) system and the single-tone approach. The latter may be used with a telephone dial pulsing system, as shown in figure 34. Control pulses are sent serially, at

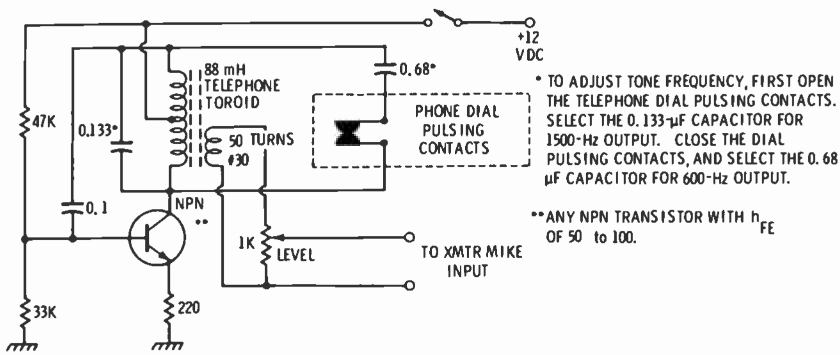


Figure 34

SECODE-TYPE 600/1500-Hz OSCILLATOR

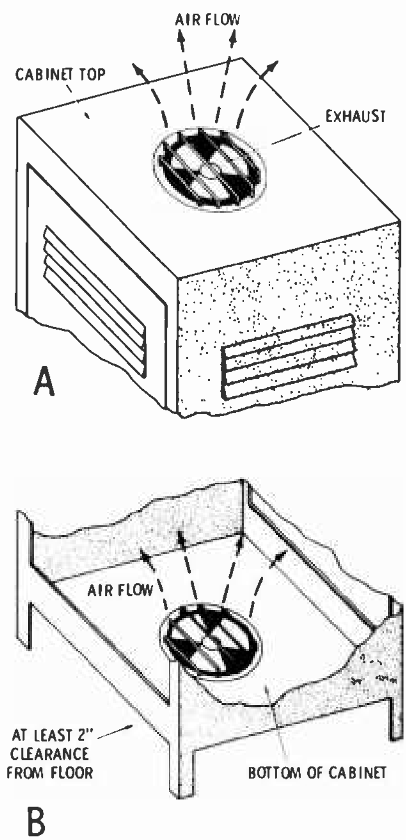


Figure 35

METHODS FOR MOUNTING VENTILATING FANS

A—Top-mounted exhaust fan. B—Bottom-mounted forced-air type.

a rate of about 10 pulses per second to initiate a command function at the repeater.

The Repeater The repeater is a receiver-transmitter combination capable of duplex operation. That is, the receiver must be capable of functioning regardless of whether the transmitter is activated or not. Since the repeater equipment must run continuously (probably in a remote spot without air conditioning) it must be well ventilated. Most repeaters have air continuously circulated about within the cabinet or enclosure by means of exhaust and intake fans as shown in figure 35.

Most commercial f-m units have a universal metering system which requires a sensitive meter to monitor test points for each important stage. A *Motorola* test set is shown in figure 36 and is typical of such units. Commercial transmitters and receivers are equipped with test receptacles into which the test set adapter is plugged for tune-up and checkout. For repeater service, the test meter is often incorporated directly into the equipment.

Transmitter Noise—Broadband noise may be radiated by any r-f generating equipment as the result of random noise components generated and amplified in the driver stages, which are amplified and passed on to the antenna through the relatively broad selectivity of the amplifier output circuitry. Enough noise may be radiated to degrade the performance of a nearby receiver operating several MHz away (figure 37A). Transmitter noise is bothersome as "off-channel" noise

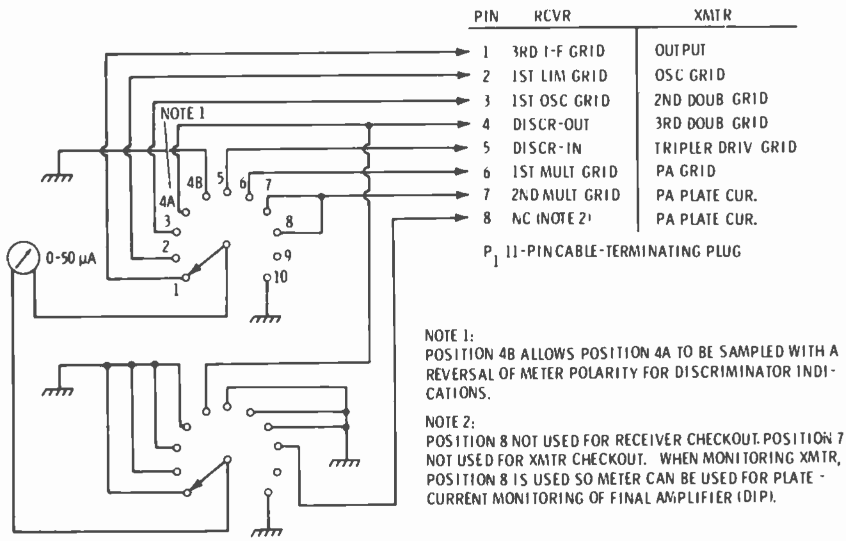


Figure 36

MOTOROLA TEST SET

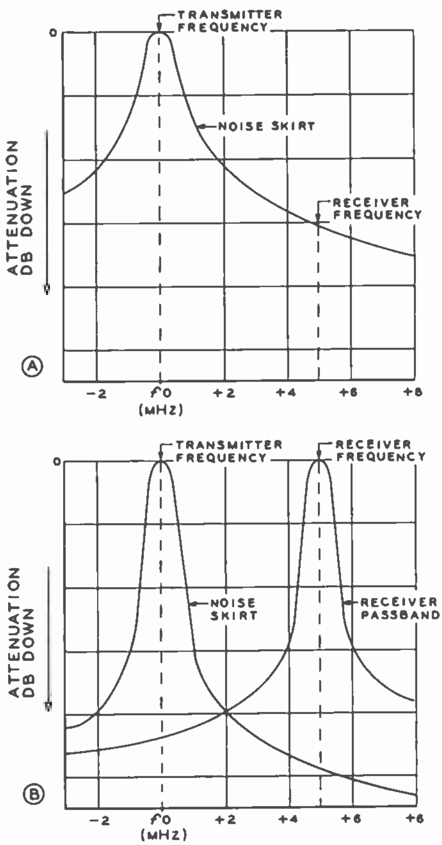


Figure 37

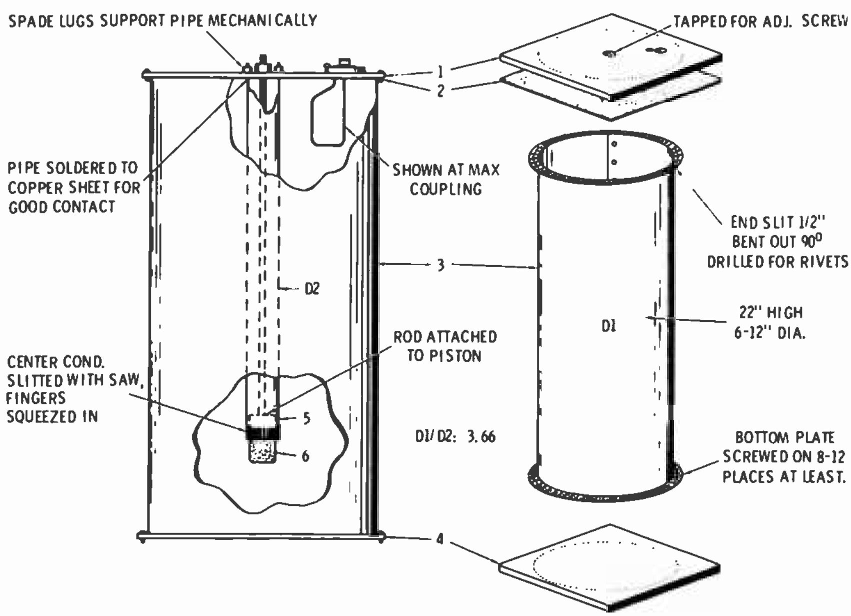
A—Broadband noise is radiated by an f-m transmitter as result of random noise components amplified and passed to antenna through relatively broad selectivity of output circuitry. Enough noise may be radiated to degrade performance of nearby receiver operating several MHz away from transmitting frequency. B—Bandpass cavity on output of transmitter and input of receiver provides sufficient attenuation and rejection of off-channel noise to protect receiver from desensitization.

which cannot be filtered out at the receiver, competing with the desired signal and reducing effective receiver sensitivity.

Receiver Desensitization—This form of interference is the result of a strong off-channel frequency signal entering the front-end of the receiver, upsetting critical voltage and current levels, and reducing receiver gain.

Intermodulation—Intermodulation is the generation of spurious frequencies in a non-linear circuit element. The undesired frequencies correspond to the sum and differences of the fundamental and harmonics of two or more frequencies passing through the element, as discussed in Chapter 16.

Intermodulation interference may occur from signals outside the normal operating



1. 3/32" END PLATE - STEEL OR OVER 1/8" ALUMINUM
2. FLASHING COPPER LINER, ANY GAUGE
3. ALUMINUM CYLINDER (0.032" OR THICKER)
4. 3/32" STEEL OR ALUMINUM END PLATE
5. COPPER PIPE - DIA: 1/3.66 x OUTSIDE DIA. OF CAVITY (NOT CRITICAL)
6. TUNING PISTON - ANY MATERIAL WITH FLASHING COPPER WRAPPED ON OUTSIDE. LENGTH TO ALLOW TRAVEL MAKING TOTAL CENTER CONDUCTOR VARIABLE FROM 17" TO 21".

NOTE: FOR PISTON ROD SCREW, USE 5/16-18 THREADED ROD. SECURE AT TOP WITH LOCKNUT.

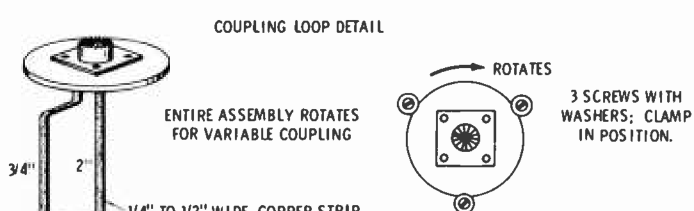


Figure 38

DESIGN DETAILS OF THE 144- TO 148-MHz CAVITY

range of the equipment to produce a product which can interfere with a desired signal.

Receiver Protection Sufficient electrical isolation between receiver and transmitter at a repeater site will protect the receiver from desensitization, intermodulation, and spurious transmitter noise. Receiver protection may be brought about by physically separating the receiver and transmitter antennas in space and by the use of a high-Q bandpass cavity at the input of the receiver to reject frequencies outside of the cavity passband (figure 38). The cavity resonator is placed in the antenna circuit in such a way as to pass the received frequency and reject the transmitted frequency. A second cavity on the output of the transmitter will reduce off-frequency transmitter noise passing to the antenna, as shown in figure 37B.

Repeater Optimization The design, control, licensing, and use of a repeater is a complex operation drawing upon many facets of electronics. Complete information on this interesting and valuable device is contained in the *Radio Amateur's F-M Repeater Handbook*, by Ken Sessions, Jr., K6MVH, and published by Editors and Engineers, Division Howard W. Sams & Co., Inc. The interested reader is referred to this book for complete and detailed information on vhf repeaters and their use.

CHAPTER FOURTEEN

Radioteletype and Specialized Transmission and Reception

Electromagnetic communication involves various modulation techniques that lie afield from the well known voice and code modulation used by the majority of radio amateurs. Chief among these specialized systems are *radioteletype*, *slow-scan television*, *broadband television*, *facsimile*, and *radio control* of remote objects. In addition, specialized space communication experiments are carried out by means of amateur radio space satellites, moon reflection, and irregular ionospheric reflection.

Radioteletype using frequency-shift keying is permitted on all amateur bands except 160 meters and is growing in popularity as a reliable, fast and accurate means of intercommunication.

Teleprinting is a form of intelligence based on a simple binary (on-off) code designed for electromechanical transmission. The code consists of d-c pulses generated by a special electric typewriter, which can be reproduced at a distance by a separate machine. The pulses may be transmitted from one machine to another by wire or by a radio circuit. When radio transmission is used, the system is termed *radioteletype* (RTTY). The name *teletype* is a registered trademark of *Teletype Corporation* and the term *teleprinter* is used in preference to the registered term.

14-1 Radioteletype Systems

The d-c pulses that comprise the teleprinter signal may be converted into three basic forms of emission suitable for radio transmission. These are: (1) *frequency-shift keying* (*FSK*), designated as F1 emission; (2) *make-break keying* (*MBK*), designated as A1 emission; and (3) *audio frequency-shift keying* (*AFSK*), designated as F2 emission.

Frequency-shift keying is achieved by varying the transmitted frequency of the radio signal a fixed amount (usually 850 Hertz or less) during the keying process. The shift is accomplished in discrete intervals designated *mark* and *space*. Both types of intervals convey information to the teleprinter. *Make-break keying* is analogous to simple c-w transmission in that the radio carrier conveys information by changing from an *on* to an *off* condition. Early RTTY circuits employed MBK equipment, which is now considered obsolete since it is less reliable than the frequency-shift technique. *Audio frequency-shift keying* employs a steady radio carrier modulated by an audio tone which is shifted in frequency according to the RTTY pulses. Other forms of information transmission may be employed by a RTTY system which also encompass translation of binary pulses into r-f signals.

The Teleprinter Code The teleprinter code consists of 26 letters of the alphabet and additional characters that accomplish machine functions, such as line feed, carriage return, bell, and upper- and lower-case shift. These special characters are required for the complete automatic process of teleprinter operation in printing received copy. Numerals, punctuation, and symbols may be taken care of in the case shift, since all transmitted letters are capitals.

The teleprinter code is made up of spaces and pulses, for transmission at 60, 67, 75 or 100 words per minute. Each character (at 60 wpm) is made up of five elements, plus a 22 millisecond *start space* and a 31 millisecond *stop pulse*. All characters are equal in total transmission time to 163 milliseconds duration to achieve machine synchronization at both ends of the RTTY circuit. Timing is usually accomplished by the use of synchonous motors in the equipment, locked to the a-c line frequency. The sequence of mark and space pulses for the letter *R* is shown in figure 1. The start space provides time for synchronization of the receiving machine with the sending machine. The stop pulse provides time for the sending mechanism as well as the receiving mechanism to properly position themselves for transmission of the following character.

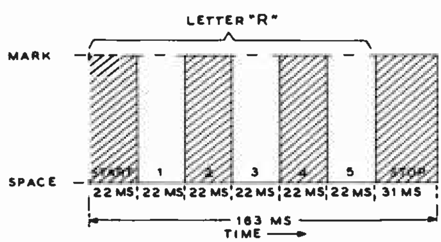


Figure 1

THE TELEPRINTER CODE

Teleprinting is based on a simple binary code made up of spaces and pulses, each of 22 milliseconds duration. Normal transmission is at the rate of 60 w.p.m. The sequence of mark and space pulses for the letter R are shown here. Start space provides time for machine synchronization and stop pulse provides time for sending and receiving mechanisms to position themselves for transmission of the following character.

The FSK system normally employs the higher radio frequency as the mark and the lower frequency as the space. This relationship often holds true in the AFSK system also. The lower audio frequency may be 2125 Hz and the higher audio frequency 2975 Hz, giving a frequency difference or shift of 850 Hz. Other, more narrow shifts (such as 170 Hz) are gradually coming into popularity in radio amateur RTTY work.

The Teleprinter The teleprinter resembles a typewriter in appearance, having a keyboard, a type basket, a carriage, and other familiar appurtenances. The keyboard, however, is not mechanically linked to the type basket or printer. When a key is pressed on the keyboard of the sending apparatus a whole code sequence for that character is generated in the form of pulses and spaces. When this code sequence is received on a remote machine, a type bar is selected

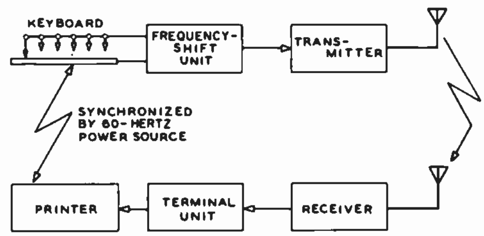


Figure 2

BLOCK DIAGRAM OF ONE-WAY-RTTY CIRCUIT

The teleprinter generates code sequence in the form of on-off pulses for the alphabet and additional special characters. Teleprinter code is transmitted at rate of 60 w.p.m. by means of frequency-shift technique. The receiving apparatus drives a mechanical printer that is usually synchronized with the keyboard by the common 60-Hz power source.

and made to print the letter corresponding to the key pressed. Synchronization of machines is accomplished by means of start and stop pulses transmitted with each character. An electromechanical device driven by the motor of the teleprinter is released when a key is pressed and transmission of the complete character is automatic.

The receiving apparatus operates in reverse sequence, being set in operation by the first pulse of a character sent by the transmitter mechanism. While each character is sent at the speed of 60 w.p.m., actual transmission of a sequence of characters may be much slower, depending on the speed of the operator. A simplified diagram of a one-way RTTY circuit is shown in figure 2.

14-2 RTTY Reception

The RTTY receiving mechanism must respond to a sequence of pulses and spaces

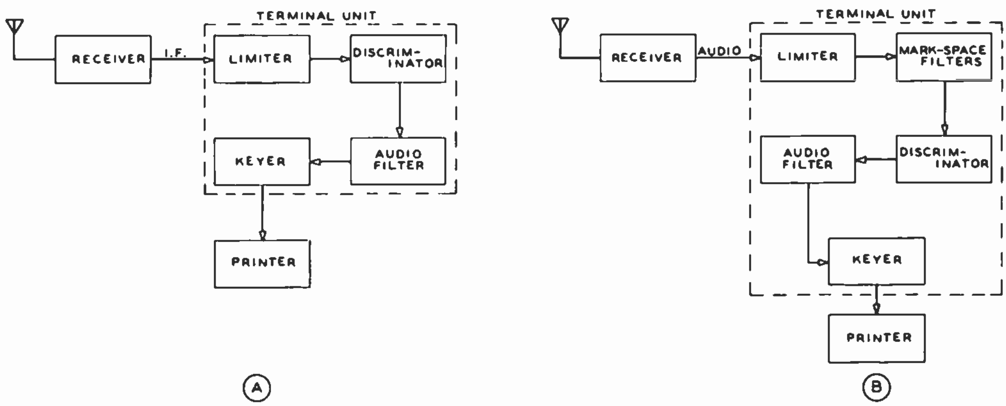


Figure 3

A shows block diagram of i-f terminal unit employing f-m discriminator technique. I-f converter requires that selectivity and interference rejection be achieved by means of selective tuned circuits of the receiver. B shows block diagram of audio-frequency terminal unit. Mark and space filters are used ahead of audio discriminator, followed by a low-pass audio filter. Beat oscillator of receiver is used to provide audio beat tones of 2125 and 2975 Hz required for nominal 850-Hz shift system.

transmitted by wire or radio. Frequency-shift keying may be demodulated by a beat-frequency technique, or by means of a discriminator as employed in f-m service. The received signal is converted into d-c pulses which are used to operate the printing magnets in the teleprinter. Conversion of RTTY signals into proper pulses is accomplished by a *receiving converter* (*terminal unit*, abbreviated TU). RTTY converters may be either i-f discriminator or audio discriminator units. A block diagram of an intermediate-frequency converter is shown in figure 3A. The RTTY signal in the i-f system of the receiver is considered to be a carrier frequency-modulated by a 22.8-Hz square wave having a deviation of plus and minus 425 Hz (for 850-Hz shift). Amplitude variations are removed by the limiter stage and the discriminator stage converts the frequency shift into a 22.8-Hz waveform, applied to the teleprinter by means of an electronic keyer. In its simplest form, the i-f converter requires that adequate selectivity and interference rejection be achieved by means of the i-f system of the receiver.

The schematic of a typical i-f RTTY converter is shown in figure 4.

A block diagram of an *audio-frequency converter* is shown in figure 3B. An audio

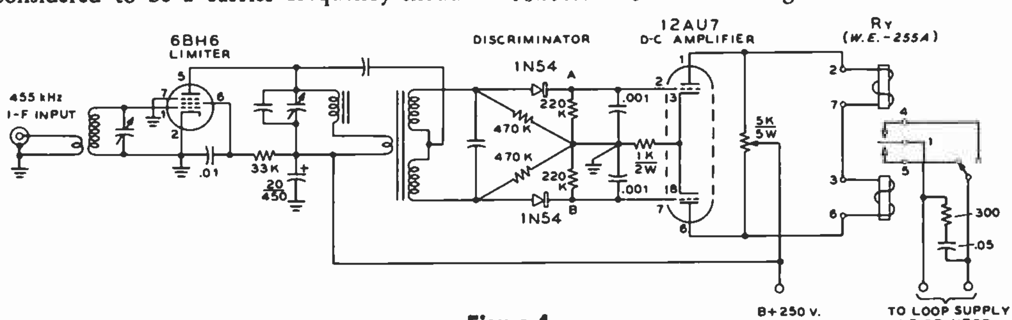


Figure 4

I-F RTTY CONVERTER

Typical i-f converter circuit illustrates this technique. Some type of indication that the RTTY signal is properly tuned is required, particularly on the hf bands. With the i-f terminal unit, a zero-center microammeter may be connected across discriminator load resistors (A-B).

limiter is followed by mark-frequency and space-frequency filters placed ahead of the discriminator stage. A low-pass filter and electronic keyer provide the proper d-c signal required by the teleprinter. The beat oscillator of the receiver is used to provide the beat tones of 2125 and 2975 Hz required in the usual 850-Hz shift system. Either frequency may be used for either mark or space, and the signals may be easily inverted by tuning the beat oscillator to the opposite side of the i-f passband of the receiver.

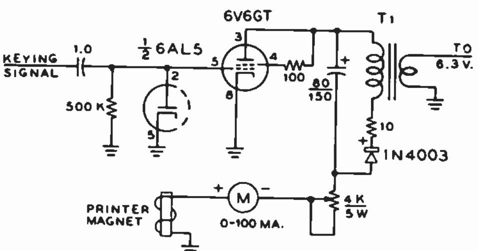


Figure 5

ELECTRONIC KEYER FOR

RTTY PRINTER

The polar relay may be eliminated and the teleprinter mechanism driven directly by a keyer such as shown here. This circuit provides loop supply and keeps the printer magnets in the ground circuit. Printer coils are placed in series for 20-ma loop operation, or in parallel for 60-ma operation. Additional printer magnets are connected either in series or parallel, to a limit of two or three before inductive effects of coils introduce undesirable side effects.

Receiving converters of both types usually include clipping and limiting stages which hold the signal at constant amplitude and converters occasionally include pulse-forming circuits which help to overcome distortion that occurs during transmission of the intelligence.

Teleprinters are actuated by electromagnets which release the motor-driven mechanism driving the type bars. The magnets require 20 or 60 milliamperes of current which may be obtained from an electronic keyer such as the one shown in figure 5. A single teleprinter may be run as an electric typewriter on a *local loop supply* which couples the keyboard and typing mechanisms in a single circuit (figure 6).

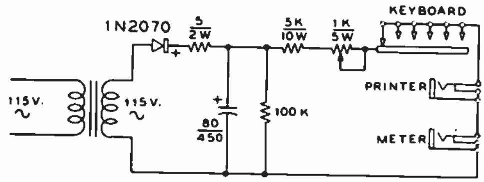


Figure 6

LOCAL LOOP SUPPLY

FOR TELEPRINTER

A single teleprinter may be run as electric typewriter on local loop supply which couples the keyboard and typing mechanism in a single circuit. Depending on how the machine is wired, the keyboard and magnets can be on plugs, or connected in series internally, with only one plug (usually "red") to the loop supply.

14-3 An Audio-Frequency RTTY Converter

This solid-state RTTY converter (or demodulator) is designed for reliable operation at a modest price. It will work either with 2125-2975 mark and space tones or the 1275-2125 mark and space required by some SSB receivers. The complete schematic is shown in figure 7. Two small op-amps and a 300-volt rated transistor are used, along with nine diodes.

The first op-amp is a high gain limiter. Reverse-connected zener diodes in the input circuit protect the amplifier against an excessive signal level. The 25K *balance* potentiometer compensates for a small degree of offset input voltage.

The output of the op-amp is fed to the discriminator filters which use surplus 88-mH toroidal inductors (T_1, T_2). Full-wave rectification and a simple RC low-pass filter remove the audio component of the signal as the shifting audio tones are converted into d-c pulses in a *slicer* stage. This op-amp takes the small voltages from the tuned filters and changes them to +10 volts for *mark* and −10 volts for *space*. Over-all gain is sufficient so that the unit will operate with shifts as low as a few cycles.

The keyer transistor (Q_1) has a 300-volt collector-emitter rating and will pass the 60 ma loop current required for teleprinters. A simple RC network in the collector-emit-

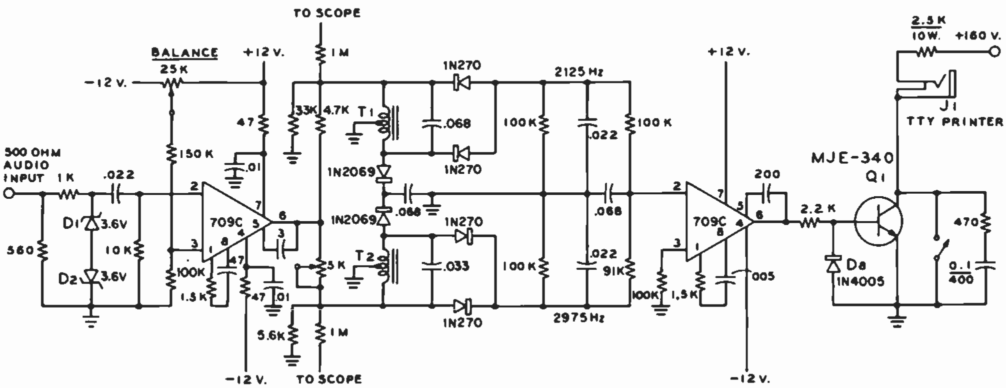


Figure 7

REPRESENTATIVE RTTV DEMODULATOR (CONVERTER)

This solid-state audio RTTY demodulator is based on a design by W6FFC (the Mainline ST-5). It uses two 709C operational amplifiers, one as an audio limiter, and the other as a trigger stage to drive the keyer transistor, which has a 300-volt collector-emitter rating. Reverse-connected zenor diodes limit the drive signal to the demomulator unit and the make and space tones are separater by tuned filters, which are built around surplus 88-mH toroid inductors (T_1, T_2).

ter circuit protects the transistor from the back-emf developed by the inductance of the selector magnets in the printer. The teleprinter keyboard may be connected in series with the printer magnets, both seriesed through jack J_1, if desired.

14-4 Frequency-Shift Keying

The keyed d-c voltage from the teleprinter is used to operate a keyer circuit to shift the transmitter carrier back and forth in frequency in accord with the mark and space intelligence of the RTTY code. Frequency-shift keying (FSK) may be accomplished by varying the frequency of the transmitter oscillator in a stable manner between two chosen frequencies. The amount of shift must be held within close tolerances as the shift must match the frequency difference between the selective filters in the receiving terminal unit. The degree of frequency shift of the oscillator is, of course, multiplied by any factor of multiplication realized in succeeding doubler stages of the transmitter. A simple diode switch suitable for many variable-frequency oscillators is shown in figure 8. Older systems often made use of a reactance tube to obtain an adjustable shift.

RTTY Duty Cycle The *duty cycle* during an RTTY transmission is unity; that is, the average-to-peak power ratio is one. Most amateur transmitters, particularly SSB equipments, are designed with a speech duty cycle in mind and must be derated for RTTY service. Generally speaking, the duty cycle for RTTY is equal to twice the plate dissipation rating of the tube or tubes in the linear amplifier stage. Thus, if the linear amplifier has a pair of, say, 6LQ6 sweep tubes, having a plate dissipation rating of 30 watts each for continuous service, the maximum input to the amplifier for RTTY

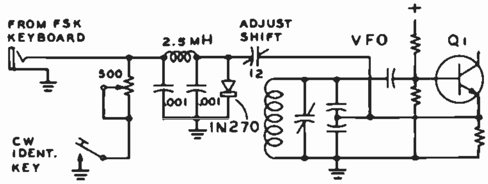


Figure 8

DIODE KEYER FOR FREQUENCY-SHIFT KEYING OF VFO

A simple diode switch may be used to vary the frequency of the transmitter in a stable manner between two chosen frequencies. The amount of shift must match the frequency difference between the selective filters in the receiving demodulators unit.

service is limited to about 60 watts per tube, or a total of 120 watts.

Auxiliary RTTY Equipment RTTY transmission by pre-punched tape is made possible by means of a *transmitter-distributor* (T-D) unit. This is an electromechanical device which senses perforations in a teleprinter tape and translates this information into electrical impulses of the five-unit teleprinter code at a constant speed (55-65 w.p.m. in the amateur radio service). The information derived from the punched tape by contact fingers is transmitted in the proper time sequence by a commutator-distributor driven at a constant speed by a synchronous motor (figure 9). Used in conjunction with the T-D is a *tape perforator* which punches the teleprinter code in a paper tape. The perforator operates mechanically from a teleprinter keyboard for originating messages. A *reperforator* may be connected to receiving equipment to "tape" an incoming message for storage or retransmission.

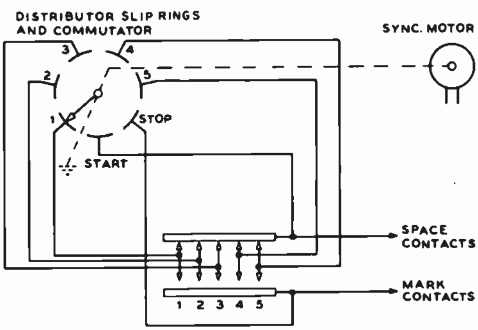


Figure 9

TRANSMITTER-DISTRIBUTOR (T-D) UNIT

T-D unit is electromechanical device which senses perforations in a teleprinter tape and translates this information into the electrical impulses of the teleprinter code. Information derived from the tape by contact fingers is transmitted in proper time sequence by a commutator-distributor driven by a constant-speed motor.

Audio Frequency-Shift Keying Audio frequency-shift keying (AFSK) is often used by radio amateurs on the vhf bands in order to avoid the problems of holding close radio-frequency stability. An audio oscillator is employed to generate a 2125-Hz tone (mark) and a 2975-Hz tone (space) when driven by the keyboard of a teleprinter, or by a tape T-D unit. The audio signal is then applied to the modulator of the vhf transmitter and the resulting amplitude-modulated signal is detected and put to use by an audio converter of the type shown in figure 7. The beat oscillator in the receiver is not used for this form of reception. AFSK is permitted only on those amateur bands on which A2 emission is authorized. A simple AFSK oscillator circuit is shown in figure 10.

Obtaining Teleprinter Machines Sources available to radio amateurs include several nonprofit RTTY societies, established in various areas of the United States for the purpose of disposing of teleprinter equipment discarded by commercial services. These societies can be contacted through active RTTY amateurs. The commercial services, including the Bell Telephone Company, generally cannot dispose of used equipment directly to radio amateurs. Commercial services should not be contacted regarding used teleprinters. Many radio amateurs, active in RTTY, rebuild machines from junked or damaged equipment at nominal cost. These amateurs are also an excellent source of maintenance support.

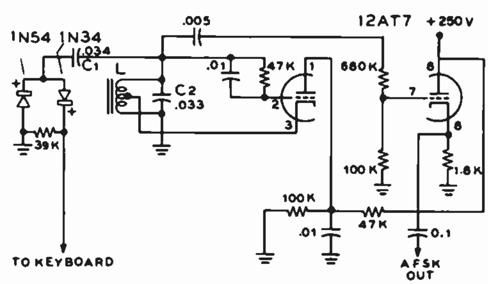


Figure 10

AFSK OSCILLATOR

Audio frequency-shift keying is often used on vhf bands to avoid problems of holding close radio-frequency stability. The L-C_1 circuit is tuned to 2975 Hz (with keyboard open). Closing the keyboard parallels capacitor C_1 and lowers the oscillator frequency to 2125 Hz. The coil L is an 88 mH toroid (with about 44" of wire removed). Capacitors C_1 and C_2 are high quality paper or mylar. Compression mica capacitors may be used as padders to place the oscillator on the correct frequencies.

14-5 Slow-Scan Television

Slow-scan television (*SSTV*) is a narrow-band system for transmitting video images approved by the F.C.C. for use in various amateur bands. Signal bandwidth of an SSTV image is limited to 3 kHz. This transformation is commonly accomplished by converting the video information to a varying tone which is fed into the audio system of an amateur transmitter. Either a-m, SSB or f-m transmission may be used. SSB is used for SSTV on the h-f bands and f-m on the vhf bands. Because of the restricted bandwidth, the video signal may be received on a communication receiver and may be preserved on an audio tape recorder running at 3¾ inches per second, or more.

A representative SSTV signal consists of a 1500-Hz tone which is shifted down to 1200-Hz for sync information and modulated upward to 2300-Hz for video (picture) information. The 1500-Hz frequency represents the *black level* and the 2300-Hz frequency is the *white level*, with tones in between giving shades of gray. The sync pulse durations are 5 milliseconds for the horizontal and 30 milliseconds for the vertical. The scanning sequence is left to right and top to bottom. Normally, 120 lines are scanned per frame, with an aspect ratio of 1:1. For 60-Hz areas, the horizontal sweep rate is 15 Hz and the vertical sweep rate is 6 to 8 seconds. Since picture transmission time is only a few seconds, it permits rapid alternation of voice and picture transmission over the same circuit.

Picture Transmission Slow-scan picture transmission evolved out of facsimile transmitting techniques wherein video images were transmitted over wire lines. A block diagram of a typical SSTV picture generator is shown in figure 11. The system uses a cathode-ray tube "flying spot" scanner to develop a 120-line picture, scanned once every 8 seconds or so. The image (in this case a negative or a transparency) is directly scanned by the *raster image* projected from a very short-persistence cathode-ray tube. The flying spot sweeps quickly across the face of the CRT and does not leave a "tail" of undecayed brightness behind it as does a medium- or long-persistence tube. The spot faintly illuminates a pickup device (or *scanner*), which is usually a *photomultiplier* tube. The small photocathode current is amplified over 40,000 times by the secondary emission action of the tube. The output voltage of the photomultiplier tube is thus a video signal whose instantaneous amplitude fol-

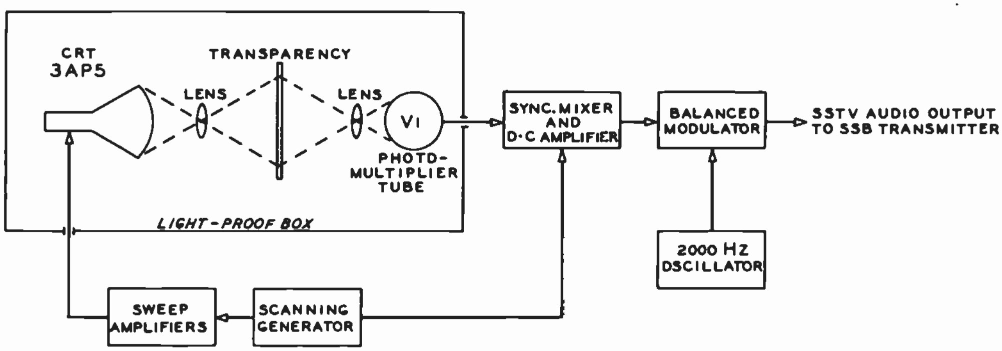


Figure 11

SLOW SCAN TV (SSTV) PICTURE GENERATOR

Cathode-ray tube (3AP5) serves as a "flying spot" scanner to develop a 120-line picture, scanned once every eight seconds or so. The image (a negative or transparency) is directly scanned by the raster image projected from the short-persistence cathode-ray tube. The scanned image is picked up by a photomultiplier tube, the output voltage of which is a video signal whose instantaneous amplitude follows the variations in picture brightness as the transparency is scanned.

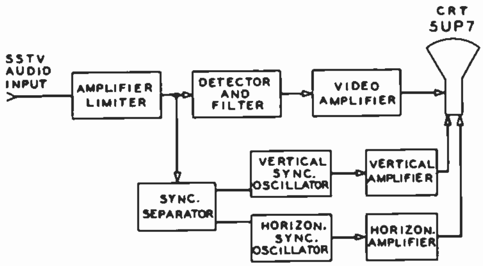


Figure 12

SSTV RECEIVING ADAPTER

Picture is developed on a long-persistence (5UP7) cathode-ray tube. SSTV signal is amplified, limited, detected, and used to intensity modulate the CRT. Sync pulses are separated from the SSTV signal to provide scanning information, including blanking and retrace signals.

lows the variations in picture brightness as the transparency is scanned.

The photo pickup assembly is contained within a light-tight box. The raster signal is derived from a scanning generator which supplies the vertical and horizontal sweep pulses.

The video signal from the photomultiplier tube is passed through a d-c amplifier and into a balanced modulator which mixes the video signal with the 2000-Hz subcarrier oscillator and suppresses the video signal, whose components lie between 0 and 1000 Hz. The output of the modulator is the sideband signal which is applied directly to the SSB transmitter as an audio signal.

Picture Reception The audio-frequency output of a communications receiver may be used to receive the SSTV signal. A block diagram of a typical SSTV adapter is shown in figure 12. A long-persistence 5UP7 cathode-ray tube (CRT) is used. The slow-scan audio signal from the receiver is amplified, limited, and fed to a detector and a low-pass filter which removes the audio components above 1000 Hz. The resulting video signal is used to intensity-modulate the long-persistence CRT. The output of the amplifier-limiter is also fed to a sync separator which separates the sync pulses from the composite sync and video signal. The 30-ms and 5-ms pulses are separated for the two scanning signals in an integrator. Blanking and retrace circuitry are excited at this point and the pulses drive multivibrators which supply the deflection signals to the picture tube.

CHAPTER FIFTEEN

Amplitude Modulation and Audio Processing

When voice, music, video, or other intelligence is superimposed on a radio frequency carrier by means of a corresponding variation in the *amplitude* of the radio frequency output of a transmitter, *amplitude modulation* is the result. Telegraph keying of a c-w transmitter is the simplest form of amplitude modulation, while video modulation in a television transmitter represents a highly complex form.

Low-level amplitude modulation (a-m) is commonly used to generate an SSB signal, the a-m wave being passed through a highly selective filter to remove the carrier and unwanted sideband. Systems for modulating the amplitude of a *high-level* carrier envelope in accordance with voice, music, or similar types of complicated audio waveforms are many and varied, and will be discussed in this chapter.

15-1 Sidebands

Modulation is essentially a form of *mixing*, or *combining*, already covered in a previous chapter. To transmit voice at radio frequencies by means of amplitude modulation, the voice frequencies are mixed with a radio-frequency carrier so that the voice frequencies are converted to radio-frequency *sidebands*.

Even though the amplitude of radio-frequency voltage representing the composite signal (resultant of the carrier and sidebands, called the *envelope*) will vary from zero to twice the unmodulated signal value during full modulation, the amplitude of the *carrier* component does not vary. Also, as long as the amplitude of the modulating voltage does not vary, the amplitude of the sidebands will remain constant. For this to be apparent, however, it is necessary to measure the amplitude of each component with a highly selective filter. Otherwise, the measured

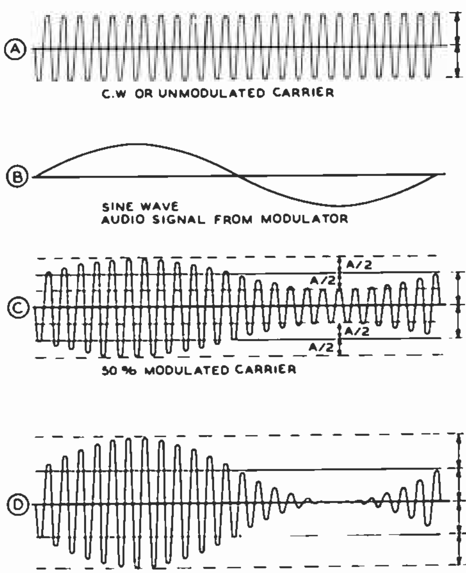


Figure 1

AMPLITUDE-MODULATED WAVE

Top drawing A represents an unmodulated carrier wave; B shows the audio output of the modulator. Drawing C shows the audio signal impressed on the carrier wave to the extent of 50 percent modulation; D shows the carrier with 100 percent amplitude modulation.

power or voltage will be a *resultant* of two or more of the components, and the amplitude of the resultant will vary at the modulation rate.

If a carrier frequency of 5000 kHz is modulated by a pure tone of 1000 Hz, or 1 kHz, two sidebands are formed: one at 5001 kHz (the sum frequency) and one at 4999 kHz (the difference frequency). The frequency of each sideband is independent of the amplitude of the modulating tone, or *modulation percentage;* the frequency of each sideband is determined only by the frequency of the modulating tone. This assumes, of course, that the transmitter is not modulated in excess of its linear capability.

When the modulating signal consists of multiple frequencies, as is the case with voice or music modulation, two sidebands will be formed by each modulating frequency (one on each side of the carrier), and the radiated signal will consist of a *band* of frequencies. The *bandwidth,* or *channel,* taken up in the frequency spectrum by a conventional double-sideband amplitude-modulated signal, is equal to twice the highest modulating frequency. For example, if the highest modulating frequency is 5000 Hz, then the signal (assuming modulation of complex and varying waveform) will occupy a band extending from 5000 Hz below the carrier to 5000 Hz above the carrier.

Frequencies up to at least 2500 Hz, and preferably 3500 Hz, are necessary for good speech intelligibility. If a filter is incorporated in the audio system to cut out all frequencies above approximately 3000 Hz, the bandwidth of an a-m signal can be limited to 6 kHz without a significant loss in intelligibility. However, if harmonic distortion is introduced subsequent to the filter, as would happen in the case of an overloaded modulator or overmodulation of the carrier, new frequencies will be generated and the signal will occupy a band wider than 6 kHz.

15-2 Mechanics of Modulation

A c-w or unmodulated r-f carrier wave is represented in figure 1A. An audio-frequency sine wave is represented by the curve of figure 1B. When the two are combined or "mixed," the carrier is said to be amplitude modulated, and a resultant similar to 1C or 1D is obtained. It should be noted that under modulation, each half cycle of r-f voltage differs slightly from the preceding one and the following one; therefore at no time during modulation is the r-f waveform a pure sine wave. This is simply another way of saying that during modulation, the transmitted r-f energy no longer is confined to a single radio frequency.

It will be noted that the *average* amplitude of the peak r-f voltage, or modulation envelope, is the same with or without modulation. This simply means that the modulation is symmetrical (assuming a symmetrical modulating wave) and that for distortionless modulation the upward modulation is limited to a value of twice the unmodulated carrier wave amplitude because the amplitude cannot go below zero on downward portions of the modulation cycle. Figure 1D illustrates the maximum obtainable distortionless modulation with a sine modulating wave, the r-f voltage at the peak of the r-f cycle varying from zero to twice the unmodulated value, and the r-f power varying from zero to four times the unmodulated value (the power varies as the square of the voltage).

While the average r-f *voltage* of the modulated wave over a modulation cycle is the same as for the unmodulated carrier, the average *power* increases with modulation. If the radio-frequency power is integrated over the audio cycle, it will be found with 100 percent sine-wave modulation the average r-f power has increased 50 percent. This additional power is represented by the sidebands, because, as previously mentioned, the carrier power does not vary under modulation. Thus, when a 100-watt carrier is modulated 100 percent by a sine wave, the total r-f power is 150 watts—100 watts in the carrier and 25 watts in each of the two sidebands.

Modulation Percentage So long as the *relative proportion* of the various sidebands making up voice modulation is maintained, the signal may be received and detected without distortion. However, the higher the average amplitude of the sidebands, the greater the audio signal produced at the receiver. For this reason it is desirable to increase the *modulation percentage,* or degree of modulation, to the point where maximum peaks just hit 100 percent. If the modulation percentage is increased so that the peaks exceed this value, distortion is introduced, and if carried very far, bad interference to signals on nearby channels will result.

Modulation Measurement The amount by which a carrier is being modulated may be expressed either as a *modulation factor*, varying from zero to 1.0 at maximum modulation, or as a percentage. The percentage of modulation is equal to 100 times the modulation factor. Figure 2A shows a carrier wave modulated by a sine-wave audio tone. A picture such as this might be seen on the screen of a cathode-ray oscilloscope with sawtooth sweep on the horizontal plates and the modulated carrier impressed on the vertical plates. The same carrier without modulation would appear on the oscilloscope screen as figure 2B.

The percentage of modulation of the positive peaks and the percentage of modulation of the negative peaks can be determined separately from two oscilloscope pictures such as shown.

The modulation factor of the positive peaks may be determined by the formula:

$$M = \frac{E_{max} - E_{car}}{E_{car}}$$

The factor for negative peaks may be determined from the formula:

$$M = \frac{E_{car} - E_{min}}{E_{car}}$$

In the above two formulas E_{max} is the maximum carrier amplitude with modulation and E_{min} is the minimum amplitude; E_{car} is the steady-state amplitude of the carrier without modulation.

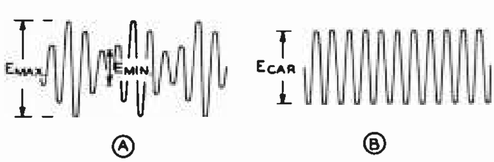


Figure 2

GRAPHICAL DETERMINATION OF MODULATION PERCENTAGE

The procedure for determining modulation percentage from the peak voltage points indicated is discussed in the text.

If the modulating voltage is symmetrical, such as a sine wave, and modulation is accomplished without the introduction of distortion, then the percentage modulation will be the same for both negative and positive peaks. However, the distribution and phase relationships of harmonics in voice and music waveforms are such that the percentage modulation of the negative modulation peaks may exceed the percentage modulation of the positive peaks, or vice versa. The percentage modulation when referred to without regard to polarity is an indication of the average of the negative and positive peaks.

Modulation Capability The *modulation capability* of a transmitter is the maximum percentage to which that transmitter may be modulated before spurious sidebands are generated in the output or before the distortion of the modulating waveform becomes objectionable. The highest modulation capability which *any* transmitter may have on the *negative* peaks is 100 percent. The maximum permissible modulation of many transmitters is less than 100 percent, especially on positive peaks.

Speech Waveform Dissymmetry The manner in which the human voice is produced by the vocal cords gives rise to a certain dissymmetry in the waveform of voice sounds when they are picked up by a good quality microphone. This is especially pronounced in the male voice, and more so on certain voice sounds than on others. The result of this dissymmetry in the waveform is that the voltage peaks on one side of the average value of the wave will be considerably greater, often two or three times as great, as the voltage excursions on the other side of the zero axis. The *average* value of voltage on both sides of the wave is, of course, the same.

As a result of this dissymmetry in the male voice waveform, there is an optimum polarity of the modulating voltage that must be observed if maximum sideband energy is to be obtained without negative peak clipping and generation of *splatter* on adjacent channels.

The use of the proper polarity of the incoming speech wave in modulating a transmitter can afford an increase of approximately two to one in the amount of speech audio power which may be placed on the carrier of an amplitude-modulated transmitter for the same amount of sideband splatter.

15-3 Systems of Amplitude Modulation

There are many different systems and methods for amplitude-modulating a carrier, but most may be grouped under three general classifications: (1) *variable-efficiency* systems in which the average input to the stage remains constant with and without modulation and the variations in the efficiency of the stage in accordance with the modulating signal accomplish the modulation; (2) *constant-efficiency* systems in which the input to the stage is varied by an external source of modulating energy to accomplish the modulation; and (3) so-called *high-efficiency* systems in which circuit complexity is increased to obtain high plate-circuit efficiency in the modulated stage without the requirement of an external high-level modulator. The various systems under each classification have individual characteristics which make certain ones best suited to particular applications.

Variable-Efficiency Modulation Since the *average* input remains constant in a stage employing variable-efficiency modulation, and since the average power output of the stage increases with modulation, the additional average power output from the stage *with* modulation must come from the plate dissipation of the tubes in the stage. Hence, for the best relation between tube cost and power output, the tubes employed should have as high a plate dissipation rating per dollar as possible.

The plate efficiency in such an amplifier is doubled when going from the unmodulated condition to the peak of the modulation cycle. Hence, the unmodulated efficiency of such an amplifier *must always be less than 45 percent*, since the maximum peak efficiency obtainable in a conventional amplifier is in the vicinity of 90 percent. Since the peak efficiency in certain types of amplifiers will be as low as 60 percent, the unmodulated efficiency in such amplifiers will be in the vicinity of 30 percent.

There are many systems of efficiency modulation, but they *all* have the general limitation discussed in the previous paragraph —so long as the carrier amplitude is to remain constant with and without modulation, the efficiency at carrier level must be not greater than one-half the peak modulation efficiency, if the stage is to be capable of 100-percent modulation.

The Class-B A-M Linear Amplifier This is the simplest practicable type amplifier for an amplitude-modulated wave or a single-sideband signal. The system requires that excitation, grid bias, and loading must be carefully controlled to preserve the linearity of the stage. Also, the grid circuit of the tube, in the usual application where grid current is drawn on peaks, presents a widely varying value of load impedance to the source of excitation. Hence it is necessary to include some sort of *swamping resistor* to reduce the effect of grid-impedance variations with modulation. If such a swamping resistance across the grid tank is not included, or is too high in value, the positive modulation peaks of the incoming modulated signal will tend to be flattened with resultant distortion of the wave being amplified.

Since a class-B a-m linear amplifier is biased to *extended* cutoff with no excitation (the grid bias at extended cutoff will be approximately equal to the plate voltage divided by the amplification factor for a triode, and will be approximately equal to the screen voltage divided by the grid-screen μ factor for a tetrode or pentode) the plate current will essentially flow in 180-degree pulses. Due to the relatively large operating angle of plate current flow the theoretical peak plate efficiency is limited to 78.5 percent, with 65 to 70 percent representing a range of efficiency normally attainable.

The carrier power output from a class-B linear amplifier of a normal 100 percent modulated a-m signal will be about one-half the rated plate dissipation of the stage, with optimum operating conditions. The peak output from a class-B linear, which represents the maximum-signal output as a single-sideband amplifier, or peak output with a 100 percent a-m signal, will be about twice the plate dissipation of the tubes in the stage. Thus the carrier-level input power to a class-B linear should be about 1.5 times the rated plate dissipation of the stage.

Adjustment of a Class-B A-M Linear Amplifier With grid bias adjusted to the correct value, and with provision for varying the excitation voltage to the stage and the loading of the plate circuit, a fully modulated signal is applied to the grid circuit of the stage. Then with an oscilloscope coupled to the output of the stage, excitation and loading are varied until the stage is drawing the normal plate input and the output waveshape is a good replica of the input signal. The adjustment procedure normally will require a succession of approximations, until the optimum set of adjustments is attained. Then the modulation being applied to the input signal should be removed to check the linearity. With modulation removed, in the case of a 100 percent a-m signal, the input to the stage should remain constant, and the peak output of the r-f envelope should fall to one-half the value obtained on positive modulation peaks.

Class-C A-M Grid Modulation One effective system of efficiency modulation for communications work is class-C control-grid bias modulation.

Class-C grid modulation requires high plate voltage on the modulated stage if maximum output is desired. The plate voltage is normally run about 50 percent higher than for maximum output with plate modulation.

The driving power required for operation of a grid-modulated amplifier under these conditions is somewhat more than is required for operation at lower bias and plate voltage, but the increased power output obtainable overbalances the additional excitation requirement. Actually, almost half as much excitation is required as would be needed if the same stage were to be operated as a class-C plate-modulated amplifier. A resistor across the grid tank of the stage serves as *swamping* to stabilize the r-f driving voltage. At least 50 percent of the output of the driving stage should be dissipated in this swamping resistor under carrier conditions.

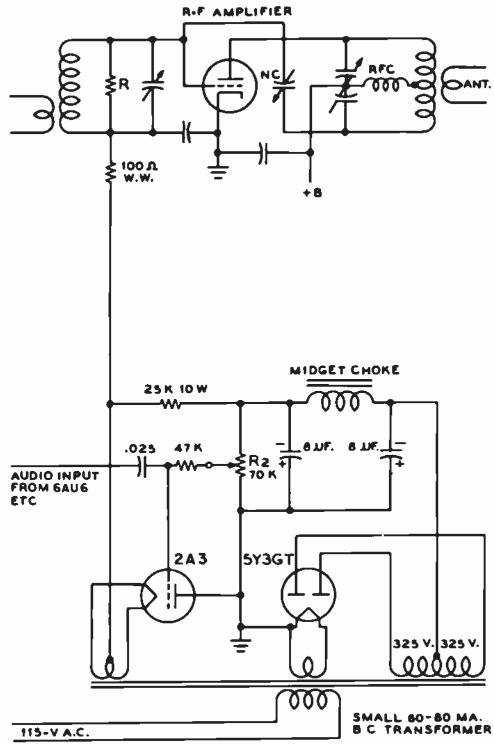


Figure 3

GRID-BIAS MODULATOR CIRCUIT

A comparatively small amount of audio power will be required to modulate the amplifier stage 100 percent. An audio amplifier having 20 watts output will be sufficient to modulate an amplifier with one kilowatt input. Proportionately smaller amounts of audio will be required for lower-powered stages. However, the audio amplifier that is being used as the grid modulator should, in any case, either employ low-plate-resistance tubes such as 2A3's, employ degenerative feedback from the output stage to one of the preceding stages of the speech amplifier, or be resistance loaded with a resistor across the secondary of the modulation transformer. This provision of low driving impedance in the grid modulator is to ensure good regulation in the audio driver for the grid-modulated stage because the grid impedance of the stage varies widely over the audio cycle.

A practical circuit for obtaining grid-bias modulation is shown in figure 3. The modulator and bias regulator tube have been combined in a single 2A3 tube.

The regulator-modulator tube operates as a cathode-follower. The average d-c voltage on the control grid is controlled by the 70,000-ohm wirewound potentiometer and this potentiometer adjusts the average grid bias on the modulated stage. However, a-c signal voltage is also impressed on the control grid of the tube and since the cathode follows this a-c wave the incoming speech wave is superimposed on the average grid bias, thus effecting grid-bias modulation of the r-f amplifier stage. An audio voltage swing is required on the grid of the 2A3 of approximately the same peak value as will be required as bias-voltage swing on the grid-bias modulated stage.

With the normal amount of comparatively tight antenna coupling to the modulated stage, an unmodulated carrier efficiency of 40 percent can be obtained, with substantially distortion-free modulation up to practically 100 percent.

Tuning the Grid-Bias Modulated Stage The most satisfactory procedure for tuning a stage for grid-bias modulation of the class-C type is as follows. The amplifier should first be neutralized, and any possible tendency toward parasitics under any condition of operation should be eliminated. Then the antenna should be coupled to the plate circuit, the grid bias should be run up to the maximum available value, and the plate voltage and excitation should be applied. The grid-bias voltage should then be reduced until the amplifier draws the approximate amount of plate current it is desired to run, and modulation corresponding to about 80 percent is then applied. If the plate current kicks up when modulation is applied, the grid bias should be reduced; if the plate meter kicks down, increase the grid bias.

When the amount of bias voltage has been found (by adjusting the fine control (R_2) on the bias supply) where the plate meter remains constant with modulation, it is more than probable that the stage will be drawing either too much or too little input. The antenna coupling should then be either increased or decreased (depending on whether the input was too little or too much, respectively) until the input is more nearly the correct value. The bias should then be readjusted until the plate meter remains constant with modulation as before. By slight jockeying back and forth of antenna coupling and grid bias, a point can be reached where the tubes are running at rated plate dissipation, and where the plate milliammeter on the modulated stage remains substantially constant with modulation.

Screen-Grid Modulation Amplitude modulation may be accomplished by varying the screen-grid voltage in a class-C amplifier which employs a pentode, beam tetrode, or other type of screen-grid tube. The modulation obtained in this way is not especially linear as the impedance of the screen grid with respect to the modulating signal is nonlinear. However, *screen-grid modulation* does offer other advantages and the linearity is quite adequate for communications work.

There are two significant and worthwhile advantages of screen-grid modulation for communications work: (1) The excitation requirements for an amplifier which is to be screen modulated are not at all critical, and good regulation of the excitation voltage is not required. The normal rated grid-circuit operating conditions specified for class-C c-w operation are quite adequate for screen-grid modulation. (2) The audio modulating power requirements for screen-grid modulation are relatively low.

A screen-grid modulated r-f amplifier operates as an efficiency-modulated amplifier, the same as does a class-B linear amplifier and a grid-modulated stage. Hence, *plate circuit* loading is relatively critical as in any efficiency-modulated stage, and must be adjusted to the correct value if normal power output with full modulation capability is to be obtained. As in the case of any efficiency-modulated stage, the operating efficiency at the peak of the modulation cycle will be between 70 and 80 percent, with efficiency at the carrier level (if the stage is operating in the normal manner with full carrier) about half of the peak-modulation value.

Screen-Grid Impedance Instead of being linear with respect to modulating voltage, as is the plate circuit of a plate-modulated class-C amplifier, the screen grid presents approximately a square-law impedance to the modulating signal over the region of signal excursion where the screen is positive with respect to ground.

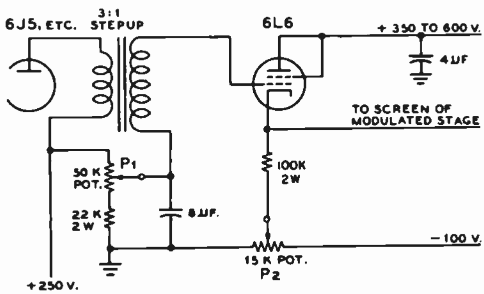


Figure 4

CATHODE-FOLLOWER SCREEN-MODULATION CIRCUIT

Another factor which must be considered in the design of a screen-modulated stage, if full modulation is to be obtained, is that the power output of a screen-grid stage with zero screen voltage is still relatively large. Hence, if anything approaching full modulation on negative peaks is to be obtained, the screen potential must be made negative with respect to ground on negative modulation peaks. In the usual types of beam tetrode tubes the screen potential must be 20 to 50 volts negative with respect to ground before cutoff of output is obtained. This condition further complicates the problem of obtaining good linearity in the audio modulating voltage for the screen-modulated stage, since the screen voltage must be driven negative with respect to ground over a portion of the cycle. Hence the screen draws *no* current over a portion of the modulating cycle, and over the major portion of the cycle when the screen does draw current, it presents approximately a square-law impedance.

The cathode-follower modulator circuit of figure 4 is capable of giving good quality screen-grid modulation, and in addition the circuit provides convenient adjustments for the carrier level and the output level on *negative* modulation peaks. This latter control (P_2), allows the amplifier to be adjusted in such a manner that negative-peak clipping cannot take place, yet the negative modulation peaks may be adjusted to a level just above that at which sideband splatter will occur.

The voltage applied to the plate of the cathode follower should be about 100 volts greater than the rated screen voltage for the tetrode tube as a c-w class-C amplifier. Then potentiometer (P_1) in figure 4 should be adjusted until the carrier-level screen voltage on the modulated stage is about one-half the rated screen voltage specified for the tube as a class-C c-w amplifier. The current taken by the screen of the modulated tube under carrier conditions will be about one-fourth the normal screen current for c-w operation.

The loading of the modulated stage should be adjusted until the input to the tube is about 50 percent greater than the rated plate dissipation of the tube or tubes in the stage. If the carrier-level screen voltage value is correct for linear modulation of the stage, the loading will have to be somewhat *greater* than that amount of loading which gives maximum output from the stage. The stage may then be modulated by applying an audio signal to the grid of the cathode-follower modulator, while observing the modulated envelope on an oscilloscope.

Note that the correct plate current for an efficiency-modulated amplifier is only slightly less than the out-of-resonance plate current of the stage. Hence carrier-level screen voltage must be low so that the out-of-resonance plate current will not be too high, and relatively heavy antenna coupling must be used so that the operating plate current will be near the out-of-resonance value, and so that the operating input will be slightly greater than 1.5 times the rated plate dissipation of the tube or tubes in the stage. Since the carrier efficiency of the stage will be only 35 to 40 percent, the tubes will be operating with plate dissipation of approximately the rated value without modulation.

Speech Clipping in the Modulated Stage The maximum r-f output of an efficiency-modulated stage is limited by the maximum permissible plate-voltage swing on positive modulation peaks. In the modulation circuit of figure 4 the *minimum* output is limited by the minimum voltage which the screen will reach on a negative modulation peak, as set by potentiometer P_2. Hence the screen-grid-modulated stage, when using the modulator of figure 4, acts effectively as a speech clipper, provided the modulating signal amplitude is not too much more than that value which will accomplish full modulation. With correct adjustments of the operating conditions of the stage it can be made to clip positive and negative modulation peaks symmetrically. However, the inherent peak-clipping ability of the stage should *not* be relied upon as a means of obtaining a large amount of speech compression, since excessive audio distortion and excessive screen current on the modulated stage will result.

Suppressor-Grid Modulation Still another form of efficiency modulation may be obtained by applying the audio modulating signal to the suppressor grid of a pentode class-C r-f amplifier (figure 5). Basically, *suppressor-grid modulation* operates in the same general manner as other forms of efficiency modulation; carrier plate-circuit efficiency is about 35 percent, and antenna coupling must be rather heavy. However, suppressor-grid modulation has one sizeable disadvantage, in addition to the fact that pentode tubes are not nearly so widely used as beam tetrodes which of course do not have the suppressor element. This disadvantage is that the screen-grid current to a suppressor-grid modulated amplifier is rather high. The high screen current is a natural consequence of the rather high negative bias on the suppressor grid, which reduces the plate-voltage swing and plate current with a resulting increase in the screen current.

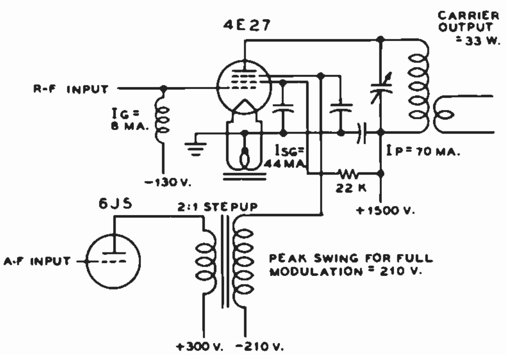


Figure 5

AMPLIFIER WITH SUPPRESSOR-GRID MODULATION

Recommended operating conditions for linear suppressor-grid modulation of a 4E27/5-125B stage are given on the drawing.

In tuning a suppressor-grid modulated amplifier, the grid bias, grid current, screen voltage, and plate voltage are about the same as for class-C c-w operation of the stage. But the suppressor grid is biased negatively to a value which reduces the plate-circuit efficiency to about one-half the maximum obtainable from the particular amplifier, with antenna coupling adjusted until the plate input is about 1.5 times the rated plate dissipation of the stage. It is important that the input to the screen grid be measured to make sure that the rated screen dissipation of the tube is not being exceeded. Then the audio signal is applied to the suppressor grid. In the normal application the audio voltage swing on the suppressor will be somewhat greater than the negative bias on the element. Hence suppressor-grid current will flow on modulation peaks, so that the source of audio signal voltage must have good regulation.

15-4 Input Modulation Systems

Constant-efficiency variable-input modulation systems operate by virtue of the addition of external power to the modulated stage to effect the modulation. There are two general classifications that come under this heading; those systems in which the additional power is supplied as audio-frequency energy from a modulator (usually called *plate-modulation systems*) and those systems in which the additional power to effect modulation is supplied as direct current from the plate supply.

Under the former classification comes *Heising* modulation (probably the oldest type of modulation to be applied to a continuous carrier), class-B plate modulation, and cathode modulation. These types of modulation are by far the easiest to get into operation, and they give a very good ratio of power input to the modulated stage to power output; 65 to 80 percent efficiency is the general rule. It is for these two important reasons that these modulation systems, particularly class-B plate modulation, are at present the most popular for a-m communications work.

Modulation systems coming under the second classification have been widely applied to broadcast work. There are quite a few systems in this class. Two of the more widely used are the *Doherty* linear amplifier, and the *Terman-Woodyard* high-efficiency grid-modulated amplifier. Both systems operate by virtue of a carrier amplifier and a peak amplifier connected together by electrical quarter-wave lines. They will be described later in this section.

Plate Modulation Plate modulation is the application of the audio power to the *plate circuit* of an r-f amplifier. The r-f amplifier must be operated class C for this type of modulation in order to obtain a radio-frequency output which changes in exact accord with the variation in plate voltage. *The r-f amplifier is 100 percent modulated when the peak a-c voltage from the modulator is equal to the d-c voltage applied to the r-f tube.* The positive peaks of audio voltage increase the instantaneous plate voltage on the r-f tube to *twice* the d-c value, and the negative peaks reduce the voltage to zero.

The instantaneous plate *current* to the r-f stage also varies in accord with the modulating voltage. The peak alternating current in the output of a modulator must be equal to the d-c plate current of the class-C r-f stage at the point of 100 percent modulation. This combination of change in audio voltage and current can be most easily referred to in terms of *audio power in watts.*

In a sinusoidally modulated wave, the antenna current increases approximately 22 percent for 100 percent modulation with a pure tone input; an r-f meter in the antenna circuit indicates this increase in antenna current. The *average power* of the r-f wave increases 50 percent for 100 percent modulation, the efficiency remaining constant.

This indicates that in a plate-modulated radiotelephone transmitter, the audio-frequency channel must supply this additional 50 percent increase in average power for sine-wave modulation. If the power input to the modulated stage is 100 watts, for example, the *average power* will increase to 150 watts at 100 percent modulation, and this additional 50 watts of power must be supplied by the *modulator* when plate modulation is used. The actual antenna power is a constant percentage of the total value of input power.

By properly matching the plate impedance of the r-f tube to the output of the modulator, the ratio of voltage and current swing to d-c voltage and current is automatically obtained. The modulator should have a peak voltage output equal to the average d-c plate voltage on the modulated stage. The modulator should also have a *peak* power output equal to the d-c plate input power to the modulated stage.

The *average* power output of the modulator will depend on the type of waveform. If the amplifier is being Heising modulated by a class-A stage, the modulator must have an average power output capability of one-half the input to the class-C stage. If the modulator is a class-B audio amplifier, the average power required of it may vary from one-quarter to more than one-half the class-C input depending on the waveform. However, the *peak* power output of any modulator must be equal to the class-C input to be modulated.

Heising Modulation *Heising* modulation is the oldest system of plate modulation, and usually consists of a class-A audio amplifier coupled to the r-f amplifier by means of a modulation choke, as shown in figure 6.

The d-c plate voltage and plate current of the r-f amplifier must be adjusted to a value which will cause the plate impedance to match the output of the modulator, since the modulation choke gives a 1-to-1 coupling ratio. A series resistor, bypassed for audio frequencies by means of a capacitor, must be connected in series with the plate of the r-f amplifier to obtain modulation up to 100 percent. The peak output voltage of a class-A amplifier does not reach a value equal to the d-c voltage applied to the amplifier and, consequently, the d-c plate voltage impressed across the r-f tube must be reduced to a value equal to the maximum available a-c peak voltage if 100% modulation is to be obtained.

Class-B Plate Modulation High-level class-B plate modulation is the least expensive method of plate modulation. Figure 7 shows a conventional class-B plate-modulated class-C amplifier.

The statement that the modulator output power must be one-half the class-C input for

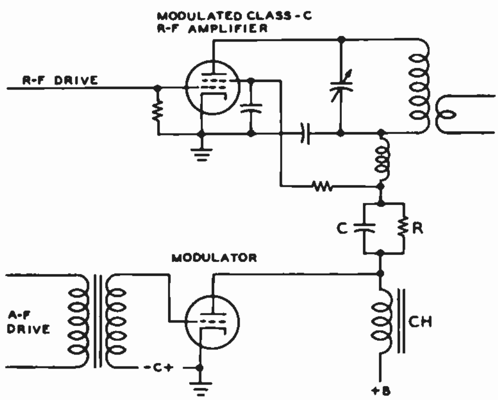


Figure 6

HEISING PLATE MODULATION

This type of modulation was the first form of plate modulation. It is sometimes known as "constant-current" modulation. Because of the effective 1:1 ratio of the coupling choke, it is impossible to obtain 100 percent modulation unless the plate voltage to the modulated stage is dropped slightly by resistor R. The capacitor (C) merely bypasses the audio around R, so that the full a-f output voltage of the modulator is impressed on the class-C stage.

100 percent modulation is correct only if the waveform of the modulating power is a *sine wave.* Where the modulator waveform is unclipped speech waveforms, the average modulator power for 100 percent modulation is considerably less than one-half the class-C input. A detailed discussion of modulation transformer calculations is given in Chapter Six.

Power Relations in Speech Waveforms It has been determined experimentally that the ratio of peak-to-average power in a speech waveform is approximately 4 to 1 as contrasted to a ratio of 2 to 1 in a sine wave. This is due to the high harmonic

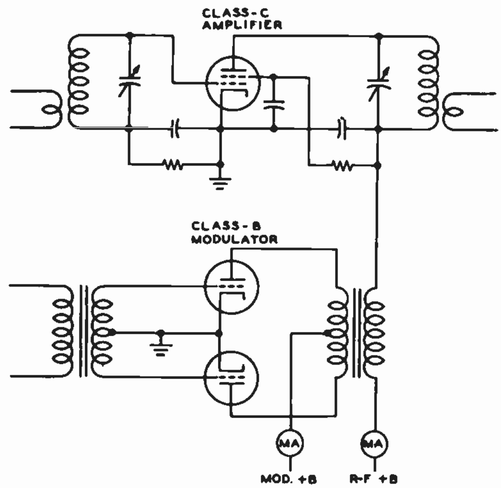


Figure 7

CLASS-B PLATE MODULATION

This type of modulation is the most flexible in that the loading adjustment can be made in a short period of time and without elaborate test equipment after a change in operating frequency of the class-C amplifier has been made.

content of such a waveform, and to the fact that this high harmonic content manifests itself by making the wave unsymmetrical and causing sharp peaks or "fingers" of high energy content to appear. Thus for unclipped speech, the *average* modulator plate current, plate dissipation, and power output are approximately one-half the sine wave values for a given *peak* output power.

For 100 percent modulation, the *peak* (instantaneous) audio power must equal the class-C input, although the average power for this value of peak varies widely depending on the modulation waveform, being greater than 50 percent for speech that has been clipped and filtered, 50 percent for a sine wave, and about 25 percent for typical unclipped speech tones.

Plate-and-Screen Modulation When *only* the plate of a screen-grid tube is modulated, it is difficult to obtain high-percentage linear modulation under ordinary conditions. The plate current of such a stage is not linear with plate voltage. However, if the screen is modulated simultaneously with the plate, the instantaneous screen voltage drops in proportion to the drop in the plate voltage, and linear modulation can then be obtained. Four satisfactory circuits for accomplishing combined plate and screen modulation are shown in figure 8.

The screen r-f bypass capacitor (C_2) should not have a greater value than 0.005 μfd, preferably not larger than 0.001 μfd. It should be large enough to bypass effectively all r-f voltage without short-circuiting high-frequency audio voltages. The plate bypass capacitor can be of any value from 0.002 μfd to 0.005 μfd. The screen-dropping resistor (R_1) should reduce the applied high voltage to the value specified for operating the particular tube in the circuit.

Figure 8C shows another method which uses a third winding on the modulation transformer, through which the screen grid is connected to a low-voltage power supply. The ratio of turns between the two output windings depends on the type of screen-grid tube which is being modulated. Normally it will be such that the screen voltage is being modulated 60 percent when the plate voltage is receiving 100 percent modulation.

If the screen voltage is derived from a dropping resistor (*not* a divider) that is bypassed for r.f. but not a.f., it is possible to secure quite good modulation by applying modulation only to the plate. Under these conditions, the screen tends to modulate itself, the screen voltage varying over the audio cycle as a result of the screen impedance increasing with plate voltage, and de-

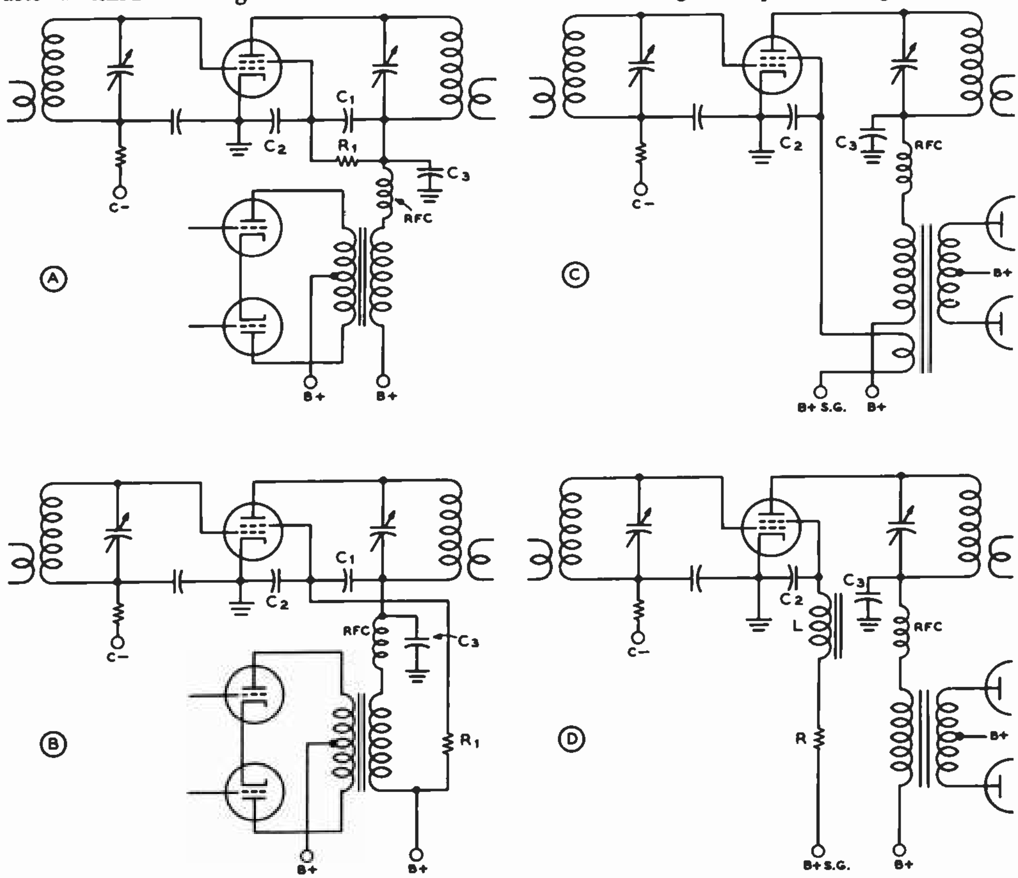


Figure 8

PLATE MODULATION OF A BEAM-TETRODE OR SCREEN-GRID TUBE

These alternative arrangements for plate modulation of tetrodes or pentodes are discussed in detail in the text. The arrangements shown at B or D are recommended for most applications.

creasing with a decrease in plate voltage. This circuit arrangement is illustrated in figure 8B.

A similar application of this principle is shown in figure 8D. In this case the screen voltage is fed directly from a low-voltage supply of the proper potential through choke L. A conventional filter choke having an inductance from 10 to 20 henrys will be satisfactory for L.

To afford protection of the tube when plate voltage is not applied but screen voltage is supplied from the exciter power supply, when using the arangement of figure 8D, a resistor of 3000 to 10,000 ohms can be connected in series with choke L. In this case the screen supply voltage should be at least 1.5 times as much as is required for actual screen voltage, and the value of resistor is chosen such that with normal screen current the drop through the resistor and choke will be such that normal screen voltage will be applied to the tube. When the plate voltage is removed the screen current will increase greatly and the drop through resistor R will increase to such a value that the screen voltage will be lowered to the point where the screen dissipation on the tube will not be exceeded. However, the supply voltage and value of resistor R must be chosen carefully so that the maximum rated screen dissipation cannot be exceeded. The maximum possible screen dissipation using this arrangement is equal to: $W = E^2/4R$ where E is the screen supply voltage and R is the combined resistance of the resistor in figure 8D and the d-c resistance of the choke (L). It is wise, when using this arrangement to check, using the above formula, to see that the value of W obtained is less than the maximum rated screen dissipation of the tube or tubes used in the modulated stage. This same system can of course also be used in figuring the screen supply circuit of a pentode or tetrode amplifier stage where modulation is not to be applied.

The modulation transformer for plate-and-screen modulation, when utilizing a dropping resistor as shown in figure 8A, is similar to the type of transformer used for any plate-modulated transmitter. The combined screen and plate current is divided into the plate voltage in order to obtain the class-C amplifier load impedance. The peak audio power required to obtain 100-percent modulation is equal to the d-c power input to the screen, screen resistor, and plate of the modulated r-f stage.

Cathode Modulation *Cathode modulation* offers a workable compromise between the good plate efficiency but expensive modulator of high-level plate modulation, and the poor plate efficiency but inexpensive modulator of grid modulation. Cathode modulation consists essentially of a mixture of the two.

The efficiency of the average well-designed plate-modulated transmitter is in the vicinity of 75 to 80 percent, with a compromise at perhaps 77.5 percent. On the other hand, the efficiency of a good grid-modulated transmitter may run from 28 to perhaps 40 percent with the average falling at about 34 percent. Now since cathode modulation consists of simultaneous grid and plate modulation, in phase with each other, we can theoretically obtain any efficiency from about 34 to 77.5 percent from our cathode-modulated stage, depending on the relative percentages of grid and plate modulation.

Since the system is a compromise between the two fundamental modulation arrangements, a value of efficiency approximately half way between the two would seem to be the best compromise. Experience has proved this to be the case. A compromise efficiency of about 56.5 percent, roughly half way between the two limits, has proved to be optimum. Calculation has shown that this value of efficiency can be obtained from a cathode-modulated amplifier when the audio-frequency modulating power is approximately 20 percent of the d-c input to the cathode-modulated stage.

Figure 9 shows the circuit of such a modulator, designed to cathode-modulate a class-C amplifier using high-mu triodes.

Cathode Modulation of Tetrodes Cathode modulation has not proved too satisfactory for use with beam tetrode tubes. This is a result of the small excitation and grid-swing requirements for such tubes, plus the fact that some means for holding the screen voltage at the potential of the cathode as far as audio is concerned is usually necessary. Because of these factors, cathode modulation is not recommended for use with tetrode r-f amplifiers.

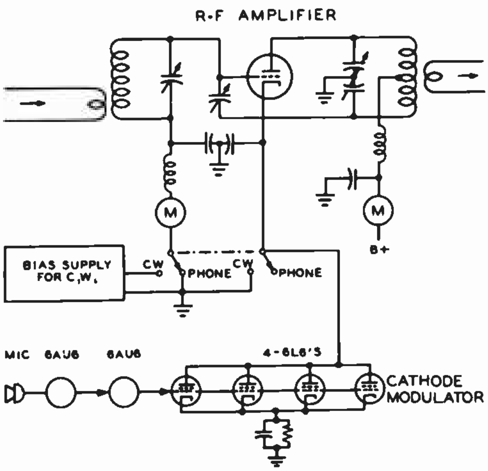


Figure 9

CATHODE-MODULATOR INSTALLATION SHOWING PHONE-CW TRANSFER SWITCH

15-5 The Doherty and the Terman-Woodyard Modulated Amplifiers

These two amplifiers will be described together since they operate on very similar principles. Figure 10 shows a greatly simplified schematic diagram of the operation of both types. Both systems operate by virtue of a *carrier tube,* (V_1 in both figures 10 and 11) which supplies the unmodulated carrier, and whose output is reduced to supply negative peaks, and a *peak tube,* (V_2) whose function is to supply approximately half the positive peak of the modulation cycle and whose additional function is to lower the load impedance on the carrier tube so that it will be able to supply the other half of the positive peak of the modulation cycle.

The peak tube is able to increase the output of the carrier tube by virtue of an impedance-inverting line between the plate circuits of the two tubes. This line is designed to have a characteristic impedance of one-half the value of load into which the carrier tube operates under the carrier conditions. Then a load of one-half the characteristic impedance of the quarter-wave line is coupled into the output. By experience with quarter-wave lines in antenna-matching circuits we know that such a line will vary the impedance at one end of the line in such a manner that the geometric mean between the two terminal impedances will be equal to the characteristic impedance of the line. Thus, if we have a value of load of *one-half* the characteristic impedance of the line at one end, the other end of the line will present a value of *twice* the characteristic impedance of the lines to carrier tube V_1.

This is the situation that exists under the carrier conditions when the peak tube merely floats across the load end of the line and contributes no power. Then as a positive peak of modulation comes along, the peak tube starts to contribute power to the load until at the peak of the modulation cycle it is contributing enough power so that the impedance at the load end of the line is equal to R, instead of the $R/2$ that is presented under the carrier conditions. This is true because at a positive modulation peak (since it is delivering full power) the peak tube subtracts a negative resistance of $R/2$ from the load end of the line.

Now, since under the peak condition of modulation the load end of the line is terminated in R ohms instead of $R/2$, the impedance at the *carrier-tube* will be *reduced* from $2R$ ohms to R ohms. This again is due to the impedance-inverting action of the line. Since the load resistance on the carrier tube has been reduced to half the carrier value, its output at the peak of the modulation cycle will be doubled. Thus we have the necessary condition for a 100-percent modulation peak; the amplifier will deliver four times as much power as it does under the carrier conditions.

On negative modulation peaks the peak tube does not contribute; the output of the carrier tube is reduced until, on a 100 percent negative peak, its output is zero.

The Electrical Quarter-Wave Line While an electrical quarter-wave line (consisting of a pi network with the inductance and capacitance units having a reactance equal to the characteristic impedance of the line) does have the desired impedance-inverting effect, it also has the undesirable effect of introducing a 90° phase shift across such a line. If the shunt elements are capacitances, the phase shift across the line lags by 90°; if they are inductances, the phase shift leads by 90°. Since there is an undesirable phase shift of 90° between the plate circuits of the carrier and peak tubes, an equal and opposite phase shift must be introduced in the exciting voltage of the grid circuits of the two tubes so that the resultant output in the plate circuit will be in phase. This additional phase shift has been indicated in figure 10 and a method of obtaining it has been shown in figure 11.

Comparison Between Doherty and Terman-Woodyard Amplifiers The difference between the Doherty linear amplifier and the Terman-Woodyard grid-modulated amplifier is the same as the difference between any linear and grid-modulated stages. Modulated r.f. is applied to the grid circuit of the Doherty linear amplifier with the carrier tube biased to cutoff and the peak tube biased to the point where it draws substantially zero plate current at the carrier condition.

In the Terman-Woodyard grid-modulated amplifier the carrier tube runs class-C with comparatively high bias and high plate efficiency, while the peak tube again is biased so that it draws almost no plate current. Unmodulated r.f. is applied to the grid circuits of the two tubes and the modulating voltage is inserted in series with the fixed bias voltages. From one-half to two-thirds as much *audio* voltage is required at the grid of the peak tube as is required at the grid of the carrier tube.

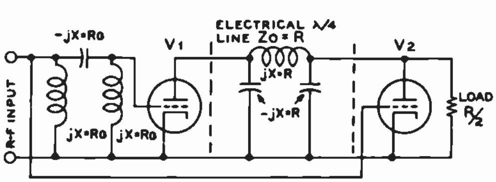


Figure 10

DIAGRAMMATIC REPRESENTATION OF THE DOHERTY LINEAR

Operating Efficiencies The resting carrier efficiency of the grid-modulated amplifier may run as high as is obtainable in any class-C stage—80 percent or better. The resting carrier efficiency of the linear will be about as good as is obtainable in any class-B amplifier—60 to 70 percent. The over-all efficiency of the bias-modulated amplifier at 100 percent modulation will run about 75 percent; of the linear—about 60 percent.

In figure 11 the plate tank circuits are detuned enough to give an effect equivalent to the shunt elements of the quarter-wave "line" of figure 10. At resonance, coils L_1 and L_2 in the grid circuits of the two tubes have each an inductive reactance equal to the capacitive reactance of capacitor C_1. Thus we have the effect of a pi network consisting of shunt inductances and series capacitance. In the plate circuit we want a phase shift of the same magnitude but in the opposite direction; so our series element is inductance L_3 whose reactance is equal to the characteristic impedance desired of the network. Then the plate tank capacitors of the two tubes (C_2 and C_3) are increased an amount past resonance, so that they have a capacitive reactance equal to the inductive reactance of the coil L_3. It is quite important that there be no coupling between the inductors.

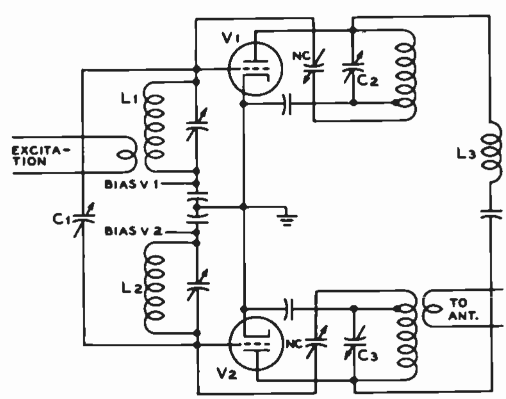


Figure 11

SIMPLIFIED SCHEMATIC OF A "HIGH-EFFICIENCY" AMPLIFIER

The basic system, comprising a "carrier" tube (V_1) and a "peak" tube (V_2) interconnected by lumped-constant quarter-wave lines, is the same for either grid-bias modulation or for use as a linear amplifier of a modulated wave.

Other High-Efficiency Modulation Systems Many other high-efficiency modulation systems have been described since about 1936. The majority of these, however, have received little application either by commercial interests or by amateurs. Nearly all of these circuits have been published in the *I.E.E.E. Proceedings* and the interested reader can refer to them in back copies of that journal.

15-6 Speech Clipping

Speech waveforms are characterized by frequently recurring high-intensity peaks of very short duration. These peaks will cause overmodulation if the *average* level of modulation on loud syllables exceeds approximately 30 percent. Careful checking into the nature of speech sounds has revealed that these high-intensity peaks are due primarily to the vowel sounds. Further research has revealed that the vowel sounds add little to intelligibility, the major contribution to intelligibility coming from the consonant sounds such as *v*, *b*, *k*, *s*, *t*, and *l*. Measurements have shown that the power contained in these consonant sounds may be down 30 db or more from the energy in the vowel sounds in the same speech passage. Obviously, then, if we can increase the relative energy content of the consonant sounds with respect to the vowel sounds it will be possible to understand a signal modulated with such a waveform in the presence of a much higher level of background noise and interference. Experiment has shown that it is possible to accomplish this desirable result simply by cutting off or *clipping* the high-intensity peaks and thus building up in a relative manner the effective level of the weaker sounds.

Such clipping theoretically can be accomplished simply by increasing the gain of the speech amplifier until the average level of modulation on loud syllables approaches 90 percent. This is equivalent to increasing the speech power of the consonant sounds by about 10 times or, conversely, we can say that 10 db of clipping has been applied to the voice wave. However, the clipping when accomplished in this manner *will produce higher order sidebands known as "splatter,"* and the transmitted signal would occupy a relatively tremendous spectrum width. So another method of accomplishing the desirable effects of clipping must be employed.

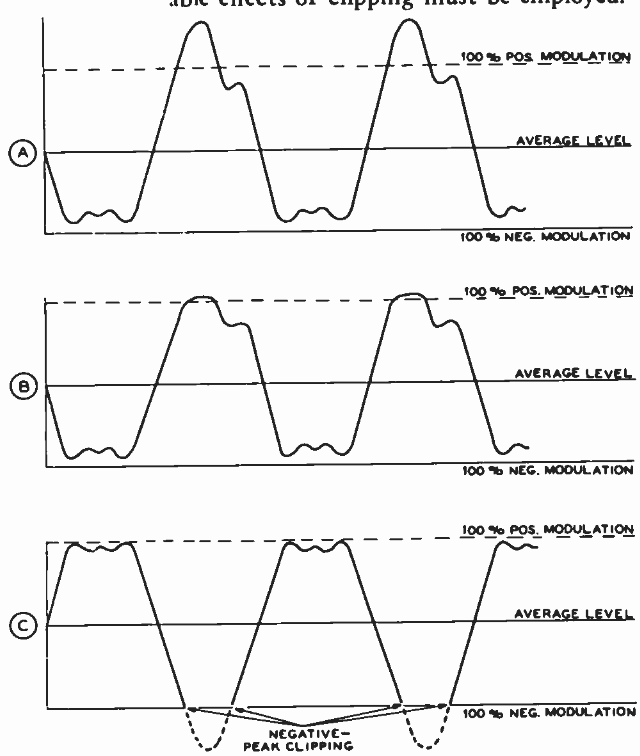


Figure 12

SPEECH-WAVEFORM AMPLITUDE MODULATION

Showing the effect of using the proper polarity of a speech wave for modulating an a-m transmitter. A shows the effect of proper speech polarity on a transmitter having an upward modulation capability of greater than 100 percent. B shows the effect of using proper speech polarity on a transmitter having an upward modulation capability of only 100 percent. Both these conditions will give a clean signal without objectionable splatter. C shows the effect of the use of improper speech polarity. This condition will cause serious splatter due to negative-peak clipping in the modulated-amplifier stage.

A considerable reduction in the amount of splatter caused by a moderate increase in the gain of the speech amplifier can be obtained by phasing the signal from the speech amplifier to the amplitude-modulated transmitter such that the high-intensity peak occurs on *upward* or positive modulation. Overloading on positive modulation peaks produces less splatter than the negative-peak clipping which occurs with overloading on the negative peaks of modulation. This aspect of the problem has been discussed in more detail in the section on *Speech Waveform Dissymmetry* earlier in this chapter. The effect of deriving proper speech polarity from the speech amplifier is shown in figure 12.

A much more desirable and effective method of obtaining speech clipping is actually to employ a clipper circuit in the earlier stages of the speech amplifier, and then to filter out the objectionable distortion components by means of a sharp low-pass filter having a cutoff frequency of approximately 3000 Hz. Tests on *clipper-filter* speech systems have shown that 6 db of clipping on voice is just noticeable, 12 db of clipping is quite acceptable, and values of clipping from 20 to 25 db are tolerable under such conditions that a high degree of clipping is necessary to get through heavy interference. A signal with 12 db of clipping doesn't sound quite *natural* but it is not unpleasant to listen to and is much more readable than an unclipped signal in the presence of strong interference.

The use of a clipper-filter in the speech amplifier of an a-m transmitter, to be completely effective, requires that phase shift between the clipper-filter stage and the final modulated amplifier be kept to a minimum. However, if there is phase shift after the clipper-filter the system does not completely break down. The presence of phase shift merely requires that the audio gain following the clipper-filter be reduced to the point where the *cant* apparent on the clipped speech waves still cannot cause overmodulation. This effect is illustrated in figures 13 and 14.

The *cant* appearing on the tops of the square waves leaving the clipper-filter centers about the clipping level. Hence, as the frequency being passed through the system is lowered, the amount by which the peak of the *canted* wave exceeds the clipping level is increased.

Phase-Shift Correction In a normal a-m transmitter having a moderate amount of phase shift the *cant* applied to the tops of the waves will cause overmodulation on frequencies below those for which the gain following the clipper-filter has been adjusted unless remedial steps have been taken. The following steps are advised:

1. Introduce bass *suppression* into the speech amplifier *ahead* of the clipper-filter.
2. *Improve* the low-frequency response characteristic insofar as it is possible in the stages *following* the clipper-filter.

If a cathode-ray oscilloscope is available the modulated envelope of the a-m transmitter should be checked with 30- to 70-Hz sawtooth waves on the horizontal axis. If the upper half of the envelope appears in general the same as the drawing of figure 13C, all is well and phase-shift is not excessive. However, if much more slope appears on the tops of the waves than is illustrated in this figure, it will be well to apply the second step in compensation in order to ensure that sideband splatter cannot take place and to afford a still higher average percentage of modulation. This second step consists of the addition of a high-level *splatter suppressor* such as is illustrated in figure 15.

The use of a high-level splatter suppressor after a clipper-filter system will afford the result shown in figure 16 since such a device will not permit the negative-peak clipping which the wave cant caused by audio-system phase shift can produce. The high-level splatter suppressor operates by virtue of the fact that it will not permit the plate voltage on the modulated amplifier to go completely to zero regardless of the incoming signal amplitude. Hence negative-peak clipping with its attendant splatter cannot take place. Such a device can, of course, also be used in a transmitter which does not incorporate a clipper-filter system. However, the full increase in average modulation level without serious distortion, afforded by the clipper-filter system, will not be obtained.

A word of caution should be noted at this time in the case of tetrode-final modulated amplifier stages which afford screen-voltage modulation by virtue of a tap or a separate winding on the modulation transformer such as is shown in figure 8C of this chapter. If such a system of modulation is in use, the high-level splatter suppressor shown in figure 15 will not operate properly since negative-peak clipping in the modulated stage can take place when the screen voltage goes too low.

Clipper Circuits Two effective low-level clipper circuits are shown in figures 17 and 18. The circuit of figure 17 is transistorized, with a modified input circuit suitable for use with high-impedance (crystal) microphones having an average output level of about 10 millivolts, peak-to-peak. Three amplifier stages boost the microphone level to about 5 volts peak-to-peak and the output of the last stage is fed to a double-diode clipper, utilizing a pair of germanium diodes. A maximum of 12 to 14 db of clipping may be achieved with this circuit, and the two-stage speech amplifier must therefore be considered as a part of the clipper circuit since it compensates for the loss of gain incurred in the clipping process. A simple RC lowpass filter starts to round-off the waveform at about 2.5 kHz. The output level of about 0.5 volt peak-to-peak is ample to drive most speech amplifiers with gain to spare. The degree of clipping is adjusted by variation of the microphone level in conjunction with the proper setting of the *gain* potentiometer.

The circuit of figure 18 has an *adjust clipping* control in addition to the *adjust gain* potentiometer. The gain control determines the modulation level of the transmitter. This control should be set so that overmodulation is impossible, regardless of the amount of clipping used. Once the gain control has been set, the clipping control may be used to set the modulation level to any percentage below 100 percent. As the modulation level is decreased, more and more clipping is introduced into the circuit, until a full 12 to 14 db of clipping is used. This means the gain control may be ad-

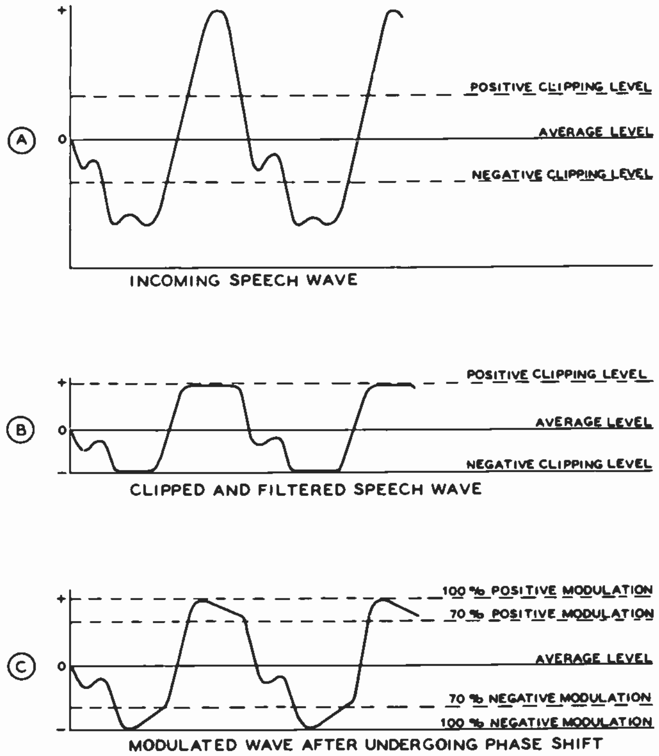


Figure 13

ACTION OF A CLIPPER-FILTER ON A SPEECH WAVE

Drawing A shows the incoming speech wave before it reaches the clipper stage. B shows the output of the clipper-filter, illustrating the manner in which the peaks are clipped and then the sharp edges of the clipped wave removed by the filter. C shows the effect of phase shift in the stages following the clipper-filter and the manner in which the a-m transmitter may be adjusted for 100-percent modulation of the "canted" peaks of the wave, the sloping top of the wave reaching about 70% modulation.

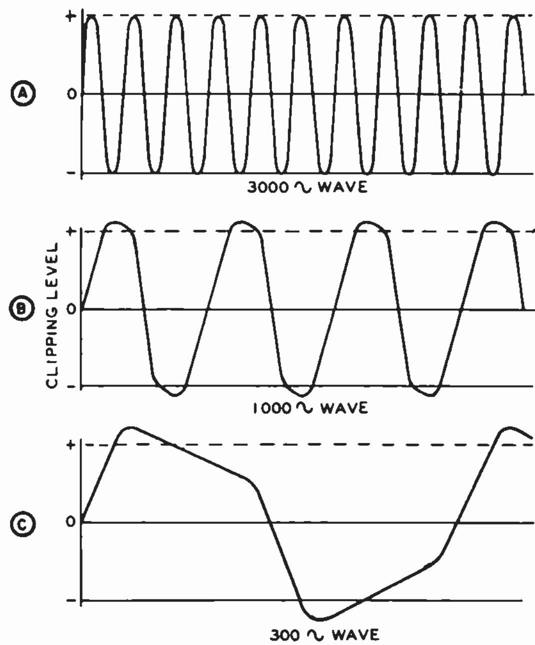


Figure 14

ILLUSTRATING THE EFFECT OF PHASE SHIFT AND FILTERED WAVES OF DIFFERENT FREQUENCY

Sketch A shows the effect of a clipper and a filter having a cutoff of about 3500 Hz on a wave of 3000 Hz. Note that no harmonics are present in the wave so that phase shift following the clipper-filter will have no significant effect on the shape of the wave. B and C show the effect of phase shift on waves well below the cutoff frequency of the filter. Note that the "cant" placed on the top of the wave causes the peak value to rise higher and higher above the clipping level as the frequency is lowered. It is for this reason that bass suppression before the clipper stage is desirable. Improved low-frequency response following the clipper-filter will reduce the phase shift and therefore the "canting" of the wave at the lower voice frequencies.

vanced some 12 db past the point at which the clipping action started. Clipping action should start at about 90-percent modulation when a sine wave is used for circuit adjustment purposes. In all cases, the use of a monitor oscilloscope to adjust clipping level is highly recommended.

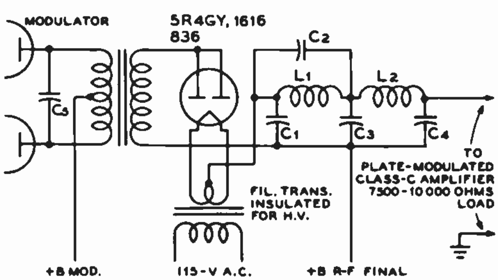


Figure 15

HIGH-LEVEL A-M SPLATTER SUPPRESSOR

This circuit is effective in reducing splatter caused by negative-peak clipping in the modulated amplifier stage. The use of a two-section filter as shown is recommended, although either a single m-derived or a constant-k section may be used for greater economy. Suitable chokes, along with recommended capacitor values, are available from several manufacturers.

High-Level Filters Even though we may have cut off all frequencies above 3 or 3.5 kHz through the use of a filter system such as shown in the circuits of figures 17 and 18, higher frequencies may again be introduced into the modulated wave by distortion in stages following the speech amplifier. Harmonics of the incoming audio frequencies may be generated in the driver stage for the modulator; they may be generated in the plate circuit of the modulator; or they may be generated by nonlinearity in the modulated amplifier itself.

Regardless of the point in the system following the speech amplifier where the high audio frequencies may be generated, these frequencies can still cause a broad signal to be transmitted even though all frequencies above 3000 or 3500 Hz have been cut off in the speech amplifier. The effects of distortion in the audio system *following* the speech amplifier can be eliminated quite effectively through the use of a *post-modulator* filter. Such a filter may be used between the modulator plate circuit and the r-f amplifier which is being modulated (figure 19).

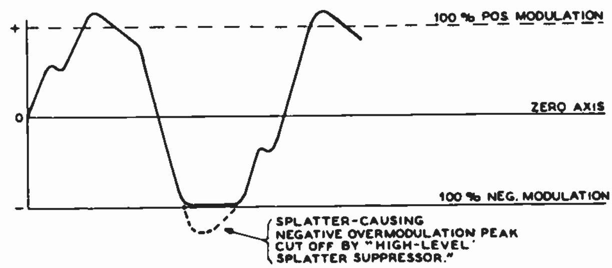
15-7 Speech Compression

Volume compression or a form of automatic gain control may be used to maintain constant voice intensity over a large range

Figure 16

ACTION OF HIGH-LEVEL SPLATTER SUPPRESSOR

A high-level splatter suppressor may be used in a transmitter without a clipper-filter to reduce negative-peak clipping, or such a unit may be used following a clipper-filter to allow a higher average modulation level by eliminating the negative-peak clipping which the wave cant caused by phase shift might produce.



of audio input to the speech system of a voice transmitter. This is accomplished by making the system gain a function of the average variations in speech amplitude. Practical systems rectify and filter the speech signal as it passes through the speech amplifier and apply the d-c component of the signal to a gain-control element in the amplifier. The compression system usually has a time constant such that the control voltage is held at a steady value between syllables and words. Simple compressors usually exhibit an attack time of about 10 milliseconds and a release time of 300 milliseconds or longer. Compression range of the order of 20 to 35 decibels is realizable in practical circuits, corresponding roughly to the dynamic range of the human speaking voice. Reverberation and background noise usually limit the practical compression range to 15 decibels or so.

A basic compression amplifier is shown in figure 20. A sample of the audio signal

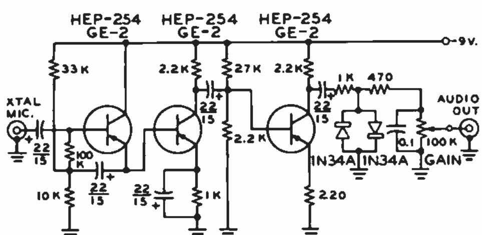


Figure 17

SPEECH CLIPPER FOR USE WITH CRYSTAL MICROPHONE

This simple clipper/amplifier may be inserted between microphone and existing speech amplifier. Power is supplied by a 9-volt transistor radio battery. Transistors are either Motorola (HEP type) or General Electric (GE type).

is taken and rectified to provide a negative control voltage which fluctuates with *average* voice level. The compression control voltage is applied as bias to the control grid of a variable-μ pentode amplifier. Compression is substantially proportional to the average input signal and thus holds the output level at a constant level. Rise time is regulated by the choice of R_2 and C_2, while release time is controlled by R_1 and C_1.

A simple and inexpensive compression circuit suitable for amplitude modulation or SSB is shown in figure 21. A two stage 12AX7 preamplifier is used, the input circuit of which is shunted with a silicon transistor working as a d-c amplifier whose gain is inversely proportional to the audio output voltage of the amplifier. The compression amplifier is suitable for use with a dynamic microphone having an impedance as high as 0.05 megohm.

A driving signal of about 0.8 volts r.m.s. at point *A* will overcome the threshold level of the system and an audio input of 10 mV at the microphone jack will produce about 10 db of compression. Rise time is about 30 mS and release time is about 100 mS, both of which are controlled by the 1-pf capacitor in the base circuit of the transistor.

A solid-state compressor/amplifier is shown in figure 22. It is designed to be used with a dynamic microphone having an impedance in the range of 500 to 2000 ohms, and provides a compression range of approximately 20 db.

Compression is achieved by change of gain brought about by variation of the emitter bypass capacitance in the first-stage transistor. With the emitter load impedance about twice the value of the collector load,

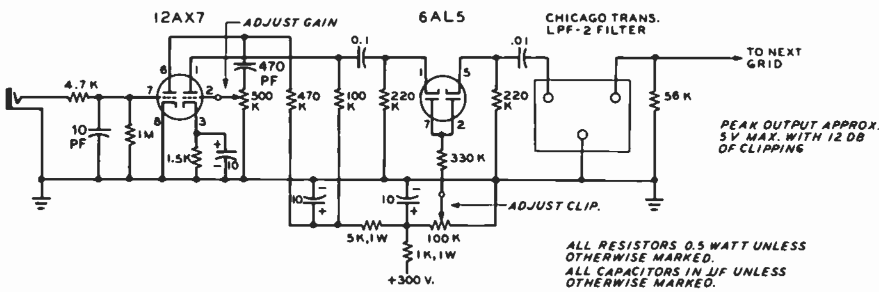


Figure 18

CLIPPER FILTER USING 6AL5 STAGE

and with unbypassed emitter, the first stage gain is about 0.5. In series with the emitter bypass capacitor (C_1) is a variable resistance composed of a diode network. Control voltage derived from the output of the amplifier is applied to the diode which, in effect, isolates bypass capacitor C_1 from signal voltages,

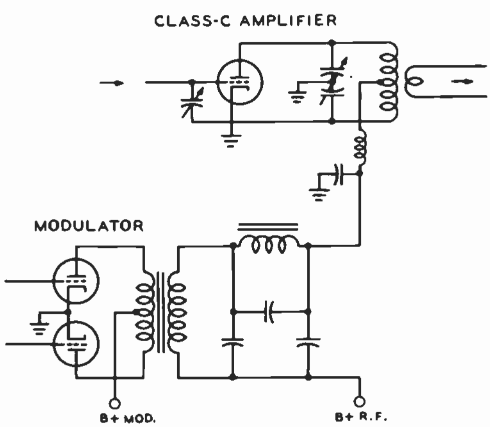


Figure 19

ADDITIONAL HIGH-LEVEL LOW-PASS FILTER TO FOLLOW MODULATOR WHEN A LOW-LEVEL CLIPPER FILTER IS USED

Suitable choke, along with recommended capacitor values, is available from several manufacturers.

effectively lowering the stage gain. Maximum stage gain is approximately the ratio of the collector load resistance (2.7K) to the forward resistance of the cathode control diode (D_1) and minimum stage gain is about 0.5.

Small coupling capacitors are used be-

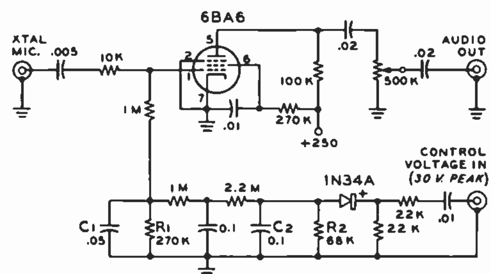


Figure 20

SPEECH COMPRESSOR FOR USE WITH CRYSTAL MICROPHONE

This basic volume compressor derives control voltage from a high-level stage in the existing speech amplifier. A signal of about 30 volts peak is required to provide up to 20 decibels of compression. Compressor is designed to be placed between microphone and station amplifier.

tween amplifier stages to limit the low-frequency response of the system.

A similar amplifier having somewhat higher gain and lower distortion is shown in figure 23. A FET is used for the control element.

Bass Suppression Most of the power represented by ordinary speech (particularly the male voice) lies below 1000 Hz. If all frequencies below 400 or 500 Hz are eliminated or substantially attenuated, there is a considerable reduction in power but insignificant reduction in intelligibility. This means that the speech level may be increased considerably without overmodulation or overload of the audio system. In addition, if speech processing is used, attenuation of the lower audio frequencies before

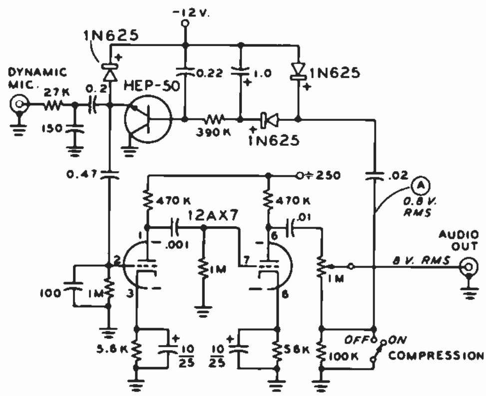


Figure 21

TRANSISTOR-CONTROLLED SPEECH COMPRESSOR

A single transistor (Motorola HEP-type) acts as a d-c amplifier, receiving its driving signal from point A. Maximum compression of 20 decibels is achieved using a dynamic microphone.

the clipper or compressor will reduce phase shift and canting of the clipper output.

A simple method of bass suppression is to reduce the size of the interstage coupling capacitors in a resistance-coupled amplifier. Figure 24 shows the frequency characteristics caused by such a suppression circuit. A second simple bass-suppression circuit is to place a small iron-core filter choke from grid to ground in a speech-amplifier stage, as shown in figure 25.

Modulated-Amplifier Distortion The systems described in the preceding paragraphs will have no effect in reducing a broad signal caused by nonlinearity in the modulated amplifier or in linear-amplifier stages. Even though the modulating waveform impressed on the modulated stage may be distortion free, if the modulated amplifier is nonlinear, distortion will be generated in the amplifier. The only way in which this type of distortion can be corrected is by making the modulated amplifier more linear. Degenerative feedback which includes the modulated amplifier in the loop will help in this regard.

Plenty of grid excitation and high grid bias will go a long way toward making a plate-modulated class-C amplifier linear, although such operating conditions will make more difficult the problem of TVI reduction. If this still does not give adequate linearity, the preceding buffer stage may be modulated 50 percent or so at the same time and in the same phase as the final amplifier. The use of a grid resistor to obtain the majority of the bias for a class-C stage will improve its linearity.

15-8 High Level Modulation

Tetrode Modulators In regard to the use of tetrodes, the advantages of these tubes have long been noted for use in modulators having from 10 to 100 watts output. The 6AQ5, 6L6GC and 6146 tubes have served well in providing audio power outputs in this range. The higher-power tetrodes such as the 813, and 4-250A, and the zero-bias triodes such as the 3-400Z are popular as high-level audio amplifiers. The beam tetrodes offer the advantages of low driving power (even down to zero driving power for many applications) as compared to the high driving-power requirements of the usual triode tubes having equivalent power-output capabilities.

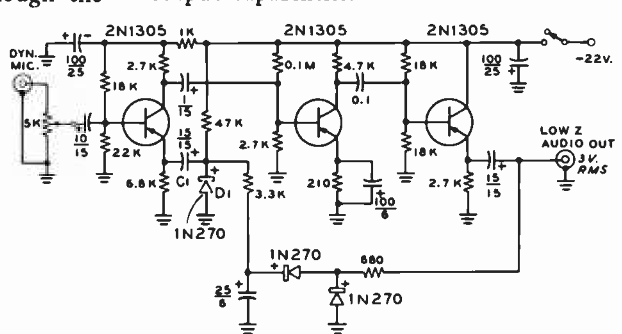


Figure 22

SOLID-STATE COMPRESSOR AMPLIFIER FOR DYNAMIC MICROPHONE

Compression is brought about by variation of emitter bypass capacitor C, in the first-stage transistor. Variable-resistance network is driven by two 1N270 diodes as a voltage doubler of output signal taken from emitter of the third-stage 2N1305 emitter follower.

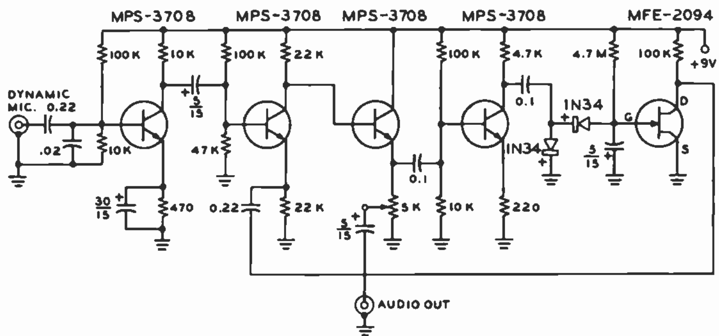


Figure 23

COMPRESSOR AMPLIFIER USING FET IN CONTROL LOOP

Emitter bypass variation of second amplifier stage is controlled by FET amplifier, with signal taken from drain element. MPS and MFE transistors by Motorola.

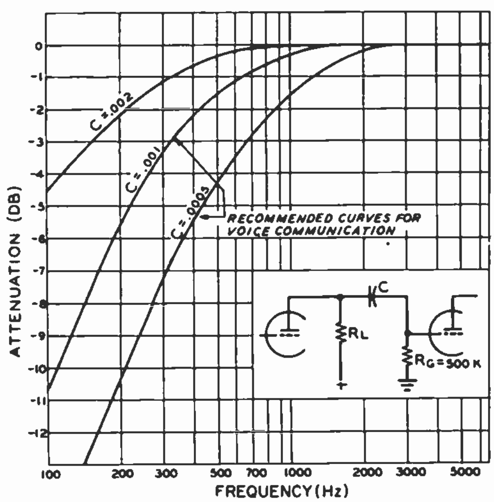


Figure 24

BASS-ATTENUATION CHART

Frequency attenuation caused by various values of coupling capacitor with a grid resistor of 0.5 megohm in the following stage ($R_G > R_L$)

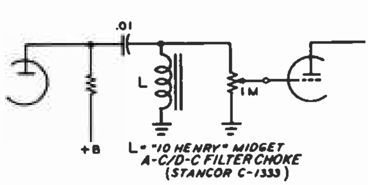


Figure 25

USE OF PARALLEL INDUCTANCE FOR BASS SUPPRESSION

On the other hand, beam-tetrode tubes require both a screen-voltage power supply and a grid-bias source. So it still is expedient in many cases to use zero-bias triodes or even low-μ triodes in many modulators for the medium-power and high-power range. A list of suggested modulator combinations for a range of power output capabilities is given later in this chapter.

A 10-Watt Modulator This compact modulator is well suited for portable or mobile operation since it may be run from a 6- or 12-volt primary supply (figure 26). A 12AX7 two-stage speech amplifier drives a 6C4 *hot-cathode* phase inverter, in which a proportion of the output voltage is developed across a cathode load resistor, out of phase with the plate signal. Matched cathode and plate-load resistors ensure that the output signals from the 6C4 stage are equal in amplitude. Two 6AQ5 tetrodes are used in the class AB_1 modulator stage, delivering about 10 watts of audio power. A simple negative-feedback circuit from the secondary of the modulation transformer to the cathode of the second speech amplifier stage smooths the audio response, and provides improved regulation of the audio output voltage. The capacitor across the secondary winding of the modulation transformer is part of the feedback circuit, reducing any tendency of the circuit to oscillate at the higher audio frequencies.

It is important that the feedback be properly phased. If the amplifier oscillates,

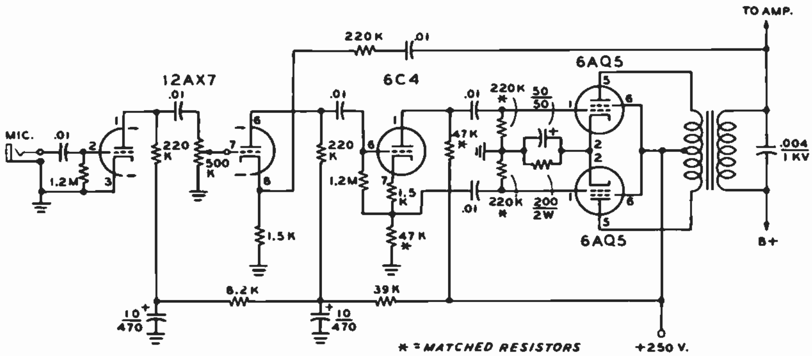


Figure 26

A 10-WATT 6AQ5 MODULATOR

either the primary or secondary connections of the output transformer should be reversed. With a 250-volt plate supply the 6AQ5 tubes should have a bias of about −15 volts and a resting plate current of about 70 ma. At full power output, plate current will rise to about 80 ma or so. Maximum-signal cathode current is about 95 ma. For 12-volt operation, the filament sections of the 12AU7 should be wired in series. The 6AQ5's may be series connected, or 12AQ5 tubes used. The 6C4 requires a series filament resistor to operate on 12 volts.

10- to 120-Watt Modulator with Beam-Power Tubes It is difficult to surpass the capabilities of the reliable beam-power tube when an audio power output of 10 to 120 watts is required of a modulator. A pair of 6L6GC tubes operating in such a modulator will deliver good plate-circuit efficiency, require only a very small amount of driving power, and impose no serious grid-bias problems. Included on the chassis of the modulator shown in figure 27 are the speech amplifier, the driver and modulation transformers for the output tubes, and a plate-current milliammeter. The power supply has not been included. The 6AU6A pentode first stage is coupled through the volume control to the grid of a 6C4 phase inverter. The output of the phase inverter is capacitively coupled to the grids of a 12AU7A which acts as a push-pull driver for the output tubes. Transformer coupling is used between the driver stage and the grids of the output tubes so that the output stage may be operated either as a class-AB_1 or class-AB_2 amplifier.

The Output Stage Either 6L6, 6L6GC, or 807 tubes may be used in the output stage of the modulator. As a matter of fact, either 6V6-GT or 6F6-G tubes could be used in the output stage if somewhat less power output is required.

Listed in Table 1 are a group of recommended operating conditions for different tube types in the output stage of the modulator. In certain sets of operating conditions the tubes will be operated class AB_1, that is with increased plate current with signal but with no grid current. Other operating conditions specify class-AB_2 operation, in which

Table 1.

Recommended Operating Conditions for Modulator of Figure 27 for Different Tube Types

Tubes V_1 V_2	Class	Plate Volts (E)	Screen Volts (C)	Grid Bias (D)	Plate-To-Plate Load (Ohms)	Plate Current (MA)	Power Output (Watts)
6V6GT	AB_1	250	250	−15	10,000	70-80	10
6V6GT	AB_1	285	285	−19	8,000	70-95	15
6L6	AB_1	360	270	−23	6,600	85-135	27
6L6	AB_2	360	270	−23	3,800	85-205	47
807	AB_1	600	300	−34	10,000	35-140	56
807	AB_1	750	300	−35	12,000	30-140	75
807	AB_2	750	300	−35	7,300	30-240	120

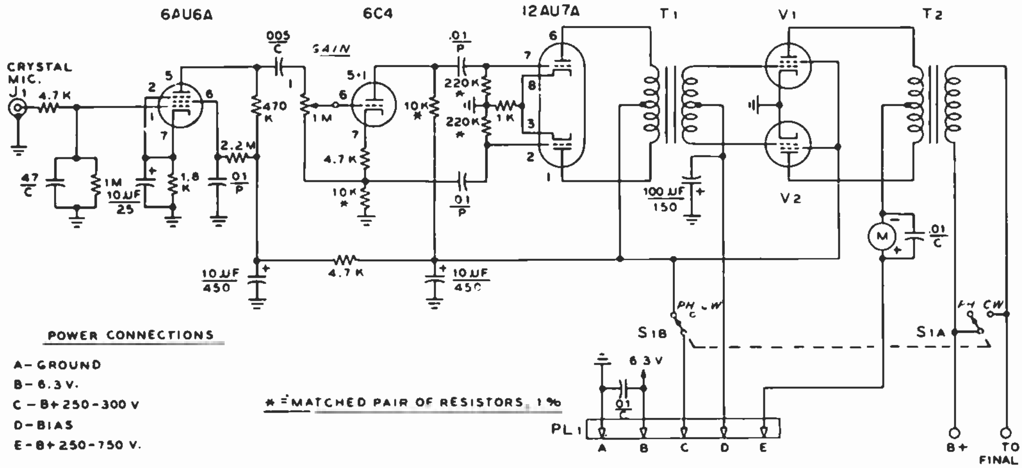


Figure 27

SCHEMATIC OF MEDIUM-POWER MODULATOR

M—0-250 d-c milliammeter
T_1—Driver transformer. Stancor A-4701, or UTC S-10
T_2—Modulation transformer
60-watt level = Stancor A-3893, or UTC S-10
125-watt level = Stancor A-3894

the plate current increases with signal and grid current flows on signal peaks.

A High-Power Modulator with Beam Tetrodes Listed in Table 2 are representative operating conditions for various tetrode tubes providing power levels up to 840 watts of audio. Complete operating data on these tube types may be obtained from the manufacturer. In the case of the 4-250A and 4-400A, both class AB_1 and AB_2 data are given. These tubes are electrically similar except that the 4-400A has greater plate dissipation. Class AB_1 operation of the modulator is recommended since it places less demand on the driver stage, and a simple transformer-coupled voltage amplifier may be used, such as the one shown in the circuit of figure 27.

Because of the power level involved and the design of the external-anode tube, the 4X150A/4CX250B tubes must be forced-air cooled in this application. It is recommended that the 813, 4-250A, and 4-400A tubes be convection cooled with a small fan. The modulator may be checked for correct operation as described in the next section.

Table 2.

Recommended Operating Conditions for Tetrode Tubes for 400- to 800-Watt Power Level

Tubes V_1, V_2	Class	Plate Volts E	Screen Volts C	Grid Bias D	Plate-to-Plate Load (ohms)	Plate Current (ma)	Power Output (watts)
813	AB_2	2500	750	−95	17,000	36 to 360	650
4-125A	AB_2	2500	350	−43	22,000	90 to 260	400
4-250A	AB_1	2500	300	−51	10,900	120 to 500	840
4-400A	AB_2		600	−110	11,400	120 to 430	625
4X150A 4CX250B 4CX300A	AB_1	2000	350	−55	9,500	200 to 500	600

15-9 General Purpose Triode Class-B Modulator

High level class-B modulators with power output in the 125- to 500-watt level usually make use of triodes such as the 811, 805, or 810 tubes with operating plate voltages between 750 and 2000. Figure 28 illustrates a general-purpose modulator unit designed for operation in this power range. The size of the modulation transformer will of course be dependent on the amount of audio power developed by the modulator.

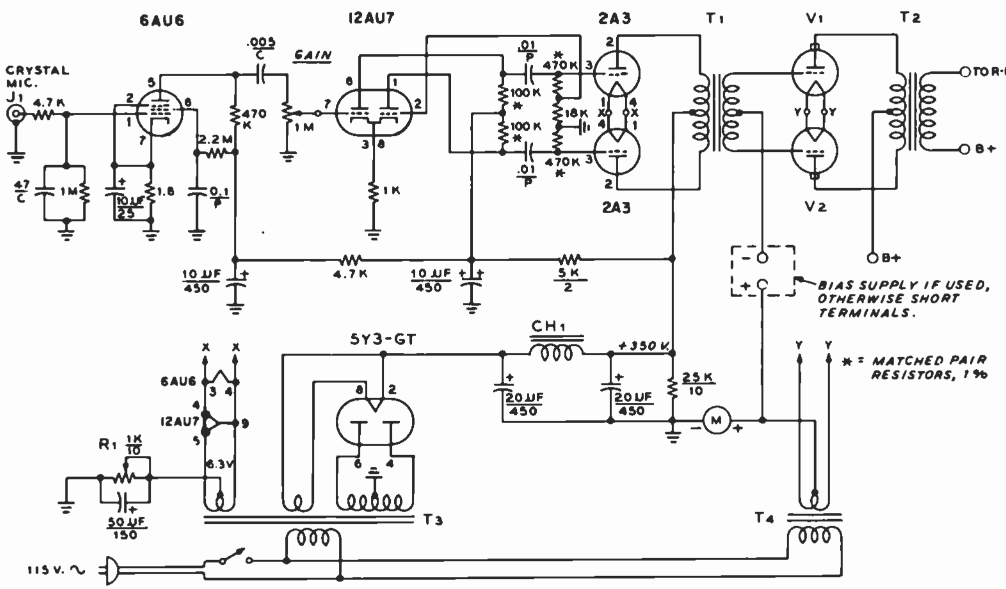


Figure 28

SCHEMATIC OF GENERAL PURPOSE MODULATOR

M—0-500 ma.
T_1—Driver transformer. Stancor A-4761
T_2—"Poly-pedance" Modulation transformer.
300-watt rating, Stancor A-3898
500-watt rating, Stancor A-3899
T_3—360-0-360 volts, 150 ma. Stancor PC-8410
T_4—Suitable for tubes used.
For 811-A's = 6.3 volt, 8 amp. Stancor P-6308
For 810's = 10 volt, 10 amp. Stancor P-6461
CH_1—14 henry, 100 ma. UTC S-19. Stancor C-1001
R_1—1K, 10 watts, adjustable. Set for plate current of 80 ma. (no signal) to 2A3 tubes (approximately 875 ohms).
V_1, V_2—See Table 3.

Circuit Description of General Purpose Modulator The modulator unit shown in figure 28 is complete except for the high-voltage supply required by the modulator tubes. A speech amplifier suitable for operation with a crystal microphone is included on the chassis along with its own power supply. A 6AU6 is used as a high-gain preamplifier stage resistance coupled to a 12AU7 phase inverter. The audio level is controlled by a potentiometer in the input grid circuit of the 12AU7 stage. Push-pull 2A3 low-μ triodes serve as the class-B driver stage. The 2A3's are coupled to the grids of the modulator tubes through a conventional multipurpose driver transformer. Cathode bias is employed on the driver stage which is capable of providing 12 watts of audio power for the grid circuit of the modulator.

For c-w operation the secondary of the class-B modulation transformer is shorted out and the filament and bias circuits of the modulator are disabled.

Table 3.

Suggested Operating Conditions For General-Purpose Modulator

Tubes V_1, V_2	Plate Voltage	Grid Bias (Volts)	Plate Current (Ma)	Plate-To-Plate Load (Ohms)	Sine Wave Power Output (Watts)
811-A	750	0	30-350	5,100	175
811-A	1000	0	45-350	7,400	245
811-A	1250	0	50-350	9,200	310
811-A	1500	−4.5	32-315	12,400	340
805	1250	0	148-400	6,700	300
805	1500	−16	84-400	8,200	370
810	2000	−50	60-420	12,000	450
810	2500	−75	50-420	17,500	500

Modulator Adjustment When the modulator has been wired and checked, it should be tested before being used with an r-f unit. A satisfactory test setup is shown in figure 29. A common ground lead should be run between the speech amplifier and the modulator. A number of 1000-ohm 100-watt resistors are connected in series and placed across the high-voltage terminals of the modulator unit to act as an audio load. Bias should be adjusted to show the indicated value from grid terminal to ground as measured with a high-resistance voltmeter. If an oscilloscope is available, it should be coupled to point "A" on the load resistor through a 500-pf ceramic TV capacitor of 10,000 volts rating. The case of the oscilloscope should be grounded to the common ground point of the modulator.

The listed plate voltage is now applied to the modulator, and bias is adjusted for proper resting plate current.

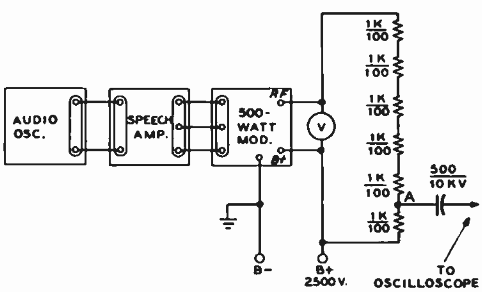


Figure 29

TEST SETUP FOR 500-WATT MODULATOR

Be extremely careful during these adjustments, since the plate supply of the modulator is a lethal weapon. Never touch the modulator when the plate voltage supply is on! Be sure you employ the TV blocking capacitor between the oscilloscope and the plate-load resistors, as these load resistors are at high-voltage potential! If a high-resistance a-c voltmeter is available that has a 2000-volt scale, it should be clipped between the high-voltage terminals of the modulator, directly across the dummy load. Do not touch the meter when the high-voltage supply is in operation! An audio oscillator should be connected to the audio input circuit of the exciter-transmitter and the audio excitation to the high-level modulator should be increased until the a-c voltmeter across the dummy-load resistor indicates an rms reading that is equal to 0.7 (70%) of the plate voltage applied to the modulator. If the modulator plate voltage is 2500 volts, the a-c meter should indicate 1750 volts developed across the 6000-ohm dummy-load resistor. This is equivalent to an audio output of 500 watts. Under these conditions, the oscilloscope may be used to observe the audio waveform of the modulator when coupled to point "A" through the 10,000 volt coupling capacitor.

When the frequency of the audio oscillator is advanced above 3500 Hz the output level of the modulator as measured on the a-c voltmeter should drop sharply indicating that the low-pass audio network is functioning properly (if low-pass network is used).

15-10 A 15-Watt Clipper-Amplifier

The near-ultimate in "talk power" can be obtained with low-level clipping and filtering combined with high-level filtering. Such a modulation system will have real "punch," yet will sound well rounded and normal. The speech amplifier described in this section makes use of low-level clipping and filtering and is specifically designed to drive a high-level modulator.

Circuit Description The schematic of the speech amplifier-clipper is shown in figure 30. A total of six tubes, including a rectifier are employed and the unit delivers 15 watts of heavily clipped audio.

A 12AX7 tube is used as a two-stage microphone preamplifier and delivers approximately 20 volts (rms) audio signal to the 6AL5 series clipper tube. The clipping level is adjustable between 0 db and 15 db by clipping control R_2. Amplifier gain is controlled by R_1 in the grid circuit of the second section of the 12AX7. A low-pass filter having a 3500-Hz cutoff follows the 6AL5 clipper stage, with an output of 5 volts peak audio signal under maximum clipping conditions. A double-triode 12AU7 cathode-follower phase inverter follows the clipper stage and delivers a 100-volt rms

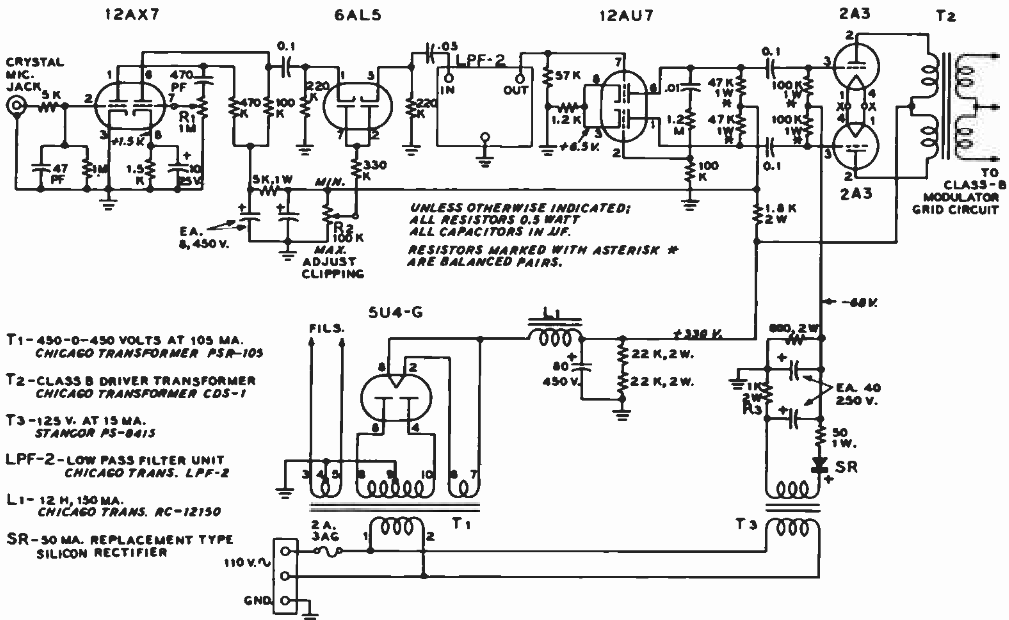


Figure 30

SCHEMATIC, 15-WATT CLIPPER-AMPLIFIER

signal to the push-pull grids of the 2A3 audio driver tubes. The 2A3 tubes operate at a plate potential of 330 volts and have a −68 volt bias voltage developed by a small diode rectifier supply applied to their grid circuit. An audio output of 15 watts is developed across the secondary terminals of the class-B driver transformer with less than 5 percent distortion under conditions of no clipping. A 5U4-G and a choke-input filter network provide unusually good voltage regulation of the high-voltage plate supply.

15-11 Auxiliary Clipper Amplifiers

Simple, compact clipper-amplifiers may be added to existing a-m equipment to provide a higher level of modulation and greater "talk power." Shown in figure 31 is a clipper-amplifier designed to be placed between the microphone and the input of a speech amplifier.

Two stages of amplification provide ample gain for the diode clippers which start to conduct at an audio level of about 0.6 volt peak. The modulation level is controlled by the *gain* potentiometer and the clipping level is set by the *adjust-clip* potentiometer in the emitter circuit of the first 2N3391 transistor. It is recommended that either clipper be adjusted for maximum performance with an oscilloscope attached to the transmitter.

15-12 Zero Bias Tetrode Modulators

Class-B zero bias operation of tetrode tubes is made possible by the application of the driving signal to the two grids of the tubes as shown in figure 32. Tubes such as the 6AQ5, 6L6GC, 807, and 813 work well in this circuit and neither a screen supply nor a bias supply is required. The drive requirements are low and the tubes operate with excellent plate circuit efficiency. The series grid resistors for the small tubes are required to balance the current drawn by the two grids, but are not needed in the case of the 813 tubes.

15-13 Pulse-Duration Modulation

A recent innovation in high-level plate

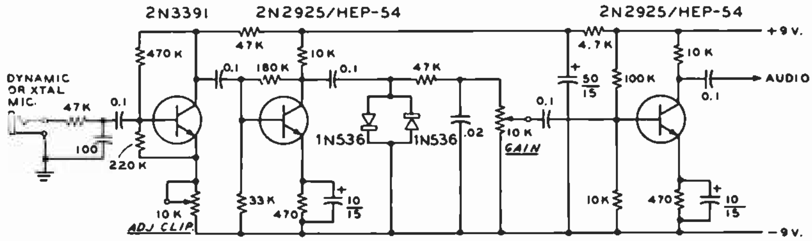


Figure 31

AUXILIARY CLIPPER/AMPLIFIER

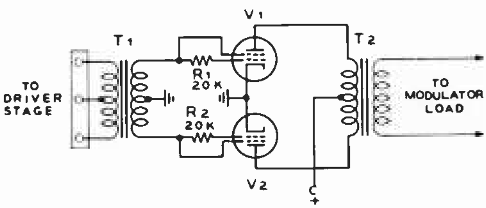


Figure 32

ZERO BIAS TETRODE MODULATOR ELIMINATES SCREEN AND BIAS SUPPLIES

Low driving power and simplicity are key features of this novel modulator. Tubes ranging in size from 6AQ5's to 813's may be employed in this circuit.

T_1—Class-B driver transformer
T_2—Modulation transformer
V_1, V_2—6AQ5, 6L6, 807, 813, etc.
R_1, R_2—Not used with 813

modulation is the *pulse-duration modulation* technique wherein the modulator tube is operated in a saturated switching mode and is placed in series with the r-f power tube.

The plate modulator in a conventional a-m transmitter operates in a linear mode that may be compared to an analog system. In the pulse-duration modulator, the modulator operates in a switching mode that may be compared to a digital computer, having two conditions; *off* and *on*. Audio information is contained in the duration of the *on* pulse.

Audio amplitude is determined by the duty cycle of the modulator tube. A square-wave signal of about 70 kHz is pulse-width modulated by the audio signal, whose amplitude causes the symmetry of the square wave to vary. The audio signal is imposed on the 70-kHz square wave train at a low level and the resulting signal is amplified to the modulating level. The square-wave component is then filtered out to leave the amplified audio voltage, plus a d-c component that is the modulated plate voltage for the class-C amplifier. This technique eliminates the need of a modulation transformer and modulation choke.

A block diagram of the *Gates* VP-100 pulse-duration modulated a-m transmitter is shown in figure 33.

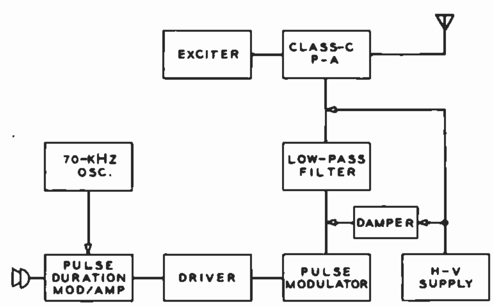


Figure 33

GATES PULSE-DURATION MODULATION SYSTEM

The audio signal is combined with a 70-kHz square-wave signal and processed to produce a modulated pulse-width modulated train which is amplified and applied to the cathode of the class-C r-f amplifier through a low-pass filter that removes the 70-kHz signal and its sidebands, thereby recovering the original audio. The modulator tube acts like a variable resistance whose value varies with the amplitude and frequency of the applied audio signal. The driver stages for the modulator are simple "on"-"off" switches. A damper diode is connected between the output of the modulator and the high-voltage supply to conduct when the modulator does not.

CHAPTER SIXTEEN

Radio Interference (RFI)

The problem of interference to other equipment is best approached by the philosophy discussed in Chapter Seventeen. By correct design procedure, spurious harmonic generation in radio transmitters may be held to a minimum. The remaining problem is twofold: to make sure that the residual harmonics generated by the transmitter are not radiated, and to make sure that the fundamental signal of the transmitter does not overload the other equipment by reason of the proximity of one to the other.

In an area of high TV-signal field intensity the TVI problem is capable of complete solution with routine measures both at the amateur transmitter and at the affected receivers. But in fringe areas of low TV-signal field strength the complete elimination of TVI is a difficult and challenging problem. The fundamentals illustrated in Chapter Seventeen must be closely followed, and additional antenna filtering of the transmitter is required.

16-1 Types of Television Interference

There are three main types of TVI which may be caused singly or in combination by the emissions from an amateur transmitter. These types of interference are:

1. Overloading of the TV set by the transmitter fundamental
2. Impairment of the picture by spurious emissions
3. Impairment of the picture by the radiation of harmonics

TV-Set Overload Even if the amateur transmitter were perfect and had no harmonic radiation or spurious emissions whatever, it still would be likely to cause overloading to TV sets whose antennas were within a few hundred feet of the transmitting antenna. This type of overloading is essentially the same as the common type of BCI encountered when operating a medium-power or high-power amateur transmitter within a few hundred feet of the normal broadcast receiver. The field intensity in the immediate vicinity of the transmitting antenna is sufficiently high so that the amateur signal will get into the BC or TV set either through overloading of the front end, or through the i-f, video, or audio systems. A characteristic of this type of interference is that it always will be eliminated when the transmitter temporarily is operated into a

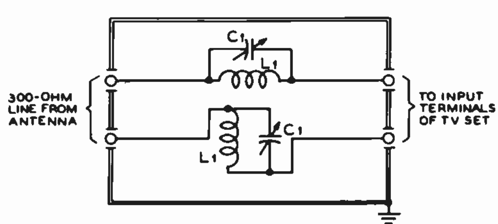


Figure 1

TUNED TRAP FOR THE TRANSMITTER FUNDAMENTAL

This trap has proven to be effective in eliminating the condition of general blocking as caused by a 50-MHz transmitter in the vicinity of a TV receiver. The tuned circuits L_1 C_1 are resonated separately to the frequency of transmission. The adjustment may be done at the station, or it may be accomplished at the TV receiver by tuning for minimum interference on the TV screen.

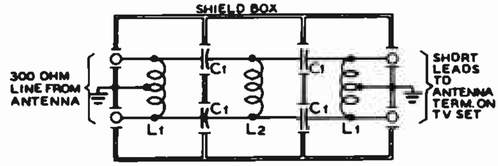
dummy antenna. Another characteristic of this type of overloading is that its effects will be substantially continuous over the entire frequency coverage of the BC or TV receiver. Channels 2 through 13 will be affected in approximately the same manner.

With the overloading type of interference, the problem is simply to keep the *fundamental* of the transmitter out of the affected receiver. Other types of interference may or may not show up when the fundamental is *taken out of the TV set* (they probably will appear), but at least the fundamental *must* be eliminated first.

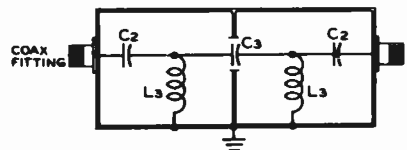
The elimination of the transmitter fundamental from the TV set is normally the only operation performed on or in the vicinity of the TV receiver. After the fundamental has been eliminated as a source of interference to reception, work may then be begun on or in the vicinity of the transmitter toward eliminating the other two types of interference.

Taking Out the Fundamental Signal More or less standard BCI-type practice is most commonly used in taking out fundamental interference. Wavetraps and filters are installed, and the antenna system may or may not be modified so as to offer less response to the signal from the amateur transmitter. In regard to a comparison between wavetraps and filters, the same considerations apply as have been effective in regard to BCI for many years; wavetraps are quite effective when properly installed and adjusted, but they must be readjusted whenever the band of operation is changed, or even when moving from one extreme end of a band to the other. Hence, wavetraps are not recommended except when operation will be confined to a relatively narrow portion of one amateur band. However, figure 1 shows a suitable trap system, especially effective at 50 MHz.

High-Pass Filters High-pass filters in the antenna lead of the TV set have proven to be quite satisfactory as a means of eliminating TVI of the overloading type. In many cases when the interfering transmitter is operated only on the bands below 30 MHz, the use of a high-pass filter in the antenna lead has completely eliminated all TVI. In some cases the installation of a high-pass filter in the antenna



Ⓐ FOR 300-OHM LINE, SHIELDED OR UNSHIELDED



Ⓑ FOR 50-75 OHM COAXIAL LINE

Figure 2

HIGH-PASS TRANSMISSION LINE FILTERS

The arrangement at A will stop the passing of all signals below about 45 MHz from the antenna transmission line into the TV set. Coils L_1 are each 1.2 microhenrys (17 turns No. 24 enam. closewound on ¼-inch dia. polystyrene rod) with the center tap grounded. It will be found best to scrape, twist, and solder the center tap before winding the coil. The number of turns each side of the tap may then be varied until the tap is in the exact center of the winding. Coil L_2 is 0.6 microhenry (12 turns No. 24 enam. closewound on ¼-inch dia. polystyrene rod). The capacitors should be about 16.5 pf, but either 15- or 20-pf ceramic capacitors will give satisfactory results. A similar filter for coaxial antenna transmission line is shown at B. Both coils should be 0.12 microhenry (7 turns No. 18 enam. spaced to ½ inch on ¼-inch dia. polystyrene rod). Capacitors C_2 should be 75-pf midget ceramics, while C_3 should be a 40-pf ceramic.

transmission line and an a-c line filter of a standard variety has proven to be completely effective in eliminating the interference from a transmitter operating in one of the high-frequency amateur bands.

Designs for high-pass filters are given in figures 2 and 3. In most cases the filters may be constructed in one of the small shield boxes which are on the market. Input and output terminals may be standard connectors, or the inexpensive type of terminal strips usually used on BC and TV sets may be employed. Coaxial terminals should of course be employed when a coaxial feed line is used to the antenna. In any event, the leads from the filter box to the TV set should be very short, including both the antenna lead and the ground lead to the box itself. If the leads from the box to the set have much length, they may pick up enough signal to nullify the effects of the high-pass filter.

Blocking from 50-MHz Signals Operation on the 50-Mc. amateur band in an area where channel 2 is in use for TV imposes a special problem in the matter of blocking. The input circuits of most TV sets are sufficiently broad so that an amateur signal on the 50-MHz band will ride through with little attenuation. Also, the normal TV antenna will have quite a large response to a signal in the 50-MHz band, since the lower limit of channel 2 is 54-MHz.

High-pass filters of the normal type simply are not capable of giving sufficient attenuation to a signal whose frequency is so close to the necessary passband of the filter. Hence, a resonant circuit element, as illustrated in figure 1, must be used to trap out the amateur field at the input of the TV set. The transmitter operating frequency will have to be near the lower frequency limit of the 50-MHz band to obtain adequate rejection of the amateur signal while still not materially affecting the response of the receiver to channel 2.

Elimination of Spurious Emissions All spurious emissions from amateur transmitters (ignoring harmonic signals for the time being) must be eliminated to comply with FCC regulations. But in the past many amateur transmitters have emitted spurious signals as a result of key clicks, parasitics, and overmodulation transients. In most cases the operators of the transmitters were not aware of these emissions since they were radiated only for a short distance and hence were not brought to his attention. But with one or more TV sets in the neighborhood it is probable that such spurious signals will be brought quickly to the attention of the operator.

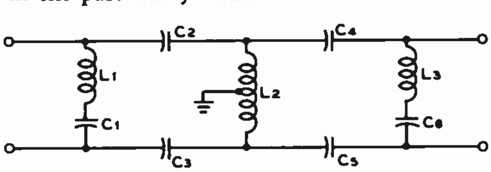


Figure 3

SERIES-DERIVED HIGH-PASS FILTER

This filter is designed for use in the 300-ohm transmission line from the TV antenna to the TV receiver. Nominal cut-off frequency is 36 MHz and maximum rejection is at about 29 MHz.
C_1,C_6—15-pf zero-coefficient ceramic
C_2,C_3,C_4,C_5—20-pf zero-coefficient ceramic
L_1 L_3—2.0 µh. About 24 turns No. 28 d.c.c. wound to 5/8" on 1/4" diameter polystyrene rod. Turns should be adjusted until the coil resonates to 29 MHz with the associated 15-pf capacitor.
L_2—0.66µh. 14 turns No. 28 d.c.c. wound to 5/8" on 1/4" dia. polystyrene rod. Adjust turns to resonate externally to 20 MHz with an auxiliary 100-pf capacitor whose value is accurately known.

16-2 Harmonic Radiation

After any condition of blocking at the TV receiver has been eliminated, and when the transmitter is completely free of transients and parasitic oscillations, it is probable that TVI will be eliminated in certain cases. Certainly general interference should be eliminated, particularly if the transmitter is a well-designed affair operated on one of the lower frequency bands, and the station is in a high-signal TV area. But when the transmitter is to be operated on one of the higher frequency bands, and particularly in a marginal TV area, the job of TVI-proofing will just have begun. The elimination of harmonic radiation from the transmitter is a difficult and tedious job which must be done in an orderly manner if completely satisfactory results are to be obtained.

First it is well to become familiar with the TV channels presently assigned, with the TV intermediate frequencies commonly used, and with the channels which will receive interference from harmonics of the various amateur bands. Figures 4 and 5 give this information.

Even a short inspection of figures 4 and 5 will make obvious the seriousness of the interference which can be caused by harmonics of amateur signals in the higher frequency bands. With any sort of reasonable precautions in the design and shielding of the transmitter it is not likely that harmonics higher than the 6th will be encountered. For this reason, the most frequently found offenders in the way of harmonic interference will almost invariably be those bands above 14 MHz.

Nature of Harmonic Interference Investigations into the nature of the interference caused by ama-

TRANSMITTER FUNDAMENTAL	2ND	3RD	4TH	5TH	6TH	7TH	8TH	9TH	10TH
7.0–7.3		21–21.9 TV I.F.			42–44 TV I.F.		56–58.4 CHANNEL 2	63–65.7 CHANNEL 3	70–73 CHANNEL 4
14.0–14.35		42–43 TV I.F.	56-57.6 CHANNEL 2	70-72 CHANNEL 4	84–86.4 CHANNEL 6	98—100.8 F-M BROADCAST			
21.0–21.45		63–64.35 CHANNEL 3	84–85.8 CHANNEL 6	105–107.25 F-M BROADCAST				189–193 CHANNELS 9 10	210–214.5 CHANNEL 13
28.0–29.7	56–59.4 CHANNEL 2	84—89.1 CHANNEL 6			168–178.2 CHANNEL 7	196—207.9 CHANNELS 10 11 12			
50.0–54.0	100–108 F-M BROADCAST		200–216 CHANNELS 11 12 13				450–486 500–540 POSSIBLE INTERFERENCE TO UHF CHANNELS		

Figure 4

HARMONICS OF THE AMATEUR BANDS

Shown are the harmonic frequency ranges of the amateur bands between 7 and 54 MHz, with the TV channels (and TV i-f systems) which are most likely to receive interference from these harmonics. Under certain conditions amateur signals in the 1.8- and 3.5-MHz bands can cause interference as a result of direct pickup in the video systems of TV receivers which are not adequately shielded.

teur signals on the TV screen, assuming that blocking has been eliminated as described earlier in this chapter, have revealed the following facts:

1. An unmodulated carrier, such as a c-w signal with the key down or an a-m signal without modulation, will give a crosshatch or herringbone pattern on the TV screen. This same general type of picture also will occur in the case of a narrow-band f-m signal either with or without modulation.
2. A relatively strong a-m or SSB signal will give in addition to the herringbone a very serious succession of light and dark bands across the TV picture.
3. A moderate strength c-w signal without transients, in the absence of overloading of the TV set, will result merely in the turning on and off of the herringbone on the picture.

To discuss condition 1 above, the herringbone is a result of the beat note between the TV video carrier and the amateur harmonic. Hence the higher the beat note the less obvious will be the resulting crosshatch. Further, it has been shown that a much stronger signal is required to produce a discernible herringbone when the interfering harmonic is as far away as possible from the video carrier, without running into the sound carrier. Thus, as a last resort, or to eliminate the last vestige of interference after all corrective measures have been taken, operate the transmitter on a frequency such that the interfering harmonic will fall as far as possible from the picture carrier. The worst possible interference to the picture from a continuous carrier will be obtained when the interfering signal is very close in frequency to the video carrier.

Isolating the Source of the Interference Throughout the testing procedure it will be necessary to have some sort of indicating device as a means of determining harmonic field intensities. The best indicator, of course, is a nearby television receiver. The home receiver may be borrowed for these tests. A portable "rabbit ears" antenna is useful since it may be moved about the transmitter site to examine the intensity of the interfering harmonics.

The first step is to turn on the transmitter and check all TV channels to determin the extent of the interference and the number of channels affected. Then disconnect the transmitting antenna and substitute a shielded dummy load, noting the change in interference level, if any. Now, remove excitation from the final stage of the transmitter, and determine the extent of interference caused by the exciter stages.

In most cases, it will be found that the interference drops materially when the transmitting antenna is removed and a dummy load substituted. It may also be found that the interference level is relatively constant, regardless of the operation of the output stage of the transmitter. In rare cases, it may be found that a particular stage in the transmitter is causing the interference and corrective measures may be applied to this stage. The common case, however, is general TVI radiating from antenna, cabinet, and power leads of the transmitter.

The first corrective measure is to properly bypass the transmitter power leads before they leave the cabinet. Each lead should be bypassed to chassis ground with a .01-μfd, 1.6-KV ceramic capacitor, or run through a 0.1-μfd, 600-volt feedthrough (*Hypass*) capacitor. If possible, the transmitter chassis should be connected to an external ground.

The next step is to check transmitter shielding. Paint should be removed from mating surfaces wherever possible and the cabinet should be made as "r-f tight" as possible in the manner discussed in Chapter 32.

16-3 Low Pass Filters

After the transmitter has been shielded, and all power leads have been filtered in such a manner that the transmitter shielding has not been rendered ineffective, the only remaining available exit for harmonic energy lies in the antenna transmission line. Hence the main burden of harmonic attenuation will fall on the low-pass filter installed between the output of the transmitter and the antenna system.

Experience has shown that the low-pass filter can best be installed externally to the main transmitter enclosure, and that the transmission line from the transmitter to the lowpass filter should be of the coaxial type. Hence the majority of low-pass filters are designed for a characteristic impedance of 52 ohms, so that RG-8/U cable (or RG-58/U for a small transmitter) may be used between the output of the transmitter and the antenna transmission line or the antenna tuner.

Transmitting-type low-pass filters for amateur use usually are designed in such a manner as to pass frequencies up to about 30 MHz without attenuation. The nominal

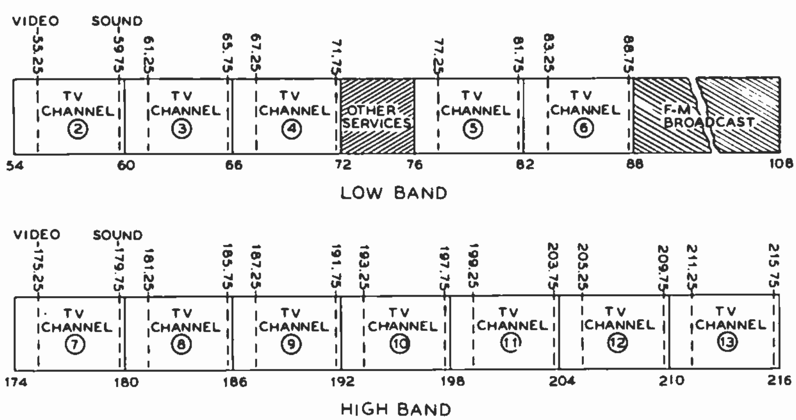


Figure 5

FREQUENCIES OF THE VHF TV CHANNELS

Showing the frequency ranges of TV channels 2 through 13, with the picture carrier and sound carrier frequencies also shown.

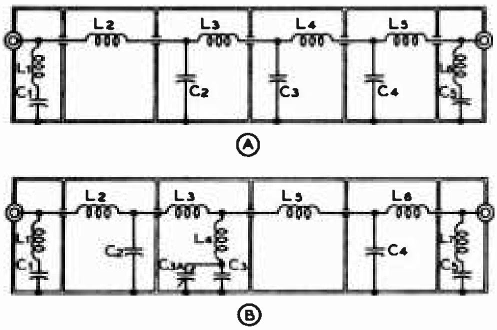


Figure 6

LOW-PASS FILTER SCHEMATIC DIAGRAMS

The filter illustrated at A uses m-derived terminating half sections at each end, with three constant-k midsections. The filter at B is essentially the same except that the center section has been changed to act as an m-derived section which can be designed to offer maximum attenuation to channels, 2, 4, 5, or 6 in accord with the constants given below. Cutoff frequency is 45 MHz in all cases. All coils, except L_4, in B above, are wound ½" I.d. with 8 turns per inch.

The A Filter

C_1, C_5—41.5 pf (40 pf will be found suitable.)
C_2, C_3, C_4—136 pf (130 to 140 pf may be used.)
L_1, L_6—0.2 μh; 3½ t. No. 14
L_2, L_5—0.3 μh; 5 t. No. 12
L_3, L_4—0.37 μh; 6½ t. No. 12

The B Filter with midsection tuned to Channel 2 (58 MHz)

C_1, C_5—41.5 pf (use 40 pf)
C_2, C_4-136 pf (use 130 to 140 pf)
C_3—87 pf (50 pf fixed and 75 pf variable in parallel.)
L_1, L_7—0.2 μh; 3½t. No. 14
L_2, L_3, L_5, L_6—0.3 μh; 5 t. No. 12
L_4—0.09 μh; 2 t. No. 14, ½" dia. ¼" long

The B Filter with midsection tuned to Channel 4 (71 MHz) All components same except that:

C_3—106 pf (use 100 pf)
L_3, L_5—0.33 μh; 6 t. No. 12
L_4—0.05 μh; 1½ t. No. 14, ⅜" dia. by ⅜" long.

The B Filter with midsection tuned to Channel 5 (81 MHz). Change the following:

C_3—113 pf (use 115 pf)
L_3, L_5—0.34 μh; 6 t. No. 12
L_4—0.033 μh; 1 t. No. 14, ⅜" dia.

The B Filter with midsection tuned to Channel 6 (86 MHz). All components are essentially the same except that the theoretical value of L_4 is changed to 0.03 μh, and the capacitance of C_3 is changed to 117 pf. (use 120 pf)

cutoff frequency of the filters is usually between 38 and 45 MHz, and *m*-derived sections with maximum attenuation in channel 2 usually are included. Well-designed filters capable of carrying any power level up to one kilowatt are available commercially from several manufacturers. Alternatively, filters in kit form are available from several manufacturers at a somewhat lower price. Effective filters may be home constructed, if the test equipment is available and if sufficient care is taken in the construction of the assembly.

Construction of Low-Pass Filters Figures 6 and 7, illustrate high-performance low-pass filters which are suitable for home construction. All are constructed in slip-cover aluminum boxes with dimensions of 17 by 3 by 2⅝ inches. Five aluminum baffle plates have been installed in the chassis to make six shielded sections within the enclosure. Feedthrough bushings between the shielded sections are *Johnson No. 135-55.*

Both the A and B filter types are designed for a nominal cutoff frequency of 45 MHz, with a frequency of maximum rejection at about 57 MHz as established by the terminating half-sections at each end. Characteristic impedance is 52 ohms in all cases. The alternative filter designs diagramed in figure 6B have provisions for an additional rejection trap in the center of the filter unit which may be designed to offer maximum rejection in channel 2, 4, 5, or 6, depending on which channel is likely to be received in the area in question. The only components which must be changed when changing the frequency of the maximum rejection notch in the center of the filter unit are inductors L_3, L_4, and L_5 and capacitor C_3. A trimmer capacitor has been included as a portion of C_3 so that the frequency of maximum rejection can be tuned accurately to the desired value. Reference to figures 4 and 5 will show the amateur bands which are most likely to cause interference to specific TV channels.

Either high-power or low-power components may be used in the filters diagramed in figure 6. With the small zero-coefficient ceramic capacitors used in the filter units of figure 6A or figure 6B, power levels up to 200 watts output may be used without danger of damage to the capacitors, *provided* the filter is feeding a 52-ohm resistive load. It may be practical to use higher levels of power with this type of ceramic capacitor in the filter, but at a power level of 200 watts on the 28-MHz band the capacitors run just perceptibly warm to the touch. As a point of interest, it is the current rating which is of significance in the capacitors used in filters such as illustrated. Since current ratings for small capacitors such as these are not readily available, it is not possible to establish an accurate power rating for such a unit. The high-power unit illustrated in figure 7, which uses *Centralab type 850S* and *854S* capacitors, has proven quite suitable for power levels up to 2 kW, PEP.

Capacitors C_1, C_2, C_4, and C_5 can be standard manufactured units with normal 5 percent tolerance. The coils for the end sections can be wound to the dimensions given (L_1, L_6, and L_7). Then the resonant frequency of the series-resonant end sections should be checked with a grid-dip meter, after the adjacent input or output terminal has been shorted with a very short lead. The coils should be squeezed or spread until resonance occurs at 57 MHz.

The intermediate *m*-derived section in the filter of figure 6B may also be checked with a grid-dip meter for resonance at the correct rejection frequency, after the hot end of L_1 has been temporarily grounded with a low-inductance lead. The variable-capacitor portion of C_3 can be tuned until resonance at the correct frequency has been obtained. Note that there is so little difference between the constants of this intermediate section for channels 5 and 6 that variation in the setting of C_3 will tune to either channel without materially changing the operation of the filter.

The coils in the intermediate sections of the filter (L_2, L_3, L_4, and L_5 in figure 6A, and L_2, L_3, L_5, and L_6 in figure 6B) may be checked most conveniently outside the filter unit with the aid of a small ceramic capacitor of known value and a grid-dip meter. The ceramic capacitor is paralleled across the small coil with the shortest possible leads. Then the assembly is placed on a cardboard box and the resonant frequency checked with a grid-dip meter.

Using Low-Pass Filters The low-pass filter connected in the output transmission line of the transmitter is capable of affording an enormous degree of harmonic attenuation. However, the filter must be operated in the correct manner or the results obtained will not be up to expectations.

In the first place, all direct radiation from the transmitter and its control and power leads must be suppressed. This subject has been discussed in the previous section. Secondly, the filter must be operated into a load impedance approximately equal to its design characteristic impedance. The filter itself will have very low losses (usually less than 0.5 db) when operated into its nominal value of resistive load. But if the filter is not terminated correctly, its losses will become excessive, and it will not present the correct value of load impedance to the transmitter.

If a filter, being fed from a high-power transmitter, is operated into an incorrect termination it may be damaged; the coils may be overheated and the capacitors destroyed as a result of excessive r-f currents. Hence it is wise when first installing a low-pass filter, to check the standing-wave ratio of the load being presented to the output of the filter with a standing-wave meter of any of

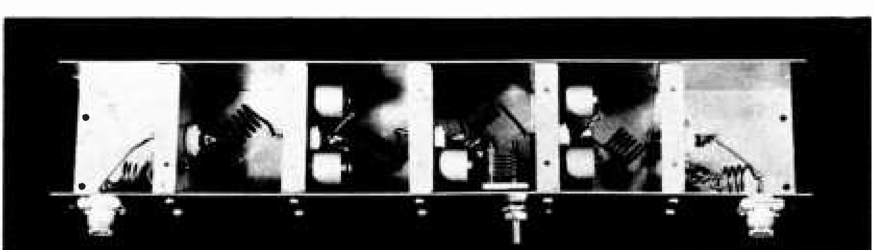


Figure 7

PHOTOGRAPH OF THE B FILTER WITH COVER REMOVED

The midsection in this filter is adjusted for maximum rejection of channel 4. Note that the main coils of the filter are mounted at an angle of about 45 degrees so that there will be minimum inductive coupling from one section to the next through the holes in the aluminum partitions. Mounting the coils in this manner was found to give a measurable improvement in the attenuation characteristics of the filter.

the conventional types. Then the antenna termination or the antenna coupling should be adjusted, with low power on the transmitter, until the s.w.r. of the load being presented to the filter is less than 2.0, and preferably below 1.5.

Half-Wave Filters A *half-wave filter* is an effective device for TVI suppression and is easily built. It offers the advantage of presenting the same value of

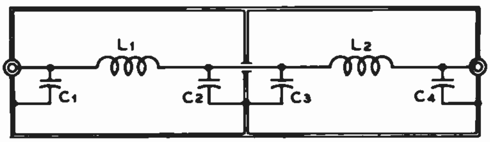


Figure 8

SCHEMATIC OF THE SINGLE-SECTION HALF-WAVE FILTER

The constants given below are for a characteristic impedance of 52 ohms, for use with RG-8/U and RG-58/U cable. Coil L_1 should be checked for resonance at the operating frequency with C_2, and the same with L_2 and C_4. This check can be made by soldering a low-inductance grounding strap to the lead between L_1 and L_2 where it passes through the shield. When the coils have been trimmed to resonance with a grid-dip meter, the grounding strap should of course be removed. This filter type will give an attenuation of about 30 db to the second harmonic, about 48 db to the third, about 60 db to the fourth, 67 to the fifth, etc., increasing at a rate of about 30 db per octave.

C_1, C_2, C_3, C_4—Silver mica or small ceramic for low power, transmitting type ceramic for high power. Capacitance for different bands is given below.

160 meters—1700 pf
80 meters— 850 pf
40 meters— 440 pf
20 meters— 220 pf
10 meters— 110 pf
6 meters— 60 pf

L_1, L_2—May be made up of sections of B&W Miniductor for power levels below 250 watts, or of No. 12 enam. for power up to one kilowatt. Approximate dimensions for the coils are given below, but the coils should be trimmed to resonate at the proper frequency with a grid-dip meter as discussed above. All coils except the ones for 160 meters are wound 8 turns per inch.

160 meters—4.2 μh; 22 turns No. 16 enam. 1" dia. 2" long

80 meters—2.1 μh; 13 t. 1" dia. (No. 3014 Miniductor or No. 12 at 8 t.p.i.

40 meters—1.1 μh; 8 t. 1" dia. (No. 3014 or No. 12 at 8 t.p.i.)

20 meters—0.55 μh; 7 t. 3/4" dia. (No. 3010 or No. 12 at 8 t.p.i.)

10 meters—0.3 μh; 6 t. 1/2" dia. (No. 3002 or No. 12 at 8 t.p.i.)

6 meters—0.17 μh; 4 t. 1/2" dia. (No. 3002 or No. 12 at 8 t.p.i.)

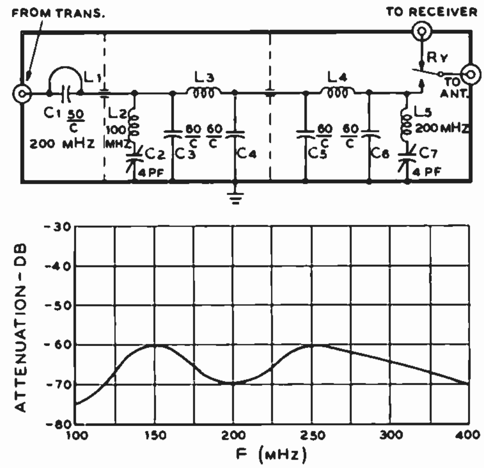


Figure 9

SIX METER TVI FILTER

C_1—50-pf Centralab 850S-50Z. Resonates with L_1 to 200 MHz.
C_2, C_7—4-pf piston capacitor. JFD type VC-4G.
C_3, C_4, C_5, C_6—60 pf. Three 20-pf capacitors in parallel. Centralab 853A-20Z.
L_1—Copper strap, 1/2" wide, 2 1/4" long, 1 7/8" between mounting holes, approximately 0.01" thick. Strap is bent in U-shape around capacitor and bolted to capacitor terminals.
L_2—11 turns #18 enam. wire, 1/4" diameter, 3/4" long, airwound. Resonates to 100 MHz with capacitor C_2.
L_3, L_4—3 turns 3/16" tubing, 1 1/4" i.d., spaced to occupy about 2 1/2". Turns are adjusted to resonate each section at 50 MHz.
L_5—6 turns #18 enam. wire, 1/4" diameter, 5/8" long, airwound. Resonates to 200 MHz with capacitor C_7.

impedance at the input terminal as appears as a load across the output terminal. The filter is a single-band unit, offering high attenuation to the second- and higher-order harmonics. Design data for high-frequency half-wave filters is given in figure 8.

A High-Power Filter for Six Meters The second and higher harmonics of a six-meter transmitter fall directly into the f-m and uhf and vhf television bands. An effective low-pass filter is required to adequately suppress unwanted transmitter emissions falling in these bands. Described in this section is a six-meter TVI filter rated at the two-kilowatt level which provides better than 75 decibels suppression of the second harmonic and better than 60 decibels suppression of higher harmonics of a six-meter transmitter (figure 9). The

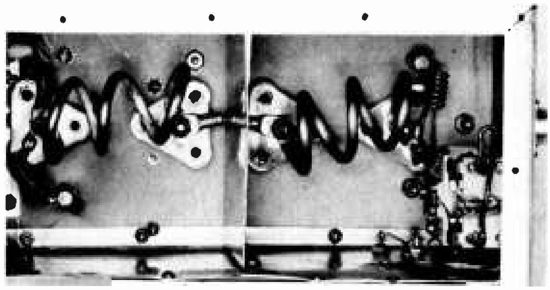


Figure 10

INTERIOR VIEW OF SIX-METER FILTER

The input compartment of the filter is at the left. The series coil is wound of copper tubing, and the connection to the output section (right) is made by a length of tubing which passes through a hole in the center shield. Series elements carry less current and employ wirewound coils. At right is antenna relay, with power leads bypassed as they leave filter compartment. Filter is set to correct frequency by adjusting the inductance of the tubing coils.

unit is composed of a half-wave filter with added end sections which are tuned to 100 MHz and 200 MHz. An auxiliary filter element in series with the input is tuned to 200 MHz to provide additional protection to television channels 11, 12, and 13.

The filter (figure 10) is built in an aluminum box measuring 4″ x 4″ x 10″ and uses *type-N* coaxial fittings. The half-wave filter coils are wound of 3/16-inch diameter copper tubing and have large copper lugs soldered to the ends. The 60-pf capacitors are made up of three 20-pf, 5kv ceramic units in parallel. A small sheet of copper is cut in triangular shape and joins the capacitor terminals and a coil lug is attached to the center of the triangle with heavy brass bolts.

The parallel-tuned 200-MHz series filter element at the input terminal is made of a length of copper strap shunted across a 50-pf, 5kv ceramic capacitor. In this particular filter, the parallel circuit was affixed to the output capacitor of the pi-network tank circuit of the transmitter and does not show in the photograph.

The filter is adjusted by removing the connections from the ends of the half-wave sections and adjusting each section to 50 MHz by spreading the turns of the coil with a screwdriver while monitoring the resonant frequency with a grid-dip oscillator. The next step is to ground the top end of each series-tuned section (C_2, L_2 and C_7, L_5) with a heavy strap. The input section is tuned to 100 MHz and the output section to 200 MHz. When tuning adjustments are completed, the straps are removed and the top of the filter box is held in place with sheet-metal screws.

16-4 Broadcast Interference

Interference to the reception of signals in the broadcast band (540 to 1600 kHz) or in the f-m broadcast band (88 to 108 MHz) by amateur transmissions is a serious matter to those amateurs living in densely populated areas. Although broadcast interference has recently been overshadowed by the seriousness of television interference, the condition of BCI is still present.

In general, signals from a transmitter operating properly are not picked up by receivers tuned to other frequencies unless the receiver is of inferior design, or is in poor condition. Therefore, if the receiver is of good design and is in good repair, the burden of rectifying the trouble rests with the owner of the interfering station. Phone and c-w stations both are capable of causing broadcast interference, key-click annoyance from the code transmitters being particularly objectionable.

Broadcast interference, as covered in this section refers primarily to standard (amplitude-modulated, 550-1600 kHz) broadcast. Interference with f-m broadcast reception is much less common, due to the wide separation in frequency between the f-m broadcast band and the more popular amateur bands, and due also to the limiting action which exists in all types of f-m receivers. Occasional interference with f-m broadcast by a harmonic of an amateur transmitter has been reported; if this condition is encountered, it may be eliminated by the procedures discussed in the first portion of this chapter under *Television Interference*.

Interference Classifications Depending on whether it is traceable directly to causes within the *station* or within the *receiver*, broadcast interference may be divided into two main classes. For example, that type of interference due to transmitter overmodulation or flat-topping is at once listed as being caused by improper operation, while an interfering signal that tunes in and out with a broadcast station is probably an indication of cross-modulation or image response in the receiver, and the poorly designed input stage of the receiver is held liable. The various types of interference and recommended cures will be discussed in the following paragraphs.

Blanketing This is not a tunable effect, but a total blocking of the receiver. A more or less complete "washout" covers the entire receiver range when the carrier is switched on. This produces either a complete blotting out of all broadcast stations, or else knocks down their volume several decibels—depending on the severity of the interference. Voice modulation causing the blanketing will be highly distorted or even unintelligible. Keying of the carrier which produces the blanketing will cause an annoying fluctuation in the volume of the broadcast signals.

Blanketing generally occurs in the immediate neighborhood (inductive field) of a powerful transmitter, the affected area being directly proportional to the power of the transmitter. Also, it is more prevalent with transmitters which operate in the 160-meter and 80-meter bands, as compared to those operating on the higher frequencies.

In the rare case where the broadcast receiver utilizes an external antenna, a simple wavetrap tuned to the frequency of the local transmitter will minimize the signal entering the receiver (figure 11). The wavetrap should be installed as close to the reciever antenna terminal as practical. Most broadcast receivers, however, dispense with an external antenna and instead use a ferrite "loopstick" antenna concealed within the receiver cabinet. Loopstick pickup at the higher frequencies is quite restricted and it is usually found that severe blanketing may be reduced by merely bypassing each side of the receiver power line to the chassis of the receiver with a pair of .01-μfd, 1.6-kV ceramic disc capacitors.

Phantoms With two strong local carriers applied to a nonlinear impedance, the beat note resulting from cross modulation between them may fall on some frequency within the broadcast band and will be audible at that point. If such a "phantom" signal falls on a local broadcast frequency, there will be heterodyne interference as well. This is a common occurrence with broadcast receivers in the neighborhood of two amateur stations, or an amateur and a broadcast station. It also sometimes occurs

1.8 MHz	1 inch No. 30 enam. closewound on 1" form	75-pf var.
3.5 MHz	42 turns No. 30 enam. closewound on 1" form	50-pf var.
7.0 MHz	23 turns No. 24 enam. closewound on 1" form	50-pf var.
14 MHz	10 turns No. 24 enam. closewound on 1" form	50-pf var.
21 MHz	7 turns No. 24 enam. closewound on 1" form	50-pf var.
28 MHz	4 turns No. 24 enam. closewound on 1" form	25-pf var.
50 MHz	3 turns No. 24 enam. spaced ½" on 1" form	25-pf var.

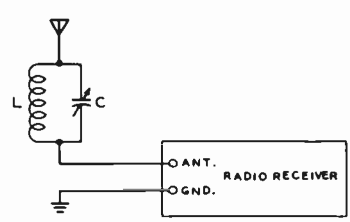


Figure 11

WAVETRAP FOR BCI

If the radio receiver has an external antenna, a parallel-tuned circuit may be placed at the antenna terminal and tuned to the frequency of the offending signal. Table shows circuit constants for amateur-band wavetraps.

when one of the stations is located in the immediate vicinity.

As an example: an amateur signal on 3514 kHz might beat with a local 2414 kHz carrier to produce a 1100-kHz phantom. If the two carriers are strong enough in the vicinity of a circuit which can cause rectification, the 1100-kHz phantom will be heard in the broadcast band. A poor contact between two oxidized wires can produce rectification.

Two stations must be transmitting simultaneously to produce a phantom signal; when either station goes off the air the phantom disappears. Hence, this type of interference is apt to be reported as highly intermittent and might be difficult to duplicate unless a test oscillator is used "on location" to simullate the missing station. Such interference cannot be remedied at the transmitter, and often the rectification takes place some distance from the receivers. In such occurrences it is most difficult to locate the source of the trouble.

It will also be apparent that a phantom might fall on the intermediate frequency of a simple superhet receiver and cause interference of the untunable variety if the manufacturer has not provided an i-f wavetrap in the antenna circuit.

This particular type of phantom may, in addition to causing i-f interference, generate harmonics which may be tuned in and out with heterodyne whistles from one end of the receiver dial to the other. It is in this manner that *birdies* often result from the operation of nearby amateur stations.

When one component of a phantom is a

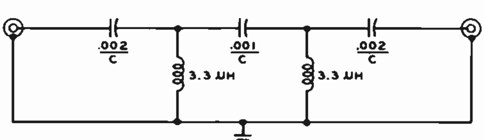


Figure 12

HYPASS FILTER FOR AMATEUR RECEIVER

This simple filter attenuates signals below 1600 kHz to reduce overload caused by strong nearby broadcast station. Filter is designed to be placed in series with coaxial line to receiver. Filter should be built in small shield box with appropriate coaxial fittings. J. W. Miller ferrite choke 74F336AP may be used for 3.3 μH inductor.

steady unmodulated carrier, only the intelligence presence on the other carrier is conveyed to the broadcast receiver.

Phantom signals almost always may be identified by the suddenness with which they are interrupted, signaling withdrawal of one party of the union. This is especially baffling to the inexperienced interference locater, who observes that the interference suddenly dissapears, even though his own transmitter remains in operation.

If the mixing or rectification is taking place in the receiver itself, a phantom signal may be eliminated by removing either one of the contributing signals from the receiver input circuit. In the case of phantom crosstalk in an amateur-band receiver, a simple high-pass filter designed to attenuate signals below 1600 kHz may be placed in the coaxial antenna lead to the receiver (figure 12). This will greatly reduce the strength of local broadcast signals, which in a metropolitan area may amount to fractions of a volt on the receiver input circuit.

A-c/d-c Receivers Inexpensive table-model a-c/d-c receivers are particularly susceptible to interference from amateur transmissions. In most cases the receivers are at fault; but this does not absolve the amateur of his responsibility in attempting to eliminate the interference.

In cases of interference to inexpensive receivers, particularly those of the a-c/d-c type, is will be found that stray receiver rectification is causing the trouble. The offending stage usually will be found to be a high-μ triode as the first audio stage following the second detector. Tubes of this type are quite nonlinear in their grid characteristic, and hence will readily rectify any r-f signal appearing between grid and cathode. The r-f signal may get to the tube as a result of direct signal pickup due to the lack of shielding, but more commonly will be fed to the tube from the power line as a result of the series heater string.

The remedy for this condition is simply to ensure that the cathode and grid of the high-μ audio tube (usually a 6AV6 or equivalent) are at the same r-f potential. This is accomplished by placing an r-f bypass capacitor with the shortest possible leads directly from grid to cathode, and then adding an impedance in the lead from the volume control to the grid of the audio tube. The impedance may be an amateur band r-f choke (such as a National R-100U) for best results, but for a majority of cases it will be found that a 47,000-ohm ½-watt resistor in series with this lead will give satisfactory operation. Suitable circuits for such an operation on the receiver are given in figure 13.

In many a-c/d-c receivers there is no r-f bypass included across the plate-supply rectifier for the set. If there is an appreciable level of r-f signal on the power line feeding the receiver, r-f rectification in the power rectifier of the receiver can cause a particularly bad type of interference which may be received on other broadcast receivers in the vicinity in addition to the one causing the rectification. The soldering of a 0.01-μfd, 1.6 kV disc ceramic capacitor directly from anode to cathode of the power rectifier (whether it is of the vacuum-tube or silicon-rectifier type) usually will bypass the r-f signal across the rectifier and thus eliminate the difficulty.

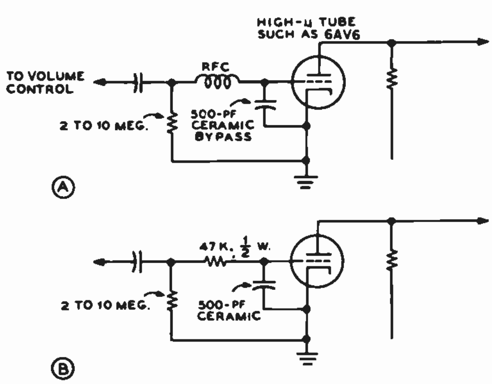


Figure 13

CIRCUITS FOR ELIMINATING AUDIO-STAGE RECTIFICATON

"Floating" Volume-Control Shafts Several sets have been encountered where there was only a slight interfering signal; but, on placing one's hand to the volume control, the signal would greatly increase. Investigation revealed that the volume control was installed with its shaft insulated from ground. The control itself was connected to a critical part of a circuit, in many instances to the grid of a high-gain audio stage. The cure is to install a volume control with *all* the terminals insulated from the shaft, and then to ground the shaft.

Image Interference In addition to those types of interference already discussed, there are two more which are common to superhet receivers. The prevalence of these types is of great concern to the amateur, although the responsibility for their existence more properly rests with the broadcast receiver.

The mechanism whereby image production takes place may be explained in the following manner: when the first detector is set to the frequency of an incoming signal, the high-frequency oscillator is operating on another frequency which differs from the signal by the number of kHz of the intermediate frequency. Now, with the setting of these two stages undisturbed, there is another signal which will beat with the high-frequency oscillator to produce an i-f signal. This other signal is the so-called *image*, which is separated from the desired signal by twice the intermediate frequency.

Thus, in a receiver with a 175-kHz intermediate frequency tuned to 1000 kHz; the h-f oscillator is operating on 1175 kHz, and a signal on 1350 kHz (1000 kHz plus 2 × 175 kHz) will beat with this 1175 kHz oscillator frequency to produce the 175-kHz i-f signal. Similarly, when the same receiver is tuned to 1450 kHz, an amateur signal on 1800 kHz can come through.

If the image appears only a few Hz or kHz from a broadcast carrier, heterodyne interference will be present as well. Otherwise, it will be tuned in and out in the manner of a station operating in the broadcast band. Sharpness of tuning will be comparable to that of broadcast stations producing the same avc voltage at the receiver.

The second variety of superhet interference is the result of harmonics of the receiver high-frequency oscillator beating with amateur carriers to produce the intermediate frequency of the receiver. The amateur transmitter will always be found to be on a frequency equal to some harmonic of the receiver hf oscillator, *plus or minus the intermediate frequency.*

As an example: when a broadcast superhet with 465-kHz intermediate frequency is tuned to 1000 kHz, its high-frequency oscillator operates on 1465 kHz. The third harmonic of this oscillator frequency is 4395 kHz, which will beat with an amateur signal on 3930 kHz to send a signal through the i-f amplifier. The 3930 kHz signal would be tuned in at the 1000-kHz point on the dial.

Some oscillator harmonics are so related to amateur frequencies that more than one point of interference will occur on the receiver dial. Thus, a 3500-kHz signal may be tuned in at six points on the dial of a nearby broadcast superhet having a 175-kHz intermediate frequency and no r-f stage.

Insofar as remedies for image and harmonic superhet interference are concerned, it is well to remember that *if* the amateur signal did not in the first place reach the input stage of the receiver, the annoyance would not have been created. It is therefore good policy to try to eliminate it by means of a wavetrap or low-pass filter. Broadcast superhets are not always the acme of good shielding, however, and the amateur signal is apt to enter the circuit through channels other than the input circuit. If a wavetrap or filter will not cure the trouble, the only alternative will be to attempt to select a transmitter frequency such that neither image nor harmonic interference will be set up on favorite stations in the susceptible receivers. The equation given earlier may be used to determine the proper frequencies.

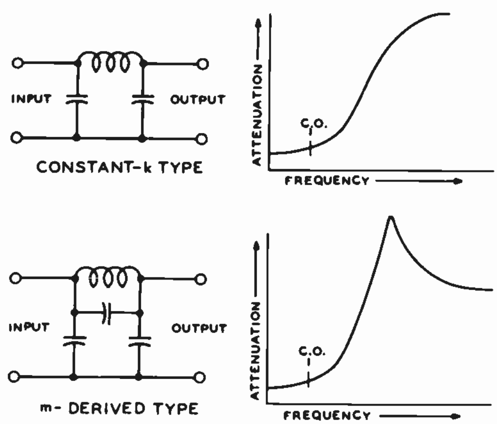


Figure 14

TYPES OF LOW-PASS FILTERS

Filters such as these may be used in the circuits between the antenna and the input of the receiver.

Low-Pass Filters The greatest drawback of the wavetrap is the fact that it is a single-frequency device; i.e., it may be set to reject at one time only one frequency (or, at best, an extremely narrow band of frequencies). Each time the frequency of the interfering transmitter is changed, every wavetrap tuned to it must be retuned. A much more satisfactory device is the *wave filter* which requires no tuning. One type, the low-pass filter, passes all frequencies below one critical frequency, and eliminates all higher frequencies. It is this property that makes the device ideal for the task of removing amateur frequencies from broadcast receivers.

A good low-pass filter designed for maximum attenuation around 1700 kHz will pass all broadcast carriers, but will reject signals originating in any amateur band. Naturally such a device should be installed only in standard broadcast receivers, never in all-wave sets.

Two types of low-pass filter sections are shown in figure 14. A composite arrangement comprising a section of each type is more effective than either type operating alone. A composite filter composed of one *k*-section and one shunt-derived *m*-section is shown in figure 15, and is highly recommended. The *m*-section is designed to have

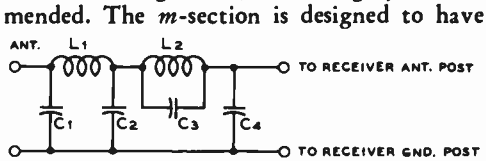


Figure 15

COMPOSITE LOW-PASS FILTER CIRCUIT

This filter is highly effective in reducing broadcast interference from all high-frequency stations, and requires no tuning. Constants for 400-ohm terminal impedance and 1600 kHz cutoff are as follows: L_1, 65 turns No. 22 d.c.c. closewound on 1½ in. dia. form. L_2, 41 turns ditto, not coupled to L_1. C_1, 250-pf fixed mica capacitor. C_2, 400-pf fixed mica capacitor. C_3 and C_4, 150-pf fixed mica capacitors, former of 5% tolerance. With some receivers, better results will be obtained with a 200-ohm carbon resistor inserted between the filter and antenna post on the receiver. With other receivers the effectiveness will be improved with a 600-ohm carbon resistor placed from the antenna post to the ground post on the receiver. The filter should be placed as close to the receiver terminals as possible.

maximum attenuation at 1700 kHz, and for that reason C_3 should be of the close-tolerance variety.

If a fixed 150-pf mica capacitor of 5 percent tolerance is not available for C_1, a compression trimmer covering the range of 125-175 pf may be substituted and adjusted to give maximum attenuation at about 1700 kHz.

16-5 Miscellaneous Interference

Stereo Interference Stereo sound systems may receive interference from strong nearby radio transmitters, both amateur and broadcast. In most cases, the interference is caused by stray pickup of the r-f signal by the interconnecting leads of the stereo system accompanied by audio rectification in the low-level stages of the system. The solution to this difficulty, in general, is to bypass and shield all speaker leads and bypass the power lead to the amplifier or tuner units. The power line leads may be bypassed to the chassis with a pair of .01-μfd, 1.6-kV ceramic capacitors. Speaker leads should be shielded (with the shield grounded) and each lead bypassed to the amplifier chassis with a .001-μfd, 600-volt ceramic capacitor. Shielded leads should also be employed between the amplifier and tuner and the phono cartridge. The framework of the turntable should be grounded to the chassis of the amplifier to reduce stray r-f pickup in the turntable wiring and equipment.

In some cases it may be necessary to install an r-f filter in the input circuit of the amplifier. A small r-f choke in series with the input lead, together with a 500-pf ceramic capacitor between leads and ground will reduce r-f pickup in the input circuit of the amplifier. In high-impedance circuits, it may be necessary to decrease the size of the capacitor to 50 pf or so to prevent loss of the higher audio frequencies.

Telephone Interference The carbon microphone of the telephone often serves as an efficient rectifier of nearby radio signals, injecting the modulation of the signal on the telephone circuit. Older (type 300) telephones merely require the installation of a .001-μfd ceramic capacitor across the terminals of the carbon microphone. Most telephone companies supply a special capacitor for this purpose on request.

The newer 500-series telephones, however, contain an automatic-level control circuit in the base which includes a thermistor unit which is sensitive to strong r-f fields. In addition to the microphone capacitor, the 500-series unit requires the installation of a pair of 2.5-mH, 100-ma r-f chokes, one in series with each side of the line, placed within the telephone base underneath the dial mechanism. This prevents r-f pickup by the telephone line from reaching the thermistor, thus eliminating the interference.

Power-Line Interference Power-line interference may reach a radio receiver by transmission along the line or by direct radiation. Typical sources of power-line interference are spark and electrostatic discharge. Spark discharge from brush-type motors, heaters for fish aquariums, thermostats on sleeping blankets, and heating pads are prolific sources of such interference. If the interfering unit can be located, bypass capacitors on the power line directly at the unit will usually suppress the noise. The noise may often be located by using a portable radio as a direction finder, homing in on the noise source. Direct power-line noise, caused by leaky insulators or defective hardware on high-voltage transmission lines is harder to pinpoint, as the noise may be carried for a considerable distance along the line. Standing waves of noise are also apparent on power lines, leading to false noise peaks that confuse the source. Many power companies have a program of locating interference and it is recommended that the amateur contact the local company office and register a complaint of power-line interference rather than to try and find it himself, since the cure for such troubles must be applied by the company, rather than the amateur.

Electrostatic discharge may be caused by intermittent contact between metallic objects in a strong electric field. Guy wires or hardware on power poles are a source of this form of interference. In addition, loose hardware on a nearby TV antenna, or the tower of the amateur antenna may cause this type of interference in the presence of a nearby power line. This type of interference is hard to pinpoint, but may often be found with the aid of a portable radio. In any event, suspected power-line interference originating on the power-line system should be left to the power-company interference investigator.

16-6 Help in Solving TVI

Some TV set manufacturers will supply high-pass TV filters at cost for their receivers or provide information on TVI reduction upon request. When writing to the manufacturer about TVI problems, supply complete details, including model and serial number of the TV set involved; the name and address of the TV set owner; the name, address, and call letters of the amateur involved; and particulars of the interference problem (channels affected, frequency of amateur transmitter, sound or picture affected, etc.) The following manufacturers can supply information and assistance:

Motorola
Consumer Product Division
9401 W. Grand Ave.
Franklin Park, Ill. 60313

Heath
Benton Harbor, Michigan 49022

Olympic International
88-89 Union Turnpike
Glendale, N.Y. 12270

RCA Sales Corp.
600 North Sherman Drive
Indianapolis, Ind. 46201

Magnavox
7 regional service centers in:
East Rutherford, N.J.
Atlanta, Georgia
Westlake, Ohio
Skokie, Ill.
Dallas, Texas
Torrance, Calif.
South San Francisco, Calif.

TMA Company (Muntz, TMA, Howard Stereo)
1020 Noel Ave.
Wheeling, Ill. 60090

Emerson
Emerson TV Sales Corp.
Jersey City, N.J. 07302

Philco Customer Service
Box 3635
Philadelphia, Penna. 19125

Zenith Service Dept.
Zenith Sales Co.
5801 West Dickens
Chicago, Ill. 60639

Sylvania Customer Service
700 Endicott St.
Batavia, N.Y. 14020

Sears
Dept. 698/731A
Staff Offices
Sears, Roebuck & Co.
925 So. Homan Ave.
Chicago, Ill. 60607

CHAPTER SEVENTEEN

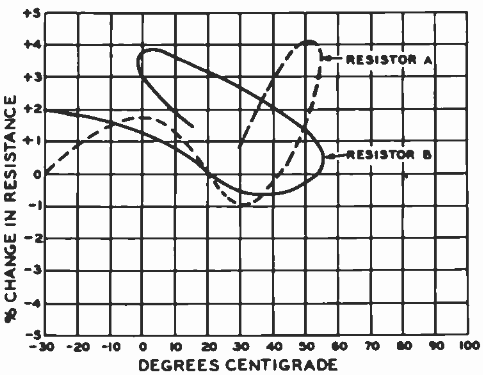
Equipment Design

The performance of communication equipment is a function of the design, and is dependent on the execution of the design and the proper choice of components. This chapter deals with the study of equipment circuitry and the basic components that go to make up this circuitry. Modern components are far from faultless. Resistors have inductance and reactance, and inductors have resistance and distributed capacitance. None of these residual attributes show up on circuit diagrams, yet they are as much responsible for the success or failure of the equipment as are the necessary and vital bits of resistance, capacitance, and inductance. Because of these unwanted attributes, the job of translating a circuit on paper into a working piece of equipment often becomes an impossible task to those individuals who disregard such important trivia. Rarely do circuit diagrams show such pitfalls as ground loops and residual inductive coupling between stages. Parasitic resonant circuits are seldom visible from a study of the schematic. Too many times radio equipment is rushed into service before it has been entirely checked. The immediate and only too apparent results of this enthusiasm are receiver instability, transmitter instability, difficulty of neutralization, r.f. wandering all over the equipment, and a general "touchiness" of adjustment. Hand in glove with these problems go the more serious ones of receiver overload, TVI, keyclicks, and parasitics. By paying attention to detail, with a good working knowledge of the limitations of the components, and with a basic concept of the actions of ground currents, the average amateur will be able to build equipment that will work "just like the book says."

The twin problems of TVI and parasitics are an outgrowth of the major problem of over all circuit design. If close attention is paid to the cardinal points of circuitry design, the secondary problems of TVI and parasitics will in themselves be solved.

17-1 Resistors

The resistance of a conductor is a function of the material, the form the material takes, the temperature of operation, and the frequency of the current passing through the resistance. In general, the variation in resistance due to temperature is directly proportional to the temperature change. With most wirewound resistors, the resistance increases with temperature and returns to its original value when the temperature drops to normal. So called composition or carbon resistors have less reliable temperature/resistance characteristics. They usually have a positive temperature coefficient, but the retrace curve as the resistor is cooled is often erratic, and in many cases the resistance does not return to its original value after a heat cycle. It is for this reason that care must be taken when soldering composition resistors in circuits that require close control of the



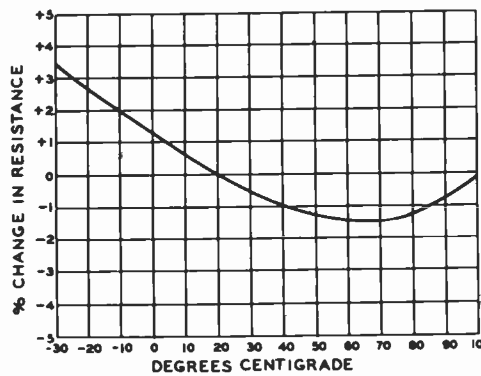


Figure 1

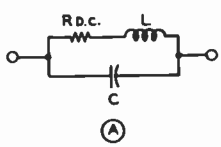
HEAT CYCLE OF UNCONDITIONED COMPOSITION RESISTORS

HEAT CYCLE OF CONDITIONED COMPOSITION RESISTORS

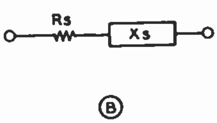
resistance value. Matched resistors used in precise circuits can be driven out of tolerance by the act of soldering them into the circuit. Long leads should be left on the resistors and long-nose pliers should grip the lead between the iron and the body of the resistor to act as a heat block. General temperature characteristics of typical carbon resistors are shown in figure 1. The behavior of an individual resistor will vary from these curves depending on the manufacturer, the size and wattage of the resistor, etc.

Inductance of Resistors Every resistor because of its physical size has in addition to its desired resistance, less desirable amounts of inductance and distributed capacitance. These quantities are illustrated in figure 2A, the general equivalent circuit of a resistor. This circuit represents the actual impedance network of a resistor at any frequency. At a certain specified frequency the impedance of the resistor may be thought of as a series reactance (X_s) as shown in figure 2B. This reactance may be either inductive or capacitive depending on whether the residual inductance or the distributed capacitance of the resistor is the dominating factor. As a rule, skin effect tends to increase the reactance with frequency, while the capacitance between turns of a wirewound resistor, or capacitance between the granules of a composition resistor tends to cause the reactance and resistance to drop with frequency. The behavior of various types of composition resistors over a

Figure 2



EQUIVALENT CIRCUIT OF A RESISTOR



EQUIVALENT CIRCUIT OF A RESISTOR AT A PARTICULAR FREQUENCY

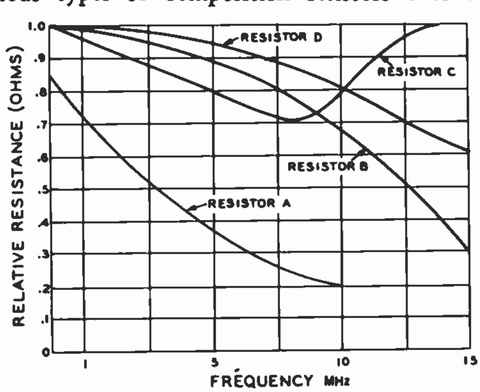


Figure 3

FREQUENCY EFFECTS ON SAMPLE COMPOSITION RESISTORS

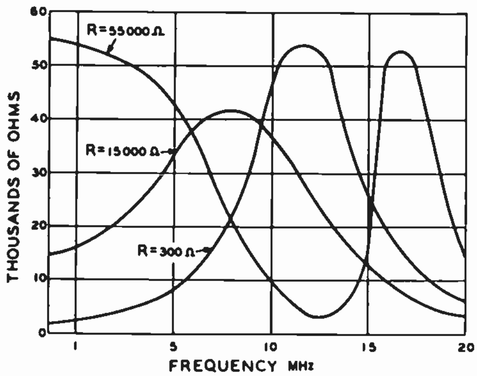


Figure 4

CURVES OF THE IMPEDANCE OF WIRE-WOUND RESISTORS AT RADIO FREQUENCIES

large frequency range is shown in figure 3. By proper component design, noninductive resistors having a minimum of residual reactance characteristics may be constructed. Even these have reactive effects that cannot be ignored at high frequencies.

Wirewound resistors act as low-Q inductors at radio frequencies. Figure 4 shows typical curves of the high-frequency characteristics of cylindrical wirewound resistors. In addition to resistance variations wirewound resistors exhibit both capacitive and inductive reactance, depending on the type of resistor and the operating frequency. In fact, such resistors perform in a fashion as low-Q r-f chokes below their parallel self-resonant frequency.

17-2 Capacitors

The inherent residual characteristics of capacitors include series resistance, series inductance and shunt resistance, as shown in figure 5. The series resistance and inductance depend to a large extent on the physical configuration of the capacitor and on the material from which it is composed. Of great interest to the amateur constructor is the series inductance of the capacitor. At a certain frequency the series inductive reactance of the capacitor and the capacitive reactance are equal and opposite, and the capacitor is in itself series resonant at this frequency. As the operating frequency of the circuit in which the capacitor is used is increased above the series-resonant frequency, the effectiveness of the capacitor as a bypassing element deteriorates until the unit is useless.

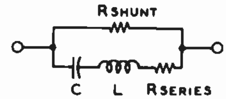


Figure 5

EQUIVALENT CIRCUIT OF A CAPACITOR

Bypass Capacitors The usual forms of bypass capacitors have dielectrics of paper, mica, or ceramic. For audio work, and low-frequency r-f work up to perhaps 2 MHz or so, the paper capacitors are satisfactory as their relatively high internal inductance has little effect on the proper operation of the circuit. The actual amount of internal inductance will vary widely with the manufacturing process, and some types of paper capacitors have satisfactory characteristics up to a frequency of 5 MHz or so.

When considering the design of transmitting equipment, it must be remembered that while the transmitter is operating at some relatively low frequency (for example, 7 MHz), there will be harmonic currents flowing through the various bypass capacitors of the order of 10 to 20 times the operating frequency. A capacitor that behaves properly at 7 MHz however, may offer considerable impedance to the flow of these harmonic currents. For minimum harmonic generation and radiation, it is obviously of greatest importance to employ bypass capacitors having the lowest possible internal inductance.

CAPACITOR		LEAD LENGTHS	RESONANT FREQ.
.02	µfd MICA	NONE	44.5 MHz
.002	µfd MICA	NONE	23.5 MHz
.01	µfd MICA	¼″	10 MHz
.0009	µfd MICA	¼″	55 MHz
.002	µfd CERAMIC	⅝″	24 MHz
.001	µfd CERAMIC	¼″	55 MHz
500	pf BUTTON	NONE	220 MHz
.0005	µfd CERAMIC	¼″	90 MHz
.01	µfd CERAMIC	½″	14.5 MHz

Figure 6

SELF-RESONANT FREQUENCIES OF VARIOUS CAPACITORS WITH RANDOM LEAD LENGTH

Mica-dielectric capacitors have much less internal inductance than do most paper capacitors. Figure 6 lists self-resonant frequencies of various mica capacitors having various lead lengths. It can be seen from inspection of this table that most mica capacitors become self-resonant in the 12- to 50-MHz region. The inductive reactance they would offer to harmonic currents of 100 MHz, or so, would be of considerable magnitude. In certain instances it is possible to deliberately series-resonate a mica capacitor to a certain frequency somewhat below its normal self-resonant frequency by trimming the leads to a critical length. This is sometimes done for maximum bypassing effect in the region of 40 to 60 MHz.

The *button-mica* capacitors shown in figure 7 are especially designed to have extremely low internal inductance. Certain types of button-mica capacitors of small physical size have a self-resonant frequency in the region of 600 MHz.

Ceramic-dielectric capacitors in general have the lowest amount of series inductance per unit of capacitance of these three universally used types of bypass capacitors. Typical resonant frequencies of various ceramic units are listed in figure 6. Ceramic capacitors are available in various voltage and capacitance ratings and different physical configurations. Standoff types such as shown in figure 7 are useful for bypassing socket and transformer terminals. Two of these capacitors may be mounted in close proximity on a chassis and connected together by an r-f choke to form a highly effective r-f filter. The inexpensive *disc* type of ceramic capacitor is recommended for general bypassing in r-f circuitry, as it is effective as a bypass unit to well over 100 MHz.

The large TV *doorknob* capacitors are useful as by-pass units for high voltage lines. These capacitors have a value of 500 pf, and are available in voltage ratings up to 40,000 volts. The dielectric of these capacitors is usually titanium dioxide. This material exhibits piezoelectric effects, and capacitors employing it for a dielectric will tend to "talk-back" when a-c voltages are applied across them.

An important member of the varied line of capacitors is the *coaxial*, or *Hypass*, type of capacitor. These capacitors exhibit superior bypassing qualities at frequencies up

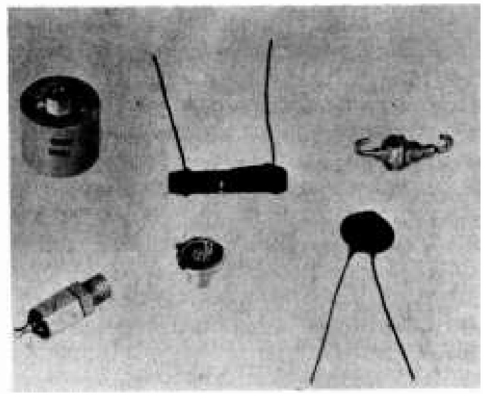


Figure 7

TYPES OF CERAMIC AND MICA CAPACITORS SUITABLE FOR HIGH-FREQUENCY BYPASSING

The Centralab 858S (1000 pf) is recommended for screen and plate circuits of tetrode tubes.

to 200 MHz and the bulkhead type is especially effective when usd to filter leads passing through partition walls between two stages.

Variable Air Capacitors Even though air is the perfect dielectric, air capacitors exhibit losses because of the inherent resistance of the metallic parts that make up the capacitor. In addition, the leakage loss across the insulating supports may become of some consequence at high frequencies. Of greater concern is the inductance of the capacitor at high frequencies. Since the capacitor must be of finite size, it will have tie rods, metallic braces, and end plates; all of which contribute to the inductance of the unit. The actual amount of the inductance will depend on the physical size of the capacitor and the method used to make contact to the stator and rotor plates. This inductance may be cut to a minimum value by using as small a capacitor as is practical, by using insulated tie rods to prevent the formation of closed inductive loops in the frame of the unit, and by making connections to the centers of the plate assemblies rather than to the ends as is commonly done. A large transmitting capacitor may have an inherent inductance as large as 0.1 microhenry, making the capacitor susceptible to parasitic resonances in the 50- to 150-MHz range of frequencies.

The question of optimum C/L ratio and capacitor plate spacing is covered in Chapter Eleven. For all-band operation of a high-power stage, it is recommended that a capacitor just large enough for 40-meter operation be chosen. (This will have sufficient capacitance for operation on all higher-frequency bands.) Then use fixed padding capacitors for operation on 80 meters. Such padding capacitors are available in air, ceramic, and vacuum types.

Specially designed variable capacitors are recommended for uhf work; ordinary capacitors often have "loops" in the metal frame which may resonate near the operating frequency.

17-3 Wire and Inductors

Any length of wire, no matter how short, has a certain value of inductance. This property is of great help in making coils and inductors, but may be of great hindrance when it is not taken into account in circuit design and construction. Connecting circuit elements (themselves having residual inductance) together with a conductor possessing additional inductance can often lead to puzzling difficulties. A piece of No. 10 copper wire ten inches long (a not uncommon length for a plate lead in a transmitter) can have a self-inductance of 0.15 microhenrys. This inductance and that of the plate tuning capacitor together with the plate-to-ground capacity of the vacuum tube can form a resonant circuit which may lead to parasitic oscillations in the vhf regions. To keep the self-inductance at a minimum, all r-f carrying leads should be as short as possible and should be made out of as heavy material as possible.

At the higher frequencies, solid enameled copper wire is most efficient for r-f leads. Tinned or stranded wire will show greater losses at these frequencies. Tank-coil and tank-capacitor leads should be of heavier wire than other r-f leads.

The best type of flexible lead from the envelope of a tube to a terminal is thin copper strip, cut from thin sheet copper. Heavy, rigid leads to these terminals may crack the envelope glass when a tube heats or cools.

Wires carrying only audio frequencies or direct current should be chosen with the voltage and current in mind. Some of the low-filament-voltage transmitting tubes draw heavy current, and heavy wire must be used to avoid voltage drop. The voltage is low, and hence not much insulation is required. Filament and heater leads are usually twisted together. An initial check should be made on the filament voltage of all tubes of 25 watts or more plate dissipation rating. This voltage should be measured right at the tube sockets. If it is low, the filament-transformer voltage should be raised. If this is impossible, heavier or parallel wires should be used for filament leads, cutting down their length if possible.

Coaxial cable may be used for high-voltage leads when it is desirable to shield them from r-f fields. RG-8/U cable may be used at d-c potentials up to 8000 volts, and the lighter RG-58/U may be used to potentials of 3000 volts. Spark plug-type high-tension wire may be used for unshielded leads, and will withstand 10,000 volts.

If this cable is used, the high-voltage leads may be cabled with filament and other low-voltage leads. For high-voltage leads in low-power exciters, where the plate voltage is not over 450 volts, ordinary radio hookup wire of good quality will serve the purpose.

No r-f leads should be cabled; in fact it is better to use enameled or bare copper wire for r-f leads and rely on spacing for insulation. All r-f joints should be soldered, and the joint should be a good mechanical junction before solder is applied.

The efficiency and Q of air coils commonly used in amateur equipment is a factor of the shape of the coil, the proximity of the coil to other objects (including the coil form), and the material from which the coil is made. Dielectric losses in so-called "air-wound" coils are low and the Q of such coils runs in the neighborhood of 300 to 500 at medium frequencies. Unfortunately, most of the transmitting-type plug-in coils on the market designed for link coupling have far too small a pickup link for proper operation at 3.5 and 7 MHz. The coefficient of coupling of these coils is about 0.5, and additional means must be employed to provide satisfactory coupling at these low frequencies. Additional inductance in series with the pickup link, the whole being reso-

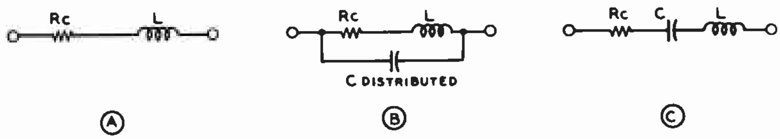


Figure 8

ELECTRICAL EQUIVALENT OF R-F CHOKE AT VARIOUS FREQUENCIES

nated to the operating frequency, will often permit satisfactory coupling.

Coil Placement For best Q a coil should be in the form of a solenoid with length from one to two times the diameter. For minimum interstage coupling, coils should be made as small physically as is practicable. The coils should then be placed so that adjoining coils are oriented for minimum mutual coupling. To determine if this condition exists, apply the following test: the axis of one of the two coils must lie

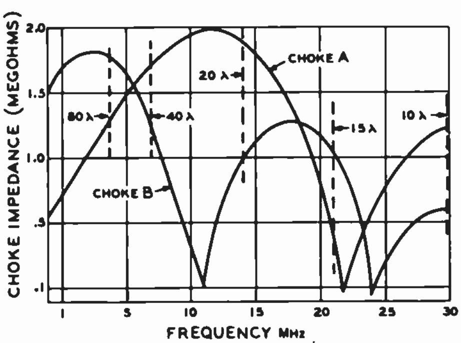


Figure 9

FREQUENCY-IMPEDANCE CHARACTERISTICS FOR TYPICAL PIE-WOUND R-F CHOKES

in the plane formed by the center turn of the other coil. If this condition is not met, there will be appreciable coupling unless the unshielded coils are very small in diameter or are spaced a considerable distance from each other.

Insulation On frequencies above 7 MHz, ceramic, polystyrene, or *Mycalex* insulation is to be recommended. Cold flow must be considered when using polystyrene. Micarta has low losses on the lower frequencies but should never be used in the field of high-frequency tank circuits.

Lucite (or *Plexiglas*), which is available in rods, sheets, or tubing, is satisfactory for use at all radio frequencies *where the r-f voltages are not especially high.* It is very easy to work with ordinary tools and is not expensive. The loss factor depends to a considerable extent on the amount and kind of plasticizer used.

The most important thing to keep in mind regarding insulation is that the best insulation is air. If it is necessary to reinforce air-wound coils to keep turns from vibrating or touching, use strips of *Lucite* or polystyrene cemented in place with epoxy.

Radio-Frequency Chokes *R-f chokes* may be considered to be special inductances designed to have a high value of impedance over a large range of frequencies. A practical r-f choke has inductance, distributed capacitance, and resistance. At low frequencies, the distributed capacitance has little effect and the electrical equivalent circuit of the r-f choke is as shown in figure 8A. As the operating frequency of the choke is raised the effect of the distributed capacitance becomes more evident until at some particular frequency the distributed capacitance resonates with the inductance of the choke and a parallel-resonant circuit is formed. This point is shown in figure 8B. As the frequency of operating is further increased the over-all reactance of the choke becomes capacitive, and finally a point of series resonance is reached (figure 8C). This cycle repeats itself as the operating frequency is raised above the series-resonant point, the impedance of the choke rapidly becoming lower on each successive cycle. A chart of this action is shown in figure 9. It can be seen that as the r-f choke approaches and leaves a condition of series resonance, the performance of the choke is seriously impaired. The condition

TABLE 1 AIRWOUND INDUCTORS

AIRWOUND INDUCTORS

COIL DIA. *INCHES*	TURNS PER INCH	B & W	AIR DUX	INDUCTANCE μH
1/2	4	3001	404T	0.16
	6	—	406T	0.40
	8	3002	408T	0.72
	10	—	410T	1.12
	16	3003	416T	2.90
	32	3004	432T	12.0
5/8	4	3005	504T	0.28
	6	—	506T	0.62
	8	3006	508T	1.1
	10	—	510T	1.7
	16	3007	516T	4.4
	32	3008	532T	18.0
3/4	4	3009	604T	0.39
	6	—	606T	0.87
	8	3010	608T	1.57
	10	—	610T	2.45
	16	3011	616T	6.40
	32	3012	632T	26.0
1	4	3013	804T	1.0
	6	—	806T	2.3
	8	3014	808T	4.2
	10	—	810T	6.6
	16	3015	816T	16.8
	32	3016	832T	68.0

COIL DIA. *INCHES*	TURNS PER INCH	B & W	AIR DUX	INDUCTANCE μH
1 1/4	4	—	1004	2.75
	6	—	1006	6.30
	8	—	1008	11.2
	10	—	1010	17.5
	16	—	1016	42.5
1 1/2	4	—	1204	3.9
	6	—	1206	8.8
	8	—	1208	15.8
	10	—	1210	24.5
	16	—	1216	63.0
1 3/4	4	—	1404	5.2
	6	—	1406	11.8
	8	—	1408	21.0
	10	—	1410	33.0
	16	—	1416	85.0
2	4	—	1604	6.6
	6	—	1606	15.0
	8	3900	1608	26.5
	10	3907-1	1610	42.0
	16	—	1616	108.0
2 1/2	4	—	2004	10.1
	6	3905-1	2006	23.0
	8	3906-1	2008	41.0
	10	—	2010	108.0
3	4	—	2404	14.0
	6	—	2406	31.5
	8	—	2408	58.0
	10	—	2410	89.0

NOTE:
COIL INDUCTANCE APPROXIMATELY PROPORTIONAL TO LENGTH. I.E., FOR 1/2 INDUCTANCE VALUE, TRIM COIL TO 1/2 LENGTH.

of series resonance may easily be found by shorting the terminals of the r-f choke in question with a piece of wire and exploring the windings of the choke with a grid-dip oscillator. Most commercial transmitting-type chokes have series resonances in the vicinity of 11 or 24 MHz.

High Power R-F Chokes By observing the series-resonant frequency of the choke, a homemade, high power r-f choke may be made very inexpensively. Representative designs are listed in Table 2. The first choke covers the 7.0- to 30-MHz frequency region with the first series resonance at 43 MHz. The choke is rated for an operating potential of 5 kV and a maximum d-c current of 2 amperes. The second choke covers the 3.5- to 30-MHz region, with the exception of the series-resonance frequency near 25 MHz. The choke is rated for 3 kV at 1 ampere. The third choke is designed for the 21- to 54-MHz region with a series resonance near 130 MHz. It has the same voltage and current ratings as the second choke.

17-4 Grounds

At frequencies of 30 MHz and below, a chassis may be considered as a fixed ground reference, since its dimensions are only a fraction of a wavelength. As the frequency is increased above 30 MHz, the chassis must be considered as a conducting sheet on which there are points of maximum current and potential. However, for the lower amateur frequencies, an object may be assumed to be at ground potential when it is affixed to the chassis.

In transmitter stages, two important current loops exist. One loop consists of the input circuit and chassis return, and the other loop consists of the output circuit and chassis return. These two loops are shown in figure 10A. It can be seen that the chassis forms a return for both the input and output circuits, and that *ground currents* flow in the chassis towards the cathode circuit of the stage. For some years the theory has been to separate these ground currents from the chassis by returning all ground leads to one

Table 2. H-F Radio-Frequency Chokes for Power Amplifiers

4000-Watt Peak Rating	
7-30 MHz:	90 turns #18 Formex, close-wound, about 4⅛" long on ¾" diam. × 6½" long teflon form. Series resonant at 43 MHz (32μH).
14-54 MHz:	43 turns #16 Formex space-wound wire diameter, about 4⅛" long on ¾" diam. × 6½" long Teflon form. Series resonant at 96 MHz (15μH) It is suggested that the form be grooved on a lathe for ease in winding.
2000-Watt PEP Rating	
3.5-30 MHz:	110 turns #26e., space-wound wire diameter, about 4" long on 1" diam. × 6" long ceramic form. Series resonant at 25 MHz. (78μH).
21-54 MHz:	48 turns #26e., space-wound wire diameter, about 1½" long on ½" × 3" long ceramic form. Or Air-Dux 432-T (B & W 3004) on wood form. Series resonant near 130 MHz. (75μH.)

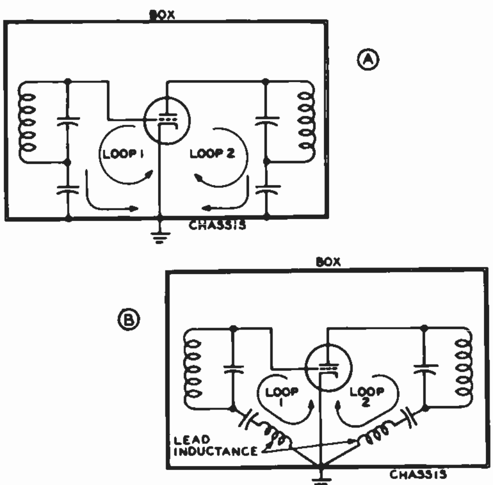


Figure 10

GROUND LOOPS IN AMPLIFIER STAGES

A. Using chassis return
B. Common ground point

point, usually the cathode of the tube for the stage in question. This is well and good if the ground leads are of minute length and do not introduce cross couplings between the leads. Such a technique is illustrated in figure 10B, wherein all stage components are grounded to the cathode pin of the stage socket. However, in transmitter construction the physical size of the components prevent such close grouping. It is necessary to spread the components of such a stage over a fairly large area. In this case it is best to ground items directly to the chassis at the nearest possible point, with short, direct grounding leads. The ground currents will flow from these points through the low inductance chassis to the cathode return of the stage. Components grounded on the top of the chassis have their ground currents flow through holes to the cathode circuit which is usually located on the bottom of the chassis, since such currents travel on the surface of the chassis. The usual "top to bottom" ground path is through the hole cut in the chassis for the tube socket. When the gain per stage is relatively low, or there are only a small number of stages on a chassis this universal grounding system is ideal. It is only in high gain stages (i-f strips) where the "gain per inch" is very high that circulating ground currents will cause operational instability.

Intercoupling of Ground Currents It is important to prevent intercoupling of various different ground currents when the chassis is used as a common ground return. To keep this intercoupling at a minimum, the stage should be completely shielded. This will prevent external fields from generating spurious ground currents, and prevent the ground currents of the stage from upsetting the action of nearby stages. Since the ground currents travel on the surface of the metal, the stage should be enclosed in an electrically tight box. When this is done, all ground currents generated inside the box will remain in the box. The only possible means of escape for fundamental and harmonic currents are imperfections in this electrically tight box. Whenever we bring a wire lead into the box, make a ventilation hole, or bring a control shaft through the box we create an imperfection. It is important that the effect of these imperfections be reduced to a minimum.

17-5 Holes, Leads, and Shafts

Large size holes for ventilation may be put in an electrically tight box provided they are properly screened. Perforated metal stock having many small, closely spaced holes is the best screening material. Copper wire screen may be used provided the screen wires are bonded together every few inches. As the wire corrodes, an insulating film prevents contact between the individual wires, and the attenuation of the screening suffers. The screening material should be carefully soldered to the box, or bolted with a spacing of not less than two inches between bolts. Mating surfaces of the box and the screening should be clean.

A screened ventilation opening should be roughly three times the size of an equivalent unscreened opening, since the screening rep-

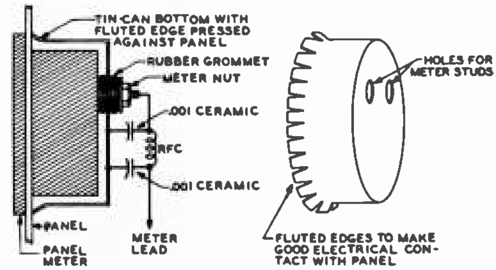


Figure 11

SIMPLE METER SHIELD

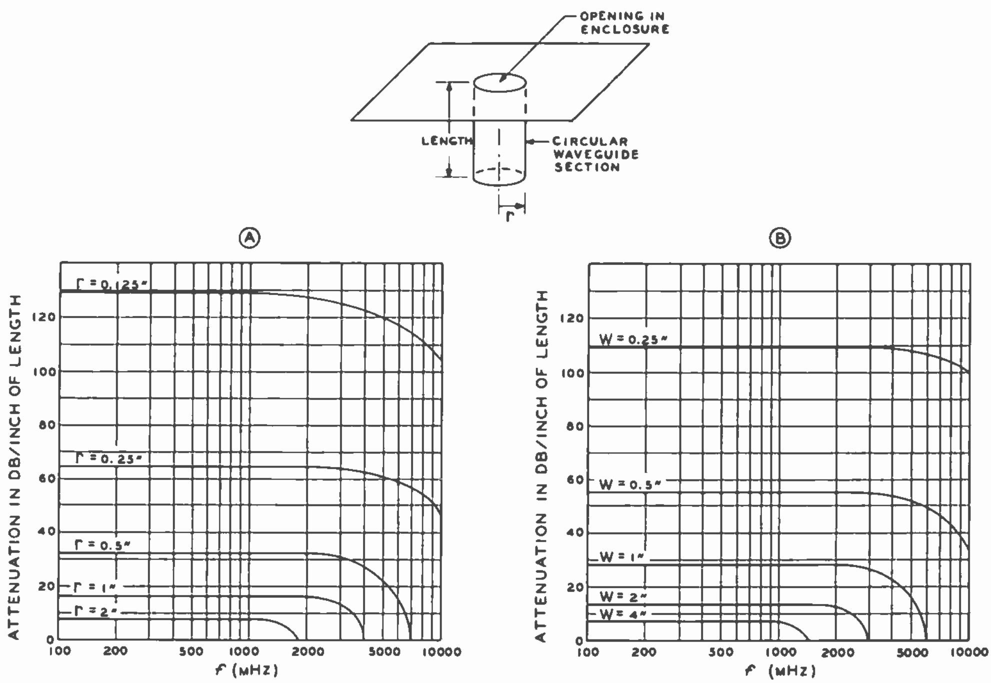


Figure 12

WAVEGUIDE-BEYOND-CUTOFF INCLOSURE OPENINGS

Waveguide section at inclosure opening can provide improved shielding efficiency. Air passes through the waveguide but r-f is attenuated to a greater degree than a simple opening can provide. Chart (A) provides attenuation in decibels/inch for circular waveguide. Chart (B) provides attenuation for rectangular waveguide for TE_{11} mode. All curves continue horizontally down to 10 MHz.

resents about a 70 percent coverage of the area. Careful attention must be paid to equipment heating when an electrically tight box is used.

Commercially available panels having half-inch ventilating holes may be used as part of the box. These holes have much less attenuation than does screening, but will perform in a satisfactory manner in all but the areas of weakest TV reception. If it is desired to reduce leakage from these panels to a minimum, the back of the grill must be covered with screening tightly bonded to the panel.

Doors may be placed in electrically tight boxes provided there is no r-f leakage around the seams of the door. Electronic weather-stripping or metal "finger stock" may be used to seal these doors. A long, narrow slot in a closed box has the tendency to act as a slot antenna and harmonic energy may pass more readily through such an opening than it would through a much larger circular hole.

Variable-capacitor or switch shafts may act as antennas, picking up currents inside the box and re-radiating them outside of the box. It is necessary either to ground the shaft securely as it leaves the box, or else to make the shaft of some insulating material.

A two- or three-inch panel meter causes a large leakage hole if it is mounted in the wall of an electrically tight box. To minimize leakage, the meter leads should be bypassed and shielded. The meter should be encased in a metal shield that makes contact to the box entirely around the meter. The connecting studs of the meter may project through the back of the metal shield. Such a shield may be made out of the end of a tin or aluminum can of correct diameter, cut to fit the depth of the meter. This complete shield assembly is shown in figure 11.

Openings for shafts, meters, and ventilation are sources of r-f leakage, and this spurious radiation may be reduced by designing the aperture through which leakage occurs as a waveguide-type attenuator.

A cutoff frequency for any waveguide is the lowest frequency at which propagation occurs without attenuation. Below cutoff, attenuation is a function of guide length and frequency. When an aperture is designed as a *waveguide below cutoff*, shielding efficiencies of a high order are achieved.

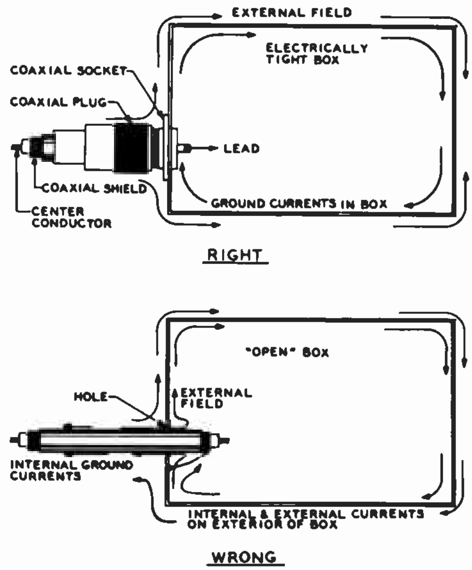


Figure 13

Use of coaxial connectors on electrically tight box prevents escape of ground currents from interior of box. At the same time external fields are not conducted into the interior of the box.

Figure 12A shows a set of design curves for circular waveguides ranging from 0.125" to 2" in radius and figure 12B shows curves for rectangular guides up to 4" in width. When the diameter or width of the opening is known, select the maximum frequency at which r-f suppression is desired. Select the appropriate curve from either chart and read attenuation in decibels per inch of length. Making the length of the waveguide three times the diameter for 100 db of attenuation and 80 db with rectangular guides is a useful design shortcut.

As an example a 1" diameter hole is required in an enclosure and 100-db transmission attenuation through the hole is desired at 100 MHz. From figure 12A attenuation is 32-db per inch at 100 MHz for radius = ½". The required length is 100/32 3.13 inches.

Pass-Through Leads Careful attention should be paid to leads entering and leaving the electrically tight box. Harmonic currents generated inside the box can easily flow out of the box on power or control leads, or even on the outer shields of coaxially shielded wires. Figure 13 illustrates the correct method of bringing shielded cables into a box where it is desired to preserve the continuity of the shielding.

Unshielded leads entering the box must be carefully filtered to prevent fundamental and harmonic energy from escaping down the lead. Combinations of r-f chokes and low-inductance bypass capacitors should be used in power leads. If the current in the lead is high, the chokes must be wound of large-gauge wire. Composition resistors may be substituted for the r-f chokes in high-impedance circuits. Bulkhead or feedthrough type capacitors are preferable when passing a lead through a shield partition. A summary of lead leakage with various filter arrangements is shown in figure 14.

Internal Leads Leads that connect two points within an electrically tight box may pick up fundamental and harmonic currents if they are located in a strong field of flux. Any lead forming a closed loop with itself will pick up such currents, as shown in figure 15. This effect is enhanced if the lead happens to be self-resonant at the frequency of the exciting energy. The solution for all of this is to bypass all internal power leads and control leads at each end, and to shield these leads their entire length. All filament, bias, and meter leads should be so treated. This will make the job of filtering the leads as they leave the box much easier, since normally "cool" leads within the box will not have picked up spurious currents from nearby "hot" leads.

Chassis Material From a point of view of electrical properties, aluminum is a poor chassis material. It is difficult to make a soldered joint to it, and all grounds must rely on a pressure joint. These pressure joints are prone to give trouble at a later date because of high resistivity caused by the formation of oxides from electrolytic action in the joint. However, the ease of working and forming the aluminum material far outweighs the electrical shortcomings, and aluminum chassis and shielding may be used with good results provided care is taken in making all grounding connections. Cadmium and zinc plated chassis are preferable from a corrosion standpoint, but are much

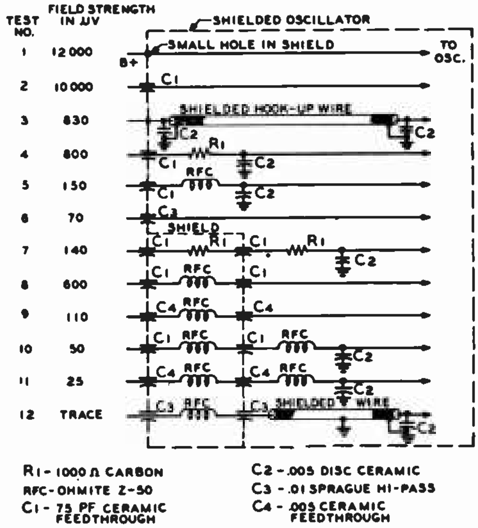


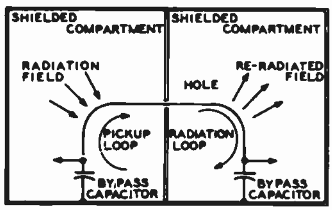
Figure 14

LEAD LEAKAGE WITH VARIOUS LEAD-FILTERING SYSTEMS

more difficult to handle in the home workshop.

17-6 Parasitic Resonances

Filament leads within vacuum tubes may resonate with the filament bypass capacitors at some particular frequency and cause instability in an amplifier stage. Large tubes of the 4-1000A and 3-1000Z type are prone to this spurious effect. In particular, an amplifier using .001-μfd filament bypass capacitors had a filament resonant loop that fell in the 7-MHz amateur band. When the amplifier was operated near this frequency, marked instability was noted, and the filaments of the tubes increased in brilliance when plate voltage was applied to the amplifier, indicating the presence of r.f. in the filament circuit. Changing the filament bypass capacitors to .01 μfd lowered the filament resonance frequency to 2.2 MHz and cured this effect. A 1-kV mica capacitor of .01 μfd used as a filament bypass capacitor on each filament leg seems to be satisfactory from both a resonant and a TVI point of view. Filament bypass capacitors smaller in value than .01 μfd should be used with caution.



Ⓐ

WRONG

ILLUSTRATION OF HOW A SUPPOSEDLY GROUNDED POWER LEAD CAN COUPLE ENERGY FROM ONE COMPARTMENT TO ANOTHER

Ⓑ

RIGHT

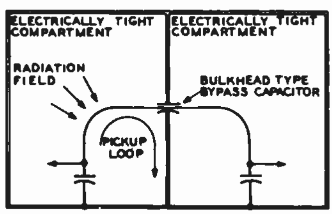
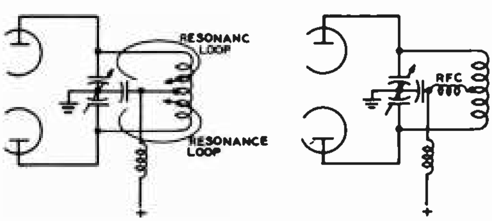


Figure 15

ILLUSTRATION OF LEAD ISOLATION BY PROPER USE OF BULKHEAD BYPASS CAPACITOR



WRONG RIGHT

Figure 16

DOUBLE RESONANCE EFFECTS IN PUSH-PULL TANK CIRCUIT MAY BE ELIMINATED BY THE INSERTION OF AN R-F CHOKE IN THE COIL CENTER TAP LEAD

Various parasitic resonances are also found in plate and grid tank circuits. Push-pull tank circuits are prone to double resonances, as shown in figure 16. The parasitic resonance circuit is usually several MHz higher than the actual resonant frequency of the full tank circuit. The cure for such a double resonance is the inclusion of an r-f choke in the center-tap lead to the split coil.

17-7 Parasitic Oscillation in R-F Amplifiers

Parasitics (as distinguished from *self-oscillation* on the normal tuned frequency of the amplifier) are undesirable oscillations either of very-high or very-low frequencies which may occur in radio-frequency amplifiers.

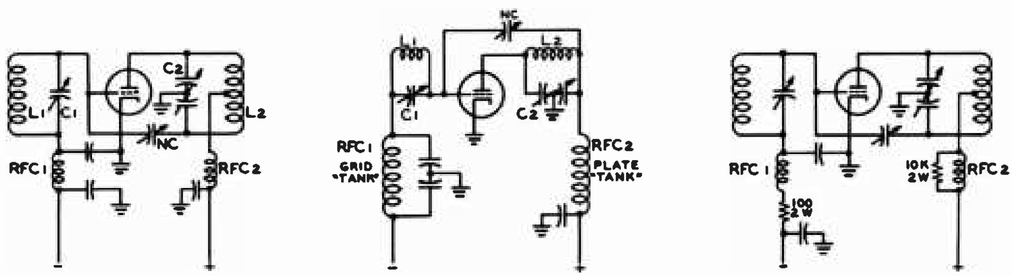
They may cause spurious signals (which are often rough in tone) other than normal harmonics, hash on each side of a modulated carrier, key clicks, voltage breakdown or flash-over, instability or inefficiency, and shortened life or failure of the tubes. They may be damped and stop by themselves after keying or modulation peaks, or they may be undamped and build up during ordinary unmodulated transmission, continuing if the excitation is removed. They may result from series- or parallel-resonant circuits of all types. Due to neutralizing lead length and the nature of most parasitic circuits, the amplifier usually is not neutralized for the parasitic frequency.

Sometimes the fact that the plate supply is keyed will obscure parasitic oscillations in a final amplifier stage that might be very severe if the plate voltage were left on and the excitation were keyed.

In some cases, a general coverage receiver will prove helpful in locating vhf spurious oscillations, but it may be necessary to check from several hundred MHz downward in frequency to the operating range. A normal harmonic is weaker than the fundamental but of good tone; a strong harmonic or a rough note at any frequency generally indicates a parasitic.

In general, the cure for parasitic oscillation is twofold: The oscillatory circuit is damped until sustained oscillation is impossible, or it is detuned until oscillation ceases. An examination of the various types of parasitic oscillations and of the parasitic oscillatory circuits will prove handy in applying the correct cure.

Low-Frequency Parasitic Oscillations One type of unwanted oscillation often occurs in shunt-fed circuits in which the grid and plate chokes resonate,



R-F CIRCUIT (A)   PARASITIC CIRCUIT FOR LOW-FREQ. OSCILLATION (B)   CURE (C)

Figure 17

LOW-FREQUENCY PARASITIC SUPPRESSION

A—Low-frequency parasitic circuit is formed by grid and plate r-f chokes and associated by-pass capacitors, as shown at B. Fundamental-frequency tank circuits have little effect on parasitic frequency. C—Parasitic circuits are "de-Q'ed" by addition of either series or parallel resistance until circuit will not sustain oscillation.

coupled through the tube's interelectrode capacitance. This also can happen with series feed. This oscillation is generally at a much lower frequency than the operating frequency and will cause additional carriers to appear, spaced from perhaps twenty to a few hundred kHz on either side of the main wave. Such a circuit is illustrated in figure 17. In this case, RFC_1 and RFC_2 form the grid and plate inductances of the parasitic oscillator. The neutralizing capacitor, no longer providing out-of-phase feedback to the grid circuit, actually enhances the low-frequency oscillation. Because of the low Q of the r-f chokes, they will usually run warm when this type of parasitic oscillation is present and may actually char and burn up. A neon bulb held near the oscillatory circuit will glow a bright yellow, the color appearing near the glass of the neon bulb and not between the electrodes.

One cure for this type of oscillation is to change the type of choke in either the plate or the grid circuit. This is a marginal cure, because the amplifier may again break into the same type of oscillation when the plate voltage is raised slightly. The best cure is to remove the grid r-f choke entirely and replace it with a wirewound resistor of sufficient wattage to carry the amplifier grid current. If the inclusion of such a resistor upsets the operating bias of the stage, an r-f choke may be used, with a 100-ohm 2-watt carbon resistor in series with the choke to lower the operating Q of the choke. If this expedient does not eliminate the condition, and the stage under investigation uses a beam-tetrode tube, negative resistance can exist in the screen circuit of such tubes. Try larger and smaller screen bypass capacitors to determine whether or not they have any effect. If the condition is coming from the screen circuit an audio choke with a resistor across it in series with the screen-feed lead will often eliminate the trouble.

Low-frequency parasitic oscillations can often take place in the audio system of an a-m transmitter, and their presence will not be known until the transmitter is checked on a receiver. It is easy to determine whether or not the oscillations are coming from the modulator simply by switching off the modulator tubes. If the oscillations are coming from the modulator, the stage in which they are being generated can be determined by removing tubes successively, starting with the first speech amplification stage, until the oscillation stops. When the stage has been found, remedial steps can be taken on that stage.

If the stage causing the oscillation is a low-level speech stage it is possible that the trouble is coming from r-f or power-supply feedback, or it may be coming about as a result of inductive coupling between two transformers. If the oscillation is taking place in a high-level audio stage, it is possible that inductive or capacitive coupling is taking place back to one of the low-level speech stages.

17-8 Elimination of VHF Parasitic Oscillations

Vhf parasitic oscillations are often difficult to locate and difficult to eliminate since their frequency often is only moderately above the desired frequency of operation. But it may be said that vhf parasitics always may be eliminated if the operating frequency is appreciably below the upper frequency limit for the tubes used in the stage. However, the elimination of a persistent parasitic oscillation on a frequency only moderately higher than the desired operating frequency will involve a sacrifice in either the power output or the power sensitivity of the stage, or in both.

Beam-tetrode stages, particularly those using 6146 or TV-style sweep tubes, will almost invariably have one or more vhf parasitic oscillations unless adequate precautions have been taken in advance. Many of the units described in the constructional section of this edition had parasitic oscillations when first constructed. But these oscillations were eliminated in each case; hence, the expedients used in these equipments should be studied. Vhf parasitics may be readily identified, as they cause a neon lamp to have a purple glow close to the electrodes when it is excited by the parasitic energy.

Parasitic Oscillations with Triodes In the case of triodes, vhf parasitic oscillations often come about as a result of inductance in the neutralizing leads. This is particularly true in the case of push-pull amplifiers. The cure for this effect will usually be found in reducing the length of the neutralizing leads and increasing their diameter. Both the reduction in length and increase in diameter will reduce the inductance of the leads and tend to raise the parasitic oscillation frequency until it is out of the range at which the tubes will oscillate. The use of straightforward circuit design with short leads will assist in forestalling this trouble at the outset.

Vhf parasitic oscillations may take place as a result of inadequate bypassing or long bypass leads in the filament, grid-return, and plate-return circuits. Such oscillations also can take place when long leads exist between the grid and the grid tuning capacitor or between the plate and the plate tuning capacitor. Sometimes parasitic oscillations can be eliminated by using iron or nichrome wire for the neutralizing lead. But in any event it will always be found best to make the neutralizing lead as short and of as heavy conductor as is practicable.

To increase losses at the parasitic frequency, the parasitic coil may be wound on 100-ohm 2-watt resistors. The "lossy" suppressor should be placed in the plate or grid lead of the tube close to the anode or grid connection, as shown in figure 18.

Parasitics with Beam Tetrodes Where beam-tetrode tubes are used in the stage which has been found to be generating the parasitic oscillation, all the foregoing suggestions apply in general. However, there are certain additional considerations involved in elimination of parasitics from beam-tetrode amplifier stages. These considerations involve the facts that a beam-tetrode amplifier stage has greater power sensitivity than an equivalent triode amplifier, such a stage has a certain amount of screen-lead inductance which may give rise to trouble, and such stages have a small amount of feedback capacitance.

Beam-tetrode stages often will require the inclusion of a neutralizing circuit to eliminate oscillation on the operating frequency. However, oscillation on the operating frequency is not normally called a parasitic oscillation, and different measures are required to eliminate the condition.

When a parasitic oscillation is found on a very high frequency, the interconnecting leads of the tube, the tuning capacitors and the bypass capacitors are involved. This type of oscillation generally does not occur when the amplifier is designed for vhf operation where the r-f circuits external to the tube have small tuning capacitors and inductors. Without tuning capacitors, the highest frequency of oscillation is then the fundamental frequency and no higher frequencies of resonance exist for the parasitic oscillation.

The vhf oscillation commonly occurs in h-f amplifiers, using the capacitors and associated grid and plate leads for the inductances of the tuned circuit. The frequency

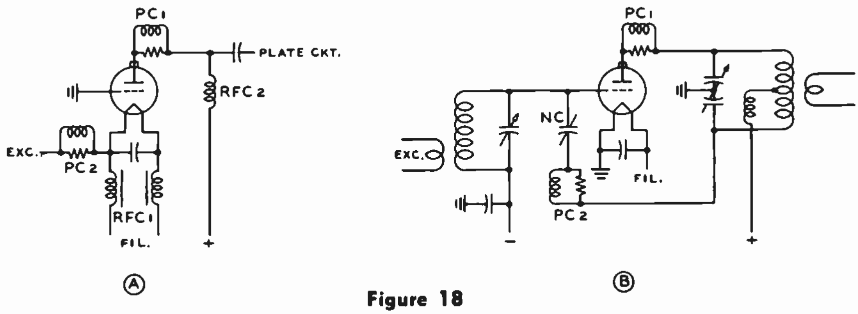


Figure 18

PARASITIC SUPPRESSION CIRCUITS

A—Plate parasitic suppressor is used in grounded-grid circuit. Filament suppressor may be added if secondary parasitic is present. B—Plate parasitic suppressor is used for grid-driven circuit, with second suppressor added in neutralizing circuit, if necessary.

of unwanted oscillation is generally well above the self-neutralizing frequency of the tube. If the frequency of the parasitic can be lowered to or below the self-neutralizing frequency, complete suppression of the parasitic will result. It is also possible to suppress the oscillation by loading the circuit so that the circuit is "lossy" at the parasitic frequency. This may be done by the use of a parasitic choke in the plate and/or grid lead of the stage in question. A parallel coil and resistor combination operates on the principle that the resistor loads the vhf circuit but is shunted by the coil for the lower fundamental frequency. The parasitic choke (figure 19) is usually made up of a noninductive resistor of about 25 to 100 ohms, shunted by three or four turns of wire, approximately one-half inch in diameter and frequently wound over the body of the resistor.

In the process of adjusting the resistor-coil combination, it may be found that the resistor runs too hot. The heat is usually caused by the dissipation of fundamental power in the resistor, which is an indication of too many turns in the suppressor coil. Just enough turns should be used to suppress the parasitic oscillation, and no more. Once the circuit is properly loaded and the parasitic suppressed, no parasitic power will be present and no power other than primary power will be lost in the resistor of the suppressor.

For medium power levels, a plate suppressor may be made of a 22-ohm, 2-watt Ohmite or Allen-Bradley composition resistor wound with 4 turns of No. 18 enameled wire. For kilowatt stages operating up to 30 MHz, a satisfactory plate suppressor may be made of three 220-ohm, 2-watt composition resistors in parallel, shunted by 3 or 4 turns of No. 14 enameled wire, ½-inch diameter and ½-inch long.

The parasitic suppressor for the plate circuit of a small tube such as the 5763, 2E26, 6146, 6LQ6, or similar type normally may consist of a 47-ohm composition resistor of 2-watt size with 4 turns of No. 18 enameled wire wound around the resistor. However, for operation above 30 MHz, special tailoring of the value of the resistor and the size of the coil wound around it will be required in order to attain satisfactory parasitic suppression without excessive power loss in the parasitic suppressor.

Tetrode Screening Isolation between the grid and plate circuits of a tetrode tube is not perfect. For maximum stability, it is recommended that the tetrode stage be neutralized. Neutralization is *absolutely necessary* unless the grid and plate circuits of the tetrode stage are each completely isolated from each other in electrically tight boxes. Even when this is done, the stage will show signs of regeneration when the plate and grid tank circuits are tuned to the same frequency. Neutralization will eliminate this regeneration. Any of the neutralization circuits described in the chapter *Generation of R-F Energy* may be used.

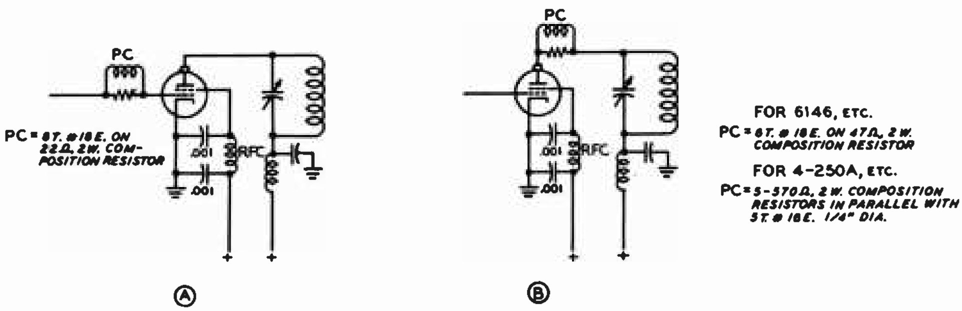


Figure 19

PLATE AND GRID PARASITIC SUPPRESSION IN TETRODE TUBES

R-C-type parasitic chokes are placed in grid (A) or plate (B) lead of tetrode and pentode tubes as shown above. Too few turns on the parasitic choke will not completely suppress the parasitic, whereas too many turns will permit the shunt resistor to absorb too much fundamental power. Five turns for the shunt coil will work well to 14 MHz. For 21 and 28 MHz, the shunt coil should be reduced to three turns.

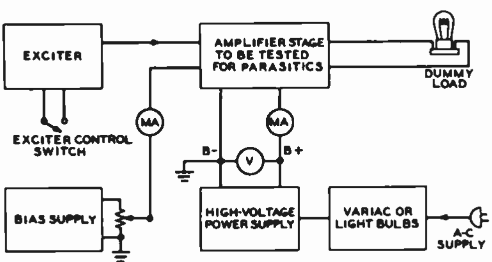


Figure 20

SUGGESTED TEST SETUP FOR PARASITIC TESTS

17-9 Checking for Parasitic Oscillations

It is an unusual transmitter which harbors no parasitic oscillations when first constructed and tested. Therefore it is always wise to follow a definite procedure in checking a new transmitter for parasitic oscillations.

Parasitic oscillations of all types are most easily found when the stage in question is running by itself, with full plate (and screen) voltage, sufficient protective bias to limit the plate current to a safe value, and no excitation. One stage should be tested at a time, and the complete transmitter should never be put on the air until all stages have been thoroughly checked for parasitics.

To protect tetrode tubes during tests for parasitics, the screen voltage should be applied through a series resistor which will limit the screen current to a safe value in case the plate voltage of the tetrode is suddenly removed when the screen supply is on. The correct procedure for parasitic testing is as follows (figure 20):

1. The stage should be coupled to a dummy load, and tuned up in correct operating shape. Sufficient protective bias should be applied to the tube at all times. For protection of the stage under test, a lamp bulb should be added in series with one leg of the primary circuit of the high-voltage power supply. As the plate-supply load increases during a period of parasitic oscillation, the voltage drop across the lamp increases, and the effective plate voltage drops. Bulbs of various sizes may be tried to adjust the voltage under testing conditions

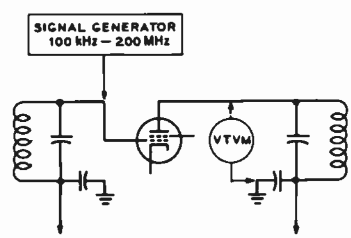


Figure 21

PARASITIC GAIN MEASUREMENT

Grid-dip oscillator and vacuum tube voltmeter may be used to measure parasitic stage gain over 100 kHz-200 MHz region.

to the correct amount. If a *Variac* or *Powerstat* is at hand, it may be used in place of the bulbs for smoother voltage control. Don't test for parasitics unless some type of voltage control is used on the high-voltage supply! When a stage breaks into parasitic oscillations, the plate current increases violently and some protection to the tube under test *must* be used.

2. The r-f excitation to the tube should now be removed. When this is done, the grid, screen, and plate currents of the tube should drop to zero. Grid and plate tuning capacitors should be tuned to minimum capacity. No change in resting grid, screen, or plate current should be observed. If a parasitic is present, grid current will flow, and there will be an abrupt increase in plate current. The size of the lamp bulb in series with the high-voltage supply may be varied until the stage can oscillate continuously, without exceeding the rated plate or screen dissipation of the tube.

3. The frequency of the parasitic may now be determined by means of an absorption wavemeter, or a neon bulb. Low-frequency oscillations will cause a neon bulb to glow yellow. High-frequency oscillations will cause the bulb to have a soft, violet glow.

4. When the stage can pass the above test with no signs of parasitics, the bias supply of the tube in question should be decreased until the tube is dissipating its full plate rating when full plate voltage is applied, with no r-f excitation. Excitation may now be applied and the stage loaded to full input into a dummy load. The signal should now be monitored in a nearby receiver which has the antenna terminals grounded or otherwise shorted out. A series of rapid dots should be sent, and the frequency spectrum for several MHz each side of the carrier frequency carefully searched. If any vestige of parasitic is left, it will show up as an occasional "pop" on a keyed dot. This "pop" may be enhanced by a slight detuning of the input or output circuit.

5. If such a parasitic shows up, it means that the stage is still not stable, and further measures must be applied to the circuit. Parasitic suppressors may be needed in both screen and grid leads of a tetrode, or perhaps in both grid and neutralizing leads of a triode stage. As a last resort, a 10,000-ohm 25-watt wirewound resistor may be shunted across the input circuit of a high powered stage. This strategy removed a keying "pop" that showed up in a commercial transmitter, operating at a plate voltage of 5000.

Test for Parasitic Tendency in Tetrode Amplifiers In most high-frequency transmitters there are a great many resonances in the tank circuit at frequencies other than the desired operating frequency. Most of these parasitic resonant circuits are not coupled to the tube and have no significant tendency to oscillate. A few, however, are coupled to the tube in some form of oscillatory circuit. If the regeneration is great enough, oscillation at the parasitic frequency results. Those spurious circuits existing just below oscillation must be found and suppressed to a safe level.

One test method is to feed a signal from a grid-dip oscillator into the grid of a stage and measure the resulting signal level in the plate circuit of the stage, as shown in figure 21. The test is made with all operating voltages applied to the tubes. Class-C stages should have bias reduced so a reasonable amount of static plate current flows. The grid-dip oscillator is tuned over the range of 100 kHz to 200 MHz, the relative level of the r-f voltmeter is watched, and the frequencies at which voltage peaks occur are noted. Each significant peak in voltage gain in the stage must be investigated. Circuit changes or suppression must then be added to reduce all peaks by 10 db or more in amplitude.

17-10 Forced Air Cooling

A large percentage of the primary power drain of a transmitter is converted to heat emitted by tubes and components. The resulting temperature rise must be held within reasonable limits to ensure satisfactory life for the equipment.

Forced-air-cooled systems may be used to remove excess heat. A typical system consists of an *air blower*, a *conduit* to guide the air to the tube or component, a *heat radiator* on the component, and an air *exhaust exit*. The resistance to the air passage through such a system is termed system *back pressure, pressure drop,* or *static pressure.* Air requirements are normally expressed as a pressure drop defined in *inches of water* (as measured by a manometer) with a corresponding volumetric air flow defined in *cubic feet per minute* (c.f.m.). A typical air-cooling system is shown in figure 22. Cooling requirements for most transmitting tubes are provided on the data sheet and air requirements and blower data for some popular tubes are given in figure 23.

Adequate cooling of the tube envelope and seals is one of the factors leading to long tube life. Deteriorating effects increase directly with the temperature of the tube envelope and seals. Even if no cooling air is specified by the technical data sheet for a particular tube, ample free space for circulation of air about the tube is required, or else air must be forced past the tube.

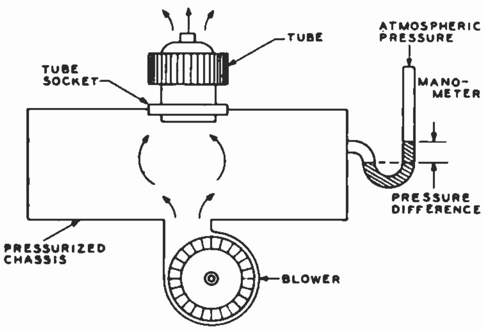


Figure 22

FORCED-AIR COOLING SYSTEM

Centrifugal blower pressurizes plenum chamber (air-tight chassis) and air is exhausted through the tube socket and anode cooler of vacuum tube. Pressure difference between plenum chamber and atmosphere is measured with manometer tube.

TUBE TYPE	AIR CFM	BACK PRESSURE	BLOWER SIZE	RPM	SOCKET CHIMNEY
3-400Z 3-500Z	13	0.20	3	1600	SK410 SK416 SK406
3-1000Z	25	0.64	3 3/4 2 1/2	3000 6000	SK510 SK516
4-1000A	25	0.64	3 3/4 2 1/2	3000 6000	SK510 SK506
4CX250B	6.4	1.12	2 1/2	6000	SK600 SK606
4CX1000A 4CX1500B	22	0.3	3	3100	SK800 SK806
5CX1500A	47	1.12	3	6000	SK840 SK806

Figure 23

COOLING REQUIREMENTS FOR TRANSMITTING TUBES

Air-system sockets and chimneys are required for high-power transmitting tubes. Complete air-cooling data for these types may be obtained from Application Engineering Department, Eimac Division of Varian, San Carlos, Calif. 94070.

As the frequency of operation of the tube is extended into the vhf region, additional cooling is usually required because of the larger r-f losses inherent in the tube structure.

Temperature-sensitive paint or crayons may be used to monitor the temperature of a tube under operating conditions. If the paint is applied to the tube envelope in a *very* thin coat, it will melt and virtually disappear at its critical temperature. After subsequent cooling, it will have a crystalline appearance indicating that the surface with which it is in contact has exceeded the critical temperature. Temperature-sensitive tapes and decals are also available to measure envelope temperature of transmitting tubes.

17-11 Conduction Cooling

The anode power dissipation density in a modern transmitting tube is extremely high and *conduction cooling* is often used to remove the heat from the tube structure.

A conduction cooling system comprises the heat source (the power tube), a *thermal link* to transfer the heat, and a *heat sink,* where the heat is removed from the system. The thermal link has the dual properties of a thermal conductor and an electrical insulator. *Beryllium oxide* (BeO) combines these propérties and is generally used for the thermal link. The BeO link may be brazed to the tube or be a detachable accessory (figure 24).

Most conduction-cooled tubes have an output capacitance which is higher than conventional air-cooled tubes due to the added capacitance between the tube anode and the heat sink, typically 6 to 10 pF. The capacitance is caused by the BeO dielectric. Below about 150 MHz, this added capacitance causes little difficulty since it can be included in the matching network design. Above 150 MHz, care in network design still permits successful operation up to the frequency limit of the tube, but attention must be

Figure 24

CONDUCTION-COOLED TUBE WITH INTEGRAL THERMAL LINK

Experimental type Y-406 tetrode makes use of beryllium oxide thermal link to transfer anode heat to an external heat sink. Link is pressed against the sink, with mating surfaces coated with silicone grease to improve interface thermal resistance. The heat sink transfers excess system heat to the surrounding atmosphere.

given to bandwidth and efficiency requirements and the physical length and configuration of the required resonating inductance as the added capacitance of the thermal link will limit the value of resonating inductance.

Normal use of electron tubes having Beryllium oxide is safe. However, BeO dust or fumes are highly toxic and breathing them can be injurious to health. Never perform work on any ceramic part of a power tube utilizing this material which could possibly generate dust or fumes. At the end of the useful life of the tube or heat sink, the BeO material should be returned prepaid to the manufacturer with written authorization for its disposal.

CHAPTER EIGHTEEN

Station Assembly and Transmitter Control

18-1 Station Layout

The amateur radio station has literally moved from the garage or home workshop into the living room during the past two decades. Gone are the black-crackle panels and the six-foot steel relay rack, and in their place are the new-design streamlined, miniaturized desk-top cabinets. Bandswitching linear amplifiers, solid-state power supplies and compact transceivers and exciters are the modern counterparts of the bulky plug-in coil class-C amplifiers, cumbersome modulators, and weighty power supplies that identified the amateur station of the late "fifties."

Station location and layout, accordingly, has undergone vast changes in the past few years and it is possible to place a high-power station within a desk, bookcase, or console assembly if space is at a premium.

Ideally, the best arrangement is for a complete room in the home or apartment to be devoted to the station, affording maximum comfort for operation, yet permitting operation and work to be done with minimum interference to family life. Regardless of the size and scope of the amateur station, however, the arrangement must be one that affords maximum efficiency, power capability, and safety to the operator. The care that has gone into the equipment and accessories must be carried over into the primary power system and control circuitry, and the control system for the rotary antenna, or antenna switching network.

Power Systems At a minimum, an amateur station will have a transceiver, or separate receiver and transmitter that will exhibit an average primary power drain of about 500 watts. In addition, an electric clock, desk light, and one or two auxiliary pieces of equipment will consume another 200 watts or so. Since the usual home outlet is rated to handle only about 600 watts maximum, the transmitting equipment, unless it is of medium power, should be powered from a separate primary outlet. In addition, there should be an outlet available for a soldering iron and one or two additional outlets provided for powering extra pieces of equipment being worked on (figure 1).

It thus becomes obvious that six to ten outlets connected to the 117-volt a-c line should be available at the operating position.

It may be practical to install an outlet strip at the rear of the operating desk so as to have the flexibility of moving the desk from one position to another. Alternatively, the outlet strip may be wall mounted behind the desk. It is inadvisable from the standpoint of safety to use a number of adapter plugs at one wall outlet to power the station equipment.

Line regulation is of importance in the amateur station installation. Poor regulation will cause the house lights to fluctuate with keying or modulation and in severe cases will cause an annoying shrinkage of the television image in a nearby receiver powered from the same line. It is good practice, therefore, to check the power capability of the house wiring before placing the full equipment load on the 117-volt primary service.

Power-Line "Standards" A confusion of power-line voltages and frequencies, as well as a multiplicity of plugs and connectors exists throughout the world. In the United States and Canada the nominal design center for consumer primary power is 117 volts, 60 Hz. Voltages between 110 and 125 are commonly encountered. In many overseas countries, 220 or 240 volts at 50 Hz may be found. In addition, unique combinations, such as 137 volts at 42 Hz, or 110 volts at 16⅔ Hz may exist as a result of special circumstances. Operation of equipment on one phase of a three-phase 240-volt power system calls for a design center of 208 volts.

Aside from the primary power complex-

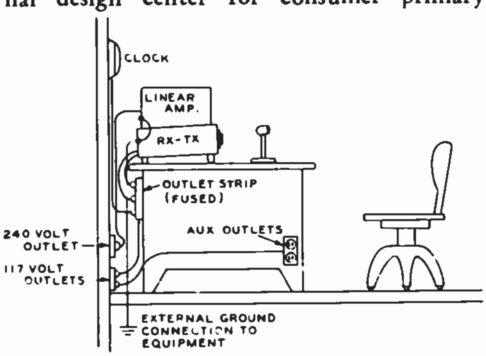


Figure 1

CONVENIENT POWER SYSTEM FOR AN AMATEUR STATION

Receiver or transceiver draws power from 117-volt a-c line through multiple-outlet strip attached to the rear of the operating desk. Additional outlets provide power for clock and other gear. A separate outlet is mounted near the front of the table for soldering iron and other temporary uses. A separate 240-volt power line is run from the main fuse box to power the linear amplifier. All equipment in the operating room is fused in series with fuses in main box. A separate 240-line is run in from the main fuse box to power the linear amplifier. All equipment in the operating room is fused in series with fuses in main box. 240-volt line is run in conduit. It is recommended that the entire house be wired in conduit, if possible, to reduce r-f pickup of the primary power lines as an aid in reduction of television interference and line noise.

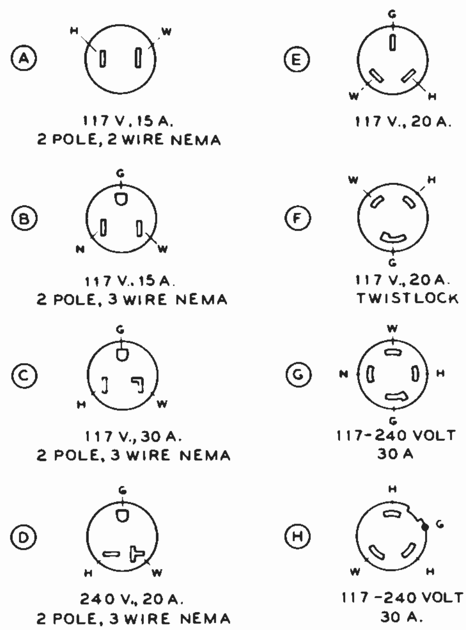


Figure 2

COMMON RECEPTACLE STANDARDS IN THE UNITED STATES

The front view of various common 117-volt and 240-volt standard receptacles is shown. 117-volt circuits have one wire (neutral) at about ground potential and the other wire (hot) above ground. The neutral wire (white W, with nickel screw terminal) is unfused while the hot wire (H, black, red, or blue with brass screw terminal) is fused. The switch should be in the hot line. The neutral is grounded at the distribution transformer and should not be grounded at any other point. Neutral is often referred to as system ground and is coded white. Equipment ground (G) is separately grounded at the electrical device and is coded green (circuits A, B, and C). 240-volt single phase receptacles are polarized so that 117-volt plugs cannot be used by error. Duplex (E) and Twistlock (F) are common industrial plugs, while the plugs of figures G and H are used with electric stoves, motors, air conditioners, etc.

ity, an endless number of plug and receptacle designs harass the experimenter. Recently, the *National Electrical Manufacturers Association* in the United States has announced standards covering general-purpose receptacles designed for the consumer wiring system, based on a design center of 117 volts, or the multiwire 240-volt, single-phase system used in many new homes.

A clear distinction is made in all specifications between *system ground* and *equipment ground*. The former, referred to as a grounded conductor, normally carries line current at ground potential. Terminals for system grounds are marked *W* and are color-coded *white*. Terminals for equipment grounds are marked *G* and are color-coded *green*. In this standard, the equipment ground carries current only during short circuit conditions.

A summary of some of the more common NEMA receptacle configurations, and other configurations still in popular use are shown in figure 2. A complete chart covering all standard NEMA plugs and receptacles may be obtained for twenty-five cents from: *The Secretary, NEMA Wiring Device Section, 155 East 44th Street, New York, N.Y., 10017.*

Checking an Outlet with a Heavy Load To make sure that an outlet will stand the full load of the entire transmitter, plug in an electric heater rated at about 50 percent greater wattage than the power you expect to draw from the line. If the line voltage does not drop more than 5 volts (assuming a 117-volt line) under load and the wiring does not overheat, the wiring is adequate to supply the transmitter. About 600 watts total drain is the maximum that should be drawn from a 117-volt *lighting* outlet or circuit. For greater power, a separate pair of heavy conductors should be run right from the meter box. For a 2-kW PEP transmitter the total drain is so great that a 240-volt "split" system ordinarily will be required. Most of the newer homes are wired with this system, as are homes utilizing electricity for cooking and heating.

With a three-wire system, be sure there is no fuse in the neutral wire at the fuse box. A neutral fuse is not required if both "hot" legs are fused, and, should a neutral fuse blow, there is a chance that damage to the radio transmitter will result.

Outlet Strips The *outlet strips* which have been suggested for installation in the baseboard or for use on the rear of a desk are obtainable from the large electrical-supply houses. These strips are quite convenient in that they are available in varying lengths with provision for inserting a-c line plugs throughout their length. In many cases it will be desirable to reduce the equipment cord lengths so that they will plug neatly into the outlet strip without an excess to dangle behind the desk.

Contactors and Relays The use of power-control relays and primary switches often will add considerably to the operating convenience of the station installation. The most practical arrangement usually is to have a main a-c line switch on the operating desk to energize or cut the power to the outlet strip on the rear of the operating desk. Through the use of such a switch it is not necessary to remember to switch off a large number of separate switches on each of the items of equipment on the operating desk.

While a single main switch is best for applying the a-c line power to the equipment, the changing over between transmit and receive can best be accomplished through the use of relays. Such a system usually involves three relays, or three groups of relays. The relays and their functions are: (1) power-control relay for the transmitter—applies the 117- or 240-volt line to the primary of the high-voltage transformer and energizes the exciter; (2) control relay for the receiver—makes the receiver inoperative by any one of a number of methods when closed, also may apply power to a keying or a modulation monitor; and (3) the antenna changeover relay—connects the antenna to the transmitter when the transmitter is energized and to the receiver when the transmitter is not operating.

Controlling Transmitter Power Output It is necessary, in order to comply with FCC regulations, that transmitter power output be limited to the minimum amount necessary to sustain communication. This requirement may be met in several ways. Many amateurs have a separate amplifier capable of relatively high power output for use when calling, or when interference is severe, using the exciter for local contacts, or when interference is absent. In most cases, the exciter acts as a driver for the amplifier stage when full power output is required.

One of the most common arrangements for obtaining two levels of power output involves the use of a plate transformer having a double primary for the high-voltage power supply. The majority of the high-power plate transformers of standard manufacture have just such a dual-primary arrangement. The two primaries are designed for use with either a 117-volt or 240-volt line. When such a transformer is to be operated from a 117-volt line, operation of both primaries in parallel will deliver full output from the plate supply. Then when the two primaries are connected in series and still operated from the 117-volt line the output voltage from the supply will be reduced approximately to one half. In the case of the normal amplifier, a reduction in plate voltage to one half will reduce the power input to the stage to one quarter.

If the transmitter is to be operated from a 240-volt line, the usual procedure is to operate the filaments from one side of the line, the low-voltage power supplies from the other side, and the primaries of the high-voltage transformer across the whole line for full power output. Then when reduced power output is required, the primary of the high-voltage plate transformer is operated from one side to center tap rather than across the whole line. This procedure places 117 volts across the 240-volt winding the same as in the case discussed in the previous paragraph. Figure 3 illustrates the two standard methods of power reduction with a plate transformer having a double primary; A shows the connections for use with a 117-volt line and B shows the arrangement for a 240-volt a-c power line to the transmitter.

When tuning the transmitter, the antenna coupling network, or the antenna system itself it is desirable to be able to reduce the power input to the final stage to a relatively low value, and it is further convenient to be able to vary the power input continuously from this relatively low input up to the full power capabilities of the transmitter. The use of a variable-ratio autotransformer in the circuit from the line to the primary of the plate transformer will allow a continuous variation in power input from zero to the full capability of the transmitter.

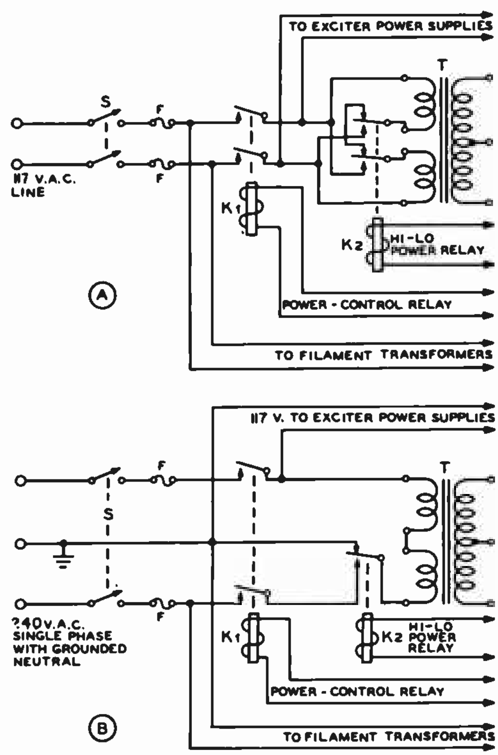


Figure 3

FULL-VOLTAGE/HALF-VOLTAGE SYSTEM OF POWER CONTROL

The circuit at A is for use with a 117-volt a-c line. Transformer T is of the standard type having two 117-volt primaries; these primaries are connected in series for half-voltage output when the power-control relay K_1 is energized but the hi-lo relay (K_2) is not operated. When both relays are energized the full output voltage is obtained. At B is a circuit for use with a standard 240-volt residence line with grounded neutral. The two relays control the output of the power supplies the same as at A.

Switching Between 1 kW d-c and 2 kW PEP A linear amplifier properly adjusted for 2-kW PEP input will often show a very low level of efficiency when the drive level and antenna loading are adjusted for 1-kW d-c input for c-w operation. Some amplifier designs solve this problem by reducing the d-c plate potential of the amplifier tube or tubes when switching from the SSB to the c-w mode. For example, a 2-kW PEP linear amplifier may be operating at a plate potential of 3 kV and a peak d-c plate current of 666 ma. Power input is 2 kW PEP and power output is, typically, 1.3 kW, PEP. Efficiency is about 65 percent. Switching to c-w, the operator drops excitation and readjusts antenna loading to provide a d-c input of 1 kW which corresponds to 3 kV at 333 ma. In most instances, amplifier efficiency will drop to about 30 percent, providing a power output of 300 watts, PEP. Unless the plate tank circuit has sufficient range to provide the proper plate load impedance for the 1-kW mode—and most simple pi- or pi-L networks do not—plate efficiency will drop badly.

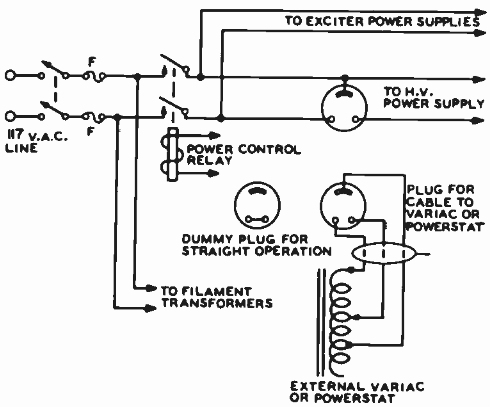


Figure 4

CIRCUIT WITH VARIABLE-RATIO AUTOTRANSFORMER

When the dummy plug is inserted into the receptacle on the equipment, closing of the power-control relay will apply full voltage to the primaries. With the cable from the Variac or Powerstat plugged into the socket the voltage output of the high-voltage power supply may be varied from zero to about 15 percent above normal.

If however, the plate potential under c-w operating conditions is dropped to about 65 percent of that employed in the SSB mode, plate efficiency will remain high in both conditions. For the above example, dropping the plate potential to about 2 kV and boosting the plate current to 500 ma will provide approximately the same degree of efficiency at the 1-kW d-c power level as will the 3-kV potential and 666 ma peak plate current at the 2-kW PEP power level. Many manufactured linear amplifiers accomplish the SSB to c-w switchover by dropping the plate potential on the amplifier tubes in the manner described. This is easily accomplished by the use of a tapped primary or secondary winding on the plate power transformer.

Variable-Ratio Autotransformers There are several types of variable-ratio autotransformers available on the market. Of these, the most common are the *Variac* manufactured by the *General Radio Company*, and the *Powerstat* manufactured by the *Superior Electric Company*. Both these types of variable-ratio transformers are excellently constructed and are available in a wide range of power capabilities. Each is capable of controlling the line voltage from zero to about 15 percent above the nominal line voltage. The maximum power-output capability of these units is available only at approximately the nominal line voltage, and must be reduced to a maximum current limitation when the output voltage is somewhat above or below the input line voltage. This, however, is not an important limitation for this type of application since the output voltage seldom will be raised above the line voltage, and when the output voltage is reduced below the line voltage the input to the transmitter is reduced accordingly.

One convenient arrangement for using a *Variac* or *Powerstat* in conjunction with the high-voltage transformer of a transmitter is illustrated in figure 4. In this circuit a heavy three-wire cable is run from a plug on the transmitter to the *Variac* or *Powerstat*. The *Variac* or *Powerstat* then is installed so that it is accessible from the operating desk so that the input power to the transmitter may be controlled during operation.

18-2 Transmitter Control Methods

When assured time delay of the proper interval and greater operating convenience are desired, a group of inexpensive a-c relays may be incorporated into the circuit to give

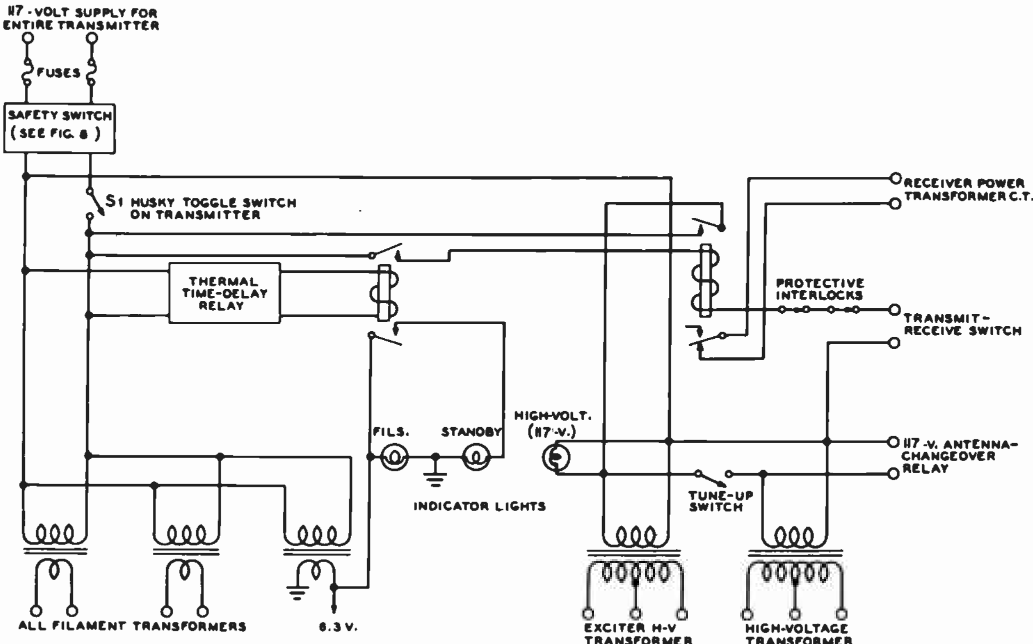


Figure 5

TRANSMITTER CONTROL CIRCUIT

Closing S_1 lights all filaments in the transmitter and starts the time-delay in its cycle. When the time-delay relay has operated, closing the transmit-receive switch at the operating position will apply plate power to the transmitter and disable the receiver. A tune-up switch has been provided so that the exciter stages may be tuned without plate voltage on the final amplifier.

a control circuit such as is shown in figure 5. This arrangement uses a 117-volt thermal (or motor-operated) time-delay relay and a dpdt 117-volt control relay. Note that the protective interlocks are connected in series with the coil of the relay which applies high voltage to the transmitter. A *tune-up switch* has been included so that the transmitter may be tuned up as far as the grid circuit of the final stage is concerned before application of high voltage to the final amplifier. Provisions for operating an antenna-changeover relay and for cutting the plate voltage to the receiver when the transmitter is operating have been included.

A circuit similar to that of figure 5 but incorporating push-button control of the transmitter is shown in figure 6. The circuit features a set of START-STOP and TRANSMIT-RECEIVE buttons at the transmitter and a separate set at the operating position. The control push buttons operate independently so that either set may be used to control the transmitter. It is only necessary to push the START button momentarily to light the transmitter filaments and start the time-delay relay in its cycle. When the standby light comes on it is only necessary to touch the TRANSMIT button to put the transmitter on the air and disable the receiver. Touching the RECEIVE button will turn off the transmitter and restore the receiver. After a period of operation it is only necessary to touch the STOP button at either the transmitter or the operating position to shut down the transmitter. This type of control arrangement is called an electrically locking push-to-transmit control system. Such systems are frequently used in industrial electronic control.

18-3 Safety Precautions

The best way for an operator to avoid serious accidents from the high-voltage supplies of a transmitter is for him to use his head, act only with deliberation, and not take unnecessary chances. However, no one

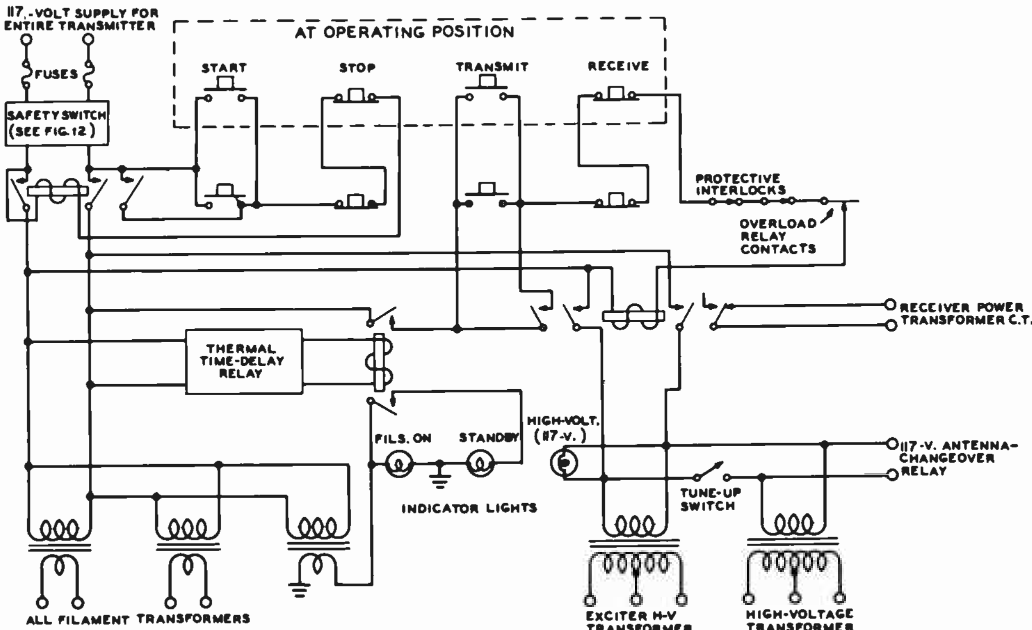


Figure 6

PUSH-BUTTON TRANSMITTER-CONTROL CIRCUIT

Pushing the START button either at the transmitter or at the operating position will light all filaments and start the time-delay relay in its cycle. When the cycle has been completed, a touch of the TRANSMIT button will put the transmitter on the air and disable the receiver. Pushing the RECEIVE button will disable the transmitter and restore the receiver. Pushing the STOP button will instantly drop the entire transmitter from the a-c line. If desired, a switch may be placed in series with the lead from the RECEIVE button to the protective interlocks; opening the switch will make it impossible for any person accidentally to put the transmitter on the air. Various other safety provisions, such as the protective-interlock arrangement described in the text have been incorporated. With the circuit arrangement shown for the overload-relay contacts, it is only necessary to use a simple normally closed d-c relay with a variable shunt across the coil of the relay. When the current through the coil becomes great enough to open the normally closed contacts the hold circuit on the plate-voltage relay will be broken and the plate voltage will be removed. If the overload is only momentary, such as a modulation peak or a tank flashover, merely pushing the TRANSMIT button will again put the transmitter on the air. This simple circuit provision eliminates the requirement for expensive overload relays of the mechanically latching type, but still gives excellent overload protection.

is infallible, and chances of an accident are greatly lessened if certain factors are taken into consideration in the design of a transmitter, in order to protect the operator in the event of a lapse of caution. If there are too many things one must watch for or keep in mind there is a good chance that sooner or later there will be a mishap; and it only takes *one*. When designing or constructing a transmitter, the following safety considerations should be given attention.

Grounds For the utmost in protection, everything of metal on the front panel of a transmitter capable of being touched by the operator should be at ground potential. This includes dial set screws, meter *zero-adjustment* screws, meter cases if of metal, meter jacks, *everything* of metal protruding through the front panel or capable of being touched or *nearly* touched by the operator. This applies whether or not the panel itself is of metal. Do not rely on the insulation of meter cases or tuning knobs for protection.

The B negative or chassis of all plate power supplies should be connected together, and to an external ground such as a waterpipe.

Exposed Wires and Components With metal-chassis construction it is possible to arrange things so as to incorporate a protective shielding housing which will not interfere with ventilation yet will prevent contact with all wires and components carrying high voltage d-c or a-c, in addition to offering shielding action.

If everything on the front panel is at ground potential (with respect to external ground) and all units are effectively housed with protective covers, then there is no danger except when the operator must reach into the interior part of the transmitter, as when neutralizing, adjusting coupling, or troubleshooting. The latter procedure can be made safe by making it possible for the operator to be *absolutely certain* that all voltages have been turned off and that they cannot be turned on either by short circuit or accident.

Combined Safety Signal and Switch The common method of using red pilot lights to show when a circuit is *on* is useless except from an ornamental standpoint. When the red pilot is not lit it *usually* means that the circuit is turned off, but it *can* mean that the circuit is on but the lamp is burned out or not making contact.

To enable you to work inside your transmitter with absolute assurance that it is impossible for you to obtain a shock except from possible undischarged filter capacitors (see following topic for elimination of this hazard), it is prudent to incorporate a device similar to that of figure 7. It is placed near the point where the main 117-volt leads enter the room (preferably near the door) and in such a position as to be inaccessible to small children. Notice that this switch breaks *both* leads; switches that open just one lead do not afford complete protection, as it is sometimes possible to complete a primary circuit through a short or accidental ground. Breaking just one side of the line may be all right for turning the transmitter on and off, but when you are going to place a hand inside the transmitter, *both* 117-volt leads should be broken.

When you are all through working your transmitter for the time being, simply throw the main switch to neutral.

For 100 percent protection, just obey the following rule: *never work on the transmitter or reach inside any protective cover except when the green pilots are glowing.* To avoid confusion, no other green pilots should be used on the transmitter; if you want an indicator jewel to show when the filaments are lighted, use amber instead of green.

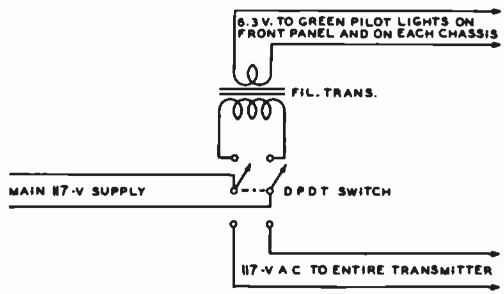


Figure 7

COMBINED MAIN SWITCH AND SAFETY SIGNAL

When shutting down the transmitter, throw the main switch to neutral. If work is to be done on the transmitter, throw the switch all the way to "pilot," thus turning on the green pilot lights on the panel and on each chassis, and ensuring that no voltage can exist on the primary of any transformer, even by virtue of a short or accidental ground.

Safety Bleeders Filter capacitors of good quality hold their charge for some time, and when the voltage is more than 100 volts it is just about as dangerous to get across an undischarged filter capacitor as it is to get across a high-voltage supply that is turned on. Most power supplies incorporate bleeders to improve regulation, but as these are generally wirewound resistors, and as wirewound resistors occasionally open up without apparent cause, it is desirable to incorporate an auxiliary safety bleeder across each heavy-duty bleeder. Carbon resistors will not stand much dissipation and sometimes change in value slightly with age. However, the chance of their opening up when run well within their dissipation rating is very small.

To make *sure* that all capacitors are bled, it is best to short each one with an insulated screwdriver. However, this is sometimes awkward and always inconvenient. One can be virtually sure by connecting auxiliary carbon bleeders across all wirewound bleeders used on supplies of 1000 volts or more. For every 500 volts, connect in series a 500,000-ohm 2-watt carbon resistor. The drain will be negligible (1 ma) and each resistor will have to dissipate only 0.5 watt. Under these conditions the resistors will last indefinitely with little chance of opening up.

Do *not* attempt to use fewer resistors by using a higher value for the resistors; not over 500 volts should appear across any single 1-watt resistor.

In the event that the regular bleeder opens up, it will take several seconds for the auxiliary bleeder to drain the capacitors down to a safe voltage, because of the very high resistance. Therefore it is best to allow 10 or 15 seconds to elapse after turning off the plate supply before attempting to work on the transmitter.

Protective Interlocks With the increasing tendency toward construction of transmitters in desk cabinets a transmitter becomes a particularly lethal device unless adequate safety provisions have been incorporated. Even with a combined safety signal and switch as shown in figure 7 it is still conceivable that some person unfamiliar with the transmitter could come in contact with high voltage. It is therefore recommended that the transmitter, whenever possible, be built into a complete metal housing or cabinet and that *all* doors or access covers be provided with protective interlocks (all interlocks must be connected in *series*) to remove the high voltage whenever these doors or covers are opened.

Relay Sequence It is important that the antenna changeover relay be activated before r-f power flows through the relay contacts. Certain VOX or key-operated sequences do not provide this protection. As a result, the contacts of the antenna relay may be damaged from making and breaking the r-f current, or eventual damage may occur to the transmitting equipment because of repeated operation without r-f load during the periods of time necessary for the antenna relay to close. The proper relay sequence can be achieved by actuating the antenna relay by the control system, then, in turn, actuating the transmitter by a separate set of control contacts on the antenna relay, as shown in figure 8A. In this manner, the antenna relay must be closed before r-f is applied to the contacts.

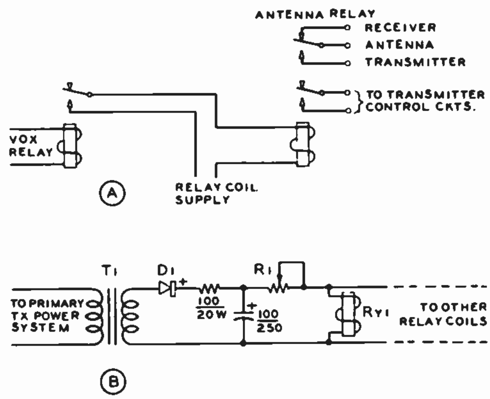


Figure 8

ANTENNA-RELAY CONTROL SYSTEM

A—Antenna relay should be actuated before r-f power flows through contacts. Extra set of contacts are used to control transmitter circuits after antenna relay closes. B—A-c relays may be operated from simple d-c power supply to reduce hum and chatter. Transformer T_1 may be a 1:1 isolation transformer of 50 watts capacity, with D_1 a 1 ampere, 600 volt p.i.v. diode. Series resistor R_1 is adjusted to provide proper relay action and may be of the order of 500 to 5000 ohms, 50 watts. Additional relay coils may be placed in parallel across coil Ry_1. Relay may be energized by applying primary power (with due regard to time-lag in filter system) or by completing secondary circuit between resistor R_1 and relay coil.

D-C Relay Operation Relays designed to operate from an a-c source are often troublesome sources of audible hum and chatter. Cleaning the relay striker and pole pieces will alleviate this annoyance somewhat, but operation of the relay from a d-c source will eliminate this difficulty. A-c relays may be operated without damage from a d-c source capable of supplying a d-c voltage equal to about 70 percent of the a-c design voltage. Thus an 85-volt d-c supply will be proper to operate 117-volt a-c relays. A suitable supply for such service is shown in figure 8B.

18-4 Transmitter Keying

The carrier from a c-w telegraph transmitter must be broken into dots and dashes for the transmission of code characters. The carrier signal is of constant amplitude while the key is closed, and is entirely removed when the key is open. When code characters are being transmitted, the carrier may be considered as being modulated by the keying. If the change from the no-output condition to full-output, or vice versa, occurs too rapidly, the rectangular pulses which form the keying characters contain high-frequency components which take up a wide frequency band as sidebands and are heard as clicks.

To be capable of transmitting code characters and at the same time not causing unnecessary interference to neighboring amateurs, the c-w transmitter *must* meet two important specifications.

1. It must have no parasitic oscillations either in the stage being keyed or in any succeeding stage.
2. It must have some device in the keying circuit capable of shaping the leading and trailing edge of the waveform.

Both these specifications must be met before the transmitter is capable of c-w operation. Merely turning a transmitter on and off by the haphazard insertion of a telegraph key in some power lead is an invitation to trouble.

Key-Click Elimination Key-click elimination is accomplished by preventing a too-rapid make and break of power to the antenna circuit, thus rounding off the keying characters so as to limit the sidebands to a value which does not cause interference to adjacent channels. Some circuits which eliminate key clicks introduce too much time lag and thereby add *tails* to the dots. These tails may cause the signals to be difficult to copy at high speeds.

Location of Keyed Stage Considerable thought should be given as to which stage in a transmitter is the proper one to key. If the transmitter is keyed in a stage close to the oscillator, the change in r-f loading of the oscillator may cause the oscillator to shift frequency with keying. This will cause the signal to have a distinct chirp. The chirp will be multiplied as many times as the frequency of the oscillator is multiplied. A chirpy oscillator that would be passable on 80 meters would be unusable on 28-MHz c.w.

Keying the oscillator itself is an excellent way to run into keying difficulties. If no key-click filter is used in the keying circuit, the transmitter will have bad key clicks. If a key-click filter is used, the slow rise and decay of oscillator voltage induced by the filter action will cause a keying chirp. This action is true of all oscillators, whether electron coupled or crystal controlled.

The more amplifier or doubler stages that follow the keyed stage, the more difficult it is to hold control of the shape of the keyed waveform. A heavily driven stage acts as a peak clipper, tending to square up a rounded keying impulse, and the cumulative effect of several such stages cascaded is sufficient to square up the keyed waveform to the point where bad clicks are reimposed on a clean signal.

A good rule of thumb is to never key back farther than one stage removed from the final amplifier stage, and never key closer than one stage removed from the frequency-controlling oscillator of the transmitter. Thus there will always be one isolating stage between the keyed stage and the oscillator, and one isolating stage between the keyed stage and the antenna. At this point the waveform of the keyed signal may be most easily controlled.

Keyer Circuit Requirements Many new design transmitters, and some of those of older design as well, use a medium-power beam tetrode tube or a zero-bias triode either as the output stage or as the driver for the output stage of a high power transmitter.

It is generally undesirable to key further down in the transmitter chain than the stage just ahead of the final amplifier. If a low-level stage, which is followed by a series of amplifiers, is keyed, serious transients may be generated in the output of the transmitter even though the keyed stage is being turned on and off very smoothly. This condition arises as a result of *pulse sharpening*, which has been discussed previously.

In a properly keyed transmitter, the output from the stage should be completely cut off when the key is up, and the time constant of the rise and decay of the keying wave should be easily controllable.

It should be possible to make the rise period and the decay period of the keying wave approximately equal. This type of keying envelope is the only one tolerable for commercial work, and is equally desirable for obtaining clean-cut and easily readable signals in amateur work.

Last, for the sake of simplicity and safety, it should be possible to ground the frame of the key, and yet the circuit should be such that placing the fingers across the key will not result in an electrical shock. In other words, the keying circuit should be inherently safe.

18-5 Cathode Keying

The lead from the cathode or center-tap connection of the filament of an r-f amplifier can be opened and closed for a keying circuit. Such a keying system opens the plate voltage circuit and at the same time opens the grid bias return lead. For this reason, the grid circuit is blocked at the same time the plate circuit is opened. This helps to reduce the backwave that might otherwise leak through the keyed stage.

The simplest cathode keying circuit is illustrated in figure 9, where a key-click filter is employed, and a hand key is used to break the circuit. This simple keying circuit is not

An electronic switch can take the place of the hand key. (figure 10). This will remove the danger of shock. At the same time, the opening and closing characteristics of the

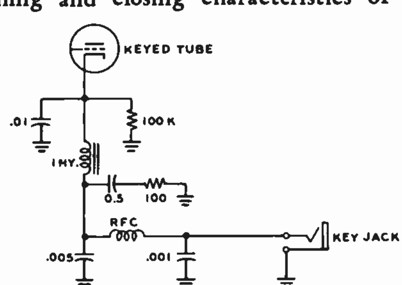


Figure 9

CENTER-TAP KEYING WITH KEY CLICK FILTER

The constants shown are suggested as preliminary values and may be changed to permit optimum amplifier keying under different operating conditions. Inductance of iron-core choke and series capacitor value are changed to alter keying characteristic. R-f choke and associated capacitors reduce transients caused by sparking at key contacts.

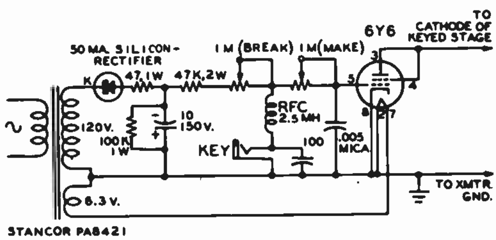


Figure 10

VACUUM-TUBE KEYER FOR CENTER-TAP KEYING

Single 6Y6 triode-connected can key 80 milliamperes. Two in parallel may be used for cathode currents under 160 ma. If softer keying is desired, increase grid capacitor to .01 µfd.

electronic switch may easily be altered to suit the particular need at hand. Such an electronic switch is called a *vacuum-tube keyer*. A low-resistance tube such as a triode-connected 6Y6 is used in this keyer. The 6Y6 acts as a very high resistance when sufficient blocking bias is applied to it, and as a very low resistance when the bias is removed. The desired amount of lag or *cushioning effect* can be obtained by employing suitable resistance and capacitance values in the grid of the keyer tube(s). Because very little spark is produced at the key, due to the small amount of power in the key circuit, sparking clicks are easily suppressed.

Because of the series resistance of the keyer tubes, the plate voltage at the keyed tube will be from 30 to 60 volts less than the power supply voltage. This voltage appears as cathode bias on the keyed tube, assuming the bias return is made to ground, and should be taken into consideration when providing bias.

18-6 Grid-Circuit Keying

Grid-circuit, or *blocked-grid keying* is another effective method of keying a c-w transmitter. A basic blocked-grid keying circuit is shown in figure 11. The time constant of the keying is determined by the RC circuit, which also forms part of the bias circuit of the tube. When the key is closed, operating bias is developed by the flow of grid current through R_1. When the key is open, sufficient fixed bias is applied to the tube to block it, preventing the stage from functioning. If an un-neutralized tetrode is keyed by this method, there is the possibility of a considerable backwave caused by r-f leakage through the grid-plate capacitance of the tube.

Certain high-μ triode tubes, such as the 811-A and the 3-500Z, automatically block themselves when the grid-return circuit is opened. It is merely necessary to insert a key and associated key-click filter in the grid-return lead of these tubes. No blocking bias supply is needed. This circuit is shown in figure 12.

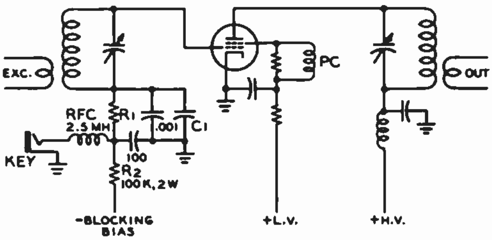


Figure 11

SIMPLE BLOCKED-GRID KEYING SYSTEM

The blocking bias must be sufficient to cut off plate current to the amplifier stage in the presence of the excitation voltage. R_1 is normal bias resistor for the tube. R_1 and C_1 should be adjusted for correct keying waveform.

A more elaborate blocked-grid keying system using a 6C4 and VR-150 is shown in figure 13. Two stages are keyed, preventing any backwave emission. The first keyed stage may be the oscillator, or a low-powered buffer. The last keyed stage may be the driver stage to the power amplifier, or the amplifier itself. Since the circuit is so proportioned that the lower-powered stage comes on *first* and goes off *last*, any keying chirp in the oscillator is not emitted on the air. Keying lag is applied to the high-powered keyed stage only.

18-7 Screen-Grid Keying

The screen circuit of a tetrode tube may be keyed for c-w operation. Unfortunately, when the screen grid of a tetrode tube is brought to zero potential, the tube still delivers considerable output. Thus it is necessary to place a negative blocking voltage on the screen grid to reduce the backwave through the tube. A suitable keyer circuit which will achieve this is shown in figure 14A.

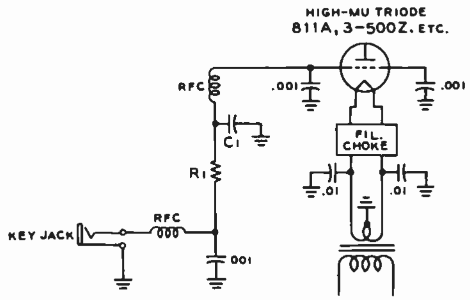


Figure 12

SELF-BLOCKING KEYING SYSTEM FOR HIGH-μ (ZERO BIAS) TRIODE

High-μ triodes such as the 811A, 572B (T-160L), 3-400Z, 3-500Z, etc. may be keyed by opening the d-c grid return circuit. Components R_1 and C_1 are adjusted for correct keying waveform. This circuit is not suited for keying a linear amplifier stage as the voltage drop across resistor R_1 provides additional grid bias to the amplifier tube.

A 6L6 is used as a combined clamper tube and keying tube. When the key is closed, the 6L6 tube has blocking bias applied to its control grid. This bias is obtained from the rectified grid bias of the keyed tube. Screen voltage is applied to the keyed stage through a screen dropping resistor and a VR-105 regulator tube. When the key is open, the 6L6 is no longer cutoff, and conducts heavily. The voltage drop across the dropping resistor caused by the heavy plate current of the 6L6 lowers the voltage on the VR-105 tube until it is extinguished, removing the screen voltage from the tetrode r-f tube. At the same time, rectified grid bias is applied to the screen of the tetrode through the 1 megohm resistor between screen and key. This voltage effectively cuts off the screen of the tetrode until the key is closed again. The RC circuit in the grid of the 6L6 tube determines the keying characteristic of the tetrode tube.

A simple keying relay for screen-grid keying is shown in figure 14B.

A more elaborate screen-grid keyer is shown in figure 15. This keyer is designed to block-grid–key the oscillator or a low-powered buffer stage, and to screen-key a medium-powered tetrode tube such as a

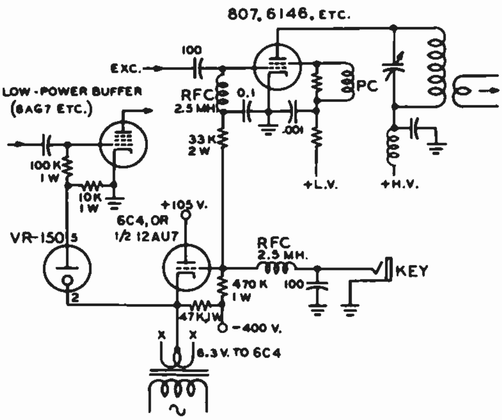


Figure 13

TWO-STAGE BLOCKED-GRID KEYER

A separate filament transformer must be used for the tube, as its filament is at a potential of −400 volts.

6146. A 6BL7 is used as the screen keyer, and a 12AU7 is used as a cathode follower and grid-block keyer. As in the figure 13 circuit, this keyer turns on the exciter a moment before the tetrode stage is turned on. The tetrode stage goes off an instant before the exciter does. Thus any keying chirp of the oscillator is effectively removed from the keyed signal.

18-8 Differential Keying Circuits

Excellent waveshaping may be obtained by a *differential keying system* whereby the master oscillator of the transmitter is turned on a moment before the rest of the stages are energized, and remains on a moment longer than the other stages. The chirp, or frequency shift, associated with abrupt switching of the oscillator is thus removed from the emitted signal. In addition, the differential keyer can apply waveshaping to the amplifier section of the transmitter, eliminating the click caused by rapid keying of the latter stages.

The ideal keying system would perform as illustrated in figure 16. When the key is closed, the oscillator reaches maximum output almost instantaneously. The following stages reach maximum output in a fashion determined by the waveshaping circuits of the keyer. When the key is released, the output of the amplifier stages starts to decay in a predetermined manner, followed shortly thereafter by cessation of the oscillator. The over-all result of these actions is to provide relatively soft "make" and "break" to the keyed signal, meanwhile preventing oscillator frequency shift during the keying sequence.

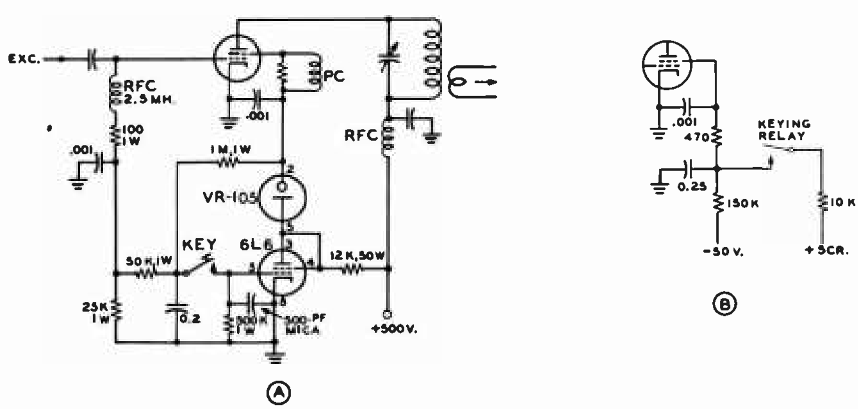


Figure 14

SCREEN-GRID KEYER CIRCUITS FOR TETRODE AND PENTODE TUBES

A—A 6L6 acts as a clamper and keyer tube. When key is open, 6L6 conducts and voltage drop across screen resistor extinguishes the voltage regulator tube, removing screen voltage from the tetrode tube. Bias is applied to screen to completely cut off the tetrode. B—Simple version of circuit A employs screen keying relay. When relay is open, negative bias voltage applied through 150K resistor effectively blocks tetrode-tube screen circuit.

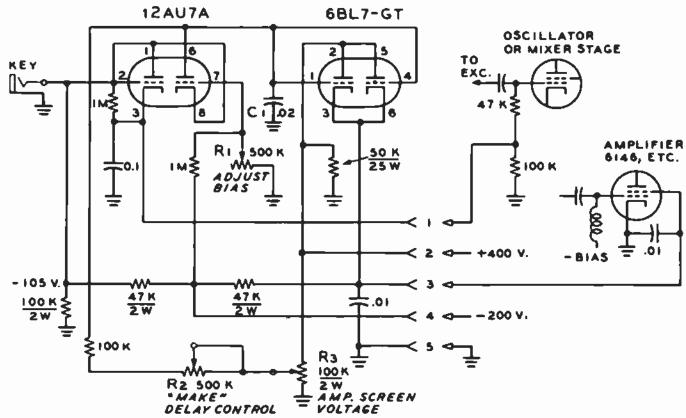


Figure 15

DIFFERENTIAL KEYING SYSTEM FOR SCREEN CIRCUIT

This keyer contains a 6BL7 series screen-voltage keyer tube and utilizes one-half of a 12AU7 as a control triode for the keyer tube. The second section of the 12AU7 is diode-connected to apply blocking bias to the 6BL7. With open key, pin 7 of the 12AU7 is adjusted to about −85 volts by means of potentiometer R_1, using vtvm. The 12AU7 and 6BL7 are now plugged in their sockets and, with the key still open, −105 to −110 volts should be observed at cathode pin 3 of the 12AU7. The screen voltage to the keyed amplifier stage (6146, etc.) should be about −50 volts. With the key closed, screen control potentiometer R_3 should be set to the proper screen voltage. Sharpness of keying on wave front is adjusted by the setting of potentiometer R_2. "Softness" of keying may be increased by raising value of capacitor C_1. The 6BL7 tube should be run from separate 6.3-volt filament supply with the filament transformer center tap connected to amplifier screen voltage lead to keep heater-cathode voltage of 6BL7 within its rating.

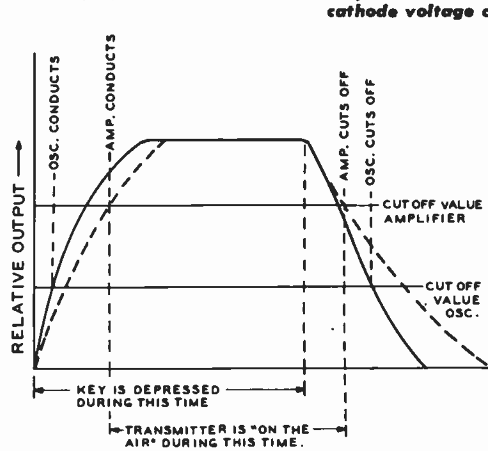


Figure 16

TIME SEQUENCE OF A DIFFERENTIAL KEYER

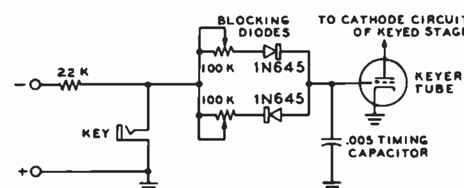


Figure 17

BLOCKING DIODES EMPLOYED TO VARY TIME CONSTANT OF "MAKE" AND "BREAK" CHARACTERISTICS OF VACUUM-TUBE KEYER

The rates of charge and decay in a typical RC keying circuit may be varied independently of each other by the blocking-diode system of figure 17. Each diode permits the charging current of the timing capacitor to flow through only one of the two variable potentiometers, thus permitting independent adjustment of the "make" and "break" characteristics of the keying system.

A practical differential keying system making use of this differential technique is shown in figure 18. A switching diode turns the oscillator on before the keying action starts, and holds it on until after the keying sequence is completed. Time constant of the keying cycle is determined by values of C and R. When the key is open, a cutoff bias of about −110 volts is applied to the screen-grid circuits of the keyed

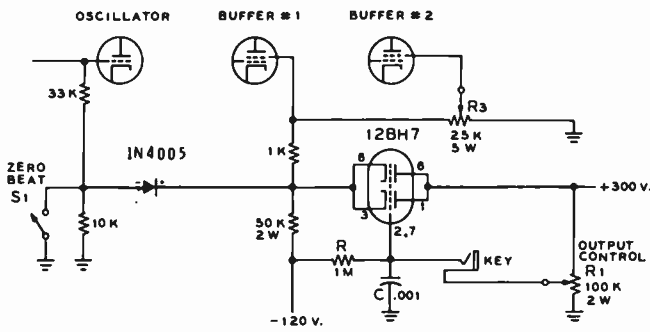


Figure 18

DIFFERENTIAL KEYING SYSTEM WITH OSCILLATOR-SWITCHING DIODE

stages. When the key is closed, the screen-grid voltage rises to the normal value at a rate determined by the time constant (RC). On opening the key again, the screen voltage returns to cutoff value at the predetermined rate.

The potentiometer (R_1) serves as an output control, varying the minimum internal resistance of the 12BH7 keyer tube, and is a useful device to limit power input during tuneup periods. Excitation to the final amplifier stage may be controlled by the screen potentiometer (R_3) in the second buffer stage. An external bias source of approximately -120 volts at 10 ma is required for operation of the keyer, in addition to the 300-volt screen supply.

Blocking voltage may be removed from the oscillator for zeroing purposes by closing switch S_1, rendering the diode switch inoperative.

A second popular keying system is shown in figure 19. Grid-block keying is used on tubes V_2 and V_3. A waveshaping filter consisting of R_2, R_3, and C_1 is used in the keying control circuit of V_2 and V_3. To avoid chirp when the oscillator (V_1) is keyed, the keyer tube V_4 allows the oscillator to start quickly—before V_2 and V_3 start conducting—and then continue operating until after V_2 and V_3 have stopped conducting. Potentiometer R_1 adjusts the "hold" time for vfo operation after the key is opened. This may be adjusted to cut off the vfo between marks of keyed characters, thus allowing rapid break-in operation.

Transmitter Keying The c-w transmitter keying comprises more than simply turning the carrier on and off by means of a telegraph key. The previous circuits are designed to provide clean crisp keying, when properly adjusted, without click or snap on the keyed waveform. The optimum keying characteristic is a highly subjective thing and "on-the-air" checks are questionable, since many amateurs hesitate to be truly critical of another amateur's signal unless it is causing objectionable interference.

Observation of the r-f waveform of the keyed signal on an oscilloscope can quickly show the operator the characteristic of his signal and comparison of the 'scope image with the signal heard in a local monitor will quickly provide good knowledge of the keyed signal. A properly keyed signal will have minimum thump on the make, and no perceptible click on the break. Illustration A (figure 20) shows a keyed signal having no control of make or break. The abrupt rise and decay of the waveform generates severe clicks and thumps that would be extremely objectionable on the air. Suitable filter circuits increase the rise and decay time of the signal (B), reducing the abrupt transition times to conservative values. Poor power-supply regulation can alter an otherwise perfect keyed waveform (C). Insufficient filter capacitance permits the power in the keyed wave to drop during long dashes, adding an unusual "yoop" to the signal. Other undesirable effects such as

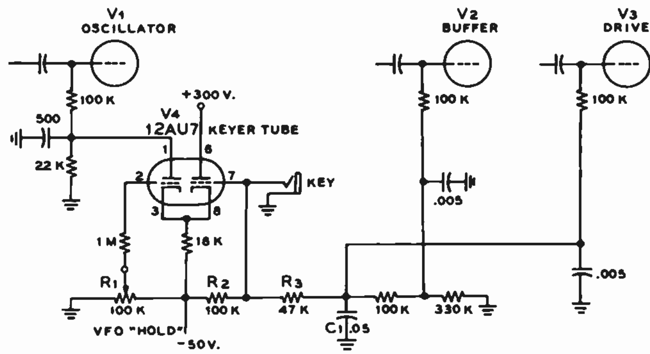


Figure 19

DIFFERENTIAL KEYER USING A 12AU7 DOUBLE TRIODE

excess ripple on the waveform may be easily observed with the use of an oscilloscope.

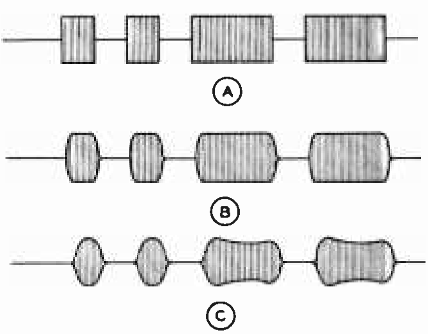


Figure 20

C-W KEYING CHARACTERISTICS

A—Abrupt rise and decay time of dot character leads to severe key clicks on make and break. B—Simple keying filter rounds dot character reducing transition time between key-open and key-closed condition. C—Poor power-supply regulation can distort keying waveform and add "yoop" to signal.

18-9 The Electronic Key

The *International Morse Code* used in radio telegraphy is made up of three elements: the *dot*, the *dash* and the *space* (see Chapter 1, Section 4). Intelligence can be transmitted at high rates of speed by using various combinations of these elements. A standard time relationship exists between the elements and between the space between words. The dot is a unit pulse and one pulse per second is termed one *baud*. The dot has a duty cycle of fifty percent, thus making the space equivalent in length to a unit pulse. The dash has a duty cycle of seventy-five percent, or three unit pulses in length. The space between words is seven unit pulses in length.

These fixed relationships between the code elements make it possible to use digital techniques to generate the timing characteristics used in an automatic electronic keying device, or *keyer*.

The representative keyer is actuated by the operator who keys at approximately correct times, the keyer functioning at precisely correct times determined by the *clock circuit* of the device.

In most keyers either an astable multivibrator or a pulse generator is used as a clock to create precise dots and dashes. The latter are made by filling in the space between two dots. Latching (memory) circuits are used so that an element, or code character, will be completed once it is initiated by the keyer paddle, or lever.

Since the transmitter following the keyer has wave-shaping circuits and possibly relay closure delay, a *weight control* may be incorporated in the keyer to vary the dot-to-space ratio.

Modern electronic keyers make use of solid-state circuitry which is admirably suited to on-off operation. A basic electronic key uses a single or dual key lever, movable in a horizontal plane and having two side contacts, much in the style of the mechanical key, or *bug*. Moving the keying paddle to the right produces a uniform string of dots and moving the paddle to the left produces a uniform string of dashes. A more

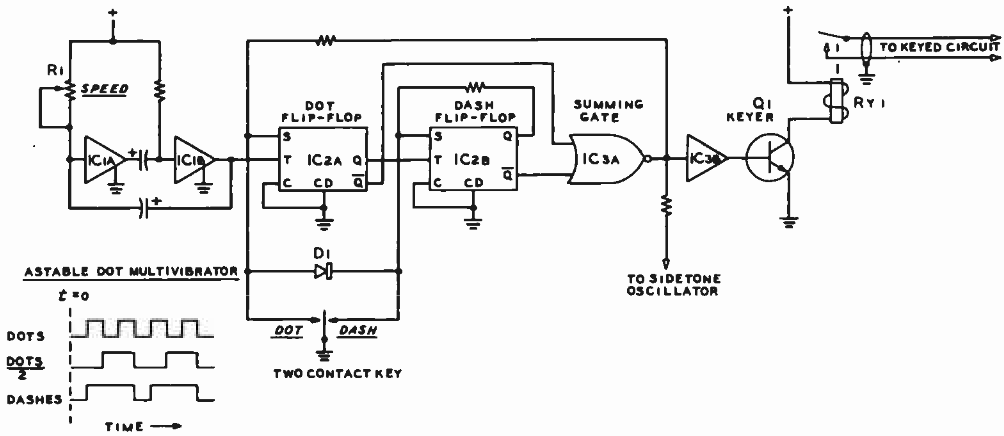


Figure 21

LOGIC FUNCTIONS OF ELECTRONIC KEYER

Astable multivibrator (IC_1) generates string of pulses (dots) with speed controlled by potentiometer R_1. Dot flip-flop sends precise square-wave dots when key contact is closed. Dash flip-flop adds long pulse to dot, forming 3-baud dash at output of summing gate. Amplifier and keying transistor drive a reed relay which controls the transmitter circuit. Dot memory, sidetone monitor, and iambic characteristic may be added to the basic keyer, if desired.

sophisticated keyer makes use of a dual *squeeze* paddle having double paddles, levers, and contacts, one set for dots and one for dashes. In one verson of this *squeeze keyer* (the *iambic* keyer), closing both paddles at once produces a string of sequential dots and dashes. This simplifies the sending of the letters having this sequence, such as C, Q, A, L, X, R, and K. Other versions of the squeeze keyer produce a string of dots or dashes when both paddles are closed. The keyer may be modified to send dots over dashes or dashes over dots when one paddle is closed after another. This action is termed *override*. Automatic dot completion is achieved by incorporating a *memory* circuit in the keyer.

A Basic Keyer The logic functions of a typical keyer are performed by silicon integrated circuits (figure 21). The pulse (dot) generator, or *clock*, is a free running multivibrator made up of two inverters (IC_{1A}, IC_{1B}) with the pulse speed controlled by potentiometer R_1. The free running, astable multivibrator allows precise spacing between the code elements as the space will always be one dot long, regardless of the sending speed. A dual flip-flop (IC_{2A}, IC_{2B}) is used as a character generator. Grounding the dot contact of the two-contact key triggers the *set* (S) input of the dot flip-flop (IC_{2A}) which then sends precise square-wave dots as long as the dot contact is closed. If the dot contact is opened before the completion of a dot, the element will be completed (dot memory).

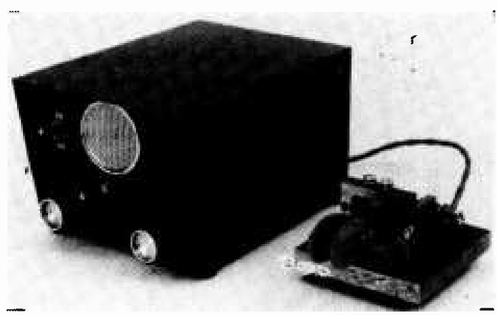


Figure 22

YSX SOLID-STATE ELECTRONIC KEY

This compact electronic key incorporates dot memory, weight control, sidetone oscillator and "squeeze key" operation. Panel controls are: primary power, speed adjust, weight and monitor-tune switch. Small p-m speaker and pilot light are mounted on panel. Terminal strip on back of unit provides connection to "squeeze key" at right.

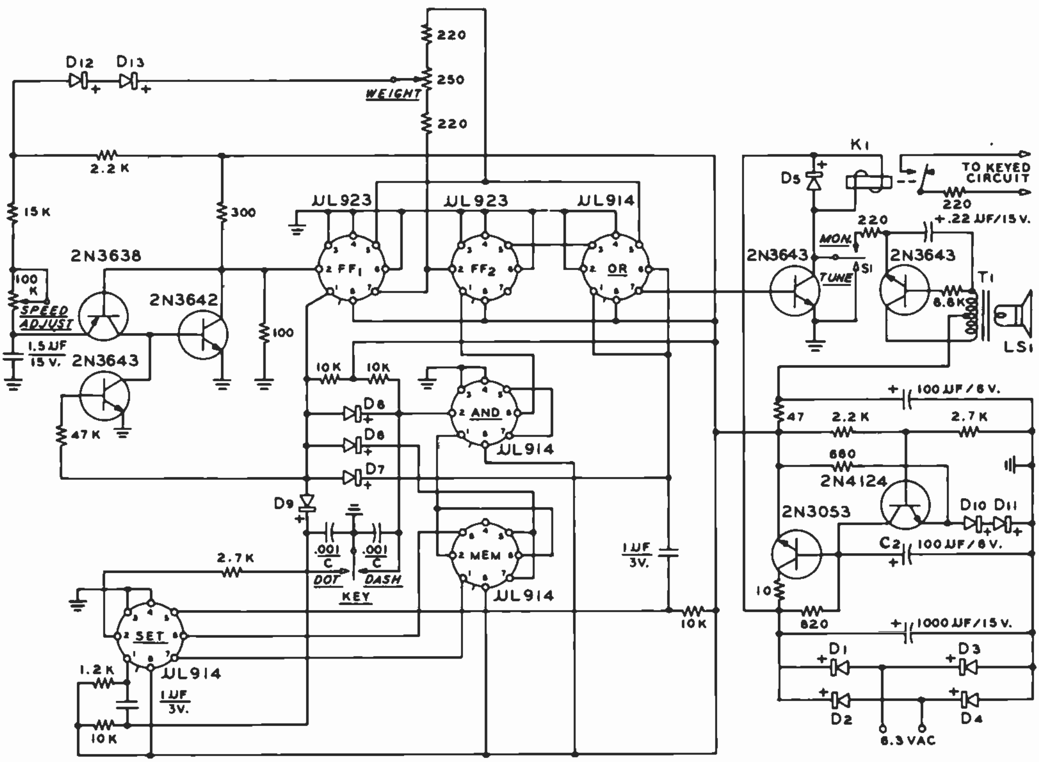


Figure 23

SCHEMATIC OF YSX KEYER

All resistors ¼-watt unless otherwise specified. 1 µf, 3 volt capacitors in set and memory circuits, as well as 1.5 µf, 15 volt capacitor in "speed adjust" circuit are nonpolarized paper or ceramic units. Diodes D_1-D_4 are 1N4001. Diodes D_5-D_9 are germanium 1N34A. Diodes D_{10}-D_{13} are 1N482. Use heat sink on 2N3053 pass transistor. Transformer T_1 is transistor output transformer, 500 ohms CT to 16 ohms. Relay is Magnecraft W102-MX1 or C-P Claire CRT-1134. Power transformer is 6.3 volt at 0.5 ampere, or more. ICs by Fairchild. Note: If keyer "double dots" on occasion, add 100K resistor between the base of the 2N3643 transistor and ground.

Grounding the dash contact of the key triggers the *set* input of the dash flip-flop (IC_{2B}) and also grounds the *set* input of the dot flip-flop through diode D_1. The dot flip-flop starts a dot, the dash flip-flop is triggered, and a second dot is initiated completing the dash element at the end of the second dot. The outputs of the flip-flops are added in a summing gate (IC_3). Once a character has started, it is impossible to alter it with the paddle and characters are self-completing.

The transmitter is actuated by a keying transistor (Q_1) employing a fast-operating relay in the collector circuit. In many instances, a *reed relay* is used. This type of relay has operate and release times of less than one millisecond and can allow good keying up to 100 words per minute. Some keyers eliminate the relay in favor of a keying transistor having a high collector-to-emitter voltage rating and a large collector current rating, thus permitting the transistor to be used to directly key cathode or grid circuits carrying up to several hundred milliamperes with an open-key voltage up to 300.

A *sidetone oscillator* or keying monitor can be driven by the keyer to provide the operator with an audible indication of the keying process.

Variation in the control logic and the use of a double paddle key permits conversion of the basic keyer to iambic keying

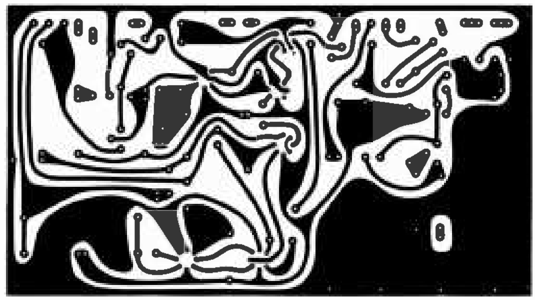


Figure 24

HALF-SIZE LAYOUT OF PRINTED-CIRCUIT BOARD

whereby grounding either the dot or the dash contact and then immediately grounding the other produces alternating dots and dashes. Another version will produce a dot or dash override sequence whereby closing both contacts simultaneously, only dots (or dashes) are generated.

18-10 The YSX Electronic Key

This compact electronic key (figure 22) is a modern solid-state version of the popular W9TO keyer first introduced in a vacuum-tube version in 1953. The keyer is designed after the configuration shown in figure 21 except that a pulse generator is used as the electronic clock instead of an astable multivibrator. The unit described in this section was adapted by W6YSX from the basic solid-state keyer designs of K3CUW and VE7BFK. It incorporates dot memory, weight control, sidetone oscillator, and optional "squeeze-key" operation. Integrated circuits and high-low level logic techniques such as described in Chapter 4 are used in this compact keyer.

Keyer Circuit Details A schematic diagram of the YSX keyer is shown in figure 23. Pulses generated in the clock circuit are timed by the *Speed Adjust*

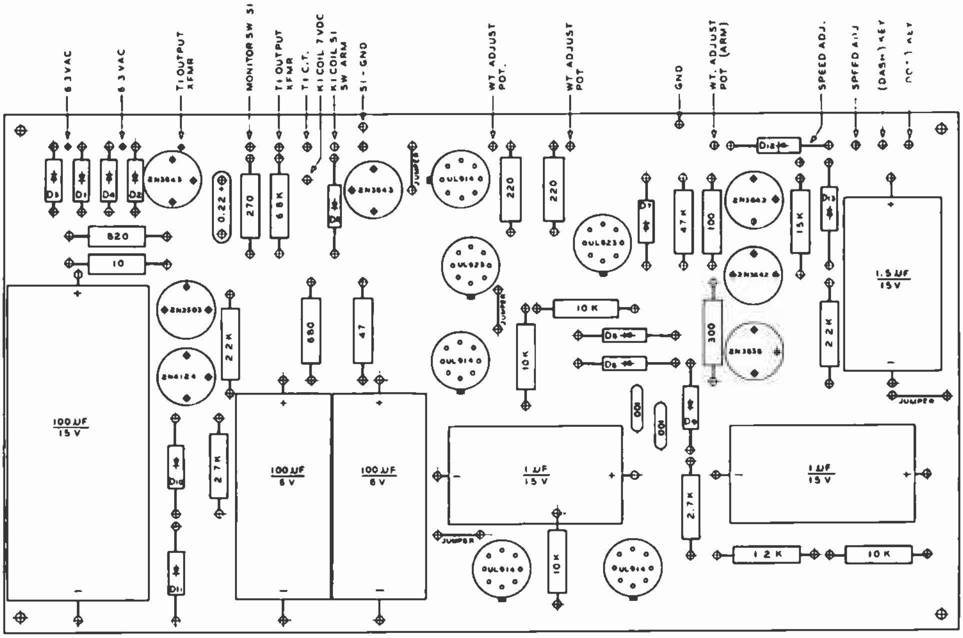


Figure 25

PLACEMENT OF COMPONENTS ON PC BOARD

potentiometer and trigger the dot flip-flop (FF_1) whenever the clock is *enabled* by the dot contact of the key paddle. When the dot contact is closed, the clock and FF_1 produce a string of dots as illustrated in figure 21. When the dash contact of the key is closed, both FF_1 and FF_2 are enabled, and the output of FF_2 resembles the lower waveform in figure 21. This sequence ensures a one-to-one dot/space ratio and a three-to-one dash/dot ratio.

When the dot contact is closed, the memory (MEM) circuit is set and the dot element formed. Once formed, a pulse from the output resets the memory so only one dot is sent, holding the dot sequence until a single element is completed.

An inexpensive reed relay (K_1) keys the transmitter, driven by a 2N3643 keying transistor. A 220-ohm resistor in series with one key lead eliminates contact sticking which may occur if a capacitive load is keyed. The relay contacts are capable of keying a load up to 30 watts. A sidetone oscillator (2N3643) is incorporated in the keyer for monitoring purposes along with a midget magnetic speaker.

The keyer is run from a small 6.3-volt filament transformer and a bridge rectifier supply. A series regulator (2N3053) provides +3.6 volts to the integrated circuits and clock. Reference voltage is derived from the forward voltage drop across diodes D_{10} and D_{11}. A 2N4124 serves as an error-signal amplifier and output voltage of the regulator is determined by the base voltage of the 2N4124. This may be adjusted over a small range by varying the value of the 2.2K resistor between base and the positive buss.

Construction Keyer The complete circuit of the YSX keyer is shown in figure 23. The various components are mounted on a printed-circuit board, whose layout is shown in figure 24. Placement of the components on the board is shown in figure 25. The circuit board, power transformer and other large components are housed in a small aluminum utility cabinet. The speaker is mounted to the front panel and the printed-circuit board is fastened to the chassis, which is cut out to receive the board (figure 26).

The 1-μF capacitors in the SET and RESET circuits should be either paper or ceramic (not electrolytic) because the voltage polarity across them reverses itself during keyer operation.

R-F Immunity Keyers may be sensitive to r-f energy coupled into the unit via the keying leads or the power line. A .001-μF, 1.6-kV ceramic capacitor connected across the primary winding of the power transformer, or from one leg to ground often helps, as well as the use of shielded leads betwen keyer and transmitter. A separate ground wire run from keyer to transmitter is a good idea. In some cases, a small r-f choke in the keying lead from relay K_1 may eliminate the trouble. The usual symptom of r-f feedback is indicated by the keyer not releasing, or by the formation of extra long dots and dashes. In rare instances, the keyer may generate an occasional dot when

Figure 26

INSIDE VIEW OF YSX KEYER

Printed circuit board is mounted over cut-out in aluminum chassis. Panel controls are at one end of assembly, with key jack and reed relay at opposite end. Filament transformer and primary circuit fuse are at edge of chassis, behind power switch.

the dash contact is closed. This is caused by high resistance in the contact circuit and cleaning the key contacts will eliminate the problem. The keyer operates cleanly at speeds between 10 and 50 words per minute.

Testing the Keyer Connect a three-conductor shielded cable between the keyer and the key paddle. Apply power and measure the supply voltages. The d-c voltage at the coil of relay K_1 should be about +6 to +8 volts (not critical) and the regulated voltage at the base of the 2N4124 transistor should be +3.6 volts. Move the *Speed Adjust* potentiometer to slow speed and close the dot contact of the paddle. A string of dots will be generated. Closing the dash contact should generate a string of dashes. Form the letter N quickly, pushing the paddle left then right immediately. The dash will self-complete and the dot will follow automatically after the proper space. Try the letters E, I, S and H, which are formed in the same manner as with a mechanical "bug" key. Now try the letters T, M, and O. Close the *Tune* switch (S_1) and observe if relay K_1 functions and the monitor is disabled.

Once proper keying conditions have been established, the keyer is ready to be connected to the transmitter. First, connect the keyer to the transmitter with a two-conductor, shielded cable and then connect the power cord of the keyer. This sequence is important to prevent possible damage to the keyer from transient voltages in the equipment.

18-11 An R-F Operated Keying Monitor

For proper sending and clean code transmission it is mandatory for the operator to monitor his signal. This may be done by copying the output of an audio oscillator that is simultaneously keyed with the transmitter. The oscillator shown in figure 27 is triggered by r-f picked up from the transmitter and thus provides an accurate replica of the keyed signal.

A unijunction transistor (2N2160) serves as a simple relaxation oscillator whose tone and volume are controlled by two potentiometers. The oscillator runs a small speaker

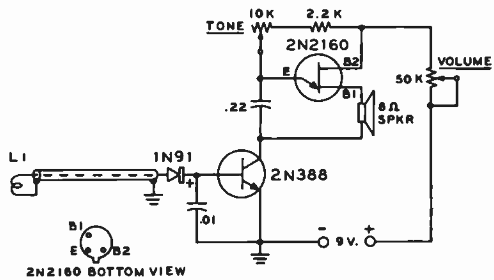


Figure 27

SCHEMATIC OF R-F ACTUATED KEYING MONITOR

and is enabled by grounding the junction of the 0.22-μF capacitor and the speaker. This is accomplished by a keying transistor (2N388) which is forward biased by a small r-f voltage developed by pickup coil L_1 and rectified by a diode.

The keying monitor may be built on a perforated circuit board and placed within an aluminum utility box. It is powered by a 9-volt transistor radio battery. The r-f pickup coil is introduced into the transmitter, in the vicinity of the tank coil of the final amplifier stage, and the trigger voltage level adjusted by moving the coil away from, or closer to, the tank inductor.

18-12 VOX Circuitry

A form of VOX (*voice-operated transmission*) is often employed in SSB operation.

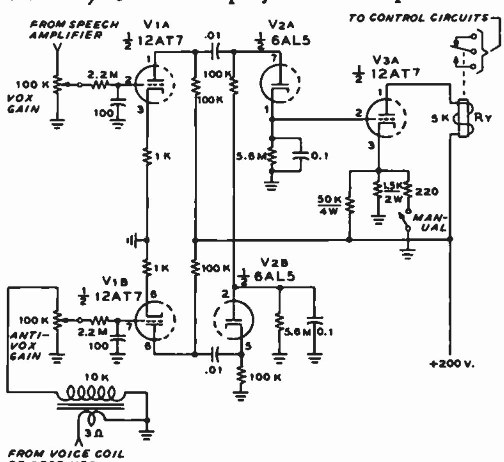


Figure 28

A REPRESENTATIVE VOX CIRCUIT

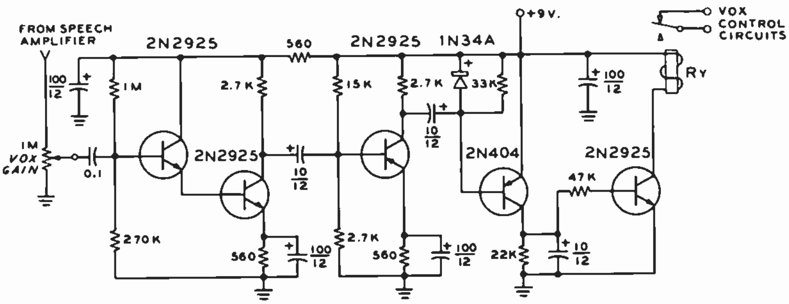


Figure 29
TRANSISTORIZED VOX

Self-contained VOX circuit may be added to existing equipment. Relay has pull-in current of about 7 ma. (Sigma 11F-1000G-SIL). Unenergized relay current is 3 ma, and actual pull-in current is 12 ma.

The VOX circuitry makes use of a transmitter control relay that is actuated by the operator's voice and is held open by an *anti-vox* circuit actuated by the audio system of the station receiver. Voice-controlled break-in operation is thus made possible without annoying feedback from the receiver speaker. A representative VOX system is illustrated in figure 28. The VOX signal voltage is taken from the speech amplifier of the SSB transmitter and adjusted to the proper amplitude by means of *VOX-gain* potentiometer. The signal is rectified by diode V_2A and the positive voice impulses are applied to the grid of the VOX relay tube (V_3A) which is normally biased to cutoff. An RC network in the VOX rectifier circuit permits rapid relay action yet delays the opening of the relay so that VOX action is sustained during syllables and between words. Delay periods of up to 0.5 second are common.

The antivox signal voltage is derived from the speaker circuit of the receiver, adjusted to the proper amplitude by the *antivox-gain* potentiometer and rectified by diode V_2B to provide a negative voice impulse which biases the vox diode (V_2A) to a nonconducting state. The relay is held in a cut-off position until a positive override signal from the VOX circuit defeats the antivox signal taken from the station receiver. The relay tube may also be actuated by the *manual* switch which drops the bias level, causing the tube to draw a heavy plate current and trip the VOX relay.

A Transistor VOX Circuit The transistor VOX circuit shown in figure 29 may be added to existing SSB equipment, or may be included in new-design equipment to provide inexpensive and compact VOX control. The unit is powered from a 9-volt transistor-radio battery, or power may be obtained from a well-filtered tap point in the station equipment.

The input impedance of the first emitter-follower stage is of sufficiently high impedance to work directly from a crystal micro-

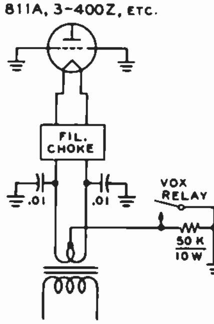


Figure 30

VOX BIAS CONTROL

Cutoff bias for grounded-grid triode may be obtained from cathode bias resistor. Action of VOX relay shorts out resistor, restoring amplifier to normal operating conditions.

phone or from the grid circuit of the first or second speech-amplifier stage in the station equipment. Two stages of high-gain RC amplifier follow the direct coupled input stage and the enhanced voice signal is rectified and clamped in a 1N34-2N404 combination, the output of which drives a 2N2925 relay-control transistor. Time delay is determined by a capacitor in the emitter circuit of the 2N404. Relay dropout is determined largely by the travel time of the relay armature.

VOX Bias Control It is desirable to completely disable a high-power linear amplifier during reception for two reasons: first, the amplifier consumes standby power unless it is biased to cutoff and, second, many amplifiers will generate "white noise" when in a normal standby condition. The white noise, or diode noise, may show up in the receiver as a loud hiss interfering with all but the loudest signals.

The circuit of figure 30 provides an automatic cutoff-bias system for a VOX-controlled amplifier stage. The resting plate current of the amplifier is passed through a 50K resistor in the filament return circuit, creating a voltage drop that is applied as cutoff bias to the tube(s). The filament circuit is raised to a positive voltage with respect to the grid, thus leaving the grid in a negative, cutoff condition. On activation of the VOX relay, a separate set of contacts short out the bias resistor, restoring the amplifier stage to normal operating condition.

CHAPTER NINETEEN

Mobile and Portable Equipment

Mobile operation is permitted on all amateur bands. Tremendous impetus to this phase of the hobby was given by the suitable design of compact mobile equipment. Complete mobile installations may be purchased as packaged units, or the whole mobile station may be home built, according to the whim of the operator.

The problems involved in achieving a satisfactory two-way installation vary somewhat with the band, but many of the problems are common to all bands. For instance, ignition noise is more troublesome on 10 meters than on 80 meters, but on the other hand an efficient antenna system is much more easily accomplished on 10 meters than on 80 meters. Also, obtaining a worthwhile amount of transmitter output without excessive battery drain is a problem on all bands.

Specialized mobile equipment is available for operation on the 2- and 6-meter bands and a small amount of mobile use is made of the 432-MHz band. The availability of surplus equipment, moreover, has stimulated f-m mobile activity, especially on 2 meters, where the use of fixed f-m repeaters placed on elevated locations has done much to enhance vhf mobile operation.

The majority of high-frequency mobile operation takes place on single sideband. The low duty-cycle of SSB equipment, as contrasted to the heavy power drain of conventional a-m gear has encouraged the use of relatively high-power sideband equipment in many mobile installations. The rigid frequency stability requirement for satisfactory SSB reception, however, has obsoleted the once-popular tuned-converter and auto-receiver combination formerly used for a-m reception. The SSB transceiver, thus, has become the universal high-frequency mobile device with the majority of use on SSB and a small but growing minority of amateurs using the transceiver for mobile c-w operation.

Portable operation is extremely popular on all h-f and vhf bands and specialized equipment for this mode of operation is available, using battery power as a primary source. To conserve battery drain, solid-state devices are commonly used and power input is limited for the same reason. Some amateurs employ gasoline driven power generators for portable and emergency service. In all cases, however, the power source is critical since even mobile power sources are limited in their ultimate capacity.

The handicap of low power in mobile and portable operation can be overcome by the ability of the operator to select his operating site in many instances. A high, clear, noise-free location will permit successful operation on the vhf bands to the line-of-sight distance and will permit contacts over thousands of miles on the h-f bands while running only a few watts of transmitter power. A good location combined with a good station antenna will permit a successful operator to compete in today's interference-full amateur bands even under the most difficult operating conditions.

19-1 Mobile and Portable Power Sources

A small transistor converter for casual listening may be run from a 9-volt battery, but larger mobile receivers, transmitters, and transceivers require power from the electrical system of the automobile. SSB equipment, with its relatively light duty cycle, is ideally suited for mobile use and demands the least primary power drain for a given radiated signal of all the common types of amateur transmission. As a result of the combination of low power requirement and enhanced communication effectiveness, SSB has supplanted amplitude modulation for mobile service on the h-f amateur bands. F-m, on the other hand, is universally used for vhf mobile service. In any case, a total equipment power drain of about 250 watts for SSB or f-m is about the maximum power that may be taken from the electrical system of an automobile without serious regard to discharging the battery when the car is stopped for *short* periods of mobile operation.

With many SSB mobile-radio installations now requiring 500 to 1000 watts peak power from the automotive electrical system, it is usually necessary to run the car engine when the equipment is operated for more than a few minutes at a time to avoid discharging the battery. Fortunately, a majority of automobiles have a 12-volt *alternator system* as standard equipment and as a result, most SSB transceivers may be run directly from the automotive electrical system without undue strain on the battery during the course of normal driving.

The Alternator A typical alternator circuit is shown in figure 1. The alternator differs from the classic generator in that it uses a rotating field to which d.c. is supplied through slip rings and carbon brushes. Field current is quite low, of the order of 3 amperes or so for many alternators. The rotating field usually has six pairs of poles, and the output of one stator winding represents six electrical cycles for each revolution of the field. The output frequency in cycles per second is one-tenth

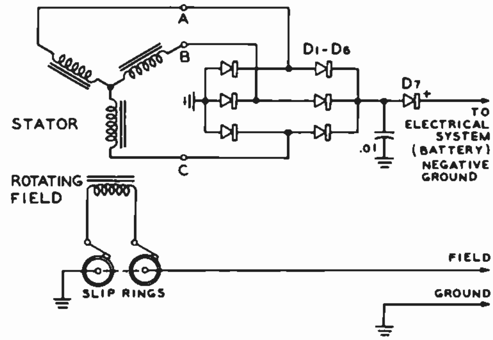


Figure 1

THREE-PHASE AUTOMOBILE ALTERNATOR

Three-phase output voltage is converted to d-c by full wave rectifier D_1, D_6. Rectifier D_7 protects rectifier assembly from transients and voltage surges in electrical system of auto.

the shaft speed expressed in revolutions per second.

The high output current of the alternator is supplied directly from the fixed stator windings in the form of three-phase current. The stator is usually connected in a *wye* (Y) configuration to an internal rectifier assembly made up of six silicon diodes which provide full-wave rectification. The ripple frequency is six times the frequency developed in one winding. Thus, at a shaft speed of 4000 r.p.m., the nominal voltage is 14, output frequency will be 400 Hz, and the ripple frequency is 2400 Hz.

The diode assembly (D_1, D_6) may be mounted on or behind the rear end-bell of the alternator, in conjunction with an isolation diode (D_7) which protects the rectifier assembly from voltage surges and helps to suppress radio noise.

The output voltage of the alternator system is a function of the shaft speed to about 5000 r.p.m. or so. Above this speed, output voltage tends to stabilize because of hysteresis losses. In any case, the alternator output is regulated through adjusting the current in the field by a mechanical voltage regulator or by a solid-state regulator. Because the reverse current through the rectifier diodes is small, the alternator is usually connected directly to the battery without the use of a cutout relay.

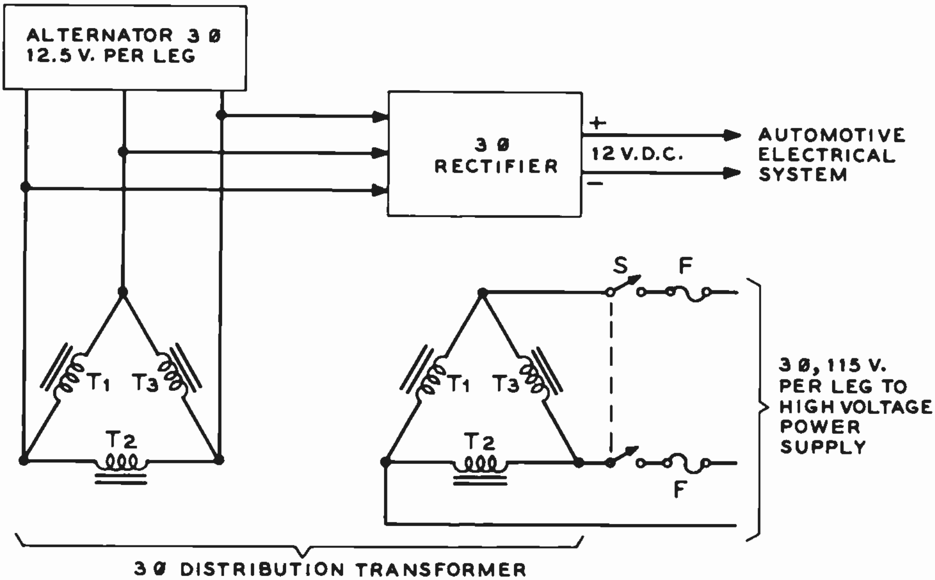


Figure 2

THREE-PHASE MOBILE POWER SYSTEM

Three-phase transformer (T_1) in delta or wye configuration provides 115 volts a.c. for operation of mobile equipment.

Using the Alternator The automotive alternator may be modified to supply 12 volts at 400 Hz for auxiliary equipment. Many alternators are capable of supplying 350 watts of power which, when the battery is charged and the auto accessories are not being used, may be employed to run the mobile gear. A diagram of such an installation is shown in figure 2. Common 60-Hz transformers may be used, or a special 400-Hz three-phase distribution transformer (figure 3). Voltage regulation of the alternator system is very good, although the frequency varies with engine speed, ranging from 100 Hz or so with the engine idling to nearly 1000 Hz at top speed. Modern power transformers, however, even though rated at 60 Hz, are capable of operating efficiently over this range of frequencies.

The schematic of an 1800-volt, 3-phase intermittent-duty SSB kilowatt power supply for mobile service is shown in figure 4. This supply is designed for use with a 1kW PEP linear amplifier using four 811A tubes or two 572B/T-160L tubes.

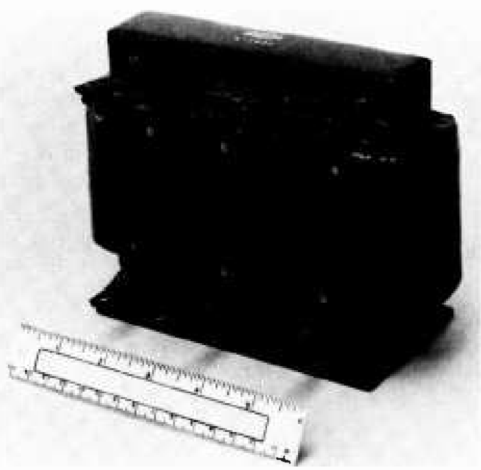


Figure 3

THREE-PHASE DELTA-WYE DISTRIBUTION TRANSFORMER

Tranex 4-1852 transformer provides 115 volts from nominal 12-volt source at 1000 watts average power. (Available from Ken Pierce Co., Box 877 Palo Alto, Calif. 94302)

A solid-state regulator is recommended for use with this supply, since mechanical relay regulators interrupt the alternator field current when the battery is fully charged, thus removing the power to the equipment. If a relay regulator is used, it should be shorted out, or otherwise disabled during mobile operation.

Batteries The voltage available at the terminals of a battery is determined by the chemical composition of the cell. Many types and sizes of batteries are available for portable radio and comunication equipment. The inexpensive *carbon-zinc* cell provides a nominal 1.5 volts and, unused, will hold a charge for about a year. The current capacity of the cell depends on the physical size of the electrodes and the composition of the electrolyte. A battery may be made up of a number of cells connected in series, providing good life under intermittent service.

Next to the carbon-zinc cell, the most commonly used unit is the *alkaline cell* (1.2 volts) which has about twice the total energy capacity per unit size as compared to the carbon-zinc cell. This cell is capable of a high discharge rate over an extended period of time and provides longer life in continuous service than does the carbon-zinc cell.

The *mercury cell* (1.34 volts) is more expensive than the previously mentioned cells, but it has an extremely long working life. In addition, the mercury cell maintains full rated voltage until just before expiration; then the voltage drops sharply. Shelf life of the mercury cell is excellent and it may be stored for long periods of time.

These three types of batteries may be recharged to some extent by reversing the chemical action by application of a reverse current to the cell. For best results, the current should be low and should have a small a-c component to provide a more even redeposit of material on the negative electrode. Recharged cells have an uncertain operating life, and the recharging cycle may vary from cell to cell.

The *nickel-cadmium (Nicad)* cell (1.25 volts) is the most expensive cell in terms of initial costs, but it may be recharged at a slow rate a number of times in reliable cycles of operation.

The wet cell, *(lead-acid)* storage battery is in near-universal use in automotive equipment The cell delivers about 2.1 volts and is rechargeable. The lead-acid cell is made of coated lead plates immersed in a solution of sulphuric acid and water. The acid content of the dielectric varies with the state of charge, which may be determined by measuring the specific gravity of the electrolyte. Generally speaking, a hydrometer reading of 1.27 indicates a fully charged cell, whereas a reading of 1.15 or below indicates the cell is in need of charging. The wet cell may be fast-charged as high as 40 amperes for a 12-volt battery, provided that care is taken to let escaping gases free themselves and

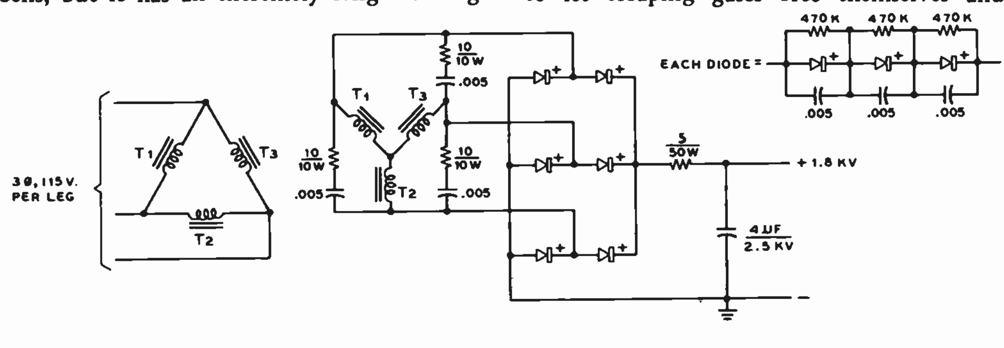


Figure 4

THREE-PHASE-MOBILE KILOWATT SUPPLY

Three-phase power from a system such as shown in figure 1 may be used to provide high voltage for mobile transmitting equipment. For 1800 volts, transformers T_1, T_2, and T_3 are 115-volt primary, 830-volt secondary (Stancor PC-8301). For 2400 volts, T_1, T_2 and T_3 are 115-volt primary, 1030-volt secondary (Stancor PC-8302). Three type 1N1697 or 1N4005 diodes are used in each stack.

provided that electrolyte temperature is held below 125° Fahrenheit.

The Nickel-Cadmium Cell The nickel-cadmium (Nicad) cell is a high-efficiency cell capable of being recharged hundreds or thousands of times in the proper circumstances. The cell has a positive nickel electrode and a negative cadmium electrode immersed in a solution of potassium hydroxide at a specific gravity of 1.300 at 72°F. The common and popular lead-acid battery does not equal the recharge ability of the nickel-cadmium battery and use of the latter is common in mobile and portable equipment and other devices where small cell size and high recharge capability are an asset.

There are two common types of nickel-cadmium batteries classified as *vented* and *nonvented*. The nonvented cell is a hermetically sealed unit which resembles a conventional dry cell in appearance. The vented cell resembles a lead-acid cell and often has a removable plug which covers a port for gas venting during the charging process.

The terminal voltage of a nickel-cadmium cell varies with the state of charge and normally runs between 1.25 and 1.30 volts on open circuit. Exact terminal voltage depends on the state of charge, the charging current, and the time of charge. The specific gravity of the electrolyte, moreover, does not change appreciably between charge and discharge, as is commonly done with lead-acid cells. At end of charge, nickel-cadmium cell voltage may drop as low as a fraction of a volt and it is possible under heavy discharge for a cell to show a negative or reversed voltage, indicating a state of extreme discharge. A terminal voltage of 1.1 volts is usually considered to be a state of complete discharge, for all practical purposes and should not be exceeded.

For standby service the nickel-cadmium cell can be maintained on a trickle charge, with the charger adjusted to maintain a terminal potential of 1.36 to 1.38 volts per cell. Following a substantial discharge, a regular charge should be given, after which the cell is placed back on trickle charge. While the overcharge tolerance is good and the cell may be left on charge for long periods of time, severe overcharge must be avoided because the cell may be destroyed by accumulation of gases within the container.

The nickel-cadmium cell may be charged by a *constant-potential process* whereby charger current is continually adjusted to maintain a constant potential of 1.55 volts across the cell. This requires a charger designed for such service, as very high current occurs at the start of charge, tapering rapidly as the charge progresses. A fully discharged cell can be completely recharged by this method in an hour or so.

The nickel-cadmium cell may also be charged by the *constant-current* process. This technique requires a charging source having an ammeter and control rheostat in the charging circuit. The cell is charged at a constant current rate. To maintain constant current, the rheostat requires adjustment during the charge period as the counter-emf of the cell rises.

The practical value of charging current varies from cell to cell and is usually specified by the manufacturer. If the extent of discharge is not known, the cell may be charged at a constant current rate until the cell voltage ceases to rise. Reasonable overcharge is not harmful as long as the electrolyte level is above the plate tops and the electrolyte temperature does not exceed 125°F.

When charging at a high rate, the nickel-cadmium cell will gas rather vigorously when approaching full charge. This gassing will cause the electrolyte level to rise above the limit line. This apparent excess electrolyte should not be removed as the level will drop back after the cell stands on open circuit following the charge. Charging disassociates water from the electrolyte which forms this gas.

The energy capability of a nickel-cadmium cell is usually rated in milliampere-hours. for small cells and ampere-hours for large ones. The rating is based on cell capability to a specific end point (usually 1.1 volts per cell) over a 10-hour period. This figure is used as the capacity of the cell and depends upon the rate of discharge. Generally speaking, the charging current is held to 10 percent of the milliampere-hour rating of a small cell and the time of charge is set at 150 percent of the time required to reestablish the maximum milliampere-hour rating of the cell. Thus a 250 milliampere-hour cell is charged at 25 milliamperes for

15 hours. This ensures that the lost energy is restored and various other losses and inefficiencies are accounted for.

19-2 Transistor Supplies

The vibrator-type of mobile supply achieves an over-all efficiency in the neighborhood of 70%. The vibrator may be thought of as a mechanical switch reversing the polarity of the primary source at a repetition rate of 120 transfers per second. The switch is actuated by a magnetic coil and breaker circuit requiring appreciable power which must be supplied by the primary source.

One of the principal applications of the transistor is in switching circuits. The transistor may be switched from an "off" condition to an "on" condition with but the application of a minute exciting signal. When the transistor is nonconductive it may be considered to be an open circuit. When it is in a conductive state, the internal resistance is very low. Two transistors properly connected, therefore, can replace the single-pole, double-throw mechanical switch representing the vibrator. The transistor switching action is many times faster than that of the mechanical vibrator and the transistor can switch an appreciable amount of power. Efficiencies in the neighborhood of 95 percent can be obtained with 28-volt primary-type transistor power supplies, permitting great savings in primary power over conventional vibrators and dynamotors.

Transistor Operation The transistor operation resembles a magnetically coupled multivibrator, or an audio-frequency push-pull square-wave oscillator (figure 5C). A special feedback winding on the power transformer provides 180-degree phase-shift voltage necessary to maintain oscillation. In this application the transistors are operated as on-off switches; i.e., they are either completing the circuit or opening it. The oscillator output voltage is a square wave having a frequency that is dependent on the driving voltage, the primary inductance of the power transformer, and the peak collector current drawn by the conducting transistor. Changes in transformer turns, core area, core material, and feedback turns ratio have an effect on the frequency of oscillation. Frequencies in common use are in the range of 120 Hz to 3500 Hz.

The power consumed by the transistors is relatively independent of load. Loading the oscillator causes an increase in input current that is sufficient to supply the required power to the load and the additional losses in the transformer windings. Thus, the over-all efficiency actually increases with load and is greatest at the heaviest load the oscillator will supply. A result of this is that an increase in load produces very little extra heating of the transistors. This feature

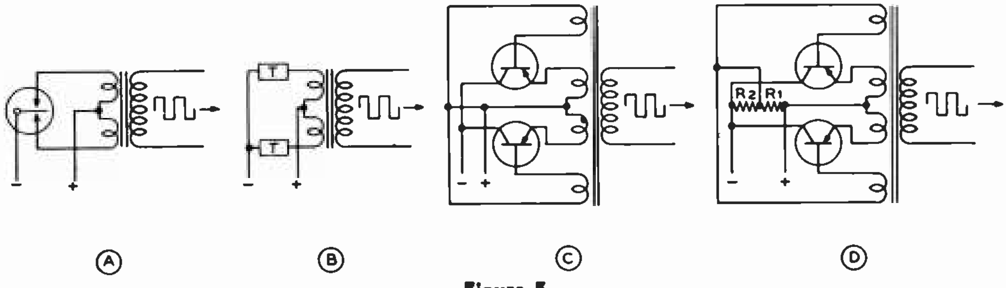


Figure 5

TRANSISTORS CAN REPLACE VIBRATOR IN MOBILE POWER SUPPLY SYSTEM

A—Typical vibrator circuit.
B—Vibrator can be represented by two single-pole single-throw switches, or transistors.
C—Push-pull square-wave "oscillator" is driven by special feedback windings on power transformer.
D—Addition of bias in base-emitter circuit results in oscillator capable of starting under full load.

means that it is impossible to burn out the transistors in the event of a shorted load since the switching action merely stops.

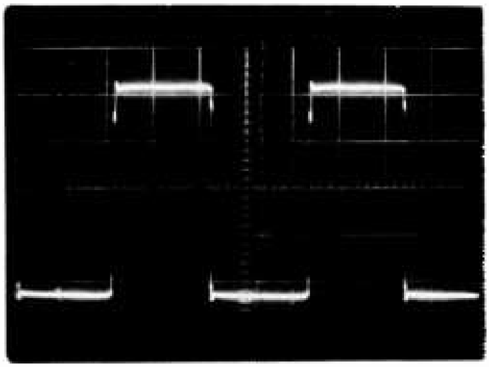


Figure 6

EMITTER-COLLECTOR WAVEFORM OF SWITCHING CIRCUIT

Square waveshape produces almost ideal switching action. Small 2-volt "spike" on leading edge of pulses may be reduced by proper transformer design. Pulse length is about 1000 microseconds and rise time is 10 microseconds.

Transistor Power Rating The power capability of the transistor is limited by the amount of heat created by the current flow through the internal resistance of the transistor. When the transistor is conducting, the internal resistance is extremely low and little heat is generated by current flow. Conversely, when the transistor is in a cut-off condition the internal resistance is very high and the current flow is extremely small. Thus, in both the "on" and "off" conditions the transistor dissipates a minimum of power. The important portion of the operating cycle is that portion when the actual switching from one transistor to the other occurs, as this is the time during which the transistor may be passing through the region of high dissipation. The greater the rate of switching, in general, the faster will be the rise time of the square wave (figure 6) and the lower will be the internal losses of the transistor. The average transistor can switch about eight times the power rating of class-A operation of the unit. Two switching transistors having 5-watt class-A power output rating can therefore switch 80 watts of power when working at optimum switching frequency.

Self-Starting Oscillators The transistor supply shown in figure 5C is impractical because oscillations will not start under load. Base bias of the proper polarity has to be momentarily introduced into the base-emitter circuit before oscillation will start and sustain itself. The addition of a bias resistor (figure 5D) to the circuit results in an oscillator that is capable of starting under full load. R_1 is usually of the order of 10 to 50 ohms while R_2 is adjusted so that approximately 100 milliamperes flow through the circuit.

The current drawn from the battery by this network flows through R_2 and then divides between R_1 and the input resistances of the two transistors. The current flowing in the emitter-base circuit depends on the value of input resistance. The induced volt-

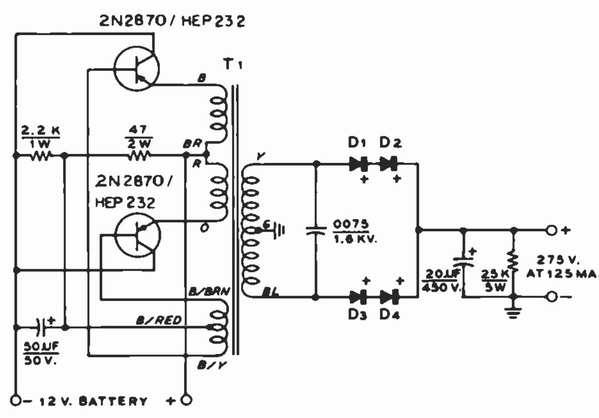


Figure 7

SCHEMATIC, TRANSISTOR POWER SUPPLY FOR 12-VOLT AUTOMOTIVE SYSTEM

T_1—Transistor power transformer. 12-volt primary, to provide 275 volts at 125 ma. Stancor DCT-1
D_1-D_4—IN4005 with .01 µfd and 100K across each diode
Use 6 amp. fuse in +12-volt lead.

age across the feedback winding of the transformer is a square wave of such polarity that it forward-biases the emitter-base diode of the transistor that is starting to conduct collector current, and reverse-biases the other transistor. The forward-biased transistor will have a very low input impedance, while the input impedance of the reverse-biased transistor will be quite high. Thus, most of the starting current drained from the primary power source will flow in R_1 and the base-emitter circuit of the forward-biased transistor and very little in the other transistor. It can be seen that R_1 must not be too low in comparison to the input resistance of the conducting transistor, or it will shunt too much current from the transistor. When switching takes place, the transformer polarities reverse and the additional current now flows in the base-emitter circuit of the other transistor.

The Power Transformer The power transformer in a transistor-type supply is designed to reach a state of maximum flux density (saturation) at the point of maximum transistor conductance. When this state is reached the flux density drops to zero and reduces the feedback voltage developed in the base winding to zero. The flux then reverses because there is no conducting transistor to sustain the magnetizing current. This change of flux induces a voltage of the opposite polarity in the transformer. This voltage turns the first transistor off and holds the second transistor on. The transistor instantly reaches a state of maximum conduction, producing a state of saturation in the transformer. This action repeats itself at a very fast rate. Switching time is of the order of 5 to 10 microseconds, and saturation time is perhaps 200 to 2000 microseconds. The collector waveform of a typical transistor supply is shown in figure 6. The rise time of the wave is about 5 microseconds, and the saturation time is 500 microseconds. The small "spike" at the leading edge of the pulse has an amplitude of about 2.5 volts and is a product of switching transients caused by the primary leakage reactance of the transformer. Proper transformer design can reduce this "spike" to a minimum value. An excessively large "spike" can puncture the transistor junction and ruin the unit.

A 35-Watt Supply The 35-watt power unit uses two inexpensive 2N2870 power transistors for the switching elements and four silicon diodes for the high-voltage rectifiers. The complete schematic is shown in figure 7. Because of the relatively high switching frequency only a single 20-μfd filter capacitor is required to provide pure direct current.

Regulation of the supply is remarkably good. No-load voltage is 310 volts, dropping to 275 volts at maximum current drain of 125 milliamperes.

The complete power package is built on an aluminum chassis-box measuring 5¼″ × 3″ × 2″. Paint is removed from the center portion of the box to form a simple heat sink for the transistors. The box therefore conducts heat away from the collector elements of the transistors. The collector of the transistor is the metal case terminal and in this circuit is returned to the nega-

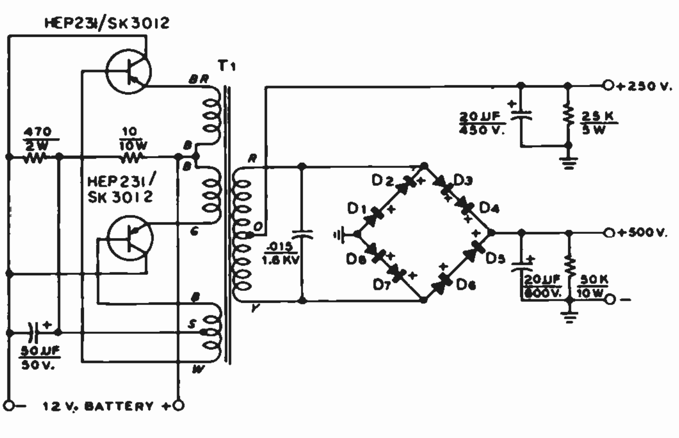


Figure 8

SCHEMATIC, 85-WATT TRANSISTOR POWER SUPPLY FOR 12-VOLT AUTOMOTIVE SYSTEM

T_1—Transistor power transformer. 12-volt primary to provide 275 volts at 125 ma. Stancor DCT-2.

D_1-D_8—1N4005 with .01 µfd and 100K across each diode.

tive terminal of the primary supply. If the negative of the automobile battery is grounded to the frame of the car the case of the transistor may be directly grounded to the unpainted area of the chassis. If the positive terminal of the car battery is grounded it is necessary to electrically insulate the transistor from the aluminum chassis, yet at the same time permit a low *thermal* barrier to exist between the transistor case and the power-supply chassis. A simple method of accomplishing this is to insert a thin mica sheet between the transistor and the chassis. *Two-mil* (0.002″) mica washers for transistors are available at many large radio supply houses. The mica is placed between the transistor and the chassis deck, and fiber washers are placed under the retaining nuts holding the transistors in place. When the transistors are mounted in place, measure the collector-to-ground resistance with an ohmmeter. It should be 100 megohms or higher in dry air. After the mounting is completed, spray the transistor and the bare chassis section with plastic *Krylon* to retard oxidation. Several manufacturers produce anodized aluminum washers that serve as mounting insulators. These may be used in place of the mica washers, if desired.

An 85-Watt Supply Figure 8 shows the schematic of a dual-voltage transistor mobile power supply. A bridge rectifier permits the choice of either 250 volts or 500 volts, or a combination of both at a total current drain that limits the secondary power to 85 watts. Thus, 500 volts at 170 milliamperes may be drawn, with correspondingly less current as additional power is drawn from the 250-volt tap.

The supply is built on an aluminum box chassis measuring 7″ × 5″ × 3″, the layout closely following that of the 35-watt supply. HEP-231 or SK3012 transistors are used as the switching elements and eight silicon diodes form the high-voltage bridge rectifier.

The transistors are affixed to the chassis in conjunction with a homemade aluminum heat sink formed from two pieces of aluminum sheet bent into channels, as shown in figure 9. Silicone grease is spread thinly between the transistors, heat sinks, and the chassis to permit better heat transfer between the various components of the assembly.

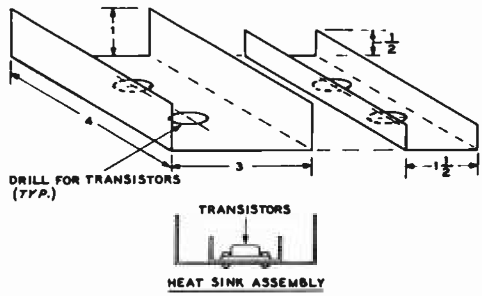


Figure 9

HOMEMADE HEAT SINK FOR POWER TRANSISTOR

A 270-Watt Transceiver Supply SSB transceivers suitable for mobile service are capable of PEP power inputs up to 250 watts or more. Shown in figure 10 is a compact triple-voltage supply capable of running many transceivers from a 12 volt d-c supply. The unit provides 900 volts at 300 milliamperes, 275 volts at 180 milliamperes, and an adjustable bias voltage of −15 to −150. Additionally, −150 volts at 40 milliamperes is available for VOX standby circuitry in auxiliary equipment.

Two heavy-duty switching transistors are used, driven by base feedback from a winding of oscillator transformer T_1. The transistors are forward-biased by a voltage divider circuit and are protected from voltage spikes by the two 1N4719 diodes. Two zener diodes (1N4746) provide transient suppression in the primary circuit of transformer T_1. A power transformer (T_2) is driven by the squarewave pulses provided by the switching circuit based on transformer T_1.

The supply is built on an aluminum chassis measuring 12″ × 6″ × 3″. The main components are mounted atop the chassis with the heat sinks mounted on one side, with the fins in a vertical position. To improve thermal conductivity, the heat sinks are bolted to a ⅛-inch thick copper plate (measuring 12″ × 6″) affixed to the side of the chassis. The transistors are in-

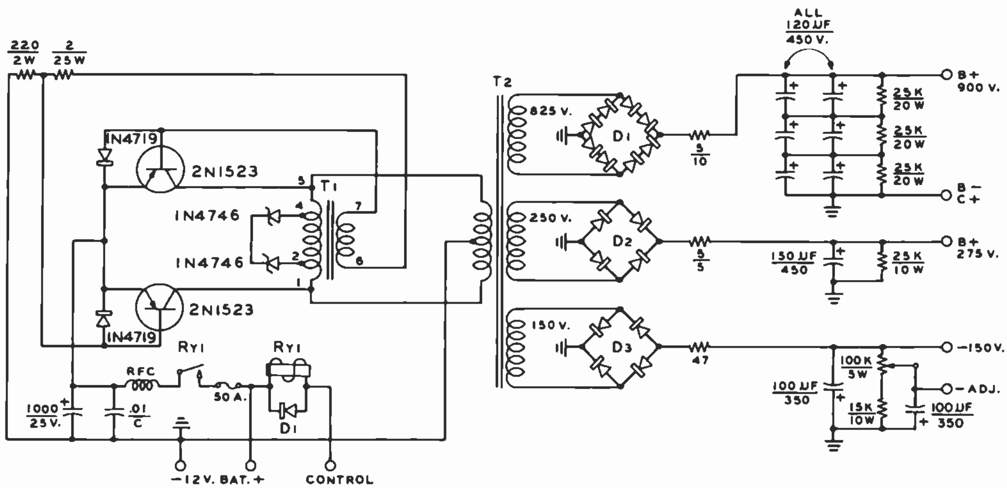


Figure 10

270-WATT MOBILE TRANSCEIVER POWER SUPPLY

D_1-D_3—Use 1N4005 diodes. Two diodes in series are used in each leg of D_1. Place 470K, 1-watt resistor and 0.1, 1.6-kV disc across each diode
RY_1—SPST contactor, 60 ampere, with 12-volt coil. Potter-Brumfield MB-3D
RFC—10 turns #10 enamel wire on 1" form
T_1—Oscillator transformer (1000 Hz). Osborne 6784 (Osborne Transformer Co., 3834 Mitchell Ave., Detroit, Michigan)
T_2—Power transformer, Osborne 21555
Heat sink—One for each 2N1523. Thermalloy 6421B, or Delco 7281366
Use Delco insulator kit 7274633 for transistors

sulated from the chassis by thin insulators coated with silicone grease.

All primary leads to the power transistors, transformer T_1, and the input terminals are wired with #6 conductors, with the negative primary circuit grounded at one point in the supply. Heavy ¼-inch battery leads run from the supply to the automobile battery. The supply should be mounted close to the battery to reduce primary voltage drop to a minimum.

A D-C to A-C Inverter For the Car or Boat Radio and electrical equipment of all kinds up to about 200 watts intermittent power consumption may be run from this compact d-c to a-c power inverter. Designed for use with 12-volt automotive systems, the inverter provides a nominal 115-volt, 60-Hz square-wave output, suitable for transformer-powered equipment, lights, or motors.

The inverter construction is straightforward, and assembly is on an aluminum chassis measuring 8″ × 6″ × 2″. A standard heat sink for the transistors is specified, however, the sink shown in figure 9 may be used. A grounded-collector circuit is

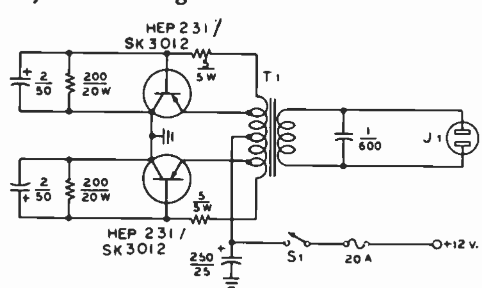


Figure 11

D-C TO A-C INVERTER FOR THE CAR

T_1—Inverter transformer. 12-volt d-c, tapped primary, 115-volt a-c, tapped secondary (Triad TY-75A)
Line Filter—J. W. Miller 5521 choke, 4 μH at 20 amperes, bypassed with 0.1-μfd capacitors on each side (12-volt circuit). J. W. Miller 7818 (115-volt circuit)
Heat Sink—Wakefield NC 623A for each transistor

used (negative ground) so the transistors need not be insulated from the heat sink or chassis. Silicon grease should be placed between the transistor, sink sections, and chassis to ensure good thermal conductivity between the units. The low-voltage primary circuit should be wired with heavy-duty flexible line cord, or stranded #12 hookup wire.

This supply is designed to start under full load, and should be turned on loaded, since unloaded operation (especially starting and stopping) may give rise to transients which may endanger the transistors.

The supply is capable of 100 watts continuous power and about twice this amount in intermittent service. Because of the square-wave output, additional line filtering may be necessary in the power line to the equipment, and a suitable line filter is tabulated in the parts list of figure 11.

Figure 12

MULTIBAND MOBILE WHIP USING HIGH-Q AIRWOUND COIL

Heavy base section provides support for adjustable loading coil. Antenna may be used over a range of about 15 kHz on 80 meters without retuning and correspondingly larger ranges on the higher frequency bands. Coil is mounted well clear of automobile body. Outer braid of coax line is grounded to bumper and to auto frame at base of antenna.

19-3 Antennas for Mobile Operation

The mobile antenna is the key to successful operation on any amateur band. Because of space limitations on the vehicle and the sweep of the vehicle body panels, the vertical whip antenna is the most popular mobile antenna, regardless of the band of operation. For h-f service, the whip takes the form of a flexible, tapered steel rod with a threaded base fitting.

Unless the whip is a resonant length (common only on the vhf, 6- and 10-meter bands) it is brought into resonance by the addition of a *loading coil* which makes up for the missing antenna length. The coil may be placed either at the base of the whip, or near the center. Over-all antenna efficiency is generally a function of the Q or circuit efficiency of the loading coil, and every effort should be made to design and use a high-Q coil, well removed from the body of the vehicle.

Antenna Mounts High-frequency whip antennas, because of their height, are usually mounted low on the vehicle, often on the rear bumper or fender as shown in figure 12. Chain or strap-type mounts are available; they clamp directly over the edges of the bumper without the need of drilling mounting holes in the vehicle. The antenna is held in position by an insulated adapter bolted to the top bracket of the mount. Sometimes a heavy spring is included in the mount to absorb the road shock.

The whip antenna must remain free and clear of the body of the vehicle. Use of a bumper mount on station wagons, trucks and vans is not recommended because the whip passes too close to the upper metal body panels of the vehicle and severe detuning of the antenna may result. In this situation, a shorter antenna mounted higher on the body or roof is recommended.

A ball mount and spring (figure 13) can be used to mount the whip antenna at an angle on the vehicle so that the antenna itself is in a vertical plane, regardless of the

Figure 13

ADJUSTABLE BASE MOUNT FOR MOBILE WHIP

Mount may be placed on automobile panel and then adjusted so that whip is vertical regardless of position of panel. Jumper wire inside spring ensures that inductance of spring does not become part of the antenna.

plane of the mount. Usual placement includes the rear deck, the side or top of the fender or (for short antennas) the top, flat portion of the roof. In the latter case, care must be taken to make sure the antenna does not strike overhead electrical wires and tree limbs.

The ball mount requires that a mounting hole be drilled in the skin of the vehicle on a relatively flat surface. Once the mount is in place, the whip is inserted in the socket and the rotary ball joint adjusted to align the whip in a vertical position.

Many amateurs hesitate to drill holes in their vehicle and are interested in an antenna mount that will not scar the body of the automobile. The trunk lip mount is a device that meets this need. The adjustable antenna mount is slipped beneath the edge of the trunk lid and bolted firmly to the groove of the car body. Enough clearance exists around the edge of most trunk lids to permit the user to bring a small coaxial cable (RG-58/U) through the gap and up to the antenna mount as shown in the illustration. Some trunk mounts fasten to the trunk lid as shown in figure 14.

A vhf whip may be clamped to the rain gutter of the vehicle by means of a gutter clamp. The mount is affixed to the outer rim of the gutter, taking care to be sure that the clamp breaks through the enamel coating of the gutter to make a good electrical contact to the body of the vehicle. Scraping off the paint at this point is a good idea. The mount is adjustable to permit placing the antenna in a vertical position.

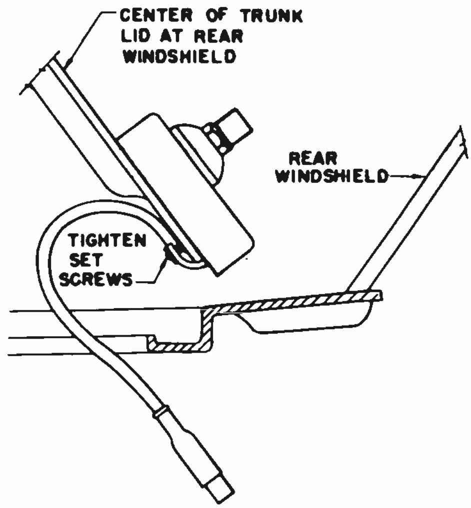


Figure 14

TRUNK-LID ANTENNA MOUNT

Antenna mount is bolted to underside of trunk lid so that auto body is not damaged by mounting holes.

Vhf Antennas In areas where vertical polarization is predominant, the vertical whip antenna is used for mobile operation. The most logical place for a vhf whip is at the center of the vehicle roof since this provides a relatively large ground-plane area and nearly omnidirectional coverage. The next best location is at or near the center of the trunk lid at the rear of the vehicle. Field-strength tests have shown that trunk-lid mounting of a 144-MHz whip antenna provides an omnidirectional pattern that is only 1 decibel less in signal strength than the same antenna in a roof-mount position.

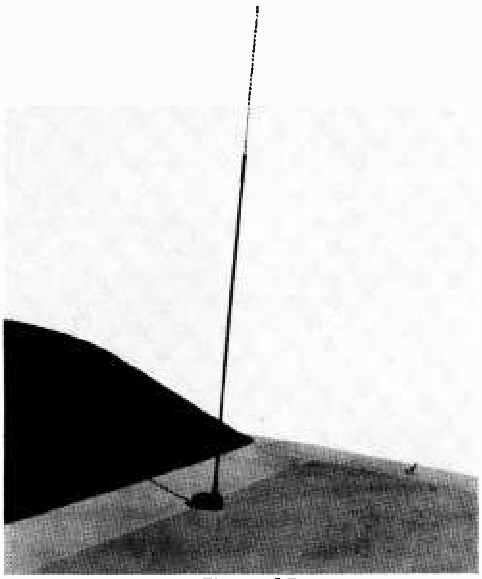


Figure 15

VHF EXTENDED WHIP EQUALS ROOF-MOUNTED GROUND PLANE

Five-eighths wave antenna mounted on rear trunk area of vehicle provides equivalent performance to quarter-wavelength ground plane mounted at center of vehicle roof.

A popular antenna for 50-MHz and 144-MHz operation is a 56-inch whip which operates as a ¼-wavelength radiator on the lower band and as a ¾-wavelength radiator on the higher band. A collapsible whip can be adjusted for minimum SWR on either band since the resonant points for each band are only a few inches apart.

A typical ⅝-wavelength whip for the 2-meter band is shown in figure 15. The whip is reduced in length to 47 inches and is base-loaded with a small coil which is mounted in the base assembly mount. Whip length is adjusted a quarter-inch at a time for lowest SWR on the transmission line to the antenna.

10-Meter Mobile Antennas The most popular mobile antenna for 10-meter operation is a rear-mounted whip approximately 8 feet long, fed with coaxial line. This is a highly satisfactory antenna, but a few remarks are in order on the subject of feed and coupling systems.

The feed-point resistance of a resonant quarter-wave rear-mounted whip is approximately 20 to 25 ohms. While the standing-wave ratio when using 50-ohm coaxial line will not be much greater than 2 to 1, it is nevertheless desirable to make the line to the transmitter exactly odd multiples of one-quarter wavelength long electrically at the center of the band. This procedure will minimize variations in loading over the band.

A more effective radiator and a better line match may be obtained by making the whip approximately 10½ feet long and feeding it with 75-ohm coax (such as RG-11/U) via a series capacitor, as shown in figure 16. The relay and series capacitor are mounted inside the trunk, as close to the antenna feedthrough or base-mount insulator as possible. The 10½-foot length applies to the over-all length from the tip of the whip to the point where the lead-in passes through the car body. The leads inside the car (connecting the coaxial cable, relay, series capacitor and antenna lead) should be as short as possible. The outer conductor of both coaxial cables should be grounded to the car body at the relay end with short, heavy conductors.

A 100-pf midget variable capacitor is suitable for C_1. The optimum setting for lowest SWR at the transmitter should be determined experimentally at the center of

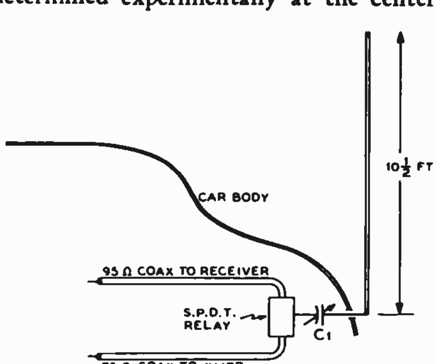


Figure 16

5/16-WAVE WHIP RADIATOR FOR 10 METERS

If a whip antenna is made slightly longer than one-quarter wave it acts as a slightly better radiator than the usual quarter-wave whip, and it can provide a better match to the antenna transmission line if the reactance is tuned out by a series capacitor close to the base of the antenna. Capacitor C_1 may be a 100-pf midget variable.

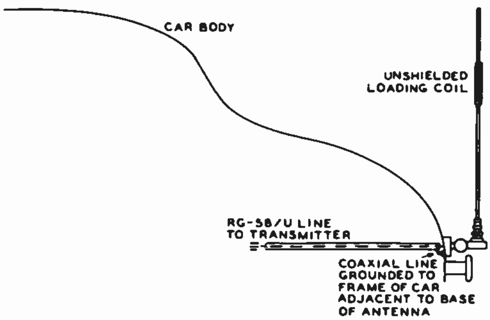


Figure 17

THE CENTER-LOADED WHIP ANTENNA

The center-loaded whip antenna, when provided with a tapped loading coil or a series of coils, may be used over a wide frequency range. The loading coil may be shorted for use of the antenna on the 10-meter band.

the band. This setting then will be satisfactory over the whole band.

If an all-band center-loaded mobile antenna is used, the loading coil at the center of the antenna may be shorted out for operation of the antenna on the 10-meter band. The usual type of center-loaded mobile antenna will be between 9 and 11 feet long, including the center-loading inductance which is shorted out. Hence such an antenna may be shortened to an electrical quarter wave for the 10-meter band by using a series capacitor as just discussed. If a pi network is used in the plate circuit of the output stage of the mobile transmitter, any reactance presented at the antenna terminals of the transmitter by the antenna may be tuned out with the pi network.

The All-Band Center-Loaded Mobile Antenna The great majority of mobile operation on the 14-MHz band and below is with center-loaded whip antennas. These antennas use an insulated bumper or body mount, with provision for coaxial feed from the base of the antenna to the transmitter, as shown in figure 17.

The center-loaded whip antenna must be tuned to obtain optimum operation on the desired frequency of operation. These antennas will operate at maximum efficiency over a range of perhaps 20 kHz on the 75-meter band, covering a somewhat wider range on the 40-meter band, and covering the whole 20-meter phone band. The procedure for tuning the antenna is as follows:

The antenna is installed, fully assembled, with a coaxial lead of RG-58/U from the base of the antenna to the place where the transmitter is installed. The rear deck of the car should be closed, and the car should be parked in a location as clear as possible of trees, buildings, and overhead power lines. Objects within 15 or 20 feet of the antenna can exert a considerable detuning effect on the antenna system due to its relatively high operating Q. The end of the coaxial cable which will plug into the transmitter is terminated in a link of 3 or 4 turns of wire. This link is then coupled to a grid-dip meter and the resonant frequency of the antenna determined by noting the frequency at which the grid current fluctuates. The coils furnished with the antennas normally are too large for the usual operating frequency, since it is much easier to remove turns than to add them. Turns then are removed, *one at a time*, until the antenna resonates at the desired frequency. If too many turns have been removed, a length of wire may be spliced on and soldered. Then, with a length of insulating tubing slipped over the soldered joint, turns may be added to lower the resonant frequency. Or, if the tapped type of coil is used, taps are changed until the proper number of turns for the desired operating frequency is found. This procedure is repeated for the different bands of operation.

Loaded Whip Antennas Short vertical antennas have low radiation resistance and high capacitive reactance and it is difficult to couple them to practical transmitter output circuitry without the necessity of resonating the antenna to the operating frequency. Experience and theory have shown that introduction of a loading coil near the center of the whip antenna can simultaneously tune the antenna to resonance and increase the radiation resistance (figure 17). Typically, an unloaded 8-foot whip has a radiation resistance of approximately 0.5 ohm at 3.9 MHz, with a capacitive input reactance of about 2000 ohms. Introduction of a high-Q center-loading coil can increase the radiation resistance to about one ohm and will cancel the large capacitive reactance of the antenna.

Ground loss resistance in the automobile and capacitance of the car body to ground have been measured to be about 10 ohms at 3.9 MHz. These radiation and loss resistances, plus the loss resistance of a typical loading coil may bring the input impedance of a typical 80-meter center-loaded whip to about 25 to 30 ohms at the resonant frequency. Over-all radiation efficiency is about five to ten percent and operational bandwidth (for a 3/1 SWR on the transmission line) is about 25 kHz when the antenna is properly matched.

The relatively low efficiency of the loaded whip antenna at the lower frequencies indicates that attention must be paid to all details of the antenna installation. The loading coil must be of the highest possible Q and all joints in the antenna system must be low resistance. To properly match the 25-ohm antenna load to a typical 50-ohm transmission line, the matching system of figure 18 may be used. The loaded whip antenna forms a portion of a network whose input impedance over a small frequency range is close to 50 ohms. The antenna is made a part of an equivalent parallel-resonant circuit in which the radiation resistance appears in series with the reactive branch of the circuit. The input impedance of such a circuit varies nearly inversely with respect to the radiation resistance of the antenna, thus the very low radiation resistance of the whip antenna may be transformed to a larger value which will match the impedance of the transmission line.

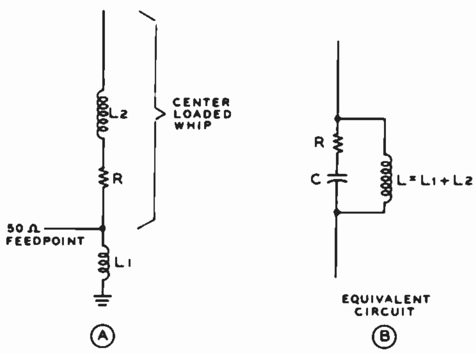


Figure 18

CENTER-LOADED WHIP ANTENNA

A—Center-loaded whip represents large loss resistance (R) which is inverse function of coil Q. High-Q coil (300 or better) provides minimum losses consistent with practical coil design. B—Equivalent circuit provides impedance match between whip antenna and 50-ohm feedpoint.

The radiation resistance of the whip antenna can be made to appear as a capacitive reactance at the feed point by shortening the antenna. In this case, this is done by slightly reducing the inductance of the center-loading coil. The inductive portion of the tuned network (L_1) consists of a small coil placed across the terminals of the antenna as shown in figure 18A. The LC ratio of antenna and matching coil determine the transformation ratio of the network when the LC product is parallel resonant at the operating frequency of the antenna.

In order to conveniently adjust this matching system, the radiation resistance of the loaded whip antenna should be known, otherwise it may be necessary to try various combinations of matching inductance and loading coil before a satisfactory value of resonant frequency SWR on the feedline is achieved. Methods of measuring the radia-

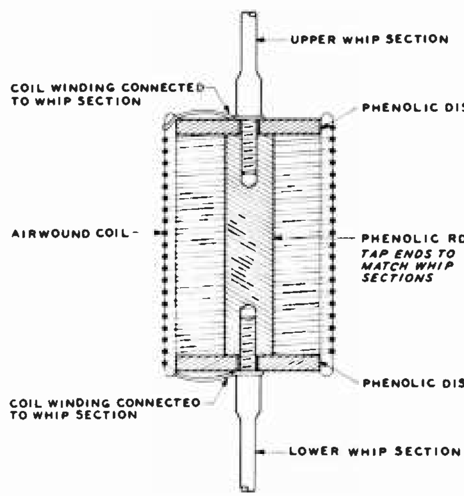


Figure 19

HIGH-Q MOBILE LOADING COIL

Efficient loading coil is assembled from section of air-wound coil stock (Air-Dux or B-W). 2½" diam coil is recommended. Approximate inductance for various bands, when used in center of 8-foot whip is: 160 meters, 700 μH; 80 meters, 150 μH; 40 meters, 40 μH; 20 meters, 9 μH; 15 meters, 2.5 μH. Complete antenna is grid-dipped to operating frequency and number of turns in coil adjusted for proper resonance.

tion resistance of the antenna are discussed in Chapter 29 of this Handbook.

Construction of a high-Q center-loading coil from available coil stock is shown in figure 19.

An SWR Meter for Mobile Use This simple reflectometer is designed to be used with mobile equipment over the 3- to 30-MHz range at power levels up to 500 watts. It may be placed in the 50-ohm coaxial transmission line to the antenna and mounted under the dash of the automobile to provide a constant check of transmitter power output and antenna operation. It is also useful for tuneup purposes, since the transmitter stages may be adjusted for maximum forward-power reading of the instrument. The circuit is bidirectional; that is, either terminal may be used for either input or output connection.

The SWR meter is constructed in an aluminum utility box measuring 4″ × 4″ × 2″ and the circuit is shown in figure 21.

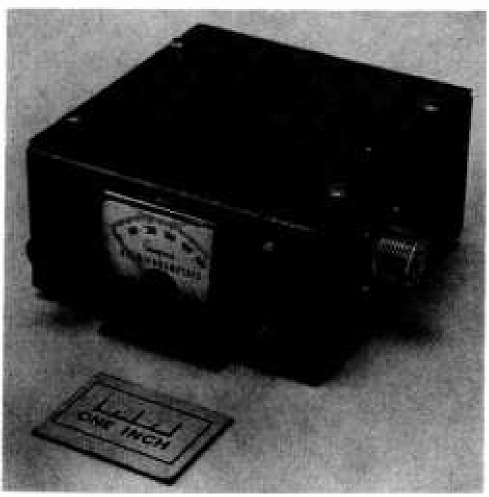


Figure 20

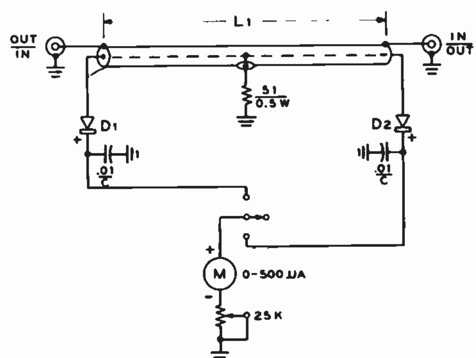
MINI-SWR METER FOR MOBILE EQUIPMENT

Inexpensive reflectometer is built in 4″ × 4″ × 2″ aluminum utility box and may be used over 3- to 30-MHz range at power levels up to 500 watts or so.

The heart of the device is a 4¾″ long pickup line made of the inner conductor of a length of RG-58A/U coaxial line and a piece of ¼-inch copper tubing, which makes a close slip fit over the polyethylene inner insulation of the line.

To assemble the pickup line, the outer jacket and braid are removed from a length of coaxial line. Before the line is passed within the tubing, the insulation is cut and removed at the center point, which is tinned. A small hole is drilled at the center of the copper-tubing section so that a connection may be made to the inner line. The line is passed through the tubing, and one lead of a 51-ohm, ½-watt composition resistor is soldered to the line at this point. The pickup line is then bent into a semicircle and the ends of the tubing are affixed to the coaxial connectors, as shown in figure 22.

Sensitivity of the SWR meter is controlled by the variable resistance in series with the meter. To check the instrument, power is fed through it to a matching dummy load and the meter switch set to read forward power. On reversal of the switch, the meter will read reflected power. In the case of a good load match, the reflected reading will be near zero, increasing in value with the degree of mismatch of the load.



D_1, D_2—1N34A
L_1—See text
M—0-500 μA, d-c, Simpson 1212

Figure 21

SCHEMATIC, MINI-SWR METER

19-4 Construction of Mobile Equipment

The following measures are recommended for the construction of mobile equipment, either transmitting or receiving, to ensure trouble-free operation over long periods:

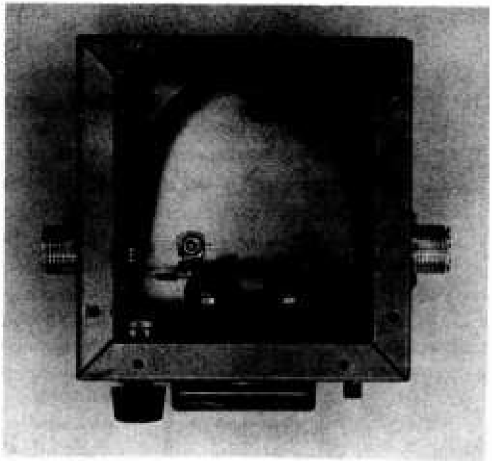


Figure 22

INTERIOR, MINI-SWR METER

Pickup line is bent in semicircle and tubing is soldered to loops of wire which connect to center pin of SO-239 coaxial receptacles. Center conductor of line is attached to diodes D_1, D_2.

Use only a stiff, heavy chassis unless the chassis is quite small.

Use lock washers or lock nuts when mounting components by means of screws.

Use stranded hookup wire except where r-f considerations make it inadvisable (such as for instance the plate tank circuit leads in a vhf amplifier). Lace and tie leads wherever necessary to keep them from vibrating or flopping around.

Unless provided with gear drive, tuning capacitors in the large sizes will require a rotor lock.

The larger-size carbon resistors and mica capacitors should not be supported from tube socket pins, particularly from miniature sockets. Use tie points and keep the resistor and capacitor "pigtails" short.

Generally speaking, rubber shock mounts are unnecessary or even undesirable with passenger car installations, or at least with full-size passenger cars. The springing is sufficiently "soft" that well constructed radio equipment can be bolted directly to the vehicle without damage from shock or vibration. Unless shock mounting is properly engineered as to the stiffness and placement of the shock mounts, mechanical-resonance "amplification" effects may actually cause the equipment to be shaken more than if the equipment were bolted directly to the vehicle.

To facilitate servicing of mobile equipment, all interconnecting cables between units should be provided with separable connectors on at least one end.

Finally, it should be remembered that the interior of the vehicle can get very hot when it is left in the sun for a period of time. Excessive heat may possibly damage solid-state devices and some crystal microphones. Try and place the mobile equipment where it will not be exposed to such heat.

Control Circuits The send-receive control circuits of a mobile installation are dictated by the design of the equipment, and therefore will be left to the ingenuity of the reader. However, a few generalizations and suggestions are in order.

Do not attempt to control too many relays, particularly heavy-duty relays with large coils, by means of an ordinary push-to-talk switch on a microphone. These contacts are not designed for heavy work, and the inductive "kick" will cause more arcing than the contacts on the microphone switch are designed to handle. It is better to actuate a single relay with the push-to-talk switch and then control all other relays, including the heavy-duty contactor for the transistor power pack with this relay.

A recommended general control circuit, where one side of the main control relay is connected to the hot 12-volt circuit, but all

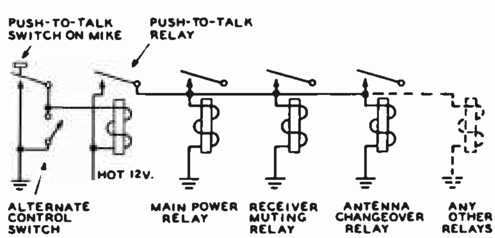


Figure 23

RELAY CONTROL CIRCUIT

Simplified schematic of the recommended relay control circuit for mobile transmitters. The relatively small push-to-talk relay is controlled by the button on the microphone or the communications switch. Then one of the contacts on this relay controls the other relays of the transmitter; one side of the coil of all the additional relays controlled should be grounded.

other relays have one side connected to the ground, is illustrated in figure 23.

When purchasing relays keep in mind that the current rating of the contacts is not a fixed value, but depends on (1) the voltage, (2) whether it is a.c. or d.c., and (3) whether the circuit is purely resistive or is inductive. If in doubt, refer to the manufacturer's recommendations. Also keep in mind that a dynamotor (if you use one) presents almost a dead short until the armature starts turning, and the starting relay should be rated at considerably more than the normal dynamotor current.

Microphones and Circuits The standardized connections for a majority of hand-held microphones provided with push-to-talk switch are shown in figure 24.

There is an increasing tendency among mobile operators toward the use of microphones having better frequency and distortion characteristics than the single-button carbon type. The high-impedance *dynamic* type is probably the most popular with the *ceramic*-crystal type next in popularity. The conventional crystal type is not suitable for mobile use since the crystal unit will be destroyed by the high temperatures which can be reached in a closed car parked in the sun in the summer time.

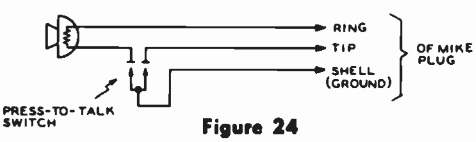


Figure 24

STANDARD CONNECTIONS FOR THE PUSH-TO-TALK SWITCH ON A HAND-HELD SINGLE-BUTTON CARBON MICROPHONE

The use of low-level microphones in mobile service requires careful attention to the elimination of common-ground circuits in the microphone lead. The ground connection for the shielded cable which runs from the transmitter to the microphone should be made at only one point, preferably directly adjacent to the input of the first tube or transistor in the speech amplifier. The use of a low-level microphone usually will require the addition of two speech stages, but these stages will take only a milliampere or two of current.

19-5 Vehicular Noise Suppression

Satisfactory reception on frequencies above the broadcast band usually requires greater attention to noise-suppression measures. The required measures vary with the particular vehicle and the frequency range involved.

Most of the various types of noise that are present in a vehicle may be broken down into the following main categories:

(1) Ignition noise.
(2) Wheel static (tire static, brake static, and intermittent ground via front wheel bearings).
(3) "Hash" from voltage-regulator contacts.
(4) "Whine" from generator commutator segment make and break.
(5) Static from scraping connections between various parts of the car.

It is best to thoroughly suppress ignition noise in your car, even though ignition noise from passing vehicles make the use of a noise limiter mandatory. However, the limiter should not be given too much work to do, because at high engine speeds a noisy ignition system will tend to mask weak signals, even though with the limiter working, ignition "pops" may appear to be completely eliminated.

Another reason for good ignition suppression at the source is that strong ignition pulses contain enough energy, when integrated, to block the agc circuit of the receiver, causing the gain to drop whenever the engine is speeded up. Since the agc circuits of the receiver obtain no benefit from a noise clipper, it is important that ignition noise be suppressed enough at the source that the agc circuits will not be affected even when the engine is running at high speed.

Ignition Noise The following procedure should be found adequate for reducing the ignition noise of practically any passenger car to a level which the clipper can handle satisfactorily at any engine speed at any frequency from 500 kHz to 148 MHz. Some of the measures may already have been taken when the auto receiver was installed.

First either install a spark-plug suppressor on each plug, or else substitute resistor plugs. The latter are more effective than suppressors and on some cars ignition noise is reduced to a satisfactory level simply by installing them. However, they may not do an adequate job alone after they have been in use for a while, and it is a good idea to take the following additional measures.

Check all high-tension connections for gaps, particularly the "pinch-fit" terminal connectors widely used. Replace old high-tension wiring that may have become leaky. Complete substitution of the ignition wiring with a commercial shielded ignition system is recommended in case of severe interference.

Check to see if any of the high-tension wiring is cabled with low-tension wiring, or run in the same conduit. If so, reroute the low-tension wiring to provide as much separation as practical.

Bypass to ground the 12-volt wire from the ignition switch at each end with a 0.1-μfd molded-case paper capacitor in parallel with a .001-μfd mica or ceramic, using the shortest possible leads.

Check to see that the hood makes good ground contact to the car body at several points. Special grounding contactors are available for attachment to the hood lacings on cars that otherwise would present a grounding problem.

If the high-tension coil is mounted on the dash, it may be necessary to shield the high-tension wire as far as the bulkhead, unless it already is shielded with armored conduit.

Wheel Static Wheel static is either static electricity generated by rotation of the tires and brake drums, or is noise generated by poor contact between the front wheels and the axles (due to the grease in the bearings). The latter type of noise seldom is caused by the rear wheels, but tire static may of course be generated by all four tires.

Wheel static can be eliminated by insertion of grounding springs under the front hub caps, and by inserting "tire powder" in all inner tubes. Both items are available at radio parts stores and from most auto radio dealers.

Voltage-Regulator "Hash" Certain voltage regulators generate an objectionable amount of "hash" at the higher frequencies, particularly in the vhf range. A large bypass capacitor will affect the operation of the regulator and possibly damage the points. A small bypass can be used, however, without causing trouble. A 0.001-μfd mica capacitor placed from the field terminal of the regulator to ground with the shortest possible leads often will produce sufficient improvement. If not, a choke consisting of about 60 turns of No. 18 d.c.c. wound on a ¾-inch form can be added. This should be placed at the regulator terminal, and the 0.001-μfd bypass placed from the generator side of the choke to ground.

Generator "Whine" Generator "whine" often can be satisfactorily suppressed from 550 kHz to 148 MHz simply by bypassing the armature terminal to ground with a special "auto-radio" capacitor of 0.25 or 0.5 μfd in parallel with a 0.001-μfd mica or ceramic capacitor. The former usually is placed on the generator when an auto radio is installed, but must be augmented by a mica or ceramic capacitor with short leads in order to be effective at the higher frequencies as well as the broadcast band.

When more drastic measures are required, special filters can be obtained which are designed for the purpose. These are recommended for stubborn cases when a wide frequency range is involved. For reception over only a comparatively narrow band of frequencies, such as the 10-meter amateur band, a highly effective filter can be improvised by connecting a resonant choke between the previously described parallel bypass capacitors and the generator armature terminal. This may consist of 11 turns of No. 10 enameled wire wound on a one-inch form and shunted with an adjustable 30-pf trimmer capacitor to permit resonating the combination to the center of the ten-meter band.

When generator "whine" shows up after once being satisfactorily suppressed, the condition of the brushes and commutator should be checked. Unless a bypass capacitor has opened up, excessive "whine" usually indicates that the brushes or commutator are in need of attention in order to prevent damage to the generator.

Body Static Loose linkages in body or frame joints anywhere in the car are potential static producers when the car is in motion, particularly over a rough road. Locating the source of such noise is difficult, and the simplest procedure is to give the car a thorough tightening up in the hope that the offending poor contacts will be caught up by the procedure. The use of braided bonding straps between the various sections of the body of the car also may prove helpful.

Miscellaneous There are several other potential noise sources on a passenger vehicle, but they do not necessarily give trouble and therefore require attention only in some cases.

The heat, oil pressure, and gas gauges can cause a rasping or scraping noise. The gas gauge is the most likely offender. It will cause trouble only when the car is rocked or is in motion. The gauge units and panel indicators should both be bypassed with the 0.1-μfd paper and 0.001-μfd mica or ceramic capacitor combination previously described.

At high car speeds under certain atmospheric conditions, corona static may be encountered unless means are taken to prevent it. The receiving-type auto whips which employ a plastic ball tip are so provided in order to minimize this type of noise, which is simply a discharge of the frictional static built up on the car. A whip which ends in a relatively sharp metal point makes an ideal discharge point for the static charge, and will cause corona trouble at a much lower voltage than if the tip were hooded with insulation. A piece of *Vinylite* sleeving slipped over the top portion of the whip and wrapped tightly with heavy thread will prevent this type of static discharge under practically all conditions. An alternative arrangement is to wrap the top portion of the whip with *Scotch* brand electrical type.

Generally speaking it is undesirable from the standpoint of engine performance to use both spark-plug suppressors *and* a distributor suppressor. Unless the distributor rotor clearance is excessive, noise caused by sparking of the distributor rotor will not be so bad but that it can be handled satisfactorily by a noise limiter. If not, it is preferable to shield the "hot" lead between ignition coil and distributor rather than use a distributor suppressor.

In many cases the control rods, speedometer cable, etc., will pick up high-tension noise under the hood and conduct it up under the dash where it causes trouble. If so, all control rods and cables should be bonded to the fire wall (bulkhead) where they pass through, using a short piece of heavy flexible braid of the type used for shielding.

In some cases it may be necessary to bond the engine to the frame at each rubber engine mount in a similar manner. If a rear-mounted whip is employed, the exhaust tail pipe also should be bonded to the frame if supported by rubber mounts.

Locating Noise Sources Determining the source of certain types of noise is made difficult when several things are contributing to the noise, because elimination of one source often will make little or no apparent difference in the total noise. The following procedure will help to isolate and identify various types of noise.

Ignition noise will be present only when the ignition is on, even though the engine is turning over.

Generator noise will be present when the motor is turning over, regardless of whether the ignition switch is on. Slipping the drive belt off will kill it.

Gauge noise usually will be present only when the ignition switch is on or in the "left" position provided on some cars.

Wheel static, when present, will persist when the car clutch is disengaged and the ignition switch turned off, with the car coasting.

Body noise will be noticeably worse on a bumpy road than on a smooth road, particularly at low speeds.

19-6 A Portable Amateur Band Receiver

The availability of low priced solid-state devices and integrated circuits makes feasible the design of a compact, completely solid-state amateur band receiver for c-w and SSB reception that performs as well as or better than an equivalent receiver using conventional vacuum tubes. The advanced receiver described in this section (figure 25) is completely solid state, making use of improved MOSFET and IC devices, and covers the amateur bands between 80 and 10 meters in 500-kHz segments. The design goal was to produce a compact receiver of top-notch performance, but one not so small as to be difficult to assemble and wire, or to operate. For easy duplication, all components used in construction of the receiver are "off-the-shelf" items readily obtainable from the larger radio parts distributors. The receiver may be run from a battery power supply or from an a-c supply so it is well suited for either portable or fixed service. This receiver was designed and built by VE3GFN.

The Receiver Circuit A block diagram of the complete solid-state receiver is shown in figure 26. The circuit is basically a four-band crystal-controlled front-end converter, followed by a tunable i-f receiver which covers the fifth band (80 meters). The bandswitching front-end, or converter, is shown in detail in figure 27. This separate assembly covers the amateur bands between 7 MHz and 29 MHz, with allowance in design for out-of-band coverage, as well as coverage as high as 30 MHz, or more. Using a *Motorola* MFE-3006 high-frequency MOSFET device in the tunable r-f amplifier stages results in high gain and good circuit stability. The r-f amplifier circuitry does not require neutralization, while permitting AGC (automatic-gain-control). voltage to be applied to the front end, a feature very necessary in solid-state receivers. The dual-gate feature of the MFE-3006 allows a separation of these functions, the incoming signal being applied to gate 1 of the MOSFET and the agc control voltage to gate 2 of the device.

Laboratory measurements taken on the receiver provide the following data on performance. *Sensitivity*: Less than 1 microvolt for a 10-decibel signal-plus-noise to noise

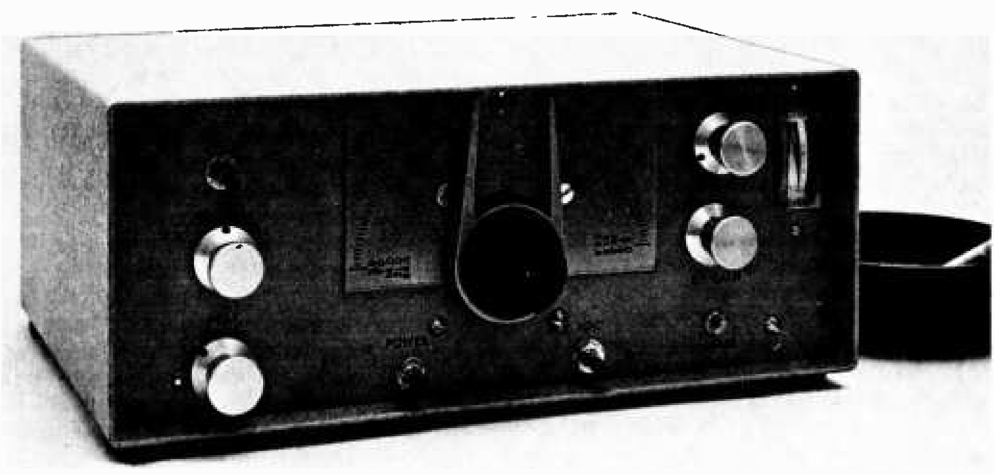


Figure 25

A SOLID-STATE AMATEUR BAND RECEIVER

This advanced communication receiver covers all amateur bands between 80 and 10 meters. It uses 3 MOSFETs, 5 FETs, 5 transistors, 2 ICs, and 3 hot carrier diodes. Measuring only 10" × 4" (panel size) and 7" deep, the solid-state receiver provides excellent reception of SSB and c-w signals, combined with exceptional strong signal overload capability.

Panel controls (l. to r.) are: Sideband selector switch (S_1); bandswitch; peak preselector (C_1) power switch (S_2); AGC switch (S_3); phone jack (J_1) insulated from the panel; r-f gain potentiometer (R_2); audio gain control (R_1); and signal-strength meter (M).

The main tuning dial is calibrated every 100 kHz, with 5-kHz markers and is made of a panel mask (figure 34). The pointer window is cut from a piece of ¼-inch aluminum stock and has a plastic window insert epoxied to the underside of the frame. The cursor line is scratched on the rear of the window.

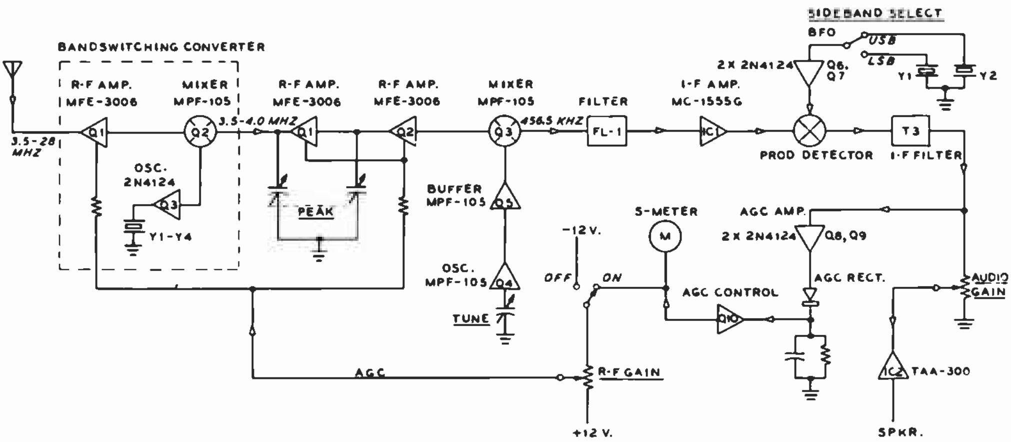


Figure 26

BLOCK DIAGRAM OF THE SOLID-STATE COMMUNICATION RECEIVER

The main portion of the receiver covers the 80-meter band (3.5—4.0 MHz) and serves as an i-f section for a bandswitching converter covering the 40-, 20-, 15-, and 10-meter bands in 500-kHz segments. The high-frequency converter unit is crystal controlled and the low-frequency variable oscillator in the 80-meter section is not switched, permitting a high degree of electrical and mechanical stability to be achieved.

I-f gain is provided by an integrated circuit module (MC-1555G) and suitable SSB selectivity is achieved by a mechanical filter. Audio agc is provided for the various r-f stages and front-end gain may be separately controlled, if desired. The complete schematic of the receiver is given in figures 27 and 29.

ratio on all bands. *Image ratio*: Better than 60 decibels on all bands, and as high as 80 decibels. *Drift*: Less than 100 Hz per hour at receiver temperature of 70°F. *Spurious Responses*: Oscillator harmonics noted at 7.0 MHz and 21.250 MHz.

The R-F Section—The tuned circuits in the high-frequency portion of the receiver are basically 20-meter circuits, which are made resonant in the other high-frequency bands by means of appropriate shunt impedances brought into the circuit by the bandswitch. For 40-meter operation, the basic tuned circuit is padded to a lower resonant frequency by means of capacitor C_1 (figure 28). For 15- and 10-meter operation, the inductance of the tuned circuit is shunted by parallel inductors (L_2 and L_3) thus effectively raising the resonant frequency of the new circuit formed by the auxiliary inductors. These tuned circuits are designed to have an essentially flat response over 500 kHz of the band in use, making a peaking control unnecessary. The 10-meter tuned circuits can be adjusted to pass any 500-kHz segment of the 10-meter band, allowing the receiver to cover the complete band, by the proper choice of local-oscillator conversion crystal and auxiliary inductor tuning.

Maximum gain is obtained from the MOSFETs in the r-f amplifier stages when gate 2 has +12 volts applied to it; however, this amount of gain has a tendency to overload the i-f system on any strong signal. Hence, provision has been made in the design of the agc system to limit the positive swing of the front-end agc input, eliminating this problem.

The Mixer-Oscillator—A 2N5459 (MPF-105) FET is used as a common-source mixer with local oscillator and received signals applied to the gate element. The crystal-controlled local oscillator is capacitively coupled to the gate and the incoming signal is inductively coupled through transformer L_4. The converter oscillator employs a 2N4124 bipolar transistor and uses an r-f choke as a broadband collector load on the lower frequencies (RFC_3). Series-connected parallel-tuned circuits provide properly selective collector loads on the two higher-frequency bands. These circuits exhibit little effect except when excited by the crystal frequency

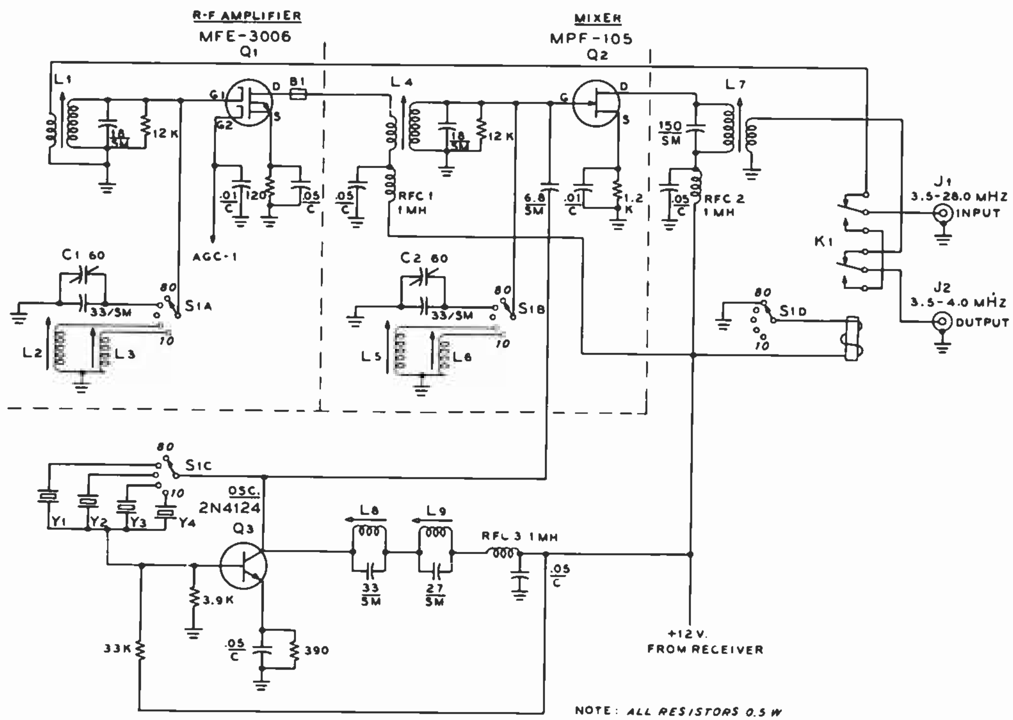


Figure 27

CONVERTER PORTION OF COMMUNICATION RECEIVER

B_1—Ceramic bead (Ferroxcube K5-001-03B or Stackpole 7D)
C_1, C_2—10 to 60-pf piston capacitors (Voltronics TM-60C, or equiv.)
J_1, J_2—Type BNC receptacles, UG-657/U
K_1—Dpdt relay, crystal-can style, 12-volt coil (Potter-Brumfield 5C-11DB or equiv.)
L_1, L_4—24 turns #32 enameled wire, closewound on ¼" diameter form. Approx. 4 μH (Q = 50). Use J. W. Miller 4500-2 (red) form, powdered iron core. Link winding is 5 turns #42 e. around "cold" end of coil
L_2, L_5—(15 meters). 20 turns #32 e., closewound on ¼" diam. form. Approx. 3.4 μH. J. W. Miller 4500-3 (green) form, powdered-iron core
L_3, L_6—(10 meters). 11 turns #32 e., as L_2. Approx. 1.4 μH
L_7—40 turns #32 e., closewound on ¼" diameter form. J. W. Miller 4500-3 (green) form, powered iron core. Tunes to 3.9 MHz. Link winding is 10 turns #32 e. around "cold" end of coil
L_8—10 turns #32 e., closewound on ¼" diameter form. J. W. Miller 4500-2 (red) form, powdered iron core. Resonates to 24.5 MHz.
L_9—15 turns #32 e., as L_8. Resonates to 17.5 MHz
RFC_{1-3}—1 millihenry. J. W. Miller 9350-44 or equiv.
S_{1A-D}—4 pole, 6 position ceramic switch. Centralab 2021 or equiv.
Y_1—3.500 MHz crystal, HC-6/U holder
Y_2—10.500 MHz, as Y_1
Y_3—17.500 MHz, as Y_1
Y_4—24.500 MHz, as Y_1

to which they are resonant. The use of tuned collector-load circuits is particularly necessary above 20 MHz or so where the common practice is to employ overtone crystals.

The schematic of the tunable 80-meter stages and low-frequency i-f section is shown in figure 29. The front end of this section of the receiver has two stages of r-f amplification using MFE-3006 MOSFETs to provide needed sensitivity and image rejection. The tuned circuits for these stages are adjustable from the panel of the receiver and provide a preselector function (*PEAK*). Good electrical isolation between the stages is necessary as the gain of this cascade

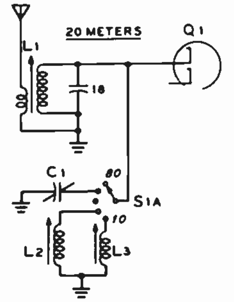


Figure 28

SIMPLIFIED R-F SWITCHING CIRCUIT

The external antenna is coupled to a resonant LC circuit for 20-meter reception. When the bandswitch is changed to 40 meters, the 20-meter circuit is padded to the lower frequency by the addition of piston capacitor C_1, placed in the circuit by switch section S_{1A}. On 15 meters, the inductance of 20-meter coil L_1 is decreased by the added shunting action of coil L_2. On 10 meters, coil L_3 is switched in the circuit. Alignment of the tuned circuit must first be done on 20 meters before the 15- and 20-meter bands are adjusted.

circuitry is considerable. To avoid cross modulation and overload, these stages are followed by an MPF-105 FET mixer (Q_3), using a common-gate circuit proven to be tolerant of high input levels.

The intermediate frequency of the receiver is 455 kHz and the frequency response of the i-f system is largely established by a mechanical filter having a passband (2.1 kHz) suitable for SSB reception. Intermediate-frequency gain is provided by a *Motorola* integrated circuit element (MC-1553G), matched to the mechanical filter by a simple transformer and resistance network.

The Product Detector—A product detector is used to provide good linearity, low insertion loss, and a minimum of beat-oscillator leakthrough into the audio system. One-half of a diode quad is used for the detector, employing 1N2970 hot-carrier diodes, resulting in excellent circuit balance. Closely matched 1K load resistors ensure minimum leakthrough while a simple low-pass audio filter (T_3) placed after the product detector attenuates all residual high-frequency products. The filter is a parallel-tuned circuit at 455 kHz offering high impedance to the intermediate frequency, and a low impedance to audio frequencies.

The local oscillator (bfo) consists of separate crystal-controlled oscillators with the outputs selected by switch S_2, feeding the input of the product detector through transformer T_2. A switch on the panel of the receiver (*SIDEBAND SELECT*) turns on one oscillator or the other for upper- or lower-sideband reception. The specified oscillator crystals should be as close to the target frequency as possible, since reduced detector output will result if one or the other of the crystals is misplaced on the slope of the filter passband. Product-detector attenuation is only about 6 decibels, which provides an audio output of nearly 10 millivolts with a 20-millivolt peak i-f signal input. Linearity of the i-f circuit and detector stages is excellent, input signals up to 300 millivolts or so being attained before distortion products in the audio signal are evident to the ear.

The audio system is a second integrated-circuit package (TAA-300) delivering almost a watt of audio power with a 10-millivolt driving signal. Speakers of 3 to 30 ohms impedance may be used, and the receiver will drive an efficient 10-inch diameter speaker with impressive results. A jack is provided on the panel for use with low-impedance earphones.

The AGC System—The agc network is novel in that the agc lines swing from positive to negative potential with increasing input signal level (figure 24). The three control lines are terminated at the arm of the *R-F GAIN* control potentiometer (R_2). One end of the potentiometer (max) is connected to the +12-volt supply line, and the other end (min) to about −3 volts when the agc switch (S_2) is off. When agc is on, the control is switched to the drain circuit of an agc control FET (Q_{10}). With no input signal, the gate of the control FET is near zero potential and the FET conducts, placing the negative end of the r-f gain control potentiometer close to ground potential. The agc lines, therefore, are at some positive potential between ground and +12 volts, depending on the setting of the potentiometer, allowing maximum receiver gain to be established, if desired. When a higher input signal level requires reduced front-end receiver gain, rectified audio of a positive polarity from the agc amplifiers (Q_8, Q_9) is applied to the gate of the control FET,

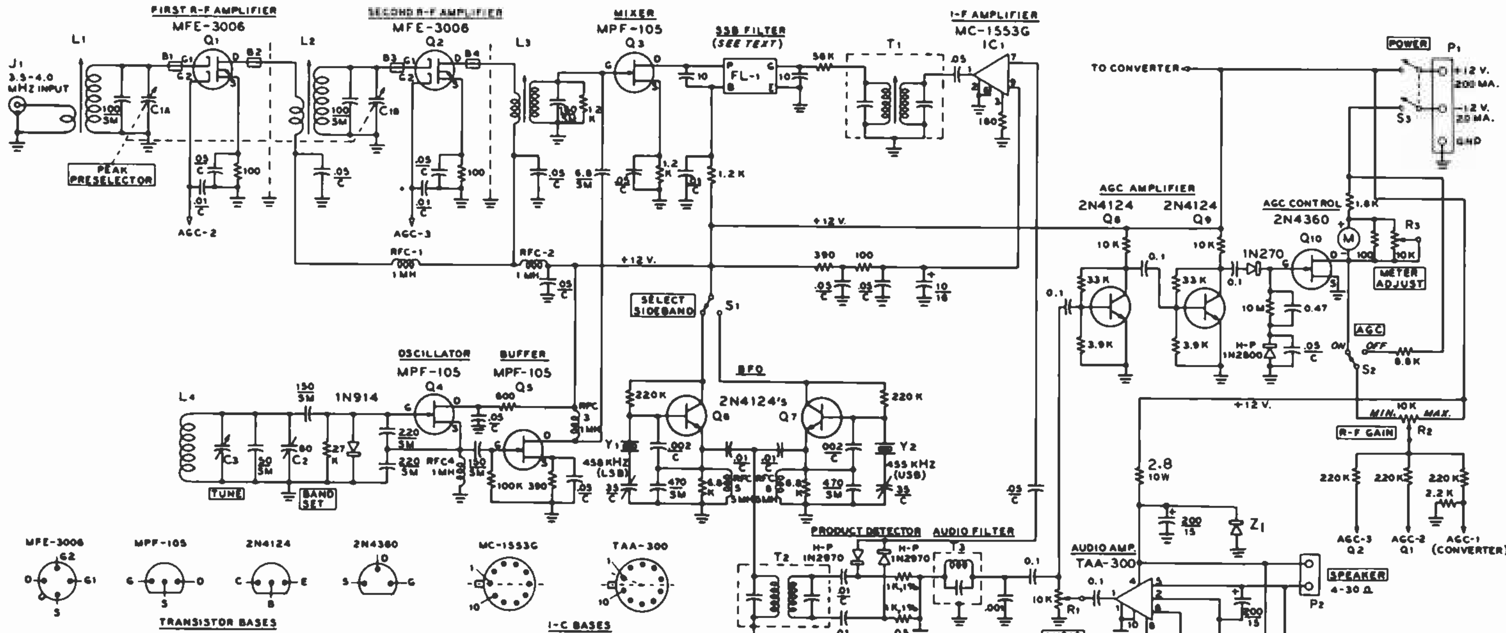


Figure 29

SCHEMATIC OF SOLID-STATE RECEIVER

B_1-B_4—Ceramic bead. Ferroxcube K5-001-003B or Stackpole 7D

C_{1A-B}—7-110 pfd per section. Miniature two section broadcast-type mica compression capacitor. Mitsumi PVC-2Z or equiv.

C_2—10- to 60-pf piston capacitor. Voltronics TM-60C or equiv.

C_3—8- to 45-pfd air capacitor. Jackson Bros. 804-50 or equiv. (Obtainable from: M. Swedgal, 258 Broadway, New York, N. Y. 10007).

FL_1—Mechanical Filter, 455-kHz center frequency, 2.1-kHz bandpass. Toyo 455-2.4C. Collins Radio Co. amateur-type filter may be used by substituting 150 pfd variable mica compression capacitors for 10 pf resonating capacitors. Also, Kokusai MF-455-15 mechanical filter may be used. See filter data for full application information

IC_1—Integrated-circuit module. Motorola MC-1553G or equiv.

IC_2—Integrated-circuit module. Phillips/Amperex TAA-300 or equiv.

L_1-L_3—40 turns #32 e. wire on 1/4" diameter form. Approx 11 μH. J. W. Miller 4500-2 (red) form, powdered iron core. Link is 10 turns #32 e. on "cold" end

L_4—40 turns #32 e., as L_1. Approx. 15 μH (see text)

T_1-T_3—Double-tuned miniature i-f transformer, 455 kHz. Armaco TR-104, or J. W. Miller 8807 or equiv.

Y_1, Y_2—Sideband-selection crystals to match filter characteristics. Type HC-6/U

Z_1—10 volt, 7 watt zener diode on 1½" square heat sink

1N2800, 1N2970—Hewlett Packard hot-carrier diodes

Teflon terminals—Sealectro FT-SM1, or equiv.

Cabinet—Hammond 1426-G (10" × 7" × 4") or equiv.

Meter—0-200 d-c microammeter

Note: All resistors ¼-watt unless otherwise specified.

reducing its conduction. Acordingly, the drain element of the FET drops toward −12 volts, taking the agc lines along with it, thus reducing front-end gain of the receiver.

The gate element of the control FET is connected to an RC circuit having a long time constant, which prevents gate voltage from changing too rapidly between c-w characteristics or between spoken syllables of an SSB signal. This circuit is designed to charge quickly when the receiver power is first applied, so that front-end gain is minimum. A diode across a portion of the time-constant circuit leaks off this charge in less than a minute, and the action can be observed on the S-meter when the receiver is first turned on.

A signal-strength meter is incorporated as part of of the agc system. The meter is connected so as to measure the current drawn by the control FET. The *METER-ADJUST* control (R_3) is set so the meter indicates full-scale current when the antenna input terminals are grounded. In operation, the *R-F GAIN* control (R_2) is set so that a small deflection of the meter (toward zero current) takes place with antenna connected but without signal input. At this point, the agc system will control receiver front-end gain in the proper manner, between near cutoff and maximum usable gain.

Power and Switching Circuits—The receiver is operated from a +12-volt 200-ma supply. In addition, −12 volts is required for agc action. The drain of the −12 volt section is only 20 milliamperes and series connected "penlite" cells may be incorporated in the receiver, if desired, for this function.

The converter portion of the receiver is switched in and out by means of a small crystal-can relay (K_1, figure 27) operated by the bandswitch. The relay is normally unenergized in all band positions except 80 meters. On this band, the relay removes the converter from the circuit and bypasses the antenna connections around the converter portion of the receiver.

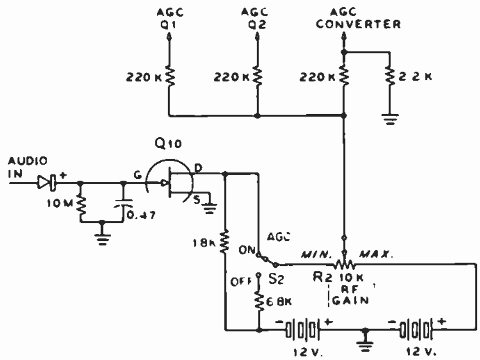


Figure 30

SIMPLIFIED AUDIO-CONTROLLED AGC SYSTEM

The three agc lines (Q_1, Q_2 and converter) are terminated at the arm of r-f gain control R_2. When agc switch S_2 is off, control voltage may be varied between +12 and −3 volts. When the agc system is on, control is switched to the drain circuit of FET Q_{10}. Agc voltage is now proportional to the audio input signal, varying between zero and +12 volts under normal conditions. A strong signal will drive the agc towards −12 volts, sharply reducing receiver gain. Maximum gain is controlled by the potentiometer.

Receiver Construction A multiband receiver such as this is a complex device and its construction should only be undertaken by a person familiar with solid-state devices in general and MOSFETs in particular, and who has built and aligned equipment approaching this complexity.

The solid-state receiver is built on a chassis within a wrap-around metal cabinet measuring 10″ × 7″ × 4″. The cabinet assembly specified comes complete with panel, chassis, and rubber mounting feet. Other cabinets of the same general configuration, of course, may be used.

General receiver assembly may be seen in the photographs and drawings. The high-frequency converter covering 40 through 10 meters is the most complex assembly and the most compact (figure 31). This unit is built in an aluminum box measuring 4″ × 2″ × 2¾″ and is mounted to the left rear of the main chassis. The converter bandswitch (S_1) is panel driven by means of an extension shaft as seen in the top-view photograph. Power and control leads are brought out through miniature feedthrough insulators mounted on the side of the box.

The variable-frequency oscillator is a second subassembly built within an aluminum box measuring 3¼″ × 2⅛″ × 1⅝″. The tuning capacitor used (C_3) is a high-

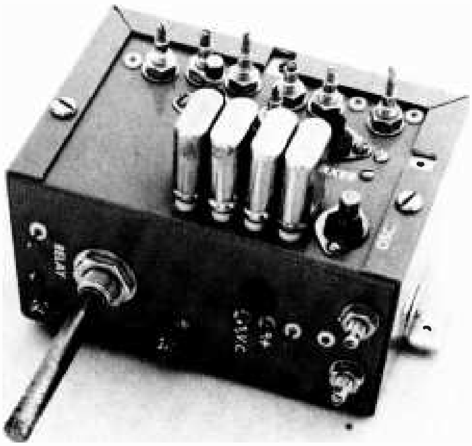


Figure 31

OBLIQUE VIEW OF CONVERTER UNIT

The converter section of the solid-state communications receiver covers the amateur bands between 80 and 10 meters and has an i-f output of 80 meters. The unit is built in a small aluminum box (4" × 2" × 2¾") with the major components mounted on the inner, U-shaped box section.

Across the rear of the assembly are the slug tuned r-f coils (l. to r.): 20-, 15-, and 10-meter coils. The 15- and 10-meter mixer coils are immediately to the right. In the right-hand corner of the box is the mixer output coil (L_7).

Along the center line of the converter unit are (l. to r.): The MFE-3006 r-f amplifier socket, the 20-meter mixer coil, and the MPF-105 mixer socket. At the front of the unit are the conversion crystals (l. to r.): 3.5 MHz, 10.5 MHz, 17.5 MHz, and 24.5 MHz. To the right of the crystals is the 2N4124 oscillator socket. Along the front section of the assembly are (l. to r.): the relay feedthrough terminal and piston capacitor C_1, bandswitch S_1, piston capacitor C_2, agc and voltage feedthrough terminals, and (at the extreme right) oscillator collector coils L_8 and L_9.

quality unit having full ball-race bearings front and back and a controlled torque. This unit provides minimum drag on the geared dial. The i-f mechanical filter is mounted to the left of the vfo assembly, with the receiver r-f stages and mixer to the left. Both the vfo and the high-frequency converter sections are built as separate units and may be tested and aligned before installation on the main receiver chassis.

The first step in construction of the solid-state receiver is to lay out the chassis, panel, tuning dial, and other major components in a "mockup" assembly to ensure that the receiver will go together without a physical conflict between the components. Figure 32 shows placement of the converter and oscillator assemblies and the i-f filter. The exact location of the vfo box behind the panel and the height of the main tuning capacitor on the side of the box are determined by the position of the tuning dial on the main panel. It is suggested that a trial panel be cut from heavy cardboard and used to support the main dial and assembly so that vfo placement may be checked before any holes are cut in the aluminum panel or the chassis. The panel is held in place by means of the various hexagonal nuts on the controls and the lower lip of the aluminum chassis is cut out to pass the dial mechanism, as shown in figure 33. Placement of the remaining components is not particularly critical, and may be done from a study of the photographs. Use of a paper template for drilling the chassis is recommended.

Receiver Wiring The receiver should be wired in an orderly manner, a stage at a time. To reduce r-f ground currents, all grounds for a single stage should be returned to that stage, preferably to a common ground point at or near the transistor socket. The gate, source, and drain bypass capacitors, for example, can all be returned to a common ground point near the transistor socket, components being grouped about the socket wherever possible, and not "stacked" above the socket, so that the latter can be reached for voltage measurements.

It is suggested that the r-f stages of the main receiver section be wired first, followed by the oscillator assembly, and then the product detector and the audio stage. The agc system, S-meter, and power wiring may be done last. A very small pencil soldering iron, miniature solder, and small diameter (No.22) hookup wire are recommended for ease in assembly. The various tuned circuits are wired and grid-dipped to frequency and the interstage shields are made up and cut to fit (a "nibbling" tool is handy here) as the work progresses. A closeup of the under-chassis r-f stages is shown in figure 33. A two-section variable mica compression-tuning capacitor is used for C_1 (*PEAK*

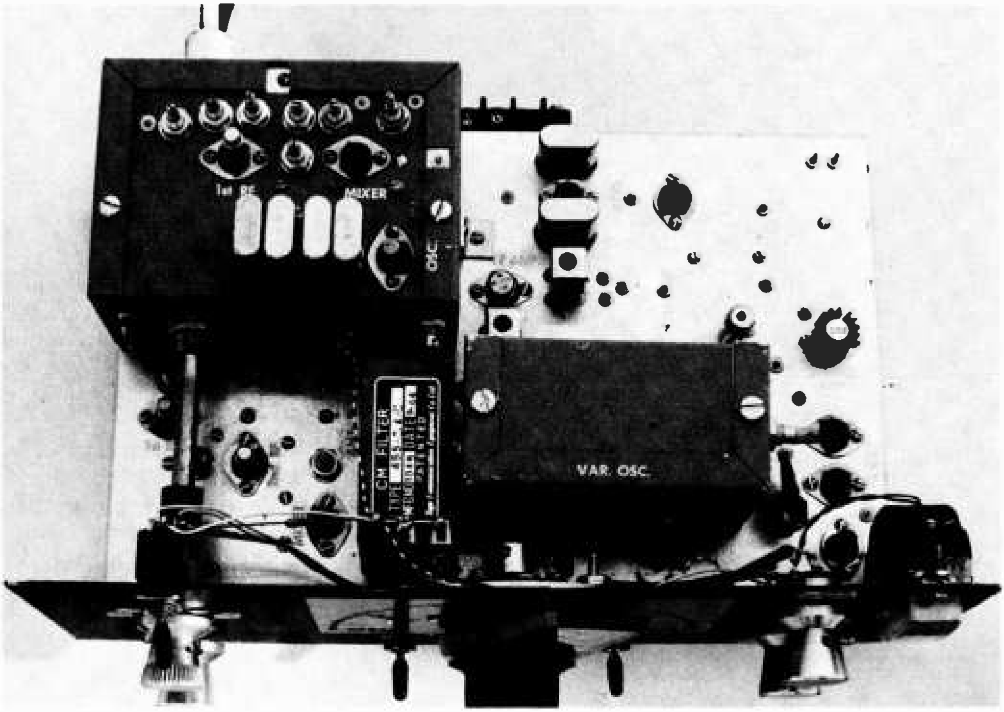


Figure 32

TOP VIEW OF RECEIVER ASSEMBLY

Placement of the major receiver components may be observed in this view. The h-f crystal-controlled converter assembly is at the left with the bandswitch extension shaft running to the front panel. At the center of the main chassis are the mechanical filter and the variable oscillator for the 80-meter portion of the receiver. Directly behind the oscillator are the i-f amplifier and the bfo stage with the associated sideband-selection crystals. At the right is the audio IC stage (with heat sink) and the "meter-adjust" potentiometer. The agc stages are in the right front corner of the receiver, with the 80-meter r-f section located at the front left corner of the chassis.

PRESELECTOR) and has an extension shaft press-fit onto the short tuning stub. The capacitor is supported from a small bracket mounted directly behind the panel.

Small shields are mounted across each MOSFET socket. The shields are cut of scrap aluminum or brass and have a mounting foot on them which is held in place by a nearby 4-40 bolt. The first r-f stage MOSFET socket (Q_1) is at the left of the photograph with the small coaxial line from the converter unit visible at the lower edge of the assembly. To the right is the second r-f stage MOSFET socket (Q_2), with the FET mixer socket above and to the right. The injection line from the vfo passes through a *Teflon* feedthrough insulator mounted in the chassis immediately behind the tuning dial and runs to the gate terminal of the FET socket.

The remainder of the construction and assembly on the main chassis is straightforward. Using ¼-watt resistors and miniature capacitors helps to keep assembly neat and compact. Liberal use of *Teflon* feedthrough insulators and terminals assists greatly in controlling underchassis "clutter."

The main tuning dial is made up of a reduction drive, a home-made pointer, and a calibrated scale etched on a piece of copper-plated circuit board of the glass-epoxy variety. The mask for the negative of the board is reproduced in figure 34. It may be photocopied from the page and used to make a negative for direct reproduction. It should be noted that in the process of etching the

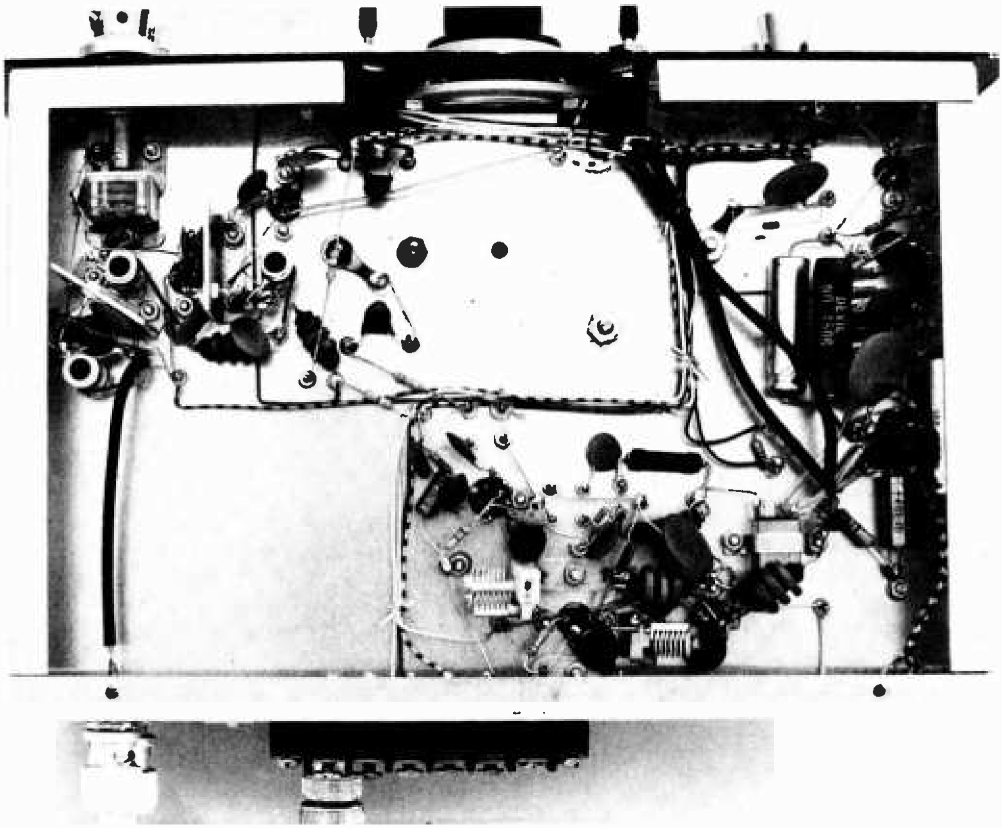


Figure 33

UNDER-CHASSIS VIEW OF SOLID-STATE COMMUNICATIONS RECEIVER

The 80-meter r-f amplifier and mixer stages are seen in the upper left corner of the chassis. The two-section variable mica compression capacitor (C_{1A-B}) is mounted to the chassis by means of a small aluminum bracket affixed behind the main panel. The capacitor is driven by a short extension shaft. An intrastage shield is placed across the first r-f amplifier MOSFET socket (Q_1) and a second similar shield is placed across the second r-f amplifier socket. The shields may be made of copper-plated circuit board, aluminum, or thin copper shim stock. The audio circuit and agc components are placed along the right-hand edge of the chassis, with the bfo, detector, and i-f components strung along the rear of the chassis area (bottom of the photograph). The two 35-pf capacitors used to adjust the frequency of the bfo crystals are supported below the chassis by their leads.

Note: The cutout at the front of the chassis is to provide room for the gear-reduction drive mounted to the panel.

board, the photoresist material should not be removed from the board after etching. It is attractive if left on and will prevent the dial surface from being corroded by fingerprints or moisture in the atmosphere.

The Converter Assembly—The general layout of the converter assembly is shown in figures 35 through 37. The MOSFETs and conversion crystals are mounted in sockets placed atop the converter box, with the various slug-tuned coils mounted at the rear of the assembly. Figure 35 shows the rear of the box with the cover removed. The r-f amplifier (Q_1) coils are at the right of the shield partition, with the mixer coils (Q2) at the left. Directly below the mixer coils is the crystal-can relay (K_1) with the coaxial leads attached to it. The various outer shields of the coaxial lines are grounded at the relay mounting bracket. Note that several *Teflon* feedthrough insulators are mounted in the L-shaped shield partition to

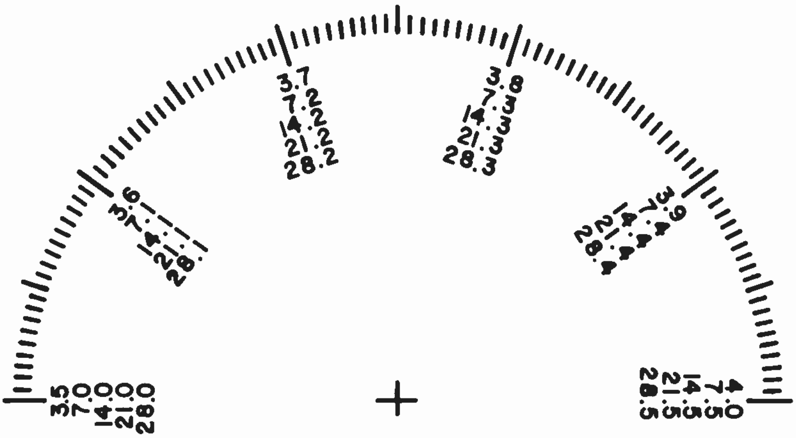


Figure 34

TUNING DIAL TEMPLATE FOR THE SOLID-STATE RECEIVER

pass power leads between the stages within the box. An oblique view of the r-f compartment is shown in figure 36. The two-section ceramic bandswitch is in the foreground, with the 40-meter piston tuning capacitor (C_1) mounted to the wall of the box in the foreground. Directly in front of the bandswitch is the feedthrough insulator for the lead to the coil of the crystal-can relay. The bandswitch is positioned to provide the shortest possible leads to the slug-tuned coils mounted adjacent to it.

An end view of the converter assembly is shown in figure 37. The relay is held in position with a small aluminum U-clamp over the body, and the opposite side of the L-shaped intrastage shield is visible.

The Variable Oscillator—The vfo is the only other separate subassembly. Layout of parts (aside from placement of the main tuning capacitor, mentioned earlier) is not critical. The components are self-supported around the capacitor using short, direct leads to prevent vibration. It is possible to build the unit in a much smaller box, but the good

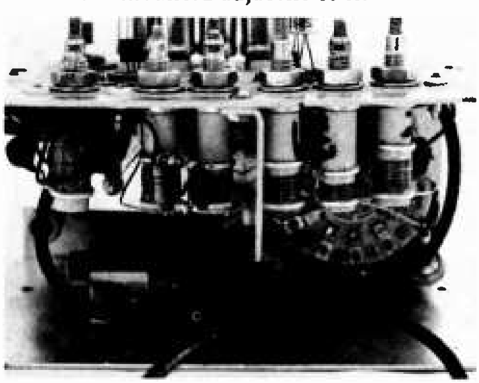


Figure 35

REAR VIEW OF CONVERTER ASSEMBLY

The r-f amplifier and bandswitch are seen at the right of the internal shield partition. R-f coils are (l. to r.): 20, 15 and 10 meters. Note Teflon feedthrough terminals mounted in the intrastage partition. The mixer stage and crystal can relay (K_1) are at the left of the partition. Mixer coils are (l. to r.): 15 and 10 meters. I-f output coil L_7 is at the extreme left.

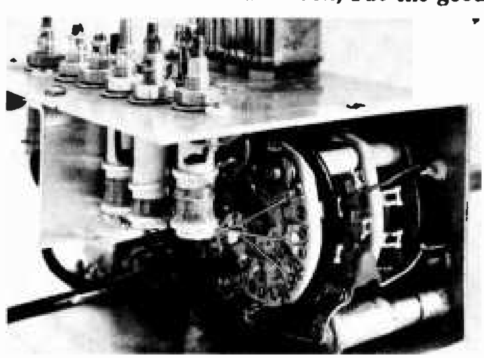


Figure 36

THE R-F AMPLIFIER

The r-f amplifier coils are in the foreground, with the bandswitch and piston capacitor (C_1) at the right. The coaxial leads run to the crystal-can relay. The outer shields of the various coaxial lines are grounded to a common point near the relay and also at the free ends in the receiver assembly. Note that coils and bandswitch have been arranged for shortest possible lead lengths.

drift characteristic (100-Hz total warmup drift) makes the larger box worthwhile. Both FET sockets are mounted on the vertical front surface of the box, with the oscillator coil (L_4) mounted to one end; and the bandset capacitor (C_2) mounted to the other end of the box.

Receiver Alignment Alignment of the receiver is not difficult if done in a systematic manner and may be done by ear alone. A quicker and better job may be achieved, however, with the use of proper instruments. The main receiver chassis is aligned first, so that a proper output indicator will be available for subsequent alignment of the converter. All alignment is done with the agc switched off. Before beginning the alignment and before power is applied to the receiver, the tuning meter should be disconnected to prevent its possible damage due to accidental overcurrent. The builder should also note the information in the transistor chapter of this Handbook regarding the handling procedures to be used with the MOSFET transistors, which are inserted toward the end of the alignment operation.

The audio portion of the receiver is tested first. A heat sink is placed over the audio IC (TAA-300) before tests are begun. A

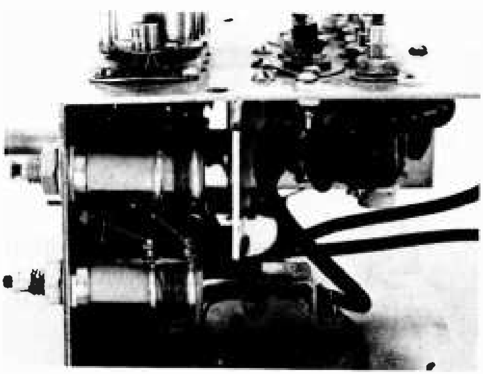


Figure 37

SIDE VIEW OF THE CONVERTER UNIT

The crystal-can relay is in the lower foreground with the 10-meter oscillator coil at the top left and the 15-meter oscillator coil at the bottom left. The internal shield (also seen in figure 33) is L-shaped and isolates the oscillator coils from the mixer coils located at the rear of the chassis deck.

1000-Hz, 10-millivolt sine-wave audio signal is applied at the arm of the *AUDIO-GAIN* potentiometer (R_1) and should result in a signal in the speaker when primary power is applied to the receiver, indicating the audio stage is working. Check the voltage at the drain of the 2N4360 agc control transistor (Q_{10}). It should be close to −12 volts. Removing the audio signal should cause it to drop to almost zero volts. This indicates that the complete agc system is working.

Next, set the *METER-ADJUST* potentiometer (R_3) for zero resistance (short circuit) and reconnect the tuning meter. With the audio signal applied again as before, adjust the meter current for minimum deflection (minimum reading). Removing the audio signal should cause the meter current to increase to a full-scale value. Although the agc is off, the system still controls the meter and it can now be used as an indicator of input signal level to the receiver. Advance the *R-F GAIN* control (R_2) fully clockwise to *Max* position. Apply a 456.5-kHz modulated signal of 1-millivolt level to the input (pin 1) of the IC i-f amplifier (MC-1553G). If the amplifier, the bfo, and the product-detector stages are working, an audio signal should be heard in the speaker. Adjust the detector filter circuit (T_3) for minimum hiss in the speaker when the audio modulation is turned *off*. Now, adjust the *AUDIO-GAIN* control (R_1) back and forth to make sure it functions properly. Apply the same r-f signal to the input of the mechanical filter and adjust i-f transformer T_1 for maximum signal in the speaker. Varying the input signal frequency above and below 456.5 kHz will provide an indication of the intermediate-frequency passband response of the receiver. Switch the bfo *SELECT-SIDEBAND* switch (S_1) to both positions to ensure that both oscillator circuits are working. Crystal alignment on the filter passband is accomplished by adjustment of the series capacitors.

The next step is to test the variable tuning oscillator. The transistors are inserted in their sockets and the oscillator tuned circuit should be adjusted to tune over the range of 3043.5 kHz to 3543.5 kHz between the extreme positions of the dial. The bandset capacitor (C_2) may be used for this adjustment, along with the slug adjustment of coil L_4. After the slug position has been determined, it should be fastened in place with a drop of cement to prevent vibration.

The tuned circuits in the r-f stages and the mixer should be adjusted to track across the 80-meter band when the *PEAK-PRESELECTOR* control is adjusted. Preliminary alignment should be done with a grid-dip oscillator with transistors Q_1, Q_2, and Q_3 *removed from their sockets.* When MOSFETs Q_1 and Q_2 are inserted in their respective sockets, a ferrite bead is slipped over the gate and drain leads of each device to suppress any tendency toward vhf parasitic oscillations. Place the peaking control (C_1) at half capacitance and apply a 10 microvolt, 3750-kHz signal at the input terminal (J_1) of the main receiver. Tune the receiver to the signal and adjust the three tuning slugs in coils L_1, L_2, and L_3 for maximum signal output. The receiver may now be used for 80-meter reception.

Converter Alignment—The high-frequency converter should now be attached to the main chassis and the various leads connected. Before the MOSFETs are placed in the sockets, the converter tuned circuits should have been grid-dipped to the approximate working frequencies. Now, the converter bandswitch is set to the 20-meter position and the main tuning dial of the receiver set to 14.250 MHz. A 10-microvolt signal at this frequency is applied to the converter input circuit, making sure that the relay K_1 is properly activated. Adjust the slug of the mixer coil (L_4) for maximum output signal, followed by adjustment of r-f coil L_1. These adjustments will not be critical due to the large bandwidth of these circuits. The converter must be first aligned on 20 meters since the tuned circuits are basically tuned to that band. Once they are aligned, do not touch them further.

The bandswitch is now placed in the 40-meter position and a 7.2-MHz signal applied to the receiver. Capacitors C_1 and C_2 are adjusted for maximum signal level. In the same fashion, a midband signal is applied to the converter for the 15- and 10-meter bands, aligning them by the slugs in the shunt coils, as before, mixer circuit first. Finally, adjust the 10-meter oscillator circuit (L_8) for best received signal on that band, then adjust the 15-meter oscillator circuit (L_9) for *minimum* received signal when a 20-meter signal is injected into the receiver. This completes alignment of the receiver.

19-7 A Solid-State 2-Meter F-M Transmitter

This reliable, compact and inexpensive f-m transmitter is suitable for portable operation from a 12-volt lantern battery or for mobile operation from the automobile electrical system. Designed by W6HDO, it is rugged and easy to assemble and cannot be damaged by a high SWR on the transmission line to the antenna. Tuning is broadbanded and fixed to cover 2 MHz of the 2-meter band, and all spurious responses and unwanted emissions are at least —35 db down from the carrier level.

Using a nominal 12-volt supply the transmitter output is 1.5 watts, more than ample for repeater or mobile-to-base service. The circuit is assembled on a printed-circuit board (figure 38) and may easily be placed in a cabinet or case of the users choice.

The Transmitter Circuit The f-m transmitter employs six low cost transistors in the r-f section plus a seventh transistor in the microphone amplifier stage, as shown in figure 39. The circuit is conventional but unlike earlier designs that proved cranky to get working and were shy of excitation, this simple transmitter uses an extra driver stage and a fundamental-frequency oscillator to permit all stages to run with ample drive level. The r-f driver transistors are emitter-biased to reduce harmonics and spurious emissions and to provide more uniform operation when transistors of different manufacture are used. The circuit is very uncritical and the builder has a wide choice of suitable transistors that may be interchanged in the driver stages.

When the tuned circuits are peaked for maximum output, the stages are automatically broadbanded to cover 2 MHz of the 144- to 148-MHz region. Thus one simple alignment for maximum power output will cover a broad range of operating frequencies.

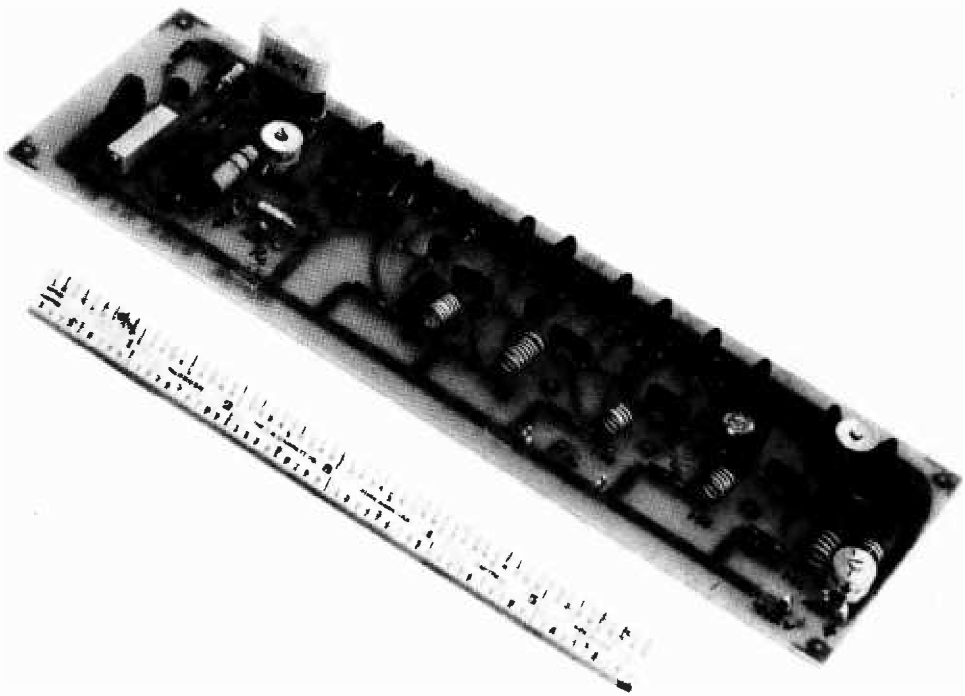


Figure 38

SOLID-STATE 2-METER F-M TRANSMITTER

This 1.5 watt transistorized f-m transmitter uses seven inexpensive transistors and works well for fixed station, portable or mobile service. Circuits are broadbanded and no critical tuning adjustments are needed. At left is audio stage, crystal oscillator, and audio gain control potentiometer. Final amplifier stage is at right with tune and load capacitors mounted to circuit board.

The oscillator stage (Q_2) uses 12-MHz type HC-6/U crystals in a parallel-resonant mode, ground for a load capacitance of 20 pfd. A variable capacitance (C_1) may be placed across the crystal holder to bring the crystal on frequency, if required. The fundamental-frequency oscillator is followed by a tripler stage (Q_3) to 36 MHz, a doubler stage (Q_4) to 72 MHz. and a final doubler (Q_5) to 144 MHz. A driver stage (Q_6) provides ample excitation for the 2N5913 amplifier. Heat sinks are placed on transistors Q_5 and Q_6, as noted in the parts list.

The microphone amplifier stage (Q_1) is designed to be used either with a carbon or dynamic microphone. Resistor R_6 supplies button current for the carbon microphone and is omitted for the dynamic unit. Frequency modulation is accomplished by two silicon diodes which pull the crystal frequency above and below the nominal center frequency. The amount of deviation is controlled by the 100K audio level potentiometer.

The antenna output terminal is placed in such a position that the shield of the coaxial antenna line may be returned to either the plus or minus power buss, as required by the electrical system ground. In addition, extra ground and tie points are mounted on the board to hold a grounding clip for the crystal and to permit substitution of a multiturn miniature potentiometer for the conventional audio gain control. Additional points are provided on the board for the experimenter who may wish to try circuit changes.

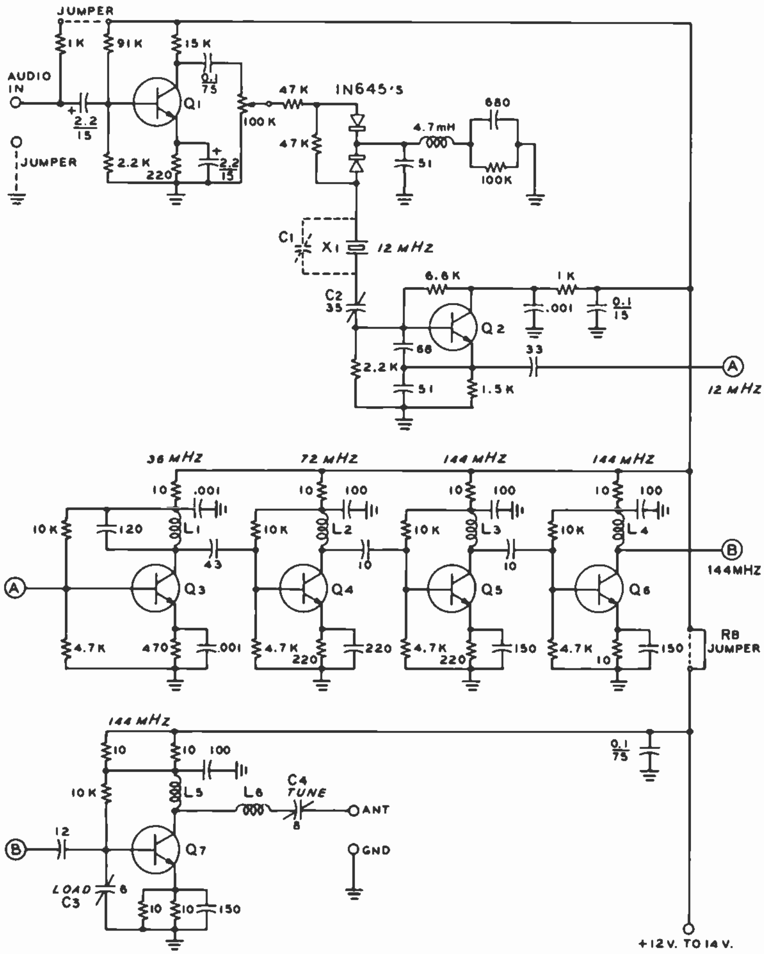


Figure 39

2-METER F-M TRANSMITTER

C_1—25-pf ceramic trimmer. JFD 25C
C_2—35-pf ceramic trimmer. JFD 35D
C_3, C_4—8 pf ceramic trimmer. JFD 8A
L_1, L_2—7 turns #22 tinned, 3/16" inside diameter. Adjust to resonance
L_3, L_4, L_5—4 turns #22 tinned, 3/16" inside diameter. Adjust to resonance
L_6—2 turns #22 tinned, 3/16" inside diameter
Q_1—2N929, 2N2926, 2N3565, 2N2721, or equivalent
Q_2-Q_5—2N708, 2N3563, 2N4124, HEP 53, MPS 3563, or equivalent
Q_6—2N3866, 2N4427, or equivalent. Heat sink with 5/8" diam, 1/4" high sink
Q_7—2N5913. Heat sink with 5/8" diam, 3/8" high (or larger) sink

Transmitter Construction The transmitter is built on a printed-circuit board measuring 8½" × 2¾". A full-size template may be obtained for 25 cents from *Editors & Engineers, 4300 West 62nd St., Indianapolis, Indiana, 46268.* Placement of the major components may be seen in the photograph. Type G-10 glass epoxy circuit board (0.060" thick) is used, as the less expensive phenolic board has high r-f loss at 144 MHz. Note that the board layout is such that the ground return lead of the various bypass capacitors forms a series-resonant, low-impedance circuit to ground which is very effective in the 144-MHz region. The circuit inductors are air-wound on a 3/16" diameter form, which is removed. The coils are soldered in position on the circuit board and tuned to resonance by compacting or expanding the coil length with a wood dowel rod.

Transmitter Adjustment The transmitter may be tuned up with a 12-volt bench power supply, using an SWR meter and a 2-watt dummy load made of composition resistors. If coil dimensions are followed, the transmitter will provide some power output without any adjustment of the inductors when the crystal is inserted and primary power applied.

To begin adjustment, the audio gain control is turned down and, starting with the first multiplier stage, the air inductors are "tweaked" to achieve maximum power output. The frequency of operation of each stage may be checked with a wavemeter, if desired. Primary power should be held to 12 volts or less during tuneup, since it is best to make sure all stages are in resonance before the voltage is raised to a nominal level of 13.5 volts. Final adjustments are made with the proper antenna attached to the transmitter, using the SWR meter to determine maximum power into the antenna.

The last step is to apply an audio signal and advance the gain control to reach the proper deviation level. Deviation depends to some extent on crystal activity and the settings of capacitors C_1 and C_2. Adjustment is best accomplished with the aid of a nearby amateur who has a receiver of the proper deviation. Normally, the parallel capacitance (C_1) will be quite small for crystals cut for 20-pfd capacitance since the residual circuit capacitance is nearly of this value. Series capacitance C_2 is set to near maximum capacity to start with, then varied slightly to provide optimum deviation with proper crystal excitation. The adjustment of capacitors C_1 and C_2 are interlocking to a degree but by "tweaking" both of them, the operator can quickly achieve a maximum deviation level for a given degree of audio input and still easily make the center-channel frequency adjustment.

When all stages are aligned, the primary voltage may be boosted to 13.5 volts. Under normal mobile service, the primary automotive electrical system may rise to over 14 volts under full charge conditions. The transmitter will stand this level of voltage with no difficulty provided the final two transistors are well ventilated and cooling air is free to move over the heat sinks.

19-8 A 70-Watt Solid-State Amplifier for 2-Meter F-M

This compact 70-watt strip-line amplifier is designed and built by W6GFS. It is in-

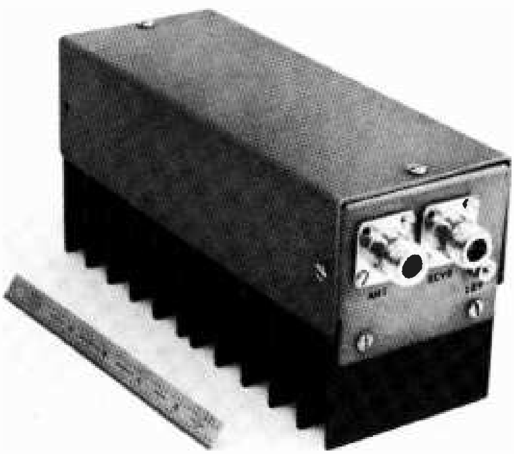


Figure 40

70-WATT 2-METER AMPLIFIER FOR F-M

Designed for use with the popular 10-watt f-m transceivers, this solid-state 2-meter amplifier delivers over 70 watts power output on the 2-meter band. Featuring broadband, fixed-tuned circuits, the amplifier requires no tuning adjustments. Large heat sink ensures cool, reliable operation. Amplifier is designed to be used with a grounded-negative, 12-volt automobile power system. Antenna and power switching is automatic.

tended to work with the popular, imported 10-watt 2-meter f-m transceivers. With a nominal 12.5-volt supply, the amplifier provides over 70 watts power output when used with one of these transceivers. This represents a power gain of 7—a worthwhile improvement!

Use of this amplifier will permit the f-m operator to "break" a repeater at an extreme distance, or under the most adverse operating conditions. Featuring broadband circuits, the amplifier requires no adjustment and may be tucked away in a corner of the automobile out of sight (figure 40).

The amplifier is designed to be cut in the coaxial line between the transceiver and the antenna and power switching is automatic. Input and output impedances are 50 ohms and the amplifier is capable of withstanding an infinite SWR on the antenna circuit without the necessity of external protection.

Amplifier Circuitry The schematic of this transistorized amplifier is shown in figure 41. A grounded-emitter, base-driven circuit is used. A *CTC* type *BM-7012* vhf power transistor is featured in a strip-line configuration. This rugged transistor is designed for vhf mobile service and has improved broadband performance by virtue of the *J-zero* package which incorporates a chip capacitance from base to emitter in the device itself.

The input impedance of the transistor (typically about one ohm at 144 MHz) is matched to the 50-ohm driving source by an L-network. The network consists of strip line L_1 and four vhf capacitors placed at the base terminal of the transistor. The base is d-c isolated from the driver by a 200-pfd series blocking capacitor in the strip line.

A second L network (L_2) transforms the output impedance of the device (typically about two ohms) to the 50-ohm output source. Again, the strip line is d-c isolated from the output line to the antenna relay and the collector voltage is shunt-fed to the device through a low resistance r-f choke (RFC_3).

To suppress low-frequency oscillations generated in common-emitter r-f amplifiers, low impedance r-f chokes (RFC_1 and RFC_3) are placed in the base and collector circuits and negative base-collector feedback is incorporated to reduce the low-frequency

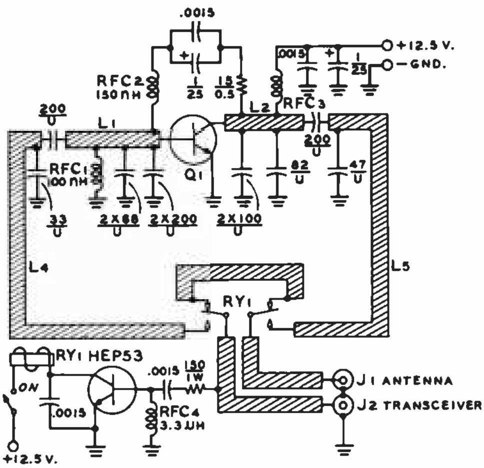


Figure 41

SCHEMATIC OF 2-METER AMPLIFIER

Capacitors marked U are vhf type mica (UNELCO (Underwood Electric Co., Inc.)
RFC_3—5 turns #18 enamel, 3/8" diameter, 3/4" long
K_1—Dpdt crystal-can relay. Electronic Specialty type 80N. 12.5-volt d-c coil
Q_1—CTC type BM-7012 power transistor
L_1, L_2—Each 3.9" long, 0.125" wide (50-ohm impedance)
Board—G-10 glass-filled epoxy
Heat sinks—Two Thermalloy 6151B or equivalent
Note: Information on components, circuit board, and a kit of parts for this amplifier may be obtained from: POWER KITS, Box 693, Cupertino, Calif. 95014

gain of the device. The feedback network consists of a series RLC circuit between base and collector. An r-f choke (RFC_2) is included to remove the feedback network at the operating frequency, and the parallel-connected capacitors in the network ensure good coupling down to the lowest frequency of possible oscillation.

The amplifier is switched in and out of the circuit by an r-f operated relay (K_1). Approximately 4 watts of drive level are necessary before the relay is activated. When the relay is open the amplifier is out of the circuit and the antenna is directly connected to the transceiver. When the transceiver is activated, the r-f signal to the amplifier is rectified and amplified by transistor Q_1 and the resulting d-c voltage is applied to the control relay. Microstrip transmission lines (L_4, L_5) from the relay to the amplifier

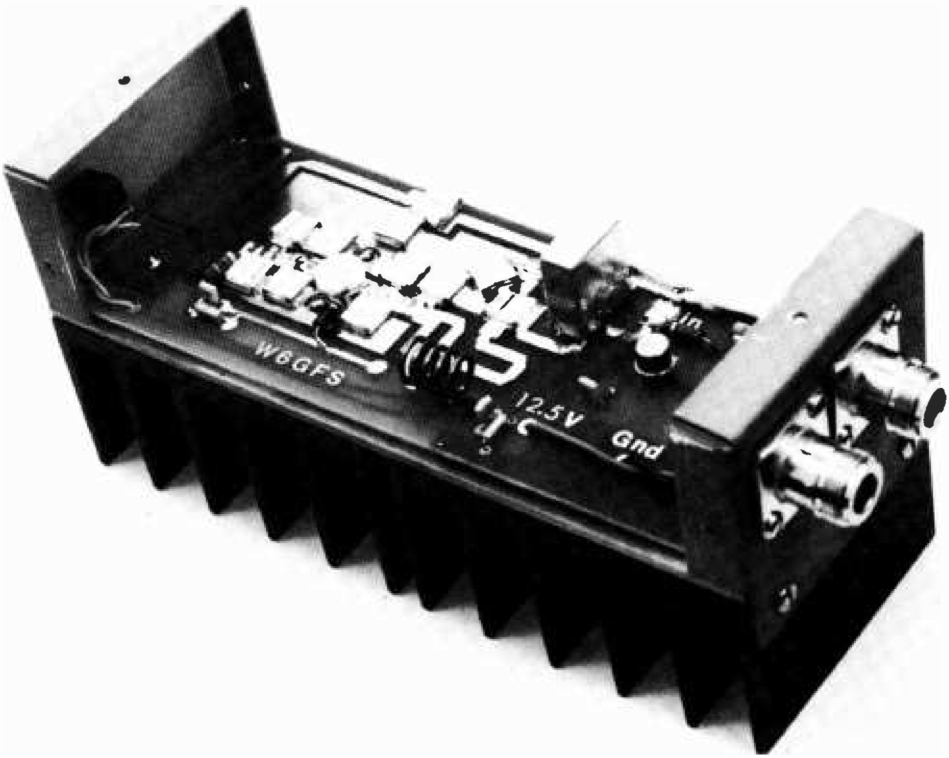


Figure 42

LAYOUT OF 2-METER AMPLIFIER

The printed-circuit board is mounted above the copper plate and heat sink assembly. The power transistor is bolted directly to the sink. In the foreground is collector choke (RFC$_3$), with the antenna relay behind it. At far left is RFC$_1$, which is plated on the circuit board.

ensures that the SWR on the line is not perturbed when the amplifier is switched into the circuit.

Amplifier Construction The 2-meter amplifier is built on a printed-circuit board of low loss, G-10 glass filled epoxy material measuring 7½″ × 2¾″ in size (figure 42). The copper foil on the underside of the board forms a ground plane for the strip-line circuitry atop the board which has a characteristic impedance of 50 ohms. The board is supported above a copper plate which, in turn, is bolted to a finned heat sink. With the dust cover in place, the amplifier measures 8¼″ × 3″ × 4″, exclusive of the coaxial fittings. The copper plate and heat sink are capable of dissipating about 70 watts while holding transistor dissipation to a safe value.

Two end plates are bolted to the heat sink which hold the coaxial fittings, the main power terminal, and the primary power switch. The dust cover is held to the end plates with sheet-metal screws.

Transistor Mounting—The technique used to mount and ground the vhf transistor is very important at high power levels. The copper plate is drilled and tapped for 4-40 screws which securely fasten the transistor to the heat sink assembly. It is important that the copper plate is flat if proper heat transfer is to be expected. A high quality silicone grease such as *GE Insulgrease* or equivalent is used on both the transistor flange and the plate. It is important that the transistor flange not be twisted or bent during installation. In addition, the major components (excluding the special *Unelco* bypass capacitors around the transistor) are soldered to the board. Note that the crystal can relay is keyed by a dark glass bead on one lead. The board is drilled with a #56 drill for the four isolated contact pads of the relay. The other relay contact pins are bent up and soldered to pads on the top of the board. The pins passing through the drilled holes are soldered to the back of the board.

The back side of the printed-circuit board is a ground plane. The ground areas adjacent to the transistor atop the board should be connected to the bottom ground plane using thin copper ground straps under each emitter lead. These straps are soldered in position before the transistor is bolted in place (see figure 14, Chapter 20).

The board is placed above the copper plate and the transistor bolted in position, with the active leads falling atop the board. Once the transistor is in position the board may be bolted to the plate using spacers on each corner bolt. The board-to-plate spacing must be adjusted so that the transistor leads are on an even plane with the surface of the board. Do not allow the leads to be bent up or down to reach the copper foil on the board since this may damage the transistor. When the board is in position, tighten the corner bolts and solder the transistor leads to the proper foil areas on the top surface of the board. The transistor emitter leads, in particular, should be grounded at the body of the transistor as lead inductance at 144 MHz is critical.

Amplifier wiring is comparatively simple as most of the r-f and control circuitry is on the printed-circuit board. The *Underwood* (Unelco) mica capacitors around the transistors are now soldered in place, the case going to the ground plane in all cases except the series blocking capacitors. The cases of these capacitors are soldered to the input lines and the tabs to the strip lines L_1 and L_2. The 200-pfd capacitors at the base terminal should be placed directly up against the transistor case to hold lead inductance to a minimum value.

Testing the Amplifier—Once the wiring has been completed, the assembly should be carefully examined for wiring errors or inadvertent shorts. In particular, the bottom of the board should be visually inspected to make sure it does not short out any connections to the copper plate. For preliminary test, SWR meters should be placed in the input and output coaxial lines and the unit connected to a 100-watt vhf dummy load. A 10-ampere d-c meter and 10-ampere fuse should be placed in the positive power lead.

The transistor is operated in a class-C mode and collector current is zero with no signal input. Thus, with switch S_1 open and drive applied, the amplifier current should be zero and the r-f driving signal should pass directly through the normally closed contacts to the antenna. When S_1 is closed and r-f drive applied, you should be able to hear relay K_1 close, indicating that the r-f sensing circuit is working. The d-c current drain should be 7 to 9 amperes, depending on the power output level of the transceiver. Keying the transceiver by opening and closing its control circuit should key the amplifier and when drive is removed, the current drain should drop to zero. Finally, the unit should be given an "on-the-air" test, and the output circuit of the transceiver peaked for maximum output from the amplifier.

CHAPTER TWENTY

Receivers, Converters, and Transceivers

Receiver construction has just about become a lost art. Excellent general-coverage and ham-band-only receivers are available on the market in many price ranges. However, even the most modest of these receivers is relatively expensive, and most of the receivers are designed as a compromise—they must suit the majority of users, and they must be designed with an eye to the price.

It is a tribute to the receiver manufacturers that they have done as well as they have. Even so, the c-w man must often pay for a high-fidelity audio system and S-meter he never uses, and the SSB operator must pay for the c-w man's narrow-band filter. For one amateur, the receiver has too much bandspread; for the next, too little. For economy's sake and for ease of alignment, low-Q coils are often found in the r-f circuits of commercial receivers, making the set a victim of crosstalk and overloading from strong local signals. Rarely does the purchaser of a commercial receiver realize that he could achieve the results he desires in a home-built receiver *if* he left off the frills and trivia which he does not need but which he must pay for when he buys a commercial product.

The ardent experimenter, however, needs no such arguments. He builds his receiver merely for the love of the game, and the thrill of using a product of his own creation.

It is hoped that the receiving equipment to be described in this chapter will awaken the experimenter's instinct, even in those individuals owning expensive commercial receivers. These lucky persons have the advantage of comparing their home-built product against the best the commercial market has to offer. Sometimes such a comparison is surprising.

When the builder has finished the wiring of a receiver it is suggested that he check his wiring and connections carefully for possible errors before any voltages are applied to the circuits. If possible, the wiring should be checked by a second party as a safety measure. Some transistors can be permanently damaged by having the wrong voltages applied to their electrodes. Electrolytic capacitors can be ruined by hooking them up with the wrong voltage polarity across the capacitor terminals. Transformer, choke, and coil windings may be damaged by incorrect wiring of the high-voltage leads.

The problem of meeting and overcoming such obstacles is just part of the game. A true radio amateur should have adequate knowledge of the art of communication. He should know quite a bit about his equipment (even if purchased) and, if circumstances permit, he should build a portion of his own equipment. Those amateurs who do such construction work are convinced that half of the enjoyment of the hobby may be obtained from the satisfaction of building and operating their own receiving and transmitting equipment.

Figure 1

COMPONENT NOMENCLATURE

CAPACITORS:

1 - VALUES BELOW 999 PF ARE INDICATED IN UNITS.
EXAMPLE: 150 PF DESIGNATED AS 150.

2 - VALUES ABOVE 999 PF ARE INDICATED IN DECIMALS.
EXAMPLE: .005 µFD DESIGNATED AS .005.

3 - OTHER CAPACITOR VALUES ARE AS STATED.
EXAMPLE: 10 µFD, 0.5 PF, *ETC.*

4 - TYPE OF CAPACITOR IS INDICATED BENEATH THE VALUE DESIGNATION.
SM = SILVER MICA
C = CERAMIC
M = MICA
P = PAPER

EXAMPLE: $\frac{250}{C}$, $\frac{.01}{P}$, $\frac{.001}{M}$

5 - VOLTAGE RATING OF ELECTROLYTIC OR "FILTER" CAPACITOR IS INDICATED BELOW CAPACITY DESIGNATION.

EXAMPLE: $\frac{10}{450}$, $\frac{20}{800}$, $\frac{25}{10}$

6 - THE CURVED LINE IN CAPACITOR SYMBOL REPRESENTS THE OUTSIDE FOIL "GROUND" OF PAPER CAPACITORS, THE NEGATIVE ELECTRODE OF ELECTROLYTIC CAPACITORS, OR THE ROTOR OF VARIABLE CAPACITORS.

RESISTORS:

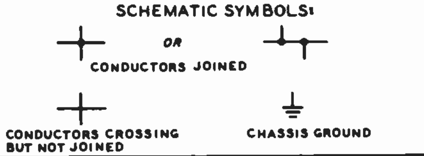
1 - RESISTANCE VALUES ARE STATED IN OHMS, THOUSANDS OF OHMS (K), AND MEGOHMS (M).
EXAMPLE: 270 OHMS = 270
4700 OHMS = 4.7 K
33,000 OHMS = 33 K
100,000 OHMS = 100 K OR 0.1 M
33,000,000 OHMS = 33 M

2 - ALL RESISTORS ARE 1-WATT COMPOSITION TYPE UNLESS OTHERWISE NOTED. WATTAGE NOTATION IS THEN INDICATED BELOW RESISTANCE VALUE.
EXAMPLE: $\frac{47\ K}{0.5}$

INDUCTORS:

MICROHENRIES = µH
MILLIHENRIES = MH
HENRIES = H

SCHEMATIC SYMBOLS:



Circuitry and Components It is the practice of the editors of this Handbook to place as much usable information in the schematic illustration as possible. In order to simplify the drawing, the component nomenclature of figure 1 is used in all the following construction chapters.

The electrical value of many small circuit components such as resistors and capacitors is often indicated by a series of colored bands or spots placed on the body of the component. Several color codes have been used in the past, and are being used in modified form at present to indicate component values. The most important of these color codes for resistors, capacitors, power transformers, chokes, i-f transformers, etc. can be found in the appendix at the end of this Handbook.

20-1 A Low-Noise DX Converter for 2 Meters

This converter was designed by K6MYC and has been in use for over 3 years in moonbounce contacts between the United States and Europe, New Zealand, and Australia. This is an improved version of the converter shown in an earlier *Radio Handbook*, providing improved noise figure and better immunity to crossmodulation and overload distortion in the presence of strong signals.

The 40673 diode-protected, insulated gate field-effect (IGFET) transistor is an N-channel depletion type silicon device that proves very effective in vhf service. The 2-meter converter described in this section (figure 2) uses two inexpensive dual gate IGFETs which combine high input resistance, low input capacitance, very low feedback capacitance, and low noise figure at very high frequencies. Neutralization is unnecessary in properly constructed r-f stages since the gate-to-source capacitance of the 40673 is very low.

The IGFET is relatively insensitive to temperature variations, making the converter completely stable under varying conditions of use. Typically, the noise figure of this converter is better than 2.0 decibels and the usable passband without tuning readjustment is 1 MHz or more. Best of all, the IGFET exhibits a large-signal crossmodulation characteristic that approaches that of a good remote-cutoff vacuum tube, thus overcoming the handicap of poor crossmodulation characteristics exhibited by bipolar transistors when used in this service.

Circuit Description The schematic of the 2-meter IGFET converter is given in figure 3. A 40673 (Q_1) is used as a low-noise unneutralized high-gain

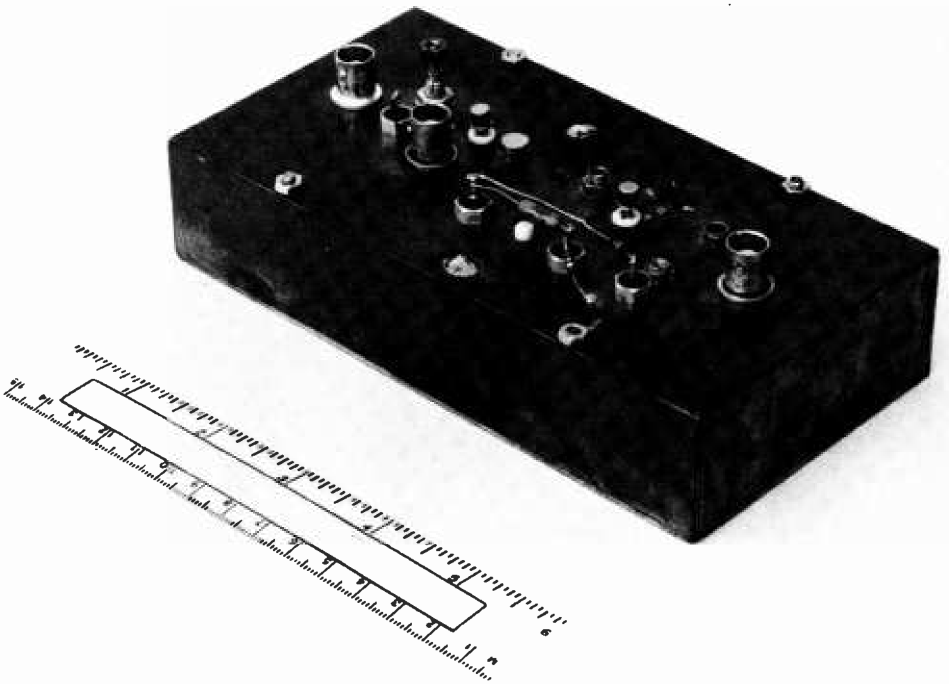


Figure 2

COMPACT IGFET CONVERTER FOR 144 MHZ

This low-noise DX converter employs two IGFET dual-gate transistors as an r-f amplifier and as a mixer. A separate local-oscillator chain is used for maximum flexibility. This converter has been used for successful 2-meter SSB moonbounce communication between Europe and the western United States.

At the right of the deck is the antenna receptacle and the r-f amplifier IGFET with input- and output-circuit tuning capacitors adjacent to the IGFET socket. To the left is the mixer stage, with the local-oscillator injection receptacle in front of the IGFET socket. The i-f output receptacle is at the extreme left of the deck. The resistors and r-f choke in the gate-2 circuit of the mixer IGFET are mounted atop the chassis at the front.

r-f amplifier in a common-source circuit. A conventional pi-network input arrangement is used to match the high input impedance of the IGFET to a low-impedance coaxial transmission line. The drain circuit of transistor Q_1 is resonated to the input frequency by a high-Q parallel-tuned circuit (C_2-L_2) which is capacitively coupled to a second 40673 diode-protected IGFET (Q_2) acting as a common-source mixer. The amplified signal is applied to a resonant circuit (C_3-L_3) and gate 1 of the mixer, and the local oscillator is coupled to gate 2. The resulting intermediate-frequency signal is taken from the drain circuit. The mixer stage is flexible enough so that the converter may be used with receivers of various intermediate-frequency tuning ranges by changing the injection frequency to the converter and readjusting output circuit L_1.

For simplicity in construction and maximum flexibility in use, the local-oscillator chain is assembled as a separate unit using bipolar transistors, and the schematic for a typical unit is shown in figure 5.

Converter Construction The IGFET converter is constructed on a piece of copper-clad (2 sides) glass epoxy circuit board measuring 6½" × 3¼". A matching shield case 1¼" deep is made of the same material soldered into a small box. Placement of the major components may be

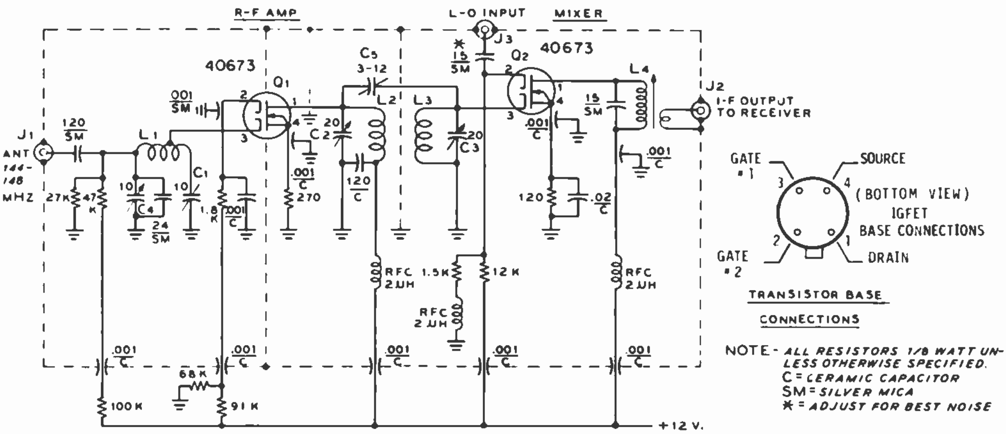


Figure 3

SCHEMATIC, 2-METER IGFET CONVERTER

C_1, C_4—10-pf. Johanson 2954 or JFD VAM-010 variable piston capacitors
C_2, C_3—20-pf. JFD-VC27G piston capacitors
C_5—12-pf variable ceramic. JFD-15D
L_1, L_2—6 turns silver-plated copper strap, 0.125" diam. × 3/4" long. Strap width is 1/16"
L_3—3 turns, same as L_1
L_4—40 turns #24 e., 1/4" diam. on slug-tuned form. J. W. Miller 40A-000CBI
J_1, J_2, J_3—Coaxial receptacles. BNC type UG-625/U

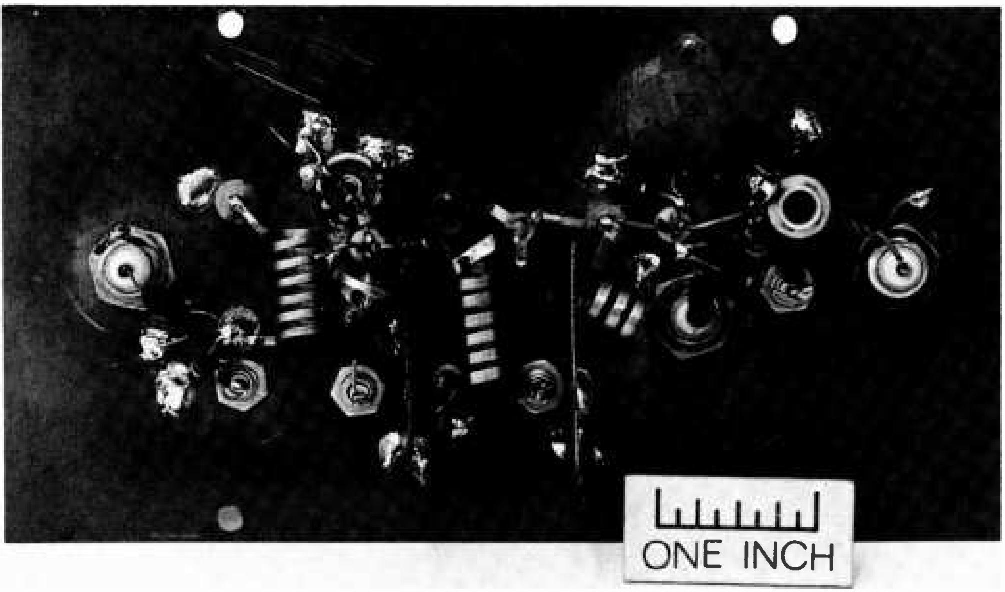


Figure 4

UNDER-CHASSIS VIEW OF IGFET CONVERTER

Placement of the major components may be seen by comparing this photograph with the underchassis drawing of figure 6. Ground connections are made directly to copper foil of the printed-circuit board. Input receptacle J_1 is at the left, and output receptacle J_2 is at the right. R-f coils are wound of silver-plated copper strap, but #14 wire may be used, if desired. Circuits are grid-dipped to frequency before IGFETs are inserted in sockets. Left interstage shield is 2" long and 3/4" high. Right shield of same height, but only 1 1/4" long. Both shields are cut from circuit-board material and soldered to the deck.

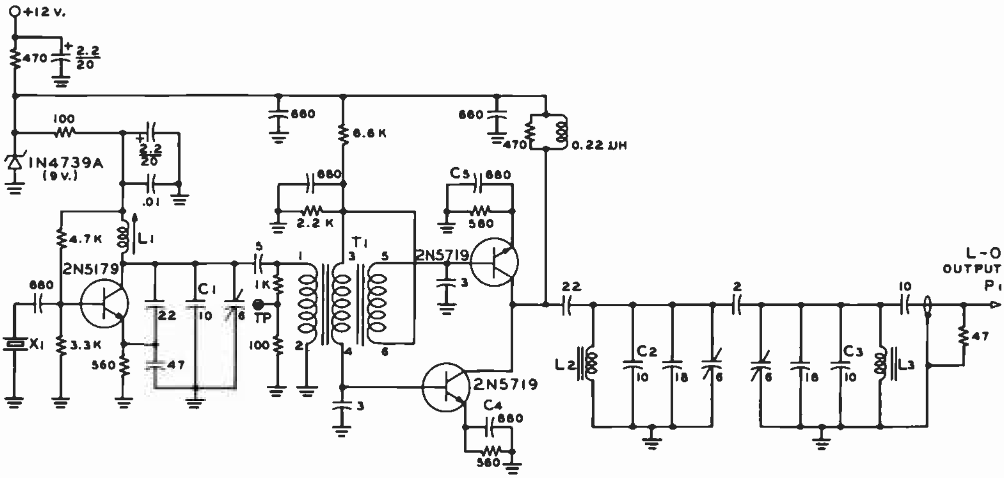


Figure 5

LOCAL-OSCILLATOR CHAIN FOR 2-METER CONVERTER

C_1, C_2, C_3—10-pf Sprague 10-TC NPO
C_4, C_5—680-pf JFD, UY-04 Unicerm capacitor
L_1—1.5 μH. Miller 4403
L_2, L_3—5 turns #30 e. on Micrometals T-12-12 core (144 MHz)
T_1—Trifilar winding of 6 turns each # 32 e. on Ferroxcube 3E2 core. (65 MHz)
X_1—Crystal, overtone, 65.000 MHz

seen in the photographs (figures 2 and 4) and layout drawing (figure 6).

A small shield measuring ¾″ × 2″ is cut out of thin copper shim stock, or of circuit-board material and is soldered in place between IGFET socket Q_1 and coil L_2 to reduce unwanted interstage coupling. A second similar shield measuring 1¼″ × ¾″ separates this coil from coil L_3. The first shield is positioned adjacent to the Q_1 socket, with the drain lead from the socket passing through a small hole drilled in the shield and on to capacitor C_2. Placement of the various small components is not particularly critical and they are grouped about the socket as suggested in the layout drawing. Gate-2 terminal of the IGFET socket is bypassed with a silver-mica button capacitor soldered on edge to the circuit board adjacent to the socket. In addition, a disc ceramic capacitor is placed next to the button mica unit. The source lead of the socket is bypassed with a similar button capacitor mounted in a hole drilled in the epoxy board.

Double bypass capacitors are placed on the source lead of mixer socket Q_2, a .02-μfd ceramic disc capacitor being used in addition to the vhf bypass capacitor in order to suppress any tendency toward a low-frequency parasitic oscillation. Other stage components are grouped closely about the IGFET socket.

Ground connections are soldered directly to the copper foil of the circuit board and practically no extra wire is required in assembling the converter. A small 15-watt "pencil" soldering iron should be used to prevent overheating the parts or the copper foil, since component density is quite high.

Converter Adjustment The IGFET device is able to withstand gate-to-source peak voltages as high as 10 and the device is protected against electrostatic discharge. It is still a good idea, however, to remove the device from the converter while work is being done on the unit. The general precautions dealing with this type of semiconductor are outlined in Chapter 4 of this handbook.

Once wiring is completed, it should be carefully checked against the schematic to be sure it is proper. Before the IGFETs are placed in their sockets, the various tuned circuits of the converter should be tuned to about 145 MHz with the aid of a grid-

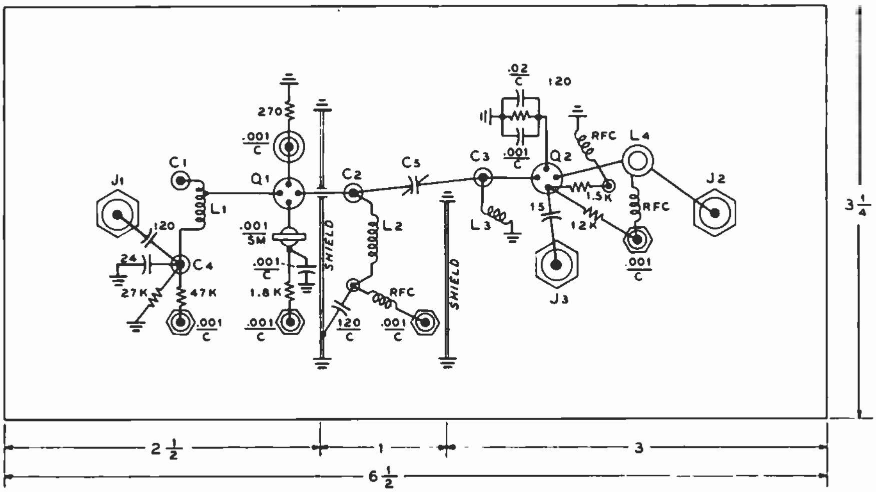


Figure 6

UNDER-CHASSIS PARTS LAYOUT FOR CONVERTER

dip oscillator. Pi-network capacitor C_4 is initially set at full capacitance and capacitor C_1 is tuned for indication of resonance as observed on the meter of the grid-dip oscillator. Spacing of the turns on coil L_1 may be adjusted slightly to permit a resonance indication at about half-capacitance setting of C_1. Coupling capacitor C_5 is set at half-capacitance and drain tuning capacitor C_2 is adjusted for resonance indication with the oscillator coupled to coil L_2. Next, the oscillator is coupled to coil L_3 and capacitor C_3 is adjusted for resonance at 145 MHz. The last step is to adjust the output inductor (L_4) for indicated resonance at the center of the chosen i-f passband.

For best noise level and optimum overload characteristic, the local-oscillator injection level is quite important. About 400 millivolts should be developed across the 50-ohm resistor at the output terminal of the oscillator unit (figure 5).

The converter is designed to be used with a 12-volt power supply having a negative ground. The supply should be well filtered and free of switching transients. Initial testing may be done with the aid of dry cells, since converter power drain is very low.

After the power supply is connected to the converter, a coaxial cable from the local-oscillator unit is connected to the injection receptacle (J_3) and a low-level mixing signal applied to the converter. Local-oscillator action may be checked in a nearby receiver, and it should be noted that the converter noise level rises when the injection signal is applied.

Since the tuned circuits are roughly in resonance, 2-meter signals should be heard when the converter is attached to an external antenna. Once the converter is found to be working, all tuned circuits are adjusted for maximum received-signal level.

Final adjustment is accomplished by re-adjusting the converter for best signal-to-noise ratio, regardless of converter gain. The noise figure is adjusted by varying the level of mixer injection voltage and by proper adjustment of the pi-network input circuit. This can best be done by listening to a weak signal, such as a remote crystal oscillator connected to a small antenna. Alternatively, a noise generator may be employed. All tuning adjustments are made with the end view of obtaining better signal to noise ratio, rather than merely greater signal strength. Exact tuning will depend on circuit loading and the reactance of the external antenna system and is best determined by experiment.

As a starter, a 50-ohm carbon resistor is used for a dummy antenna, placed across the terminals of receptacle J_1. The converter is tuned up for maximum noise level. The resistor is removed and the 50-ohm antenna system is connected in its place. If the noise level rises appreciably, the external noise is the limiting factor in vhf reception, and any receiver system improvement past this point must be accomplished by increasing the capture area (size) of the antenna and reducing the receiver passband. Adjustment of all circuits must be made for the greatest margin of received signal over receiver noise, using weak, received signals for this test.

The converter should be protected from strong input signals, such as might be generated by the station transmitter. Signal inputs of more than a half a volt or so might damage one or both IGFETs. Good input protection can be achieved by placing a pair of 1N100 germanium diodes in parallel, back to back, across antenna receptacle J_1 to ground.

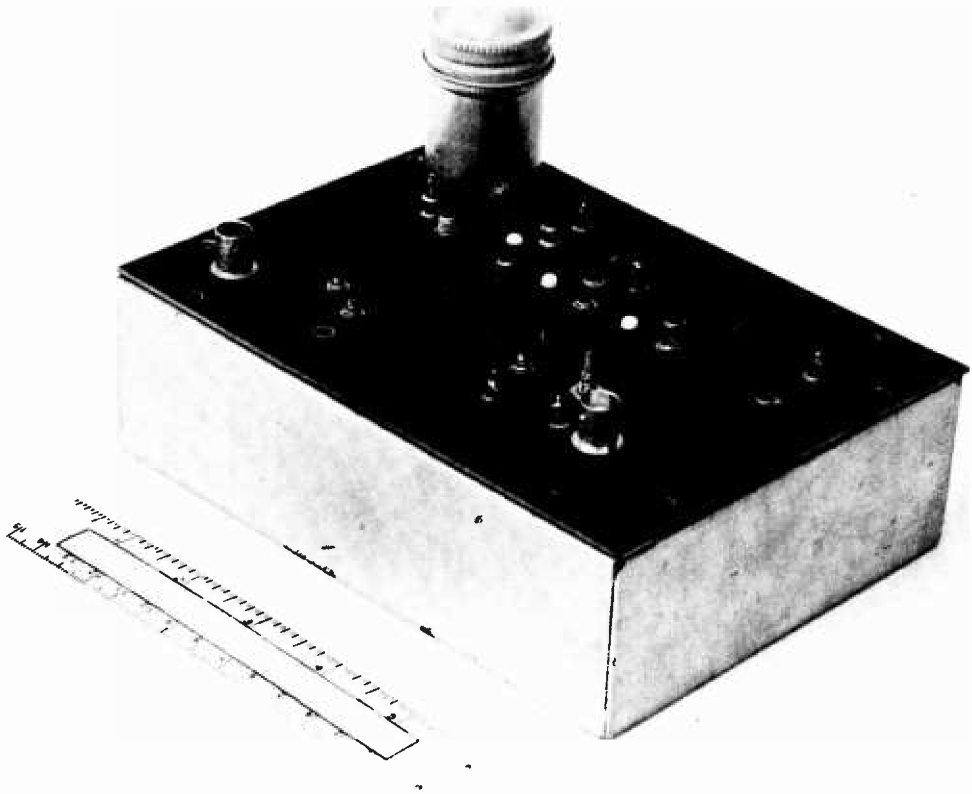


Figure 7

FET CONVERTER FOR 432 MHz

This converter has been used for extensive moonbounce communication in conjunction with the antenna-mounted FET preamplifier described in this section. The converter is built on a deck of copper-plated, phenolic circuit board and operates from a 12-volt supply. The frequency is controlled by a 51.35 MHz overtone crystal. To reduce frequency drift as a function of temperature, the crystal is mounted in the small can at the rear, made from a holder for 35-mm film. The crystal is surrounded by strips of foamed-plastic packing material to reduce temperature excursions. The antenna receptacle is in the left corner and the i-f output receptacle is in the foreground. The converter deck is supported on a small inverted aluminum chassis used for a dust cover.

20-2 A Low-Noise Converter and Preamplifier for 432 MHz

Various inexpensive FET devices will perform well as r-f amplifiers and mixers far into the vhf spectrum. This inexpensive and reliable converter for 432-MHz low-noise performance was designed and built by K6MYC for moonbounce operation and may be used either by itself, or with a remote preamplifier mounted at the receiving antenna (figure 7). The combination of a converter and remote r-f amplifier provides a superior noise figure and overcomes the troublesome problem of line loss that can degrade weak-signal reception at this frequency.

The builder can use various FETs in the r-f stage and mixer of the converter. The first choice is the *Texas Instruments* 2N5245, followed closely by the TIS-88 and the 2N4416.

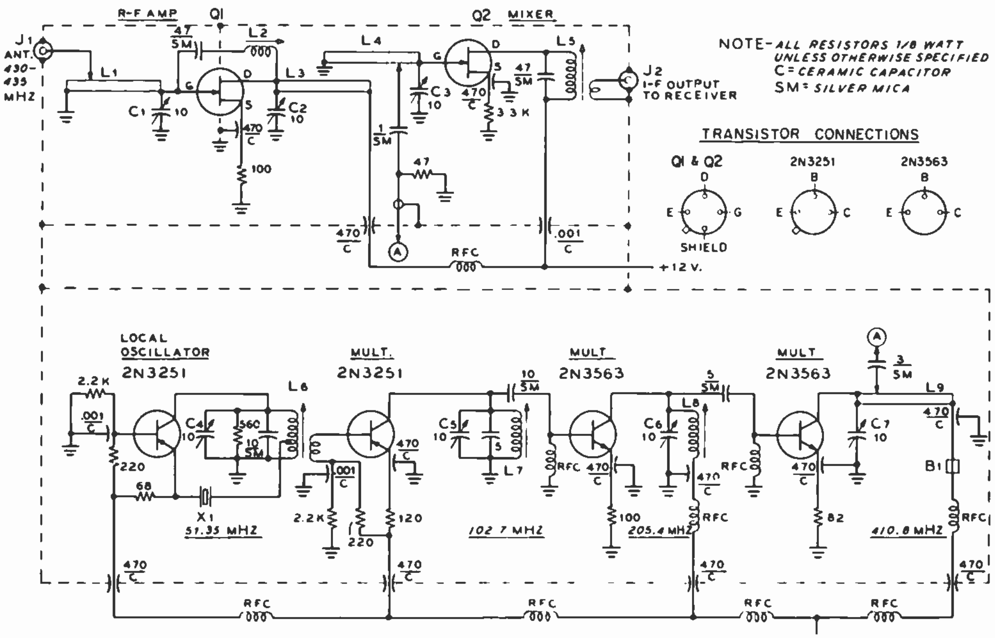


Figure 8

SCHEMATIC, 432-MHz FET CONVERTER

C_1-C_3, C_5-C_7—10-pf ceramic piston-type capacitor. (JFD 57G or Centralab 829-10)
C_4—10-pf capacitor (Johnson 160-107 or equiv.)
L_1—Copper strap, 1/4" wide × 1 1/4" long with 5/16" foot at ground end, Silver plated. Grid-dip to 432 MHz
L_2—4 turns #22 e. on 3/16" slug-tuned form (J. W. Miller 4300-7, blue)
L_3—Copper strap, 1/4" wide × 2" long. Supported at "cold" end by feedthrough capacitor. Silver plated. Grid-dip to 432 MHz
L_4—Same as L_3. Tap 1-pf capacitor at approx. midpoint of line. Adjust tap for best noise figure. Grid-dip to 432 MHz
L_5—Tune to 21.2 MHz. 15 turns #22 e. on 1/4" form. (J. W. Miller 4500-6, purple). Two-turn link made of hookup wire
L_6—8 turns #20 e., 3/8" diam. × 5/8" long. Tap 2 3/4 turns from ground end. (J. W. Miller 4400-3, green). Two-turn link made of hookup wire. Grid-dip to 51.35 MHz
L_7—4 turns #20 e., 5/16" diam. air-wound, 1/4" long. Grid-dip to 102.7 MHz
L_8—2 1/2 turns #20 e., 5/16" diam. air-wound, 1/2" long. Grid-dip to 205.4 MHz
L_9—Same as L_8. 1 1/4" long. Grid-dip to 410.8 MHz
RFC—5 μH (J. W. Miller 9340-14, or equiv.)
J_1, J_2—Coaxial receptacles, BNC type UG-625/U
B_1—Ferrite bead. Stackpole 7D (57-0180)
X_1—Crystal, 51.35 MHz, type HC-6/U
Q_1, Q_2—2N5245, TIS-88 or 2N4416

This converter provides a noise figure of better than 5 decibels, primarily determined by mixer noise.

Addition of the second remote r-f amplifier stage at the receiving antenna brings the over-all noise figure of the combination down to 3 decibels, or better.

The FET converter may be used by itself except for the most serious low-noise reception, when the addition of a remote r-f amplifier stage ahead of the converter unit becomes a necessity. The operational flexibility afforded by the use of a separate r-f amplifier cannot be overemphasized, since it permits the user to modify the system when the availability of new and better transistors or FET devices allow superior noise figure to be achieved as the state of the art advances. Some present bipolar transistors provide slightly superior noise figure at 432 MHz than that of the FET devices used; however, the problem of receiver overload from local signals or radar is ever present at 432 MHz and the use of FETs is highly recommended for the serious vhf operator.

Converter Circuitry The schematic of the 432-MHz FET converter is shown in figure 8. A 2N5245 FET is used in a neutralized common-source configuration as a strip-line r-f amplifier. This stage is followed by an inductively coupled 2N5245 as a common-source strip-line mixer stage. Both local oscillator and received signal are mixed in the gate circuit of the FET and the resulting i-f signal is taken from the drain circuit.

The local-oscillator chain is an integral part of the converter, using a 51.35-MHz fifth-overtone crystal oscillator and harmonic multipliers to generate a 410.8-MHz mixing signal. The intermediate frequency for 432-MHz reception, then, is 21.2 MHz. Bipolar transistors are used in the local-oscillator chain, two 2N3251 transistors serving as oscillator and frequency multiplier to 102.7 MHz, followed by two 2N3563 multiplier stages to 205.4 MHz and 410.8 MHz, respectively. The mixing signal is capacitively coupled to the 2N5245 FET mixer stage and oscillator coupling is adjustable to provide optimum mixer noise figure. While the converter is normally used for spot-frequency operation near 432 MHz, the tuned circuits are sufficiently broad so as to permit good reception one or two MHz removed from the frequency of adjustment, without further tuning of the converter.

Converter Construction The 432-MHz FET converter is constructed on a piece of copper-clad glass epoxy circuit board measuring 5″ × 7″. A matching shield case is made of an aluminum chassis of the same dimensions and 1½″ deep. Placement of the major components may be seen in the photographs of figures 7 and 9 and layout drawing of figure 10.

A shield plate measuring 1″ × 6″ is cut from circuit-board material and soldered along the center line of the converter, with a second shield measuring 1″ × 2″ placed across the r-f amplifier transistor to reduce intrastage coupling and to permit proper neutralization. A small opening is cut in the lower edge of the shield and the 2N5245 r-f transistor is placed in the opening, supported from the shield plate by soldering the shield lead of the transistor to the plate. The various other transistor leads go directly to the associated components, no sockets being used in the construction of the converter.

All parts should be laid out on the circuit board in a preliminary mockup before holes are drilled and assembly started. The quarter-wavelength strip-line circuits are fabricated first, since these determine the placement of other parts and the positioning of the FETs. The lines are made of short lengths of ¼-inch wide silver-plated copper strap, and are mounted 5/16″ above the circuit board. Lines L_1 and L_4 are grounded directly to the copper foil of the board and lines L_3 and L_9 are supported by means of silver-mica "button" feedthrough capacitors. The lines are supported at their high-impedance ends by piston-style tuning capacitors mounted to the circuit board.

The r-f stage neutralizing coil (L_2) is mounted in a vertical position adjacent to the intrastage shield, one terminal of the coil being soldered to the junction of capacitor C_2 and strip line L_3 and the other terminal lead passing through a small hole in the shield to terminate at the 47-pf silver-mica coupling capacitor. This, in turn, is soldered to the barrel of piston capacitor C_1. The slug of coil L_2 may be adjusted

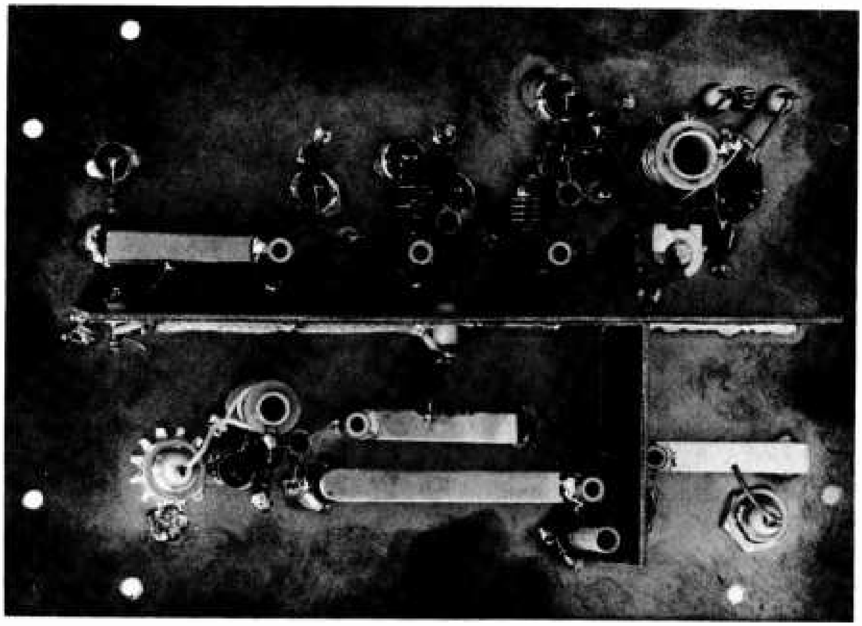


Figure 9

UNDER-CHASSIS VIEW OF CONVERTER

Placement of components may be compared with the layout drawing of figure 10. Interstage shields are made of strips of circuit-board material. Local-oscillator chain is at top of chassis with the crystal socket at the right. Antenna receptacle J_1 is at lower right, with i-f output receptacle J_2 at lower left. Transistors and FET's are mounted in an inverted position by their leads.

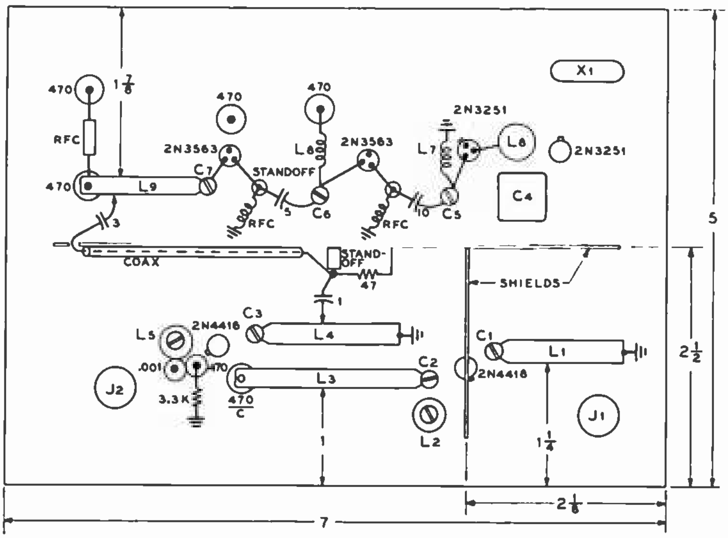


Figure 10

COMPONENT LAYOUT FOR 432 MHz CONVERTER

Tuned circuits are grid-dipped to resonance before FETs and transistors are soldered in place.

through a hole drilled in the surface of the circuit board.

All components in the r-f amplifier and mixer stages are grouped closely about the transistors which are supported by their leads in an inverted position. The positioning of strip lines L_3 and L_4 determine the coupling between the r-f amplifier and the mixer. Coupling is not critical, and an edge-to-edge spacing of ¼″ is satisfactory. The outer edge of strip line L_3 is located one inch away from the edge of the circuit board.

The local-oscillator chain is placed on the opposite of the center shield partition, with the 410.8-MHz strip-line circuit placed about 1⅞″ away from the edge of the circuit board, as shown in the layout drawing. The mixing signal is coupled through a 3-pf capacitor into a short length of miniature coaxial line which runs along the center shield back to a position near strip line L_4, at which point the line is coupled to the mixer stage through a 1-pf capacitor. The coaxial line is made up of a short length of *Teflon* insulated wire run through a small flexible copper braid. The braid is soldered to the partition and terminated at a small standoff insulator mounted on the partition in the vicinity of strip line L_4. Oscillator-chain wiring is conventional, the components being supported by their leads and several midget *Teflon* standoff insulators, as shown in the layout drawing.

Testing the Converter Upon completion, the wiring should be checked against the schematic diagram. Before the transistors are wired in place, it is suggested that the various tuned circuits be grid-dipped to the operating frequencies with the aid of a vhf grid-dip oscillator. Once the circuits are in approximate resonance, the transistors may be wired in the circuit. In order to prevent damage to the transistors, it is recommended that the transistor lead be grasped with long-nose pliers between the body of the transistor and the joint, the pliers acting in the manner of a heat sink, preventing the transistor element from being damaged by the heat of soldering. The transistors are mounted in an inverted position in each case, with the leads trimmed to about ¼″ length.

Converter adjustment follows the procedure outlined for the previously discussed units. The local-oscillator chain may be monitored in a nearby receiver for proper crystal operation. As the tuned circuits are near resonance, it is possible to peak the circuits on a local signal, while monitoring the output signal in the receiver used as an i-f strip. Preliminary alignment may be accomplished by peaking all circuits to achieve maximum signal strength.

Once the circuits are in close alignment, the converter should be adjusted for best noise figure. The various tuned circuits, including the neutralizing circuit of the r-f amplifier are all adjusted with this end view in mind. The adjustments are best accomplished with the aid of a noise generator to achieve lowest internal noise consistent with optimum gain. One accurate but time-consuming neutralization technique is to inject a 432-MHz signal into the drain circuit of the amplifier and monitor the signal in the input (gate) circuit with an auxiliary receiver. A low-level signal from a tone-modulated signal generator is coupled by a coaxial line into tuned circuit C_3-L_4. The auxiliary receiver, or converter, is coupled to the input circuit through receptacle J_1 and the tone-modulated signal monitored. The strip lines are peaked for maximum signal transfer, and the neutralizing coil (L_2) adjusted for minimum signal transfer.

Lacking a second 432-MHz receiver, another neutralizing technique is practical and somewhat quicker than the one outlined above. The converter is tuned for best noise figure with the aid of a noise meter, and system gain is then observed as the supply voltage to the r-f stage is varied from that value at which the stage was previously neutralized. When the drain voltage is varied, the reverse transfer capacitance (drain-to-gate feedback capacitance) is varied as a first-order effect. Thus, if it is found that stage gain increases as the drain voltage is increased or decreased, it is an indication that the stage is improperly neutralized. When neutralization is achieved, any change in drain voltage will result in a decrease in stage gain.

A final check on neutralization may be made by operating the converter with the input circuit short-circuited and open-circuited. The converter should remain stable under these conditions, without signs of regeneration, or "burbles" heard in the i-f strip receiver. The final check, of course, is determined by proper converter operation on weak signals, and the neutralization adjustment should be considered just one of several tools used to achieve lowest noise figure.

Frequency Stability After a period of use it was found that the received frequency drifted a sufficient amount with change of temperature so as to make very narrow-band c-w reception difficult. Satisfactory frequency stability was achieved by placing the local oscillator crystal in a metal can made from a 35-mm film holder. The can was stuffed with strips of foamed fiberglas type material to reduce temperature excursions. The crystal was thus sufficiently insulated from rapid temperature changes so as to permit good short-term stability necessary for serious c-w and SSB reception.

The Remote R-F Amplifier It was decided to place a second r-f amplifier at the receiving antenna to overcome transmission-line loss and improve reception. The schematic of the amplifier used is shown in figure 11. Essentially, this amplifier is a duplicate of the one used in the 432-MHz converter and is built in a small double-compartment brass box measuring 2″ × 3″ × 1″ in size. A brass shield is soldered across the center of the box and the FET is mounted in a small hole cut in the shield. The transistor may be easily removed and another one substituted in order to achieve the best possible noise figure. The +12-volt supply is run from the station to the antenna by a separate power lead, although the voltage may be coupled to the r-f amplifier via the center conductor of the coaxial line, if desired.

Use of a separate r-f amplifier, even if located at the operating position, is a great advantage at 432 MHz since experimental units may be built and tested without the necessity of disrupting the station receiver. As newer and better FET's and transistors appear, they may readily and easily be tested for operation in simple r-f amplifier "black boxes" built especially for transistor evaluation.

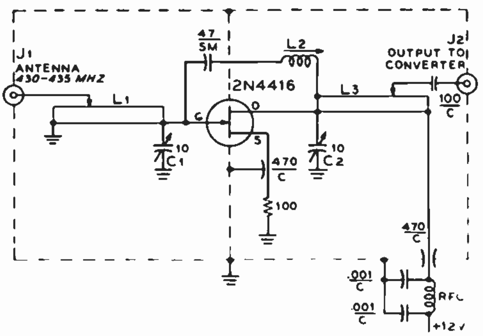


Figure 11

SCHEMATIC 432-MHz REMOTE R-F AMPLIFIER

Tuned circuits and components are the same as listed for r-f amplifier of figure 8.

20-3 A Low-Noise Preamplifier for 1296 MHz

The availability of low-noise, uhf-type solid-state devices has made practical the design and construction of a high-performance, low-noise preamplifier for 1296 MHz use. Shown in this section is a unit designed and built by W6KQG that makes use of two *Hewlett-Packard* HP-21A series NPN silicon transistors. These devices provide a typical noise figure of 3 db at 2 GHz, 4.5 db at 4 GHz, and usable gain to above 8 GHz. The gain figure, typically, of a single device is over 11 db at 2 GHz and over 6 db at 4 GHz. The transistors are available in several different style packages having various lead dimensions. Full data on these devices may be obtained from the many *Hewlett-Packard Co.* field engineering offices.

Preamplifier Circuitry and Construction A view of the 1296-MHz preamplifier is shown in figure 12 and the schematic is given in figure 13. A microstrip-line configuration is used, the unit being built on a TFE teflon fiberglas printed-circuit board of 1/32″ thickness, measuring 8″ × 5″ in size. The board is manufactured by the *3M Company* and has a dielectric constant of 2.5. The design shown is based on the use of this particular

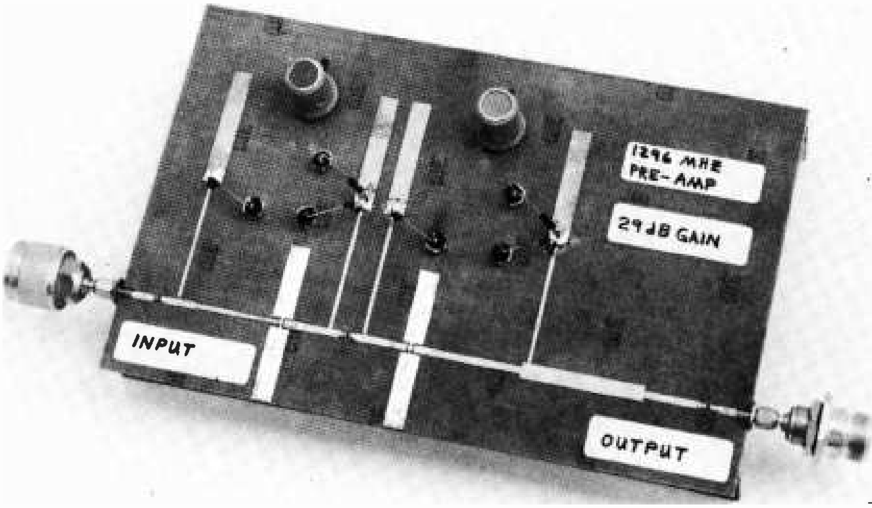


Figure 12

LOW-NOISE, SOLID-STATE 1296-MHz PREAMPLIFIER

This high gain, low noise converter is fixed-tuned and provides over 29 decibels gain over the 1.2- to 1.35-GHz frequency range. Using two Hewlett-Packard HP-21A series transistors, the unit provides a noise figure of 3 db or better at 1296 MHz. A microstrip-line design is used. The input circuit is at the left with d-c voltages fed to the transistor through quarter-wavelength microstrip lines. The transistors are connected in a grounded-emitter circuit. The emitter leads are grounded to the underside of the board and also have quarter-wavelength grounding lines placed atop the board. Ceramic chip capacitors are placed at the midpoint of the microstrip lines for d-c isolation. The amplifier is housed in an "r-f tight" aluminum cabinet to reduce r-f pickup from the nearby transmitter.

board and substitution of other board material is not recommended.

The circuit includes an input matching network (L_1, L_3), an interstage coupling network (L_4), and an output network (L_5, L_7). The preamplifier input impedance is a nominal 50 ohms and the first network transforms this value to the source impedance of the input transistor which is approximately $69.5 + j21$ ohms. Two microstrip-line transformer segments of 62 and 70 ohms respectively achieve the transformation.

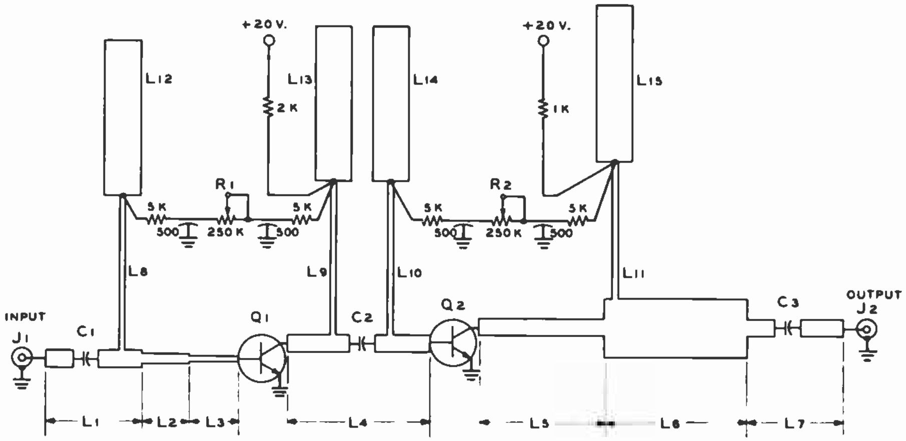
The various microstrip lines are in a nonuniform dielectric composed of air and teflon, with the average dielectric constant depending on both individual dielectric constants and the geometry and impedance of the microstrip line. An "effective wavelength" factor can be computed from these constants.

The interstage network consists of a quarter-wavelength section of 50-ohm microstrip line which delivers a near-perfect complex conjugate impedance match between the output impedance of the first stage and the input impedance of the second over the frequency region centered about 1296 MHz.

The output impedance of the second stage is approximately $115 - j82.5$ ohms and the conjugate impedance is matched to a nominal 50-ohm output termination by means of a two section microstrip-line transformer, the first section having an impedance of 50 ohms and the second 28 ohms.

In all cases, microstrip-line impedance and transformation are controlled by the choice of length and width of the line for a given board thickness and dielectric constant. PC board tapes are available in the following widths which, when used on this board material provide the impedance values given in parenthesis: 0.050″ (70 ohms), 0.062″ (62 ohms), 0.093″ (48 ohms), 0.125″ (40 ohms), 0.200″ (28 ohms), and 0.250″ (24 ohms).

The transistors are configured in the



LINE	L1	L2	L3	L4	L5	L6	L7	L8 - L11	L12 - L15
Z_0	48	62	70	48	48	28	48	70	24
L IN λ	—	.082	.085	.25	.215	.25	—	.25	.25
L IN INCHES	1.00	.542	.526	1.58	1.38	1.55	1.00	1.60	1.54

Figure 13

SCHEMATIC, 1296-MHz CONVERTER

C_1, C_2, C_3—100-pf chip capacitor about .05" × .05" × .09". American Technical Ceramics, Varidyne, or equivalent

Q_1, Q_2—Hewlett Packard HP-21A series NPN silicon transistors

Note: Emitter straps of Q_1, Q_2 are grounded to under-board foil by means of copper straps, as shown in figure 14. In addition, half-wavelength grounding lines are placed atop the chassis as shown in the photograph. Fixed resistors are 1/10 watt.

grounded-emitter mode. The dual emitter leads of each device are grounded to the copper foil on the bottom of the board, which serves as a ground plane, by the mounting technique shown in figure 14. In addition, quarter-wavelength grounding lines are placed atop the chassis to ensure that the emitter-to-ground impedance is very low. Ceramic chip capacitors are placed at the midpoint of microstrip lines L_1, L_4, and L_7 to provide d-c isolation while base and collector voltages are applied to the transistors by means of linear isolation chokes. These chokes consist of a quarter-wavelength of isolation line (L_8, L_{11}) used in conjunction with a quarter-wavelength grounding line (L_{12}, L_{15}). Isolating resistors are placed at the low-potential junction of these lines to decouple the bias potentiometers.

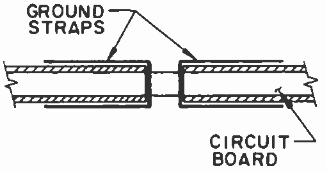


Figure 14

GROUND STRAPS FOR EMITTER TERMINALS OF TRANSISTOR

The preamplifier assembly is designed to fit within an "r-f tight," shielded cabinet to protect the devices from the strong r-f field of a nearby transmitter. All power and switching leads out of the cabinet are passed through suitable filter capacitors.

The preamplifier is fixed-tuned and provides over 29 db gain over the 1.2- to 1.35-GHz frequency range, with a noise figure better than 3 db.

A set of design notes covering the development of this amplifier may be obtained by request from the editor of this Handbook.

Figure 15

80-METER SOLID-STATE RECEIVER

This 80-meter tuner is a single conversion, solid-state receiver designed for 80-meter DX work or as an i-f strip for h-f and vhf converters. Featuring cascaded mechanical filters, IC double-balanced modulators, and a MOSFET input stage, the receiver combines excellent sensitivity with good dynamic signal range. An Eddystone dial is used to provide good bandspread and close readout. The receiver is a good building project for the beginner who wishes to start on a practical solid-state design.

20-4 A Solid-State 80-Meter Receiver

This receiver is a single-conversion, solid-state unit designed for 80-meter DX reception or for use as an i-f strip for vhf converters (figure 15). When used as a tunable i-f amplifier in conjunction with crystal-controlled converters, it serves as a basic "building block" for an excellent multiband receiving system. The receiver is designed and built by VE3GFN.

The receiver includes in its design such features as regulated power supplies, varactor diode tuning of the r-f amplifier and the bfo, a MOSFET input stage and IC double-balanced modulators for the mixer and product detector. The use of cascaded *Collins* mechanical filters for c-w reception makes this receiver virtually immune to most adjacent-channel interference; this, together with a front end that exhibits excellent overload characteristics and good suppression of stray mixing products, results in excellent over-all performance.

Due to the simple, straightforward front-end design and the use of uncritical circuitry throughout, this receiver is ideal for the builder who wishes to "cut his teeth" on solid-state construction. The unit can be aligned, from start to finish, with very basic test equipment, or even by ear alone.

While the unit shown in the photographs is quite large, the size of the receiver is completely the choice of the builder, and is dictated by the size of the available cabinet, the choice of tuning dial mechanism, and the size of the variable-frequency-oscillator tuning capacitor. The circuit itself is capable of miniaturization to a fine degree, if desired.

Additional extras, such as an S-meter, variable agc time constant, and noise limiter may be added to the basic circuit at the builder's choice.

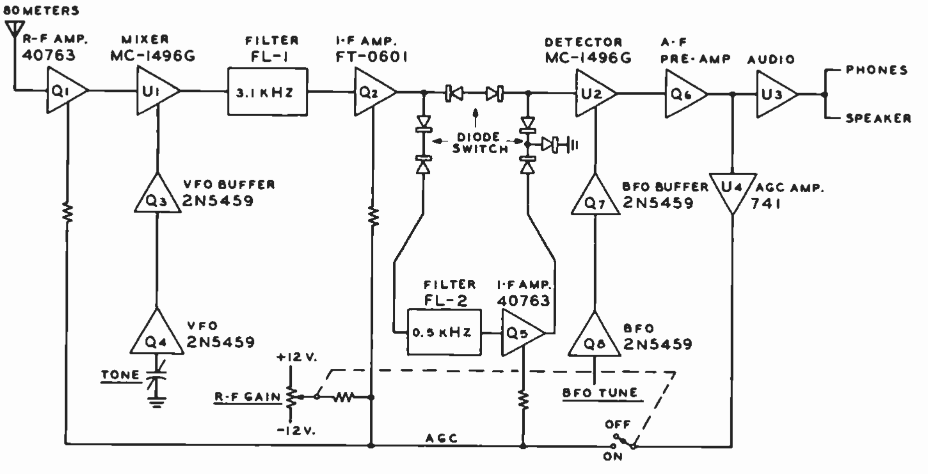


Figure 16

BLOCK DIAGRAM OF 80-METER RECEIVER

Featuring integrated circuits for mixer, detector, audio, and agc systems this compact receiver is a good beginner's project for the amateur interested in solid-state techniques. Selectivity is provided by two mechanical filters, the narrowband filter being cut into the circuit by diode switches. Audio agc is provided for the i-f and r-f stages.

The Receiver Circuit A block diagram of the receiver is shown in figure 16. The r-f module is shown in figure 17. The r-f amplifier (Q_1) employs a diode-protected, dual-gate MOSFET in a common-source configuration, with inductively coupled input and output circuits. The *RCA* 40673 or *Fairchild* FT 0601 devices can be used with equal success in this stage, and also in the i-f amplifier stages. The input and output circuits are tuned with varactor diodes, controlled by the *R-F Tune* panel adjustment (R_2). Stage tuning is quite broad and little retuning is required over the entire range. Despite this, image response is very good and no out-of-band signals can be detected under normal operation.

Stage gain is controlled by varying the potential on the second gate of the MOSFETs by means of the *RF Gain* panel adjustment (R_1). Gate #2 is bypassed to ground; this not only provides a decoupling action, but also forms part of the agc time constant. The second gate is connected through a current-limiting resistance directly to the agc circuit and through an additional 2.2-megohm series resistance to the arm of the gain control potentiometer. With agc disabled, there is a slight delay in gain change as the *RF Gain* control is varied due to the bypass capacitor being charged through the series resistance.

The mixer employs an IC double-balanced modulator (U_1, figure 18). Either the *Motorola* MC-1496G or the *Fairchild* μA *796* may be used. The stage provides about 20 db of conversion gain, a very good sinusoidal output wave, wide dynamic range, and high rejection of spurious mixing products. With the antenna disconnected no sign of spurious responses are evident throughout the tuning range.

The vfo is a JFET, with a similar device used as a buffer and impedance-matching device. While a 2N5459 is used, a 2N3819 or other similar low-power FET will also work nicely in this circuit. The oscillator is extremely stable, with no thermal compensation necessary for normal amateur operation. The tuning capacitor is a surplus unit, straight-line-frequency tuning, with tracking slots in the outer rotor plate, and excellent bearings. The best capacitor available should be used, with particular attention paid to the quality of the bearings. To ensure stability, tuned-circuit wiring is done with #12 solid wire. A piston capacitor is

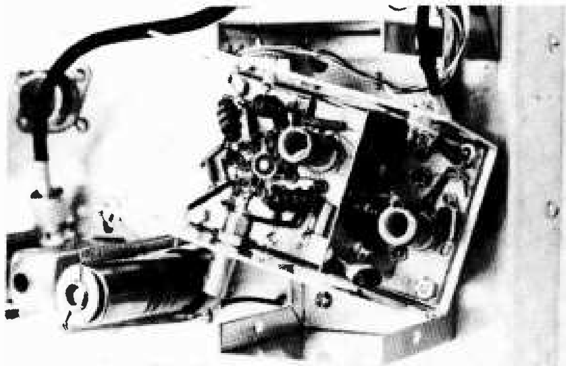


Figure 17

R-F AMPLIFIER/MIXER MODULE IN RECEIVER

The r-f amplifier and mixer module is built into a small minibox. It can be assembled and tested before it is placed in the receiver. The semiconductors are mounted upside down by their leads and are soldered directly into the circuits. Control leads are brought into the enclosure via teflon feedthrough terminals. The r-f stage is at the right of the internal shield plate and the mixer is at the left. The BNC fitting for oscillator injection is at the bottom of the box.

placed across the tuned circuit for band-edge adjustment and the oscillator coil is wound on a slug-tuned form to facilitate tracking adjustments. The vfo assembly is shown in figure 19, and the schematic is shown in figure 20.

To improve dial tracking over the c-w segment of the 80-meter band, the range of the receiver is limited to 3.5 to 3.8 MHz. The vfo will exactly track the bandspread scale of the *Eddystone* dial for the first 100 kHz and with little error for the next 100

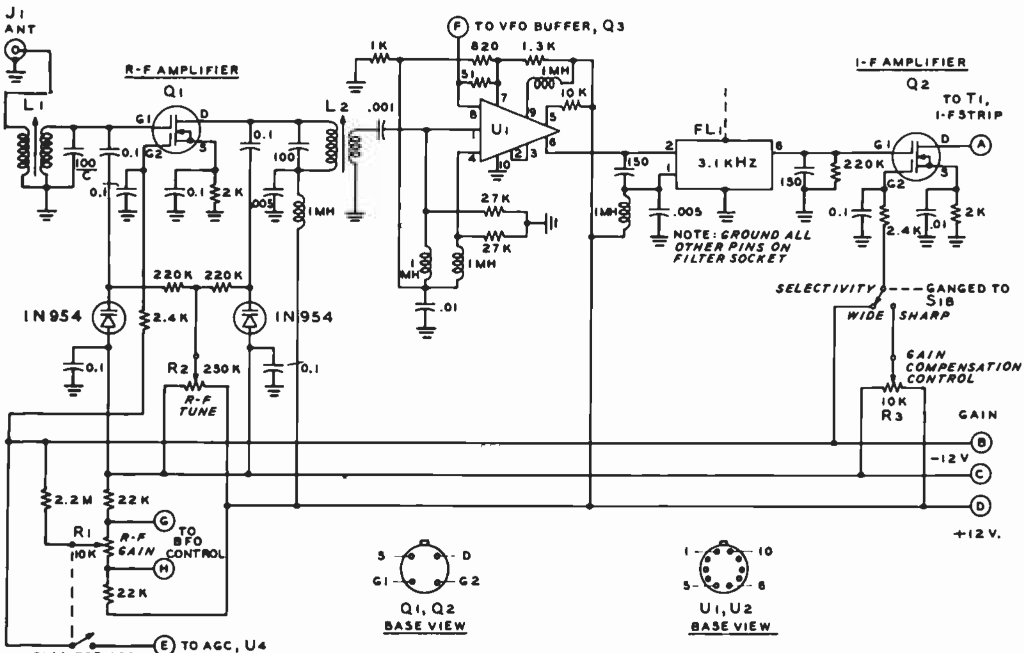


Figure 18

R-F AMPLIFIER AND MIXER MODULE

L_1, L_2—50 turns #29 e. on Cambion 1534-2-1 form, 1/4-inch diam. 17 μH., Q = 55. Link is 10 turns #27 e. on "cold" end
Q_1, Q_2—40673 or FT 0601
R-f chokes are 1 mH, 35 ma. Nytronics or Miller 153A1
U_1—μA 796 or MC 1496G
FL_1—Mechanical filter, Collins, 3.1-kHz bandwidth. Center frequency 455.0 kHz
Note: Capacitors across mechanical filter adjusted for maximum stage gain. All fixed resistors 1/2 watt.

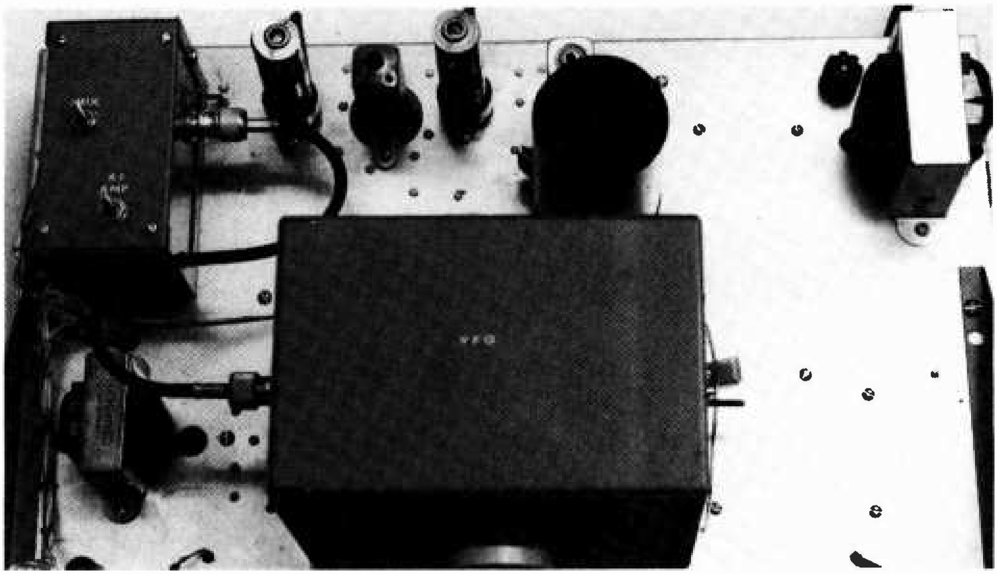


Figure 19

TOP VIEW OF THE RECEIVER

The large center box houses the vfo, most of the enclosure being taken up by the large variable capacitor. The r-f amplifier/mixer module is in the upper left-hand corner. The mechanical filters plug into 9-pin sockets along the rear of the receiver. At right is the power transformer for the positive supply, with the filter capacitor directly behind the vfo enclosure. The small control knob in the lower left corner is the gain compensation adjust for the i-f strip.

kHz. After that, calibration becomes increasingly nonlinear. Over any portion of the tuning range, however, it is possible to change frequency by a specific number of kHz by reading the bandspread scale directly. For amateur operation, the ability to shift frequency accurately is probably of more use than absolute measurement of a specific frequency. The vfo can be adjusted, if desired, so that the receiver covers the entire 500 kHz of the 80-meter band, but at the cost of less-accurate dial calibration.

The signal output of the mixer stage is coupled to a *Collins* mechanical filter having a bandwidth of 3.1 kHz and a 455-kHz center frequency (figure 21). The filter drives a 40673 IGFET amplifier which has a dual function: when using the receiver with "wide" selectivity (3.1-kHz filter only) this stage is gain-controlled by the *RF Gain* panel control. The stage serves as the i-f amplifier and drives the detector directly. When the receiver is switched to "sharp" selectivity, the amplifier drives an-

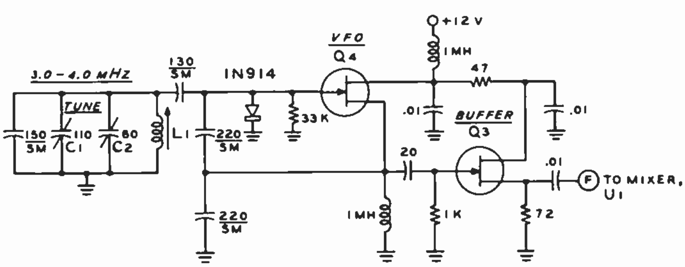


Figure 20

VFO AND BUFFER MODULE

C_1—110 pf. Surplus capacitor from "Command" transmitter or equivalent
C_2—60-pf JFD or Centralab 823-AZ
L_1—15 turns #27 e. on National XR-60 form (7 µH)
Q_3, Q_4—2N5459
Note: R-f chokes are 1mH, 35 ma
Note: Module output is 200 millivolts, peak to peak into 50-ohm load.

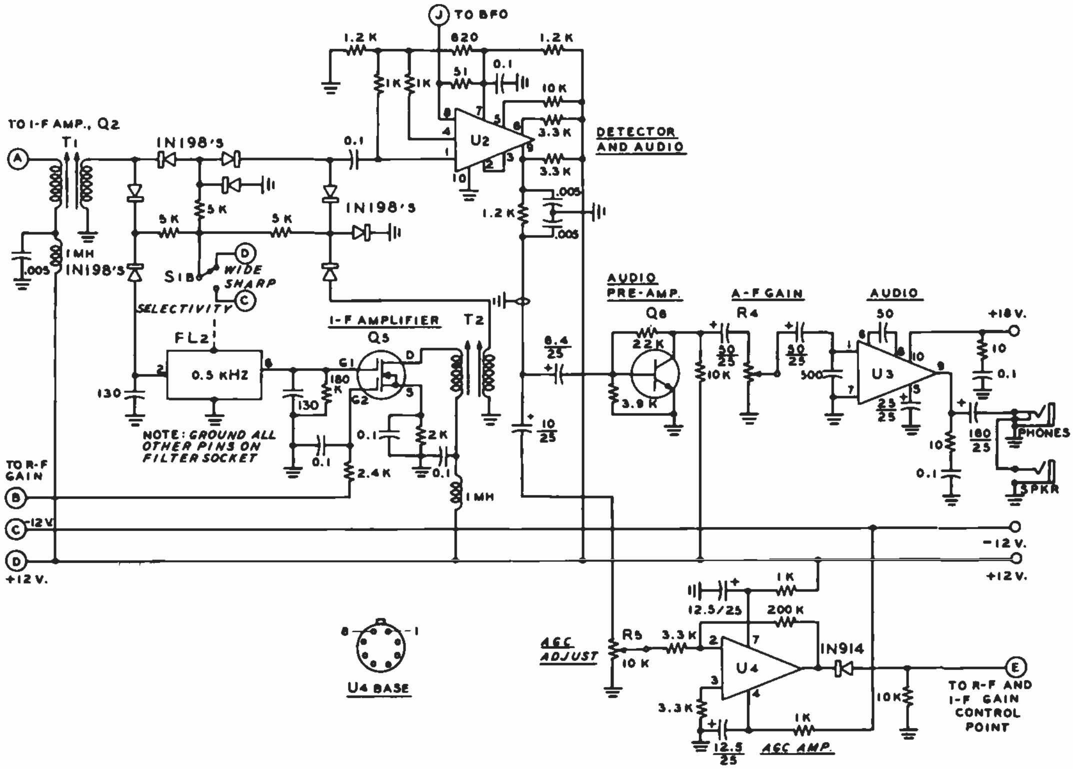


Figure 21

DETECTOR AND AUDIO MODULE

FL_2—Mechanical filter, Collins. 0.5-kHz bandwidth. Center frequency 455.0 kHz. Note: Adjust filter capacitors for maximum stage gain
U_2—μA 796 or MC 1496G
U_3—HEP 593
U_4—741 operational amplifier
T_1, T_2—455-kHz i-f transformer. Miller 12C2

other mechanical filter having a 500-Hz bandwidth. Diode gating is employed to route the i-f signal. In this case, the gain control of the amplifier is changed by means of the *Selectivity Wide/Sharp* control (S_1) to a calibration control which is adjusted so that the gain of the amplifier exactly compensates for the insertion loss of the additional mechanical filter. In this way the receiver gain is constant regardless of i-f selectivity. The "sharp" filter drives a second FET which serves as the i-f amplifier when using the "sharp" filter and which is gain-controlled from the *RF Gain* control.

Cascading the two filters results in exceptional adjacent-channel selectivity, as interfering signals outside the "broad" filter passband are attenuated over 100 db when both filters are in use. To achieve maximum rejection, shields are placed across the filter sockets to isolate inputs and outputs and wiring is routed well away from the filter area to avoid stray coupling around the filters.

Gain-control switching of the two FET amplifiers in the i-f system is done to place maximum receiver gain in front of the filters. The i-f gain is about 30 db maximum, and is an example of how receiver gain is distributed throughout the system to achieve freedom from overload and self-oscillation of high-gain stages.

The detector is a second double-balanced modulator used in a product-detector configuration (figure 22). It is extremely sensitive, with excellent noise characteristics and provides good rejection to signal feedthrough from the bfo. The high detector sensitivity

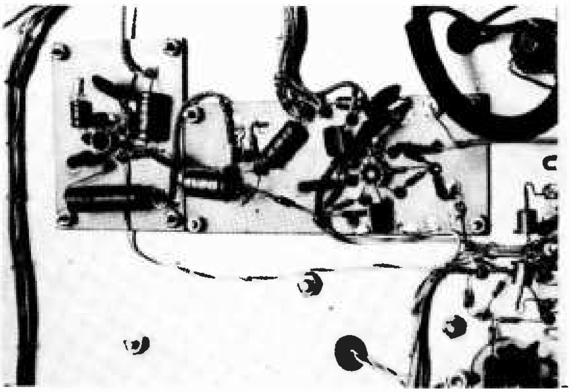


Figure 22

DETECTOR AND AUDIO-STAGE BOARDS

The printed-circuit boards are mounted under the receiver chassis after being tested separately. The detector and audio preamplifier are on the right board and the audio output stage is on a separate board, at the left. Shielded leads connect to the audio-gain control on the panel. Visible at lower right are the i-f switching diodes.

allows a minimum of i-f amplifier gain to be used.

The bfo uses a 2N5459 JFET with a similar buffer stage (figure 23). The bfo is variable in frequency as an extra convenience for c-w operation, for which the receiver is primarily intended. Frequency is controlled by a varactor diode, and tuning is done by the front panel *BFO* control. For ease of tuning, the voltage across the control is limited, restricting the tuning range to about 2 kHz either side of the 455-kHz center frequency. The slug of the bfo transformer is adjustable for proper centering of frequency (figure 24). Output from the bfo buffer stage is about 50 millivolts peak-to-peak, which is sufficient to drive the detector, yet low enough so that stray coupling to other circuits from the bfo board is minimized.

Signal output from the detector, after passing through an RC low-pass filter, drives a 2N4124 audio preamplifier, which in turn drives a HEP-593, 2-watt IC audio output amplifier. A panel jack is provided so that low-impedance earphones can be used with the receiver; the speaker being disabled when the earphones are plugged in.

The agc system controls the gain of the r-f amplifier and the common i-f amplifier stages. Audio is sampled after the panel control and amplified in a 741 operational amplifier, which saturates on any substantial signal. The op-amp is provided with an input level control to set the agc attack threshold. The output of the op-amp is rectified and the negative d-c voltage used to vary the #2 gate potential of the controlled stages, thus decreasing gain as input signal level increases.

The agc-decay time constant is a compromise between SSB and c-w operation, and the agc can be disabled by means of a switch on the rear of the *RF Gain* control.

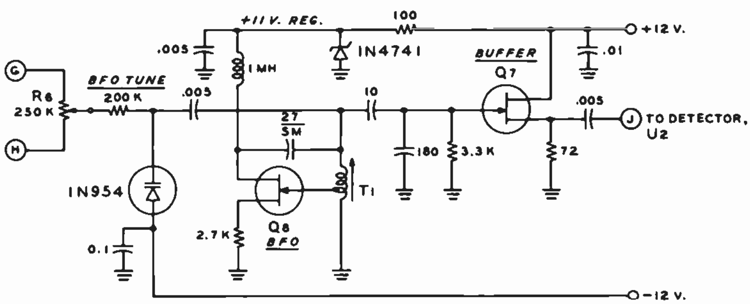


Figure 23

BFO MODULE

Q_7, Q_8—2N5459
T_1—455-kHz bfo transformer. Miller 1727

Note: Buffer output is 50 millivolts, peak to peak into 50-ohm load.

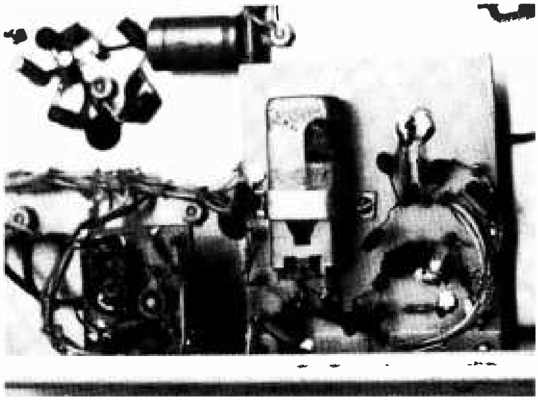


Figure 24

AGC AND BFO CIRCUIT BOARDS

At the left is the agc board with the control potentiometer mounted on it. To the right is the bfo board. The bfo transformer is held to the board with a strap placed over it.

Receiver Construction The receiver consists of a number of modules and subchassis units, with only the power supplies and i-f strip built on the main chassis. The r-f amplifier/mixer and the vfo are built as separate, shielded modules. The bfo, agc, audio, and detector are constructed on separate subchassis, but not installed in shielded compartments (figure 25).

All these subchassis assemblies are built on glass-epoxy printed-board stock, suitably

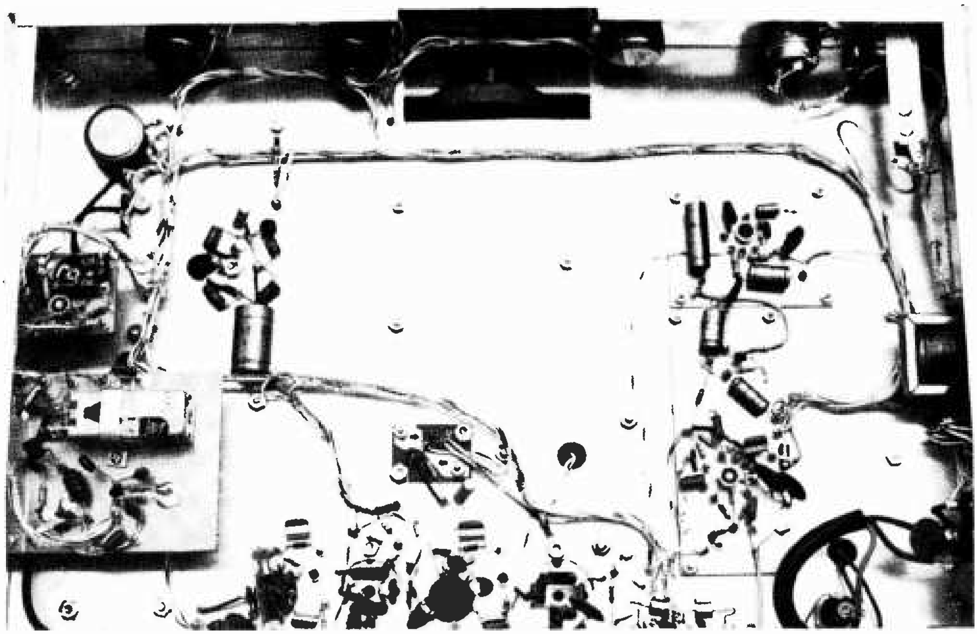


Figure 25

UNDER-CHASSIS VIEW OF THE RECEIVER

The bfo board and agc boards (figure 24) are in the lower left corner of the receiver. The detector and audio boards (figure 22) are at the right. The i-f strip is along the lower edge of the chassis; note the shield plates across the filter sockets and the arrangement of the diode gates. The components for the negative-voltage supply regulator are just above the upper corner of the bfo board, while the positive-voltage supply regulator is just above the i-f strip. The small transformer at the right of the chassis powers the dial lamps.

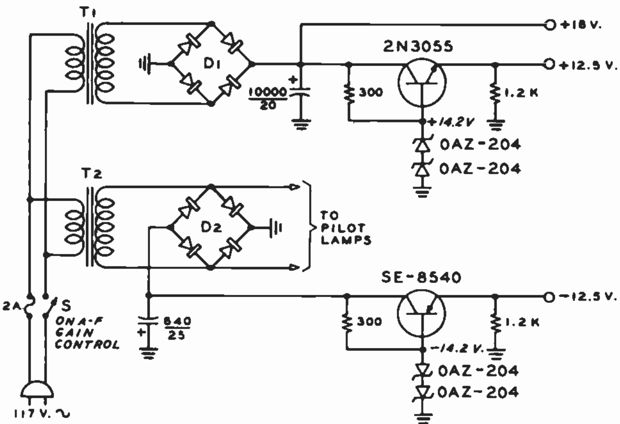


Figure 26

RECEIVER POWER SUPPLY

D_1, D_2—Bridge made of 1N4003 diodes or equivalent (100 volt, 1 amp)
T_1—12.6-volt, 2-ampere transformer
T_2—12.6-volt, 0.3-ampere transformer

cut to size, with press-fit *teflon* terminals mounted where necessary to support components. Semiconductors are soldered directly into the circuits and ground connections are made to the copper foil of the board. This assembly technique takes advantage of all the good features of p.c. boards, but eliminates the time and effort necessary for layout and fabrication of the printed board. In some cases, the teflon terminals are mounted on the main chassis itself, as in the i-f strip construction. The power-supply schematic is shown in figure 26.

All signal-carrying lines are miniature 50-ohm coaxial cable or teflon-insulated shielded wire. After all wiring is completed, the various leads are cabled and secured with plastic lacing cord.

The receiver is built on an aluminum chassis 11½″ × 16½″, custom made to fit the available cabinet. The panel is 17″ × 9″. The *Eddystone* dial requires a large slot to be cut in both chassis and panel so rigid material (1/16″ stock) is used for making the parts. This material can easily be slotted with a small saw or a nibbling tool.

Testing and Alignment The vfo is most easily adjusted for optimum tracking by the use of an auxiliary receiver monitoring the output frequency. Decreasing the vfo coil inductance raises the frequency but reduces the tuning range. Increasing the capacitance of the piston trimmer lowers the frequency The tuning range is adjusted to 3045.0—3345.0 kHz.

The power supplies should be checked for proper voltages with all stages connected. With speaker connected, the *AF Gain* at maximum and a low-level 1-kHz tone applied to the input of the a-f preamplifier, a loud tone should be heard in the speaker. Do not operate the audio stages at full level on a constant tone for a long period of time as damage may result to the output IC.

With a 455-kHz modulated signal injected at the input of the first i-f filter and the agc circuit disabled, the i-f transformers are adjusted for maximum d-c voltage at the agc output circuit while monitoring across the 10-megohm resistance with a high-impedance voltmeter, and with a 1-μfd capacitor temporarily connected across the resistor for a charge capacitor. Check the filter gating circuit and adjust the gain control of the "sharp" filter gain compensation stage for equal output level with either filter position.

A signal at 3.65 MHz is now applied to the input of the receiver and the *RF Tune* control is set to midposition. The vfo is adjusted to tune in the signal and r-f amplifier coils are peaked for maximum audio output. Hold the *RF Gain* control at a minimum position while doing this adjustment. Finally, the bfo control is adjusted for zero beat at a midposition setting.

Once the receiver is operating, the agc input-level control can be adjusted to preference. Maximum level may tend to desensitize the receiver somewhat; about one-half to three-quarters of maximum level is optimum.

When the receiver is used as a tunable i-f for high-gain converters, care should be taken to avoid crossmodulation from overload. Converter gain should be held to a minimum, or an attenuator placed between converter and receiver to keep over-all system gain to a reasonable level.

20-5 An Advanced Solid-State Deluxe Amateur Band Receiver

This receiver, designed and built by VE3GFN, takes advantage of the many recent advances in communications circuit techniques, and illustrates their application to amateur equipment (figure 27). Features such as IC power-supply regulation, varactor diode tuning of front-end circuits and variable oscillators, integrated double-balanced modulator mixers, diode switching of filters and tuned circuits, and a solid-state digital readout counter, are included in the design.

Modular construction is used as much as possible. Most of the circuits are built as separate, shielded modules, and are tested and aligned as such, completely independent

Figure 27

SOLID-STATE DELUXE AMATEUR BAND RECEIVER

This advanced receiver covers the amateur bands between 80 and 10 meters in 500-kHz segments. Featuring direct readout, varactor diode tuning, integrated circuit double-balanced modulators, and diode switching, the modularized receiver is an ideal construction project for the advanced amateur. The direct read-out escutcheon is at the upper left of the panel, with the KILOCYCLES-BAND switch directly below it. Readout is to 100 Hz. The large knob to the right is the tuning control, with the three pre-set channel switches at the right of the panel, the R-F TUNE, R-F GAIN, and A-F GAIN controls and earphone jack are along the lower edge of the panel. To the left of the main tuning control are the AC ON switch and the I-F SELECTIVITY switch. Two crystal filters provide optimum selectivity for SSB and c-w modes. A separate speaker sits atop the receiver. Construction is simplified by building the receiver in modules, each of which may be tested independently before the receiver is assembled.

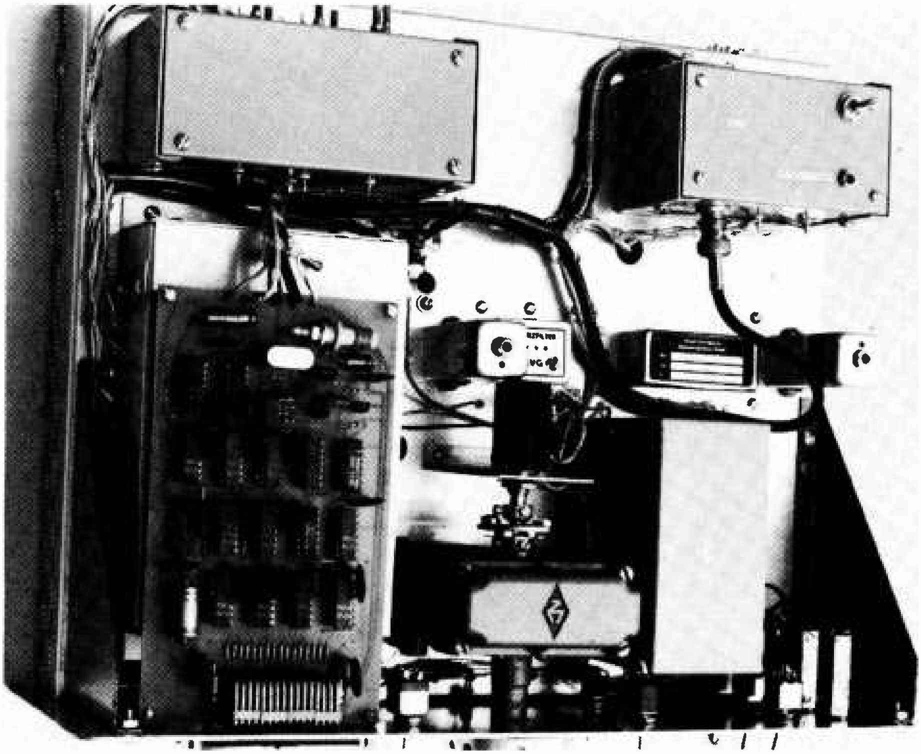


Figure 28

TOP VIEW OF RECEIVER CHASSIS

Placement of the major modules may be seen in this view. The vfo module (A) is at the right, rear of the chassis, with the heterodyne mixer module (E) in the left rear corner. The front-end bandswitching module (C) is almost completely hidden by the digital counter board (D). At the right, behind the front panel is the bfo module (F). The vfo tuning potentiometer and i-f crystal filter are at the center of the chassis. Note that the National PW-O gear reduction drive is set back from the panel to allow room to mount the various control switches. Switches and "trimpots" for the "Preset Channel" function are in the right front corner. Modules are pretested before mounting on the receiver chassis.

of the receiver system (figure 28). This technique makes system modification easy, simplifies testing and alignment, and contributes greatly to freedom from spurious mixing products and circuit radiation. Input/output specifications are provided for each module, allowing the receiver to be duplicated module by module; by meeting the various module requirements, the builder is assured that his system will function properly when assembled.

The detailed description of this receiver is along modular lines as well, with the description of each module including necessary circuit theory, construction details, and electrical specifications.

The Receiver Circuit The receiver is single conversion on all amateur bands, 80 through 10 meters; coverage of the entire 10-meter band is included (figure 29). For good stability and to avoid tracking problems, the loscal-oscillator injection voltage is derived from the mixing product of a 5.0- to 5.5-MHz *variable-oscillator module* (A) with a *crystal-oscillator module* (B), the frequency of which is changed for each band. On 20 meters only, the variable oscillator is not mixed with a crystal oscillator, but drives the signal-path mixer directly. The frequency of the variable oscillator is counted by the *digital counter and display module*

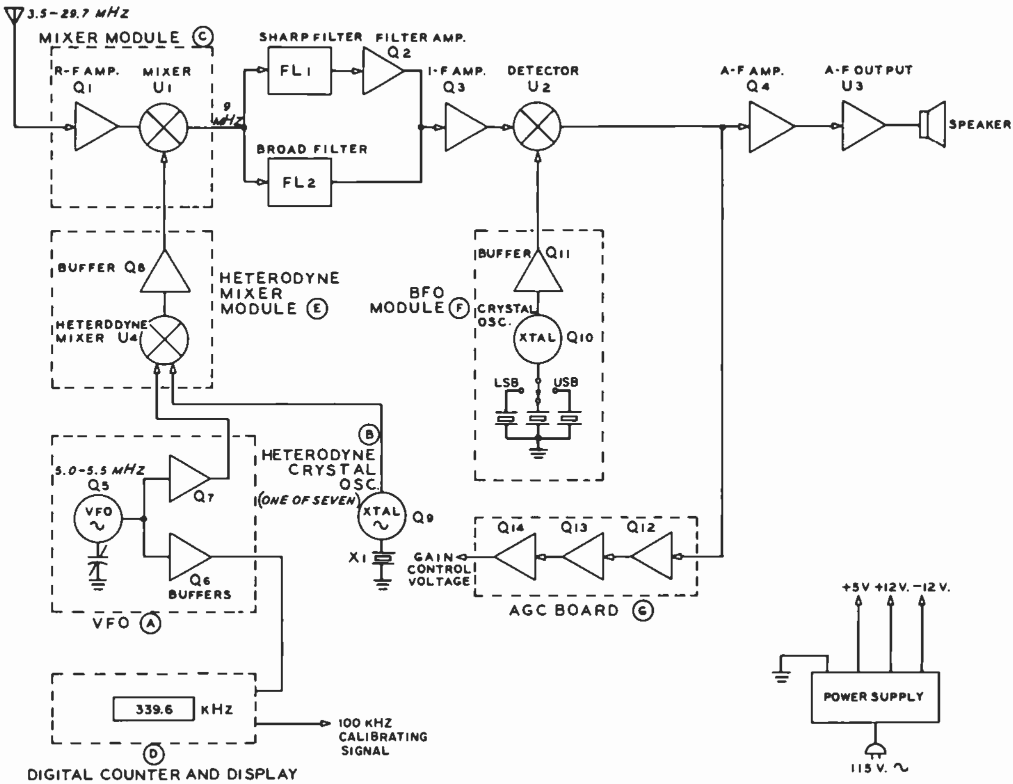


Figure 29

BLOCK DIAGRAM OF DELUXE AMATEUR BAND RECEIVER

The receiver is built and described in modules. The mixing signal is derived from a heterodyne mixer module (E). The mixing frequency is changed for each band. On 20 meters only, the variable oscillator (A) is not mixed with the heterodyne crystal oscillator (B), but drives the mixer (U_1) directly. The frequency of the variable oscillator is counted by the digital counter (D) to 100-Hz resolution and displayed as "kHz above the band edge." For 20 meters, the frequency shown is 14,339.6 kHz. Bandchanging is accomplished by a rotary switch in the mixer module. Separate i-f filters provide c-w and SSB selectivity and switchable bfo crystals provide upper and lower sideband.

(D), to 100Hz resolution, and displayed as "kHz above the band edge." While a display of exact frequency may be more convenient from the operator's standpoint, the system used is simpler, and enables the digital counter to be built and tested as separate a module as any other, completely independent of the bandswitch.

Band changing is accomplished by a rugged rotary switch built into the *front-end mixer module* (C). Extra wafers on this switch control the heterodyne crystal oscillators, the switching of the heterodyne mixer output circuits, and the variable oscillator output, through or around the heterodyne mixer system.

Signal input from the antenna is amplified by a dual-gate MOSFET r-f amplifier (Q_1) which is tuned from the front panel (*RF Tune*) by controlling the bias of varactor diodes D_1 and D_2 in the input and output tuned circuits (figure 30). The amplified antenna signal then passes through the signal-path mixer (U_1), a double-balanced IC modulator. Local-oscillator injection for this mixer comes from the *heterodyne mixer module* (E), and is the sum of the variable-oscillator frequency, and the frequency of one of the heterodyne crystal oscillators. The heterodyne mixer (U_4) has diode-gated tuned circuits in the output to control the mixing frequency. The variable oscillator

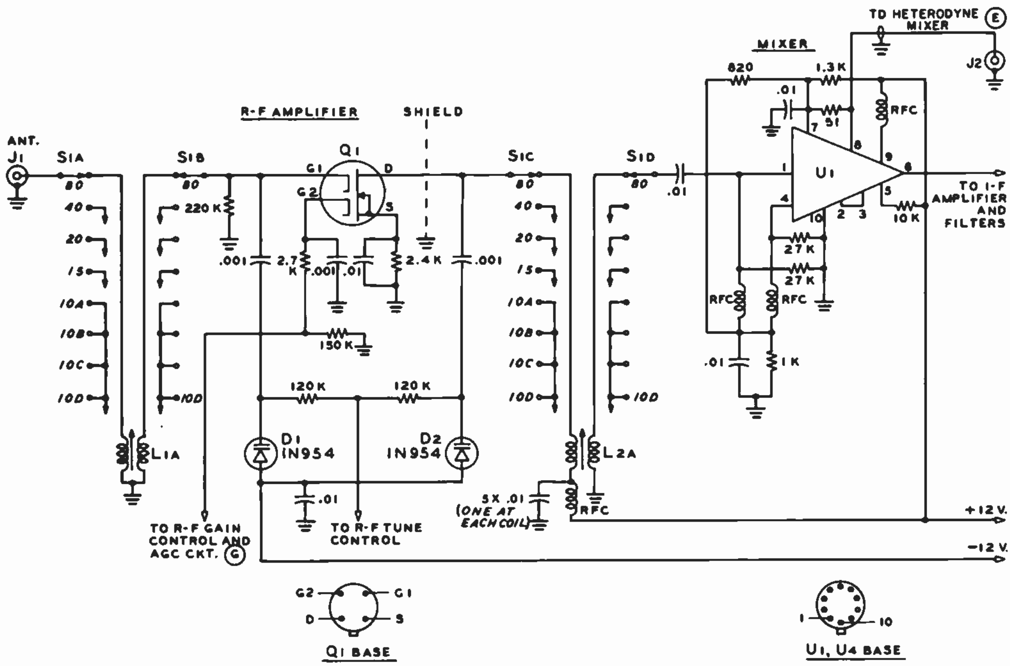


Figure 30

SCHEMATIC, FRONT-END BANDSWITCHING MODULE

J_1, J_2—BNC connector, UG-185/U
L_1, L_2—See Table 1
Q_1—RCA 40673 or Fairchild FT 0601
RFC—1-millihenry, 35-ma. J. W. Miller 10F-103A1
S_1—3 deck heavy duty rotary switch, 6 pole, 8 position. Centralab JV-9037, or equivalent
U_1—Motorola MC 1596G or Fairchild μA 796
Note: All resistors ¼ watt

(Q_5) is a JFET circuit, varactor-tuned, of high stability. Injection to the signal-path mixer from the heterodyne mixer or the variable oscillator is controlled by diode gates through the bandswitch.

The output of the signal-path mixer (U_1) is the 9.0-MHz i-f signal, which passes through one of two crystal filters; either a 2.4-kHz filter for SSB, (FL_2), or a 500-Hz filter for c-w (FL_1). The choice of filter is made by the i-f *Selectivity Broad/Sharp* panel switch, which controls diode gates that direct the i-f signal to the filters.

The sharp filter has a dual-gate MOSFET amplifier (Q_2) after its output, which is adjusted for equal system gain when using either filter; the sharp filter has more insertion attenuation, making this necessary.

The i-f signal is amplified by a dual-gate MOSFET amplifier (Q_3) common to both filters, providing up to 20 db i-f gain.

The second detector is a double-balanced modulator (U_2) used in a product-detector configuration, obtaining its beat-frequency injection from the *beat-frequency oscillator module* (F), employing a crystal oscillator (Q_{10}) whose frequency is selected by the *USB/LSB* panel switch, or the i-f *Selectivity* switch, depending on the choice of c-w or SSB. The oscillator is followed by a buffer stage (Q_{11}).

The detector output drives a high gain bipolar audio preamplifier (Q_4), which has the *AF Gain* control in its output circuit, a measure designed to increase signal-to-noise ratio.

The audio output stage (U_3) is a two-watt integrated-circuit amplifier, with its own power-supply regulator.

Frequency readout is obtained from LED devices (light emitting diodes) in the *digital counter display module* (D) which are driven by a highly stable time base and decade counters. A 100-kHz crystal is used as a standard for the count. The frequency of the variable oscillator is read to 100 Hz.

General Construction Technique Most of the circuits in this receiver are hand wired on G-10 glass epoxy printed-circuit board subchassis, using teflon press-fit terminals at the interconnection points. Ground connections are soldered directly to the copper foil. Solid-state devices are soldered directly into the circuits, with no device sockets used except in the digital counter, the only module where printed-circuit technique was necessary to simplify duplication. In a few cases (the i-f strip, for example) the main aluminum chassis was used as the construction base.

This assembly technique is ideal for high-frequency circuit work; it is quickly and easily modified, and short-lead construction is easy. Employing the copper-board subchassis utilizes most of the advantages of printed-circuit construction, but eliminates the extra time needed for artwork and board fabrication. When the final design has been completed, it can be easily adapted to the usual printed circuits if desired.

Construction of Front-End/Bandswitching Module (C)—The front-end module contains the r-f amplifier (Q_1), the signal-path mixer (U_1), associated tuned circuits, and the receiver bandswitch. The schematic is shown in figure 30.

The r-f amplifier is a dual-gate MOSFET, providing up to 20 db of r-f gain. Gain level is set by means of the *R-F Gain* panel control, which adjusts the bias on the second gate of the MOSFET device. This is done in conjunction with the automatic gain control system, and the i-f amplifier is controlled in a similar manner at the same time.

The r-f input and output circuits are tuned by means of varactor diodes D_1 and D_2, the bias (capacitance) of which is controlled by the *R-F Tune* panel control. The amplifier is stable on all bands without neutralization. While ferrite beads on the MOSFET input and output leads might contribute to inherent stability, they were found to adversely affect r-f gain at the higher frequencies.

The signal-path mixer (U_1) uses an IC as a double-balanced modulator, which provides great attenuation to undesired mixing products. This IC device is used throughout the receiver for all signal-translation applications.

The local-oscillator injection for the signal-path mixer is obtained from the heterodyne mixer module (E) on all bands except 20 meters, where the variable-frequency oscillator module drives it directly. Input to U_1 at J_2 from the mixer module should by 50-300 mV p-p of sinusoidal waveform.

Additional wafers of bandswitch S_1 control the heterodyne crystal oscillators, the gating of the output tuned circuits in the

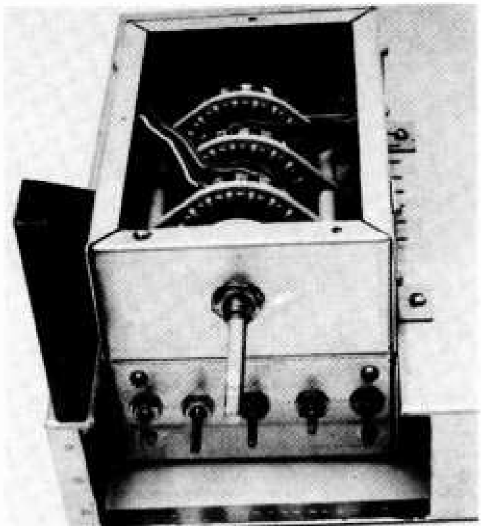


Figure 31

FRONT VIEW OF INSTALLED FRONT-END MODULE

Front-end module (C) is installed in cutout in main receiver chassis. The module is held in position by angle brackets on the sides. Module assembly is made of two aluminum chassis mounted back-to-back. Connecting terminals are on sides of the module. Digital counter board (D) mounts on top plate of module. R-f coils (L_1 series) are mounted to front of lower module chassis (left to right): 10-, 15-, 20-, 40-, and 80-meter coils.

heterodyne mixer module, and the diode gating of the variable-frequency oscillator module output.

The output of the front-end module (with the first i-f transformer connected) should be a sinusoidal waveform at 9.0 MHz, of a level about 40 db greater than the antenna signal level, with the r-f amplifier adjusted for maximum gain. Due to losses in link couplings, transformer couplings, etc., this gain figure is only a nominal one. Views of the front-end module are shown in figures 31, 32 and 33.

The bandswitching module contains the bandswitch, the r-f amplifier (Q_1) and the signal-path mixer U_1. It is the most complex and compact of the receiver modules, and its assembly will be simplified if the following step-by-step procedure (used in the construction of the prototype) is followed. See Table 1 for coil data.

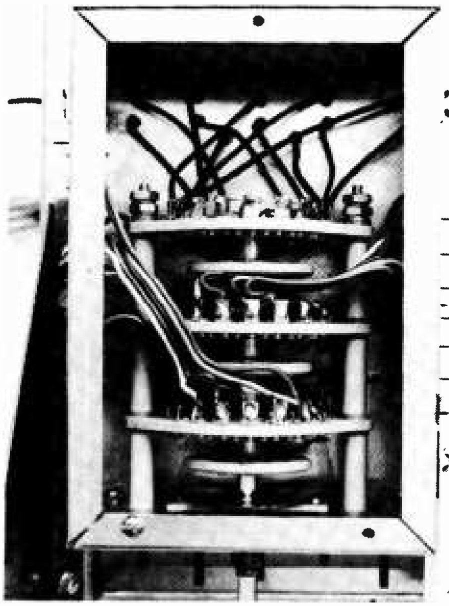


Figure 32

INTERIOR OF FRONT-END MODULE

Main bandswitch is centered in compartment with press-fit feedthrough insulators grouped near switch terminals. Wires running between switch points and underside of module (for control of crystal oscillators and heterodyne mixer gating) are kept as short as possible and laid out so as not to interfere with switch action.

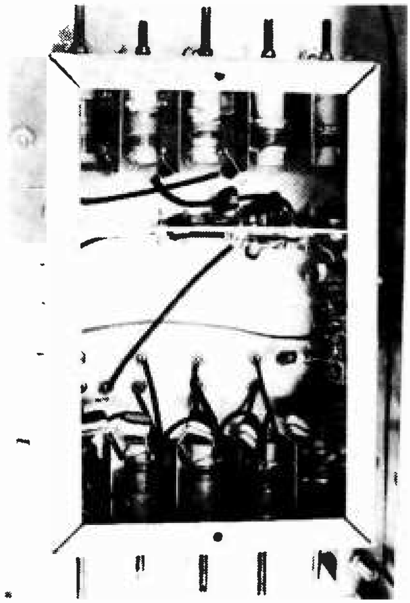


Figure 33

UNDER-CHASSIS VIEW OF FRONT-END MODULE

The horizontal shield across the enclosure contains the r-f amplifier FET and associated components. At the bottom of the compartment are the r-f input coils, with the detector coils at the top (rear) of the compartment. The signal-path mixer (U_1) is mounted on a small circuit board on the side of the enclosure. The shield sections between the coils are soldered to p.c. boards mounted on the front and back sides of the module chassis. Mounting holes for the coils are drilled through board and chassis. Note the feedthrough terminals from the bandswitch enclosure protruding through the clearance holes in the deck of the chassis.

The bandswitching mixer module (C) is built in two aluminum chassis, each 6″ × 4″ × 2″ mounted back-to-back, as seen in figure 31. The bandswitch is installed in the top chassis, and the solid-state circuits and coils in the bottom chassis. The contacts of the switch are wired to press-fit feedthrough terminals mounted in the bottom of the switch chassis; these terminals protrude through clearance holes drilled in the circuit chassis, and the coils and proper circuits are wired to them. Thus the switch is shielded from the r-f circuitry, yet leads are kept short. The bandswitching module is constructed as a separate assembly and

Table 1. R-F Amplifier and Mixer Coils (L_1, L_2)

Band (Meters)	L_1 and L_2
80	50 turns #29 e. on Cambion 1534/2/1 form, closewound. Inductance = 17 μH, Q = 55. Link = 10 turns #27 closewound on "cold" end. 100 pf connected across primary, 80 meters only.
40	35 turns # 29 e., as above. Inductance = 10 μH, Q = 85. Link = 7 turns #27, as above.
20	15 turns #29 e., as above. Inductance = 4.4 μH, Q = 70. Link = 4 turns, as above.
15	12 turns #27 e. on Cambion 1534/3/1 form, closewound. Inductance = 1.8 μH, Q = 115. Link = 3 turns as above.
10	8 turns #27 e., closewound, as above. Inductance = 0.8 μH, Q = 140. Link = 3 turns, as above.

mounted in a slot in the main chassis. Assembly of the Module is as follows:

Step 1. Cut ⅜" clearance hole in switch chassis front, center, to mount bandswitch. Do not mount the switch.

Step 2. Drill 9/64" mounting holes in all four corners of the switch chassis, allowing room for a 6-32 nut to cover the hole and clear the chassis corner. Place the two chassis back-to-back, and mark the centers for the mounting holes in the circuits chassis, using the drilled switch chassis as a template.

Step 3. Mount and secure the bandswitch in its chassis. Refer to the bottom-view photograph of the switch chassis (figure 32) for the feedthrough terminal layout. Note where the common switch arm of each wafer is on each deck, as its location requires more than a casual glance. Mark centers for the feedthrough terminals close to each wafer of the switch, being very careful of clearances when marking the terminals for the inner wafers. Now, remove the switch from its chassis, center punch the marked hole centers, bolt the two chassis together, and drill a centering hole through both chassis. Separate the chassis. The feedthrough terminals require a 9/64" hole, and the clearance holes should be enlarged to ¼".

Step 4. Install the switch in the switch chassis. Now examine the location of the feedthrough terminals, and the switch contacts to which they must be wired. The terminals for the inner wafers are almost covered by the switch, and are virtually inaccessible. These feedthrough terminals should be pre-wired (before the switch is installed) using 4" lengths of bare wire. As the switch is installed, these wires can be drawn up to the proper contacts, and wired to them, after sliding a length of insulating tubing over each lead. The switch contacts that are accessible (front and rear) should be prewired in a similar manner, and these wires run to the proper feedthrough teminals after the switch is installed. This completes the wiring of the switch chassis.

Step 5. Make up two coil-shield partition assemblies as shown in the underchassis photograph (figure 33), using the following procedure: Mount a pre-cut printed-circuit board on the front and back ends of the circuit chassis. Mark the centering holes for the five coil forms (L_1 series and L_2 series) on the outside of the chassis end pieces. Center punch and drill through both chassis and p.c. boards. Then enlarge the holes to the required size. Remove the boards and temporarily mount the coil forms, then mark the locations of the brass shield partitions. Remove the coil forms and solder the partitions into place. This method ensures proper clearance for the coil forms after the shield partitions are installed.

Step 6. Bolt the two chassis together, install all coils, and wire them to the terminals of the bandswitch.

Step 7. The r-f amplifier stage is built on the aluminum shield section which separates the r-f amplifier coils (L_1 series) from the mixer section. Install the wired r-f amplifier shield section, then install the mixer subchassis. Wire these to the proper switch terminals, and to the terminals on the side of the chassis for input, output, gain control, tune voltage, and supply lines. This completes the assembly of the Bandswitching Mixer Module.

Construction of Heterodyne Crystal Oscillator Modules (B)—The seven crystal oscillators for the heterodyne mixer are built as separate units (two to a circuit board) to avoid the bandswitching complexities and design compromises necessary in one oscillator covering 7.5 to 33.5 MHz (figures 34, 35, and 36).

The output of each oscillator should be a reasonably undistorted sinusoid, of 200-500 millivolts p-p amplitude, measured at the input (50 ohms impedance) of the heterodyne mixer (U_4). The series output attenuator circuit (10 pfd, 1K) prevents oscillator loading and eliminates any problems due to the oscillator signal being routed through the bandswitch and around the chassis.

The output of each oscillator should be measured after the series attenuator network, using an oscilloscope of at least 150-MHz bandwidth capability as an instrument of lesser bandwidth will not reveal harmonic distortion in the output. The frequency of each oscillator should be accurately checked as overtone crystals often have a penchant for operating on their second harmonic.

If a crystal does not oscillate, or if there is distortion in the output, the output tuned circuit probably requires adjustment. It may be necessary to increase the value of capacitor C_3 for the higher-frequency circuits, or that of capacitor C_2 for the lower-frequency circuits. If the crystal is sluggish, decrease the value of R_1, or eliminate R_2. If the transistor appears to be saturating, increase the value of R_1, or decrease the value of C_1. See Table 2 for coil and capacitor data.

Table 2. Heterodyne Oscillator Module—Circuit Details

Band (Meters)	C_1 (pf)	C_2 (pf)	C_3 (pf)	X_f (MHz)	$L_1A—L_1F$
80	27	100	—	7.5	35 turns #29 on Cambion 1536/2/1, L = 7.5 μH.
40	10	27	—	11.0	Same as above.
20	—	—	—		(No oscillator)
15	10	18	10	25.0	12 turns #29 as above. L = 1.5 μH.
10A	—	47	10	32.0	7 turns #29 as above. L = 0.55 μH.
Note: 10B, C, D same as 10A					

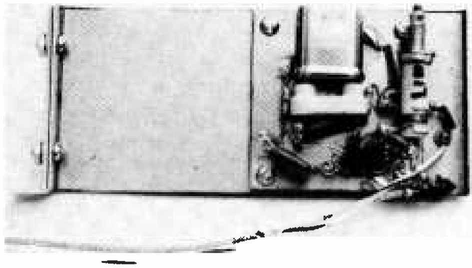


Figure 34

TYPICAL HETERODYNE CRYSTAL OSCILLATOR ASSEMBLY

The heterodyne crystal oscillator board (B) shown contains one of the seven crystal oscillator stages. The other three boards each have two oscillator stages on them and are visible in the main under-chassis view of the receiver. The oscillator slug-tuned inductor is adjustable from beneath the receiver.

Construction of Heterodyne Mixer Module (E)

The heterodyne mixer module (figures 37, 38, and 39) consists of an IC double-balanced modulator (U_4) and a JFET buffer (Q_8) to enable the mixer to drive a 50-ohm load, plus an output filter made up of seven tuned circuits and seven diode switches. The module is completely self-contained, with inputs being supplied through 50-ohm coaxial cables, and bandchanging accomplished by using one deck of the bandswitch (S_1E) to supply twelve volts to the appropriate diode switch.

The mixer/buffer output signal is 200-300 millivolts peak-to-peak of sinusoidal waveform into a 50-ohm load. The vfo injection signal (pin 1, U_4) must be 200 millivolts peak-to-peak (or less) and the crystal-oscillator injection signal (pin 8, U_4) must be 300 millivolts peak-to-peak (or less) of as sinusoidal a waveform as possible. Distortion in the input or output sine waveform increases the possibility of spurious frequencies occurring in the receiver system. The heterodyne mixer/buffer is not used on 20 meters.

Table 3. Heterodyne Mixer Module (E) Frequencies

Band (Meters)	Heterodyne Mixer Output Range (MHz)	Heterodyne Oscillator (MHz)
80	12.5 - 13.0 MHz	7.5 MHz
40	16.0 - 16.5 MHz	11.0 MHz
20	—	—
15	30.0 - 30.5 MHz	25.0 MHz
10A	37.0 - 37.5 MHz	32.0 MHz
10B	37.5 - 38.0 MHz	32.5 MHz
10C	38.0 - 38.5 MHz	33.0 MHz
10D	38.5 - 39.0 MHz	33.5 MHz

COIL DATA

80	L_4A—22 turns #29 on Cambion 1534/2/1 form. L = 5 µH. Q = 75. Note: 18-pf capacitor connected across coil.
40	L_4B—25 turns as above. L = 6 µH. Q = 75. Resonates with circuit capacitance.
15	L_4C—10 turns #27 as above. L = 1.5 µH. Q = 75.
10A—D	L_4D—L_4G—5 turns #27 as above. L = 1 µH. Q = 100. Note: 18-pf trimmer capacitor connected across each of the 10-meter coils.

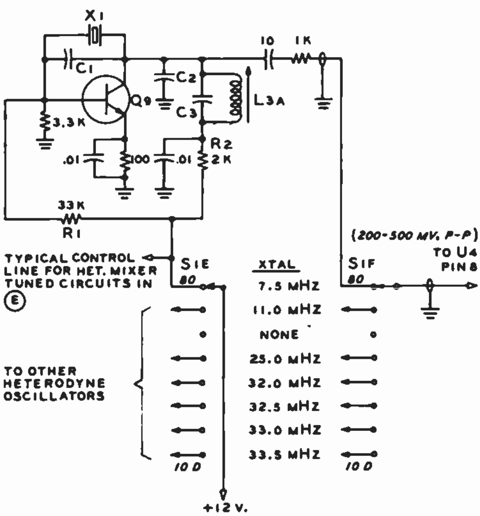


Figure 35

SCHEMATIC, HETERODYNE CRYSTAL OSCILLATOR (80 METERS)

Q_1—2N4124
Note: See Table 2 for component details. All resistors 1/4 watt

The mixer module output filter consists of seven tuned circuits (L_4A to L_4G), each of which is resonant at midband of one of the necessary injection spectrum frequencies required by the signal-path mixer as listed in Table 3.

Only one coil must be switched in the mixer output circuit at any time, and this coil completes the mixer d-c output circuit. The gating diode in series with the coil in use is forward-biased, and completes the circuit to pin 6 of the mixer (U_4), which is at an 11.3-volt level. The cathodes of the remaining switching diodes are at this same positive level, while the anodes are close to ground potential; and so are reverse-biased. The *Amperex (Philips)* BA-182 diodes (D_1—D_7) were selected for their low forward impedance and their small reverse capacitance.

A silicon diode (D_8) in series with the d-c current path to the rest of the mixer corrects the output d-c level imbalance caused by the diode junction voltage drop of the switching diodes. The series circuit of an r-f choke and 1K resistor across the mixer output (pin 6) drains additional current through the switching diodes.

The heterodyne mixer module is built into the "U" shaped portion of a minibox 2¼" × 2¼" × 5". Because of the large number of components that must be installed in this module, it is assembled using "layered" construction.

The heterodyne mixer consists of three separate sections; a chassis board, containing the components of the mixer itself (figure 39), the U-shaped section of the minibox, the sides of which have been drilled to accept the seven coils, and another chassis board, drilled to match the side of the minibox, on which components connected directly to the coils are mounted. Figure 37 showing the completed mixer module illustrates how the side of the minibox must be drilled for the coils; board containing the coil components is simply cut to match the dimensions and holes of the minibox side.

The mixer component board is wired, keeping all leads short, and using sufficient heat on the ground connections. When it is completed, mount it on the bottom of the U-section of the minibox. The board containing the coil components is then cut to size and installed on the side of the mini-

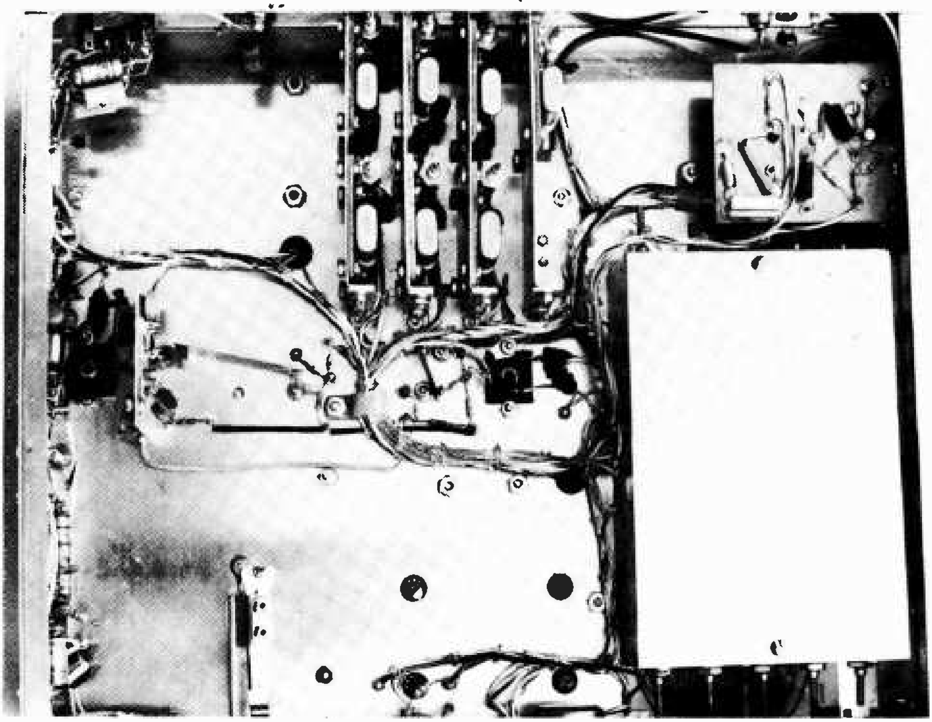


Figure 36

UNDER-CHASSIS VIEW OF RECEIVER

The heterodyne crystal oscillator boards (B) are located at the middle-rear of the main receiver chassis, with the agc board (G) to the right, in the corner. The i-f amplifier, detector (U_2), audio preamplifier (Q_1) and audio output IC (U_3) are mounted on boards along the left-hand side of the chassis. Decoupling capacitors are installed at the power connector in the upper-left corner of the chassis. The crystal filter diode gates and filter shields are just below the Heterodyne Oscillator Boards. The large Front-end Module is at lower right, recessed in a hole cut in the chassis.

Figure 37

THE HETERODYNE MIXER MODULE

The heterodyne mixer module (E) is built in an aluminum minibox. The seven coils (L_1 series) and gating-control press-fit feed-through terminals are mounted on the side of the box. The four corner screws secure the component chassis board inside the module.

box. The centers of the holes for the coils and the feedthrough terminals are then marked, and the holes drilled through both the minibox side and the coil component board.

The component board is now removed from the side of the minibox, and the coil bypass capacitors and the gating-diode biasing resistors are installed on the board. The board is then reinstalled on the side of the minibox and the coils are now installed in the module and wired in to both of the component boards. If the coils mounted nearest the bottom of the minibox are installed first, wiring will be easy. It is important to keep the diode leads as short as possible. For this reason, the 10-meter coils are mounted nearest the output pin of the mixer component board.

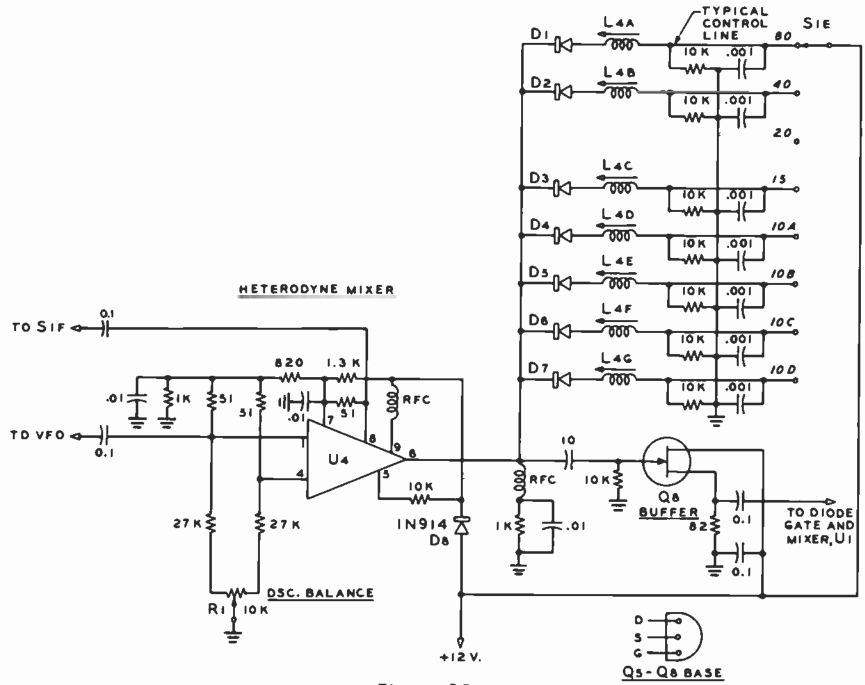


Figure 38

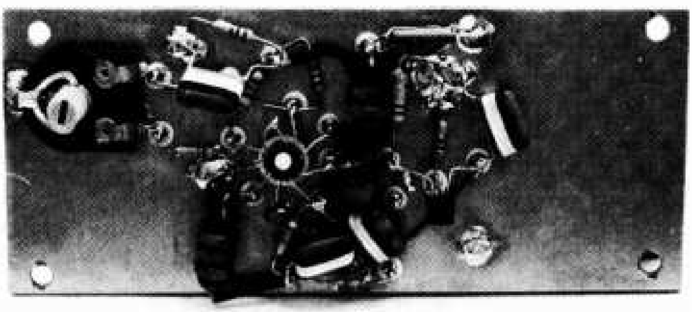
SCHEMATIC, HETERODYNE MIXER MODULE

Q_8—2N5459
U_4—Motorola MC 1496G or Fairchild μA 796
D_1-D_7—Amperex Phillips BA 182
L_{4A}-L_{4G}—See Table 3
RFC—1 millihenry, 35-ma. J. W. Miller 10F-103A1
Note: All resistors 1/4 watt

Figure 39

HETERODYNE MIXER COMPONENT BOARD

Integrated circuit U_4 is at center with the oscillator null potentiometer (R_1) at the extreme left of the board. The assembly is built as compact and flat as possible to allow clearance for the board below the coils, once the module is assembled. The buffer FET (Q_8) is at the right of the board.



To align the heterodyne mixer, inject a 5.0- to 5.5-MHz signal at the correct amplitude into one port (pin 1), and the specified frequency and amplitude to simulate the crystal oscillators into the other port (pin 8). Terminate the module output in 50 ohms, and check with a high-frequency oscilloscope and frequency meter or digital counter for the correct mixer-output frequency, and a clean sinusoidal waveform, as each tuned circuit is gated into the output by applying +12 volts to each of the gating terminals. Make sure, as each band is checked, that the correct crystal-oscillator frequency is being injected. Check each band for uniform output over the 5.0-5.5-MHz vfo range. On the 10-meter band, the output tuned circuits may interact to a certain extent, and the tuning process may have to be repeated several times. For each

Figure 40

Counter output from buffer stage Q_6 is at left (BNC connector) and oscillator inductor L_1 is atop the chassis, along with calibrating capacitor, C_1.

band, the idea is to obtain maximum output, and uniform output amplitude, at the correct frequency.

When testing the mixer, use the mixer module (pin 6) to drive the oscilloscope; do not try to probe the output circuits of the mixer itself with the oscilloscope, as even the small input capacitance of a high-frequency oscilloscope will load the tuned circuits.

Construction of Variable-Frequency Oscillator Module (A)—The vfo module consists of a voltage-controlled oscillator (Q_5), two buffer stages (Q_6 and Q_7) for the two necessary oscillator outputs, and a regulated supply for the frequency-control potentiometer that is derived from the positive and negative 12-volt rails (figures 40, 41, and 42).

The vfo itself covers 5.0 to 5.5 MHz, and is a Colpitts circuit using a JFET as the oscillating device, and a varactor diode (D_1) as the bandspread tuning capacitor. A piston trimmer (C_1) across the varactor diode circuit enables the limits of the tuning range to be accurately calibrated, once the inductance has been established. Elimination of the large plate-type variable capacitor often used in such circuits allows the vfo to be built in a much smaller enclosure than usual, and the JFET oscillator is inherently quite thermally stable.

The vfo covers a wide frequency range with the circuit constants provided, so the voltage range of the varactor diode is limited by the regulator circuits of the frequency-control potentiometer as shown in figure 43. The regulators also serve the purpose of stabilizing the tuning voltage, which directly affects the stability of the oscillator.

The tuned circuits for the vfo are wired with solid wire, with leads as short as possible; the entire vfo should be built mechanically stable and vibration proof. After construction, the output ports of the buffer stages should be checked for a sinusoidal output, of amplitudes approximating those indicated in figure 41.

The frequency range of the oscillator can be adjusted with the aid of an accurate frequency meter or digital frequency counter on the high-level output; the inductance adjustment will affect the width of the frequency variation, and the piston trimmer is used to set the lower-frequency limit of the tuning range. Adjusting each of these in turn, and checking the frequency range with each adjustment, should result in proper calibration being attained in short order.

The oscillator circuitry is built into a minibox measuring 2¼" × 2¼" × 4". The Oscillator Module is thermally coupled to the main chassis by cleaning the paint from the bottom of the module and cover-

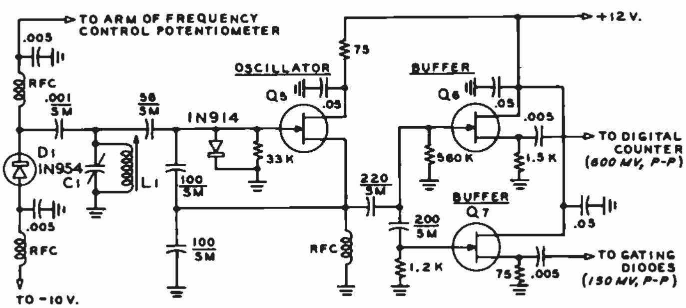


Figure 41

SCHEMATIC, VFO BUFFER MODULE

Q_5-Q_7—2N5459
C_1—Piston capacitor, 60 pf. MC 606Y
L_1—7-μH. Cambion 2419-2
RFC—1 millihenry, 35 ma. J. W. Miller 10F-103A1
Note: All resistors ¼ watt

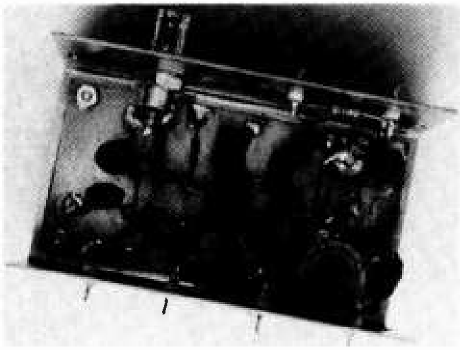


Figure 42

INTERIOR VIEW OF VFO MODULE

Components of vfo Module (A) are securely mounted to p.c. board bolted to one half of minibox. Tuned circuit (L_1-C_1) is at right, with buffer FETs just behind the BNC coaxial connector. Oscillator leads are short and heavy.

ing it with a layer of silicone grease before mounting it. By heat-sinking the module to the main chassis, excellent thermal stability is attained, even though the oscillator enclosure is quite small. No electrical temperature compensation is required.

The potentiometer used to control the frequency of the receiver is a matter of choice for the builder. Since a *National PW-O* gear reduction drive is used to drive the control potentiometer, the tuning of the receiver is very smooth, and reasonably slow even with a single-turn carbon unit of linear taper (figure 43). Using a single-turn potentiometer, a tuning rate of 35 to 60 kHz per tuning knob revolution is obtained, depending on the total coverage adjustment of the oscillator. This tuning rate enables the operator to cover the band at a fairly rapid rate, and is ideal for SSB operation. It is a bit fast, however, for tuning the band with the very selective 500-Hz c-w filter. Using a 10-turn helipot, (the *Amphenol* 2151B-104 is recommended for installation in the same space) a tuning rate of 3 to 5 kHz per knob revolution is easily attained. While it is tedious to tune the entire band at this rate, even with a "spinner" knob, it is perfect for use with the sharp filter.

By varying the total coverage adjustment of the oscillator, and by varying the voltage limits between which the tuning potentiometer wiper moves (a maximum of + and —12 volts), a wide variation in tuning rates can be attained.

Variable-Frequency Oscillator Output Gating Detail— The output of the vfo drives the heterodyne mixer module on all bands except 20 meters. The heterodyne

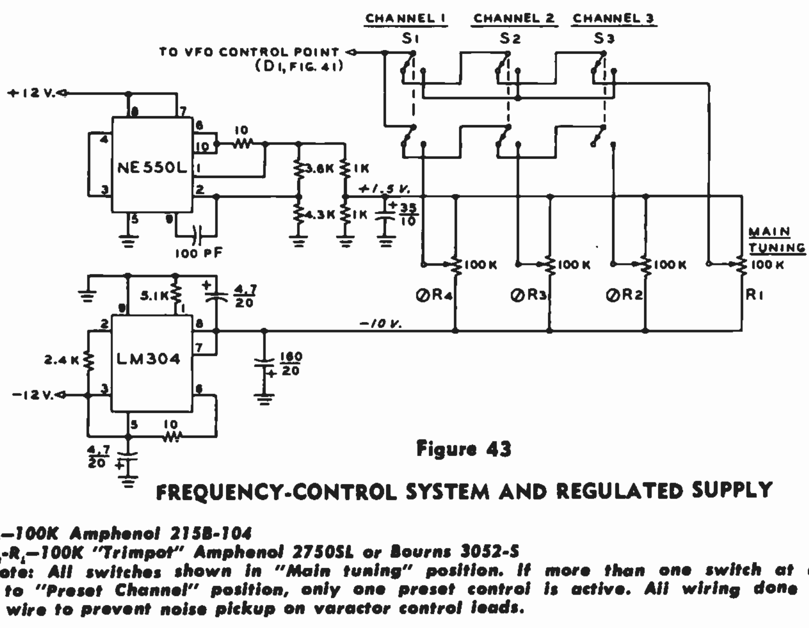


Figure 43

FREQUENCY-CONTROL SYSTEM AND REGULATED SUPPLY

R_1—100K Amphenol 215B-104
R_2-R_4—100K "Trimpot" Amphenol 2750SL or Bourns 3052-S
Note: All switches shown in "Main tuning" position. If more than one switch at a time is set to "Preset Channel" position, only one preset control is active. All wiring done with shielded wire to prevent noise pickup on varactor control leads.

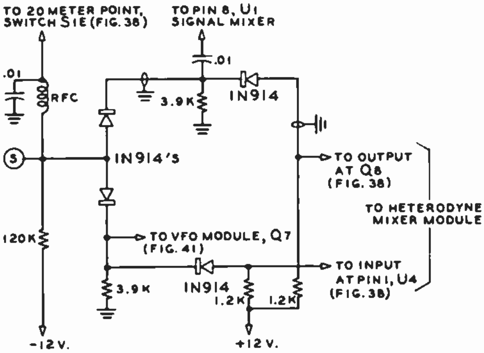


Figure 44

VFO GATING CIRCUIT

RFC—1 millihenry, 35 ma. J. W. Miller 10F-103A1
Note: All resistors 1/4 watt

mixer is driven by one of the seven crystal oscillators as well, and it drives the local-oscillator injection port of the signal-path mixer (U_4, pin 8) on all bands except 20 meters.

On 20 meters alone, the various crystal oscillators are disabled, and the vfo drives the signal-path mixer directly, while the heterodyne mixer input and output circuits are completely isolated from the system.

The circuit isolation and signal gating are accomplished by means of diode gates, which only change conduction states when the bandswitch is in the 20-meter position (figure 44). In this position, none of the crystal oscillators are supplied with d-c power; the oscillator outputs are not connected to the heterodyne mixer through the bandswitch, and the +12 volt line which normally powers the crystal oscillators now changes the status of the vfo output gates and the heterodyne mixer gates.

Diode gate status change is accomplised by changing the level at point S from —12 volts through a high impedance to +12 volts through a low impedance (the d-c resistance of the isolating choke) by means of the bandswitch. This forces the diode gates between the vfo and signal-path mixer to conduct, while at the same time reverse-biasing the diode gates at the heterodyne mixer input and output circuits.

VFO Module Frequency-Control System—The use of a varactor-diode tuned circuit in the variable oscillator allows several modes of frequency control not seen in the usual amateur receiver.

The usual mode, employing a front-panel knob ganged to a tuning element in the oscillator, is here approximated by the ten-turn potentiometer which controls the varactor bias and which is driven through a gear-reduction drive from the front panel *Main Tuning Control.*

Because the frequency of the oscillator is controlled by a varying d-c voltage, this receiver is also equipped for *channel* operation; by switching one of three switches provided, one of three preset frequencies may be selected (figure 43). These preset channels are adjustable from the front panel, by means of three panel-mounting *trimpots.* In case more than one channel switch is activated at the same time, the circuit is arranged so only one channel is actually enabled, even though all three switches may be in the channel position. Only when all three switches are in the *Main Tuning* position is frequency control by means of the panel tuning control possible. The *trimpot* channel adjusts may be easily set by a small screwdriver, while watching the digital counter; they will adjust the vfo frequency independently of band changing, of course. Once set, it is impossible to knock them out of adjustment accidentally. This feature will be found useful for net operation, for DX-ing, and for cross-frequency operation.

Remote Operation—This receiver is easily adapted for remote operation: For mobile work, the receiver can be located in the rear of the vehicle. A small control box near the driver would require only a speaker and a multiturn potentiometer to replace the main tuning potentiometer (it should be wired into the vfo module in exactly the same way). In addition, a second control to adjust the *R-F Tune* voltage is necessary, as the tuning of these circuits is quite sharp.

It is also possible to adapt this receiver to automatic tuning; a d-c sawtooth voltage of approximately one cycle/30 seconds, and ranging from —10 to +1.5 volts applied to the varactor control point in place of voltage from a tuning potentiometer will cause the receiver to scan its complete range, quickly return to the starting point, and scan again. Because of the very sharp tuning

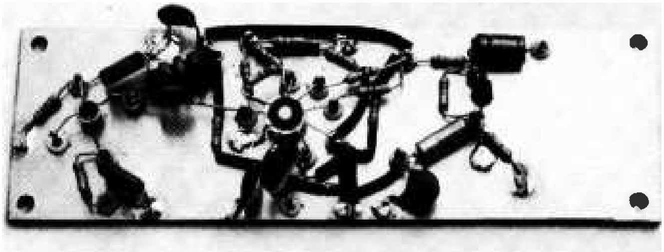


Figure 45

I-F AMPLIFIER, DETECTOR, AUDIO PREAMPLIFIER BOARD

From left-to-right are the i-f amplifier (Q_3), the detector (U_2) and the audio amplifier (Q_4) with associated components. Solid-state devices are mounted upside down to press-fit terminal insulators.

of the r-f circuits scanning over a small range is probably most practical.

I-F, Detector and Audio Circuitry—The i-f strip consists of two crystal filters with associated transformers and signal-gating circuits, an i-f amplifier common to both filters (Q3), and a gain-compensation stage (Q2) that offsets the higher insertion attenuation of the sharp i-f filter (figures 45 thru 48).

The output of the signal-path mixer (U_1, pin 6) is connected to the primary of a 9.0-MHz i-f transformer. The transformer secondary drives the gating diodes through which the i-f signal is switched to the crystal filter inputs. Because of the high insertion loss of this transformer, extra coupling is introduced by means of a small capacitor connected between primary and secondary. Conduction of the filter input/output diode gates is controlled by the *Selectivity Broad/Sharp* panel switch. The output of each crystal filter drives a similar diode gate (figure 46).

The sharp-filter output gate drives a dual-gate MOSFET amplifier stage (Q_2). This amplifier, equipped with a *trimpot* gain control, is provided to make up the gain lost due to the higher insertion loss of the sharp filter (figures 47, 48). The *trimpot* is adjusted for equal detector output when either *Sharp* or *Broad* selectivity is used. The output circuit of this amplifier is an i-f transformer which is also driven directly from

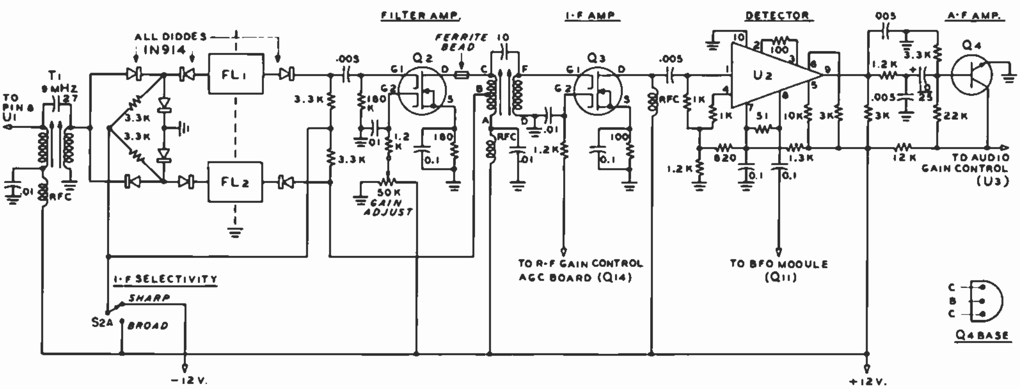


Figure 46

SCHEMATIC, I-F, DETECTOR AND FIRST AUDIO AMPLIFIER

FL_1—Crystal filter, KVG type XL-10M (Spectrum International, Box 1084, Concord, Mass. 01742)
FL_2—Crystal filter, KVG type XL-9B
Q_2, Q_3—RCA 40673 or Fairchild FT 0601
Q_4—2N4124
U_2—Motorola MC 1596G or Fairchild µA 796
T_1—9-MHz i-f transformer. J. W. Miller 1740
T_2—9-MHz i-f transformer, tapped primary. J. W. Miller 1741
RFC—1-millihenry, 35-ma. J. W. Miller 10F-103A1
Note: All resistors ¼ watt

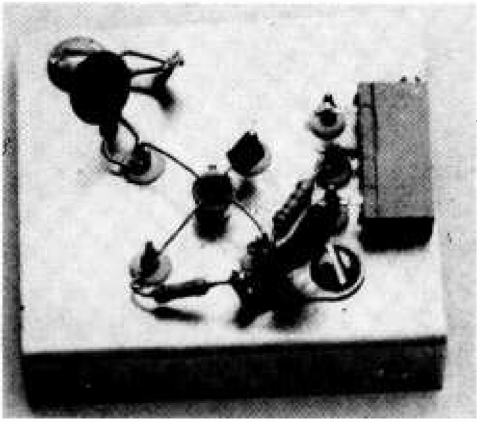


Figure 47

GAIN-COMPENSATING AMPLIFIER

The gain compensating i-f amplifier (Q_2) is mounted on a small shield bracket. Gain control is "Trimpot" seen at the right of the bracket.

the broad-filter output gate; hence this transformer is common to both filters. This section of the i-f strip is built on the main chassis, as shown in figure 28.

The second i-f transformer drives a second dual-gate MOSFET amplifier stage (Q_3), which provides up to 20 db of gain; the gain of this stage is controlled by the *R-F Gain* panel control, which also controls the gain of the r-f amplifier in the bandswitching module.

The common i-f amplifier, together with the detector and a-f preamplifier, are built on a circuit board in a manner similar to the module circuit boards (figure 45); this assembly mounted is on the right-hand side of the main chassis.

The detector is a double-balanced modulator IC (U_2), used in a product-detector configuration. It requires 50 to 300 millivolts p-p injection from the beat-frequency oscillator, applied to pin 8.

The detector drives a high-gain, bipolar transistor a-f preamplifier stage (Q_4), the output of which drives the *A-F Gain* panel control.

To test this system, connect an oscilloscope to the a-f preamplifier transistor collector. Inject 8999.0 kHz at 300 millivolts p-p from an oscillator into pin 8 of the detector. Inject 9.0 MHz at 100 millivolts p-p from a second oscillator into the primary of the input i-f transformer. Adjust the transformer slugs for maximum a-f output, check the filter gating, and adjust the gain-compensation amplifier gain control for equal audio output when either of the i-f filters is used.

The audio output stage is a HEP 593 IC amplifier (U_3), which delivers up to two watts into an 8-ohm speaker. The output connectors are so arranged that a pair of low-impedance (stereo) headphones can be plugged into the front panel jack, disabling the speaker.

The audio output stage contains its own IC power-supply regulator, mounted adjacent to the amplifier circuit board. This has been done so that the wide variations in amplifier supply current (at an audio rate) will not affect the frequency of any of the receiver oscillators by modulating their supply lines.

Construction of Beat Frequency Oscillator Module (F)—The bfo module contains the beat-frequency oscillator (Q_{10}), and a source follower (Q_{11}) which transforms both the oscillator output amplitude and impedance to the level required by the detector (figures 49 and 50).

The bfo is crystal controlled with a three-frequency capability for operating c-w, upper sideband, and lower sideband. The choice of frequency is made by applying a +12 volt level to one of the three diode-gating lines that control the crystal in use at a given time. Crystal switching is accomplished by front-panel switches *I-F Selectivity* and *USB/LSB*. When the selectivity switch is in the *Sharp* position, the bfo is set to 8999.0 kHz for use with the sharp i-f filter. When the switch is in the *Broad* position, the bfo frequency is selected by the *USB/LSB* switch.

The crystal frequency can be adjusted over a wide range above and below the nominal crystal frequency by means of the trimmer capacitor in series with each crystal. While some crystals may require shunt capacitance as well, the crystals supplied with the *KVG* filters do not.

The bfo module output should be a reasonably sinusoidal waveform of 300 millivolts p-p into 50 ohms. Once the module is constructed, the oscillator may be adjusted to each of the three beat frequencies

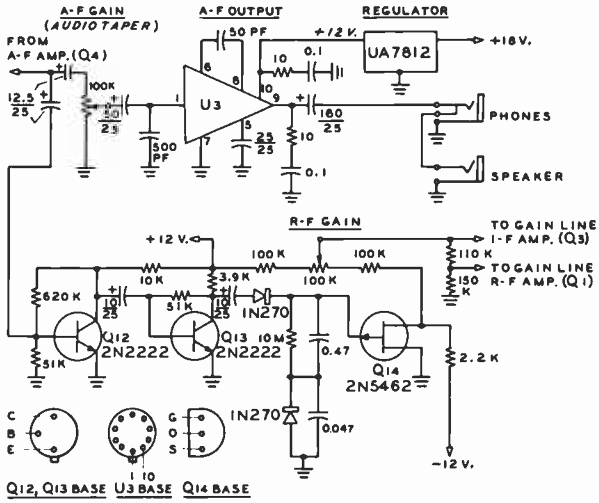


Figure 48

SCHEMATIC, AUDIO AMPLIFIER AND AGC SYSTEM

Q_{12}, Q_{13}—2N2222
Q_{14}—2N5462
U_3—HEP 593
Note: All resistors 1/4 watt

by measuring the output of the oscillator transistor at the collector with a frequency meter or digital counter connected through a 10:1 oscilloscope probe. The waveshape may be adjusted slightly, and will certainly affect the reliability of the oscillator to start oscillating when power is applied. Of the three effects, oscillator reliability must take precedence when tuning L_{12}.

Construction of Automatic Gain Control Board (G)—The agc circuit is installed and tested after the complete receiver system is in operation and aligned. Input audio to the agc is taken from the output of the detector, at the point where the audio preamplifier stage is driven (figure 48). The audio is amplified in two bipolar transistor stages (Q_{12}, Q_{13}), rectified positively, and applied to the gate of a P-channel JFET (Q_{11}), across which there is a long-time-constant circuit. The FET is operating in depletion mode; the more positive gate voltage it receives, the more negative the output voltage becomes. The FET output, tied to the bottom of the *R-F Gain* control, establishes the most negative bias level on the r-f and i-f amplifier stages in the receiver.

Advancing the *R-F Gain* control maximum clockwise has two effects: it allows the agc circuit to heavily saturate, causing the Fet output level to maintain an almost constant —12 volts. It also isolates the gain-control lines of the r-f and i-f amplifiers from this negative level. Only in a prolonged absence of input signal, under these conditions, will any change in the agc output level be noticed; and this will be a positive increase, which will tend to increase receiver gain in any case. Therefore, advancing the *R-F Gain* to maximum allows maximum receiver gain and disables agc action. To allow the agc circuit to control gain back off the *R-F Gain* to about midpoint; this allows the agc output to properly swing from —12 volts (minimum receiver gain) to ground level (maximum receiver gain).

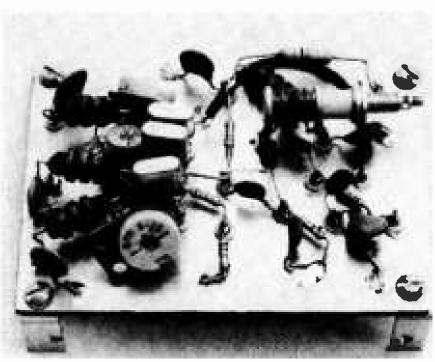


Figure 49

BFO MODULE (F)

The bfo and crystals are assembled on a p.c. board. Crystals with decoupled d-c gating lines are at the left, with the gating diodes under the r-f chokes. Crystal oscillator inductor is at the right.

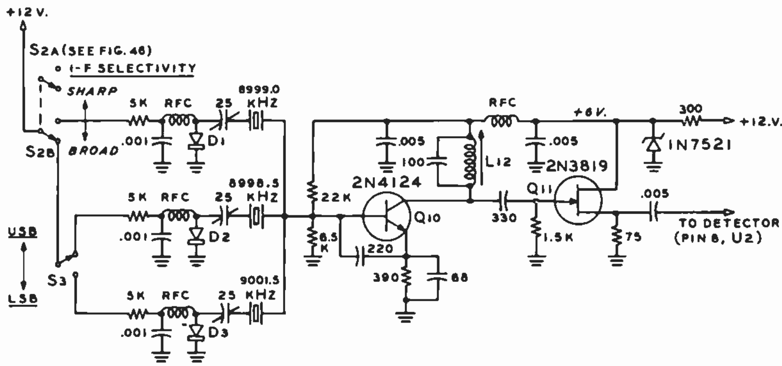


Figure 50

SCHEMATIC, BFO MODULE

D_1-D_3—Amperex Phillips BA 182
Q_{10}—2N4124
Q_{11}—2N3819
L_{12}—20 turns #29 closewound on Cambion 1535/2/1 form. L = 2 μH, Q = 70
RFC—1 millihenry, 35 ma. J. W. Miller 10F-103A1
Crystals—KVG. 8999.0 kHz, type XF-903. 8998.5 kHz, type XF-901. 9001.5 kHz, type XF-902. (Spectrum International, Box 1084, Concord, Mass. 01742)
Note: All resistors 1/4 watt

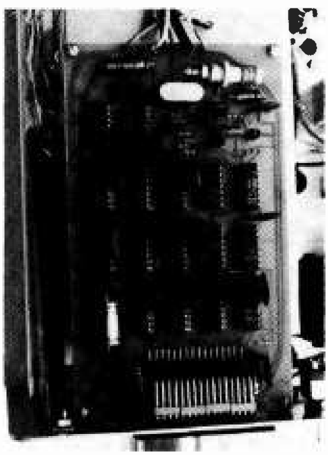


Figure 51

DIGITAL COUNTER AND DISPLAY BOARD (D)

Frequency count-out of receiver is displayed on panel-viewing LEDs (light-emitting diodes) seen at the bottom (front) of this assembly. Time base is established by 1000.00-kHz crystal at opposite end of board. ICs and other components are mounted to printed-circuit board. A full-size template of the p.c. board may be obtained by writing the publishers of this Handbook, enclosing 25 cents to cover cost of mailing.

The agc input stages are sensitive to overload; this is the reason they are driven from the detector output alone. Strong noise pulses will "hang up" the receiver gain, as will a strong signal on top of a desired weak one. The agc time-constant will be found to be quite good for SSB, but a bit slow in recovery for c-w, and it may be desirable to adjust it to one's taste after a bit of experimental listening. An alternative solution may be to add a front-panel *AGC Threshold* control, which allows front-panel adjustment to set the input audio level to the agc circuit under various operating conditions.

Construction of Digital Counter Module (D)—The digital counter is the single module built on a printed-circuit board; the only practical way of doing it. By following a board layout and figures 51, 52 ,and 53 (use the schematic as a reference, while following the board layout) it should be possible to duplicate the counter easily. Since the board layout drawings and photo are made directly from the board itself, and the schematic does not show all the many decoupling capacitors in detail, the layout and photo should guide the builder through any points of confusion. It is important that small components be used where indicated. A print of the board layout may be obtained by sending twenty-five cents to cover the cost of mailing to the Editor of this Handbook.

The frequency counter may be divided

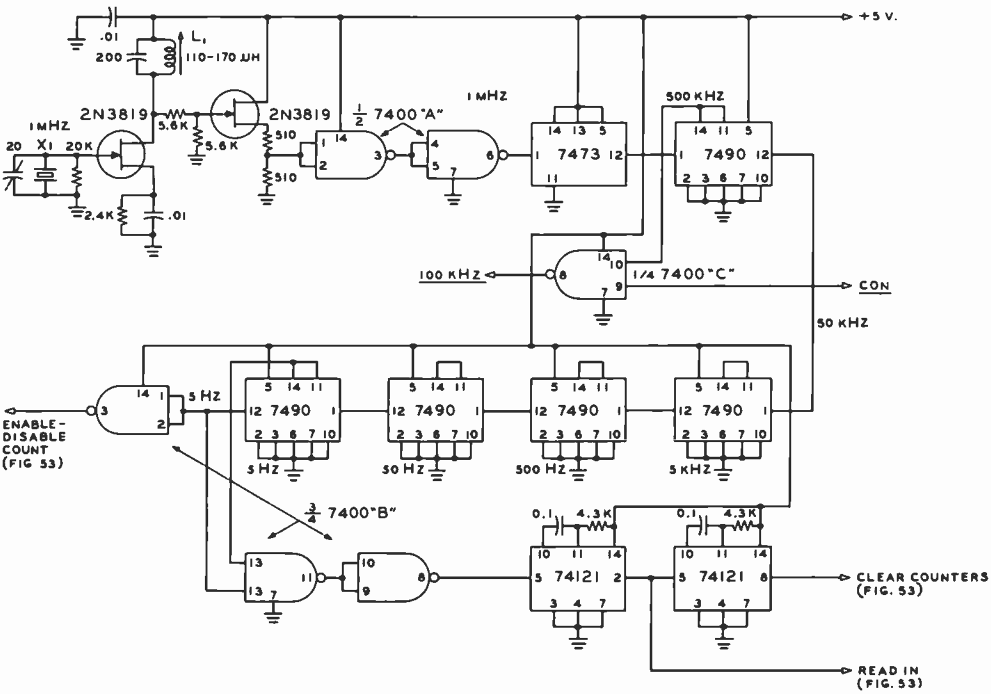


Figure 52

SCHEMATIC, COUNTER TIME-BASE MODULE

L_1—Cambion 3338-21 or equivalent
Integrated circuits—National Semiconductor DM 7400, Quad, two input NAND: National Semiconductor DM 7490 Decade counter; National Semiconductor DM 7473 Dual J-K flip-flop; National Semiconductor DM 74121 Monostable multivibrator

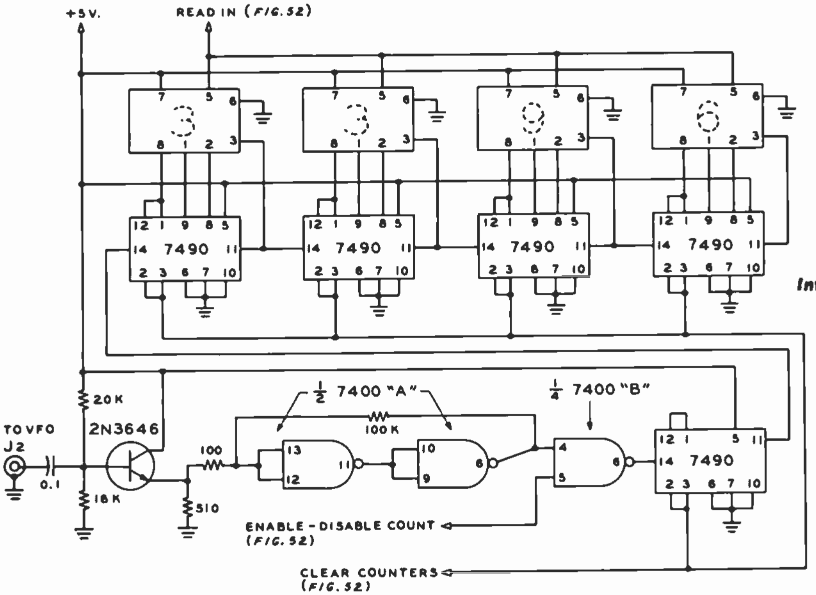
into two parts, a *time-base* and a *divider-display.*

The Time Base—The time-base consists of a 1-MHz clock oscillator, the output of which is shaped and divided down to give the needed timing pulses. The oscillator is a FET tuned-drain oscillator, crystal-controlled. The output is buffered by a FET source follower, and shaped by two NAND gates into a squarewave suitable for the inputs of the chain of decade dividers. The time-base output is a series of 5-Hz pulses, which drive the *enable/disable count* line, resulting in 5 frequency sample/updates per second of the display. The display does not blink during the count, nor can it be seen to "run up."

During the *disable* cycle, *read-in* and *clear* signals are generated using additional NAND gates, to control the operation of the *memory* circuits of the display, located inside the *readout* chips.

The Divider/Display—If this section seems to be a bit sparse, it is because the *LED Readout* integrated-circuits contain their own decoder drivers and memory circuits, making .the divide/display circuitry relatively simple to build.

The input signal to the counter is buffered by an emitter follower, squared by a *Schmitt Trigger,* and then counted down by decade counters in a manner similar to the reference clock. The decade counters count the input signal for a time determined by the time-base control, then the counter outputs are read into the memories (in binary-coded-decimal) of the readout chips. The BCD is decoded into decimal inside the readouts, and the appropriate digit is indicated on the display.



Integrated Circuits—National Semiconductor DM 7400, Quad, two input NAND. National Semiconductor DM 7490 Decade Counter; Hewlett - Packard 5082-7300 LED Digital Readout.

Figure 53

SCHEMATIC, DISPLAY MODULE

The counter module also contains the 100-kHz frequency calibrator, which is derived from the time base, and fed through a NAND gate (used to control the calibrator on/off) to the antenna input of the receiver.

The counter module has five connection points at the rear of the board:

(1) 5V—to power supply.
(2) GD—chassis ground.
(3) SIGNAL IN—input signal from the Vdc.
(4) 100 kHz—calibrator output.
(5) CON—connect to 5V to turn calibrate signal on; ground to turn calibrate signal off.

Construction of the Power Supply The power supply is built on a chassis separate from the main receiver chassis, and is housed in the speaker cabinet. An *on/off* switch is provided on the supply chassis for testing purposes; this is connected in parallel (through the power cable) with a power control switch on the receiver panel.

The supply provides three d-c rails; plus and minus twelve volts, and plus five volts (all with respect to chassis ground) and all rails are extremely well regulated, using IC regulators (figures 54 and 55).

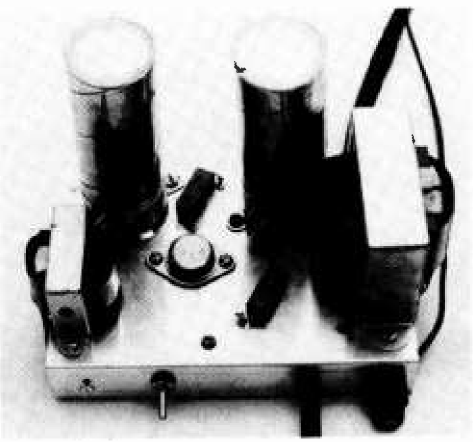


Figure 54

RECEIVER POWER SUPPLY

Filter capacitors are mounted directly above bridge rectifiers to provide short, direct leads. The two power resistors are in series with the 5-volt regulator input. The regulator is mounted between the resistors.

The negative supply is regulated by means of a positive regulator, by isolating the regulator from ground; for this reason, the neg-

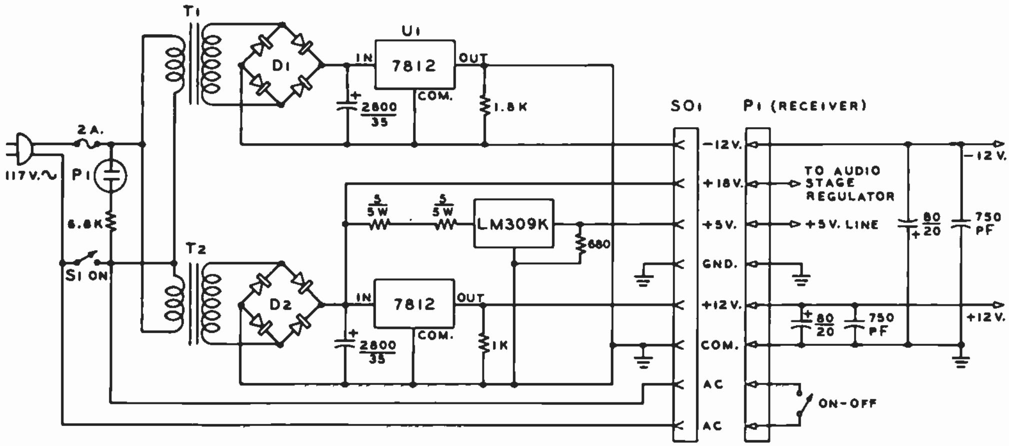


Figure 55

SCHEMATIC, RECEIVER POWER SUPPLY

T_1—12.6 volts, 0.3 ampere
T_2—12.6 volts, 2 amperes
D_1—Silicon bridge rectifier, 200 volts, p.i.v., 0.5 ampere
D_2—Silicon bridge rectifier, 200 volts p.i.v., 2 amperes
P_1—Neon pilot lamp
Integrated Circuits—Fairchild 7812; National LM 309K

ative supply regulator is not mounted/heat-sinked to the chassis; however, little current is drawn from this supply, making this procedure quite safe.

The unregulated input to the five-volt supply would be as high as eighteen volts, were it not for the series five-ohm resistors, which gradually decrease the input to the regulator with increasing load, thus minimizing the dissipation of the regulator. The load on the five-volt supply is close to one ampere and will not vary; the input voltage to the regulator under these conditions is about eight volts.

The supply is built on a 6″ × 4″ × 1″ chassis and is housed in a 8″ × 8″ × 6″ speaker cabinet. Separating the power supply from the main chassis eliminates possible hum problems, prevents component crowding, and facilitates portable operation of the receiver. If battery operation is contemplated and the digital readout is to be powered as well, it will be necessary to provide a means of shutting off the readout, except during actual frequency measurement; the high current drain of the readout will otherwise run down the supply battery quite rapidly.

It is all-important that the d-c supply lines be as free as possible from a-c ripple. Not only will audio hum rise with increasing ripple, but residual f-m hum in the output of the vfo will create intolerable audio distortion on received signals. Keep in mind that if a 10-volt change in potential across the vfo varactor tuning diode causes the frequency to change by 500 kHz, one millivolt of ripple on the varactor control line will cause 50 Hz of frequency shift, which can be easily heard in the audio, particularly when listening to a c-w signal!

Careful attention must be paid to power-supply lead lengths, lead size, and grounding when wiring the power-supply rectifier circuits; particularly the positive rectifier, which has the highest load current. The filter capacitors are mounted directly above the rectifiers, with the capacitor terminals close to the chassis, and short, heavy leads are run from the terminals directly to the bridge rectifiers. The positive rectifier is grounded directly to a solder lug, and the ground line in the receiver power cable is grounded to the same lug, to prevent ground loops. Additional decoupling capacitors are installed across the d-c input lines, directly at the receiver power connector.

With all these efforts, additional supply regulation inside the vfo module itself will be necessary, to reduce residual f-m on the vfo output to an acceptable level (details are included in the section on the vfo module).

System Alignment and Test Once the individual modules and circuit boards have been built, tested, and aligned, and the complete receiver system has been wired, the following procedure may be used to test the receiver as a system.

(1) Before connecting the receiver to the power supply, turn on the supply and check the + and —12, +5, and +18 volt lines for the proper voltage. Using a low-frequency oscilloscope, verify that ripple and noise on all lines (except the unregulated 18-volt line) is 2 millivolts, p-p or less. Now, connect the supply to the receiver, and repeat the supply tests. Note that ripple on the lines will likely increase somewhat with the increase in load current.

(2) Confirm that there is +12 volts on the common terminal of the switch deck controlling the d-c supply to the heterodyne crystal oscillators. With a 150-MHz oscilloscope, confirm that there is output from each crystal oscillator, measured at the common terminal of the oscillator output switch deck, as the bandswitch is rotated through its range.

(3) With a frequency meter or 10-MHz digital counter and oscilloscope, check for output from the vfo module of the proper waveform, amplitude, and frequency at the input to the digital counter in the receiver. With the bandswitch in the 80-, 40-, 15-, and 15-meter positions, confirm that there is vfo injection to the heterodyne mixer module. With the bandswitch in the 20-meter position, check for vfo injection directly at the signal-path mixer input, and no vfo injection to the heterodyne mixer module.

(4) With the bandswitch in the 80-, 40-, 15-, and 10-meter positions, check for oscillator injection to the signal-path mixer from the heterodyne mixer module. Check the waveform with a high-frequency meter or digital counter, and tune the vfo through its range while making this check. Output from the heterodyne mixer module should be sinusoidal, and of reasonably constant amplitude across the range on each band.

(5) With oscilloscope and counter, check for proper injection to the detector from the bfo module, while switching from one i-f filter to the other, and from USB to LSB.

(6) With a one-microvolt unmodulated signal at the antenna input, check for a comfortable audio level in headphones or speaker (*A-F Gain* at maximum) while tuning across the signal. Confirm that the digital counter in the receiver indicates the proper frequency; and keep in mind that any errors in calculation are the result of inaccuracies in the frequency being generated by the heterodyne crystal oscillator in use. During prototype testing, the receiver was able to copy 0.5-μV signals on all bands, using the speaker.

If instability is noted on any band, it may be necessary to alter wiring layout, lead dress, or add ferrite beads and/or "losser" resistors to the inputs and/or outputs of high-frequency amplifiers in the receiver. With the receiver properly shielded and grounded, no instability should be observed.

Exciters and Transceivers

The exciter is the "heart" of the amateur station and may take the form of a transmitting unit, a combination transmitter/receiver, or transceiver. Various forms of power supplies, amplifiers and accessory units combine with the basic exciter to form a complete communication system which can satisfy a wide range of needs in today's highly complex world of radio.

Several different types of medium-power exciters and transceivers for the h-f and vhf range are described in this chapter, including a state-of-the-art, broadband h-f SSB exciter, incorporating frequency synthesization and integrated circuitry; offering an interesting challenge to those amateurs experienced in solid-state techniques.

The component nomenclature outlined in figure 1 of the Receiver chapter is employed in the following sections.

21-1 A 40-Watt Broadband Exciter for 2 Meters

This broadband, 40-watt exciter/transmitter for 144-MHz operation was designed by W6ZO and built by W4HHK. Utilizing coupled interstage transformers, the unit is capable of operation over the 144- to 148-MHz range without the necessity of retuning the intermediate stages. The only tuning adjustment that is required is for the final amplifier stage and antenna circuit.

Designed for continuous-duty operation with moonbounce projects, the broadband exciter is well suited for general vhf operation. It may be plate-modulated for a-m service, keyed for c-w or phase-modulated for f-m work. The unit is stable in operation, and subharmonic and harmonic radiation are held to a minimum by the use of multiple interstage tuned circuits.

The Transmitter Circuit The transmitter circuit is shown in figure 2. A 6AS6 is used as a crystal oscillator utilizing 8-MHz fundamental frequency crystals. Crystal drive level is exceptionally low with this circuit and frequency stability is excellent. The screen voltage of the 6AS6 is regulated, and a small positive voltage is applied to the suppressor element of the tube to enhance the power gain. The oscillator is capacitively coupled to a 6CL6 tripler to the 24-MHz region. Oscillator adjustment may be accomplished by measuring the rectified grid voltage of the 6CL6 at *test point 1*, with the aid of a high-resistance voltmeter.

A double-tuned transformer is used in the interstage circuit between the 6CL6 24-MHz tripler and the 48-MHz doubler to reduce the residual 8-MHz energy which might otherwise be fed to the doubler stage. The 6CL6 doubler plate circuit is broadly resonant at 48 MHz, yet provides good re-

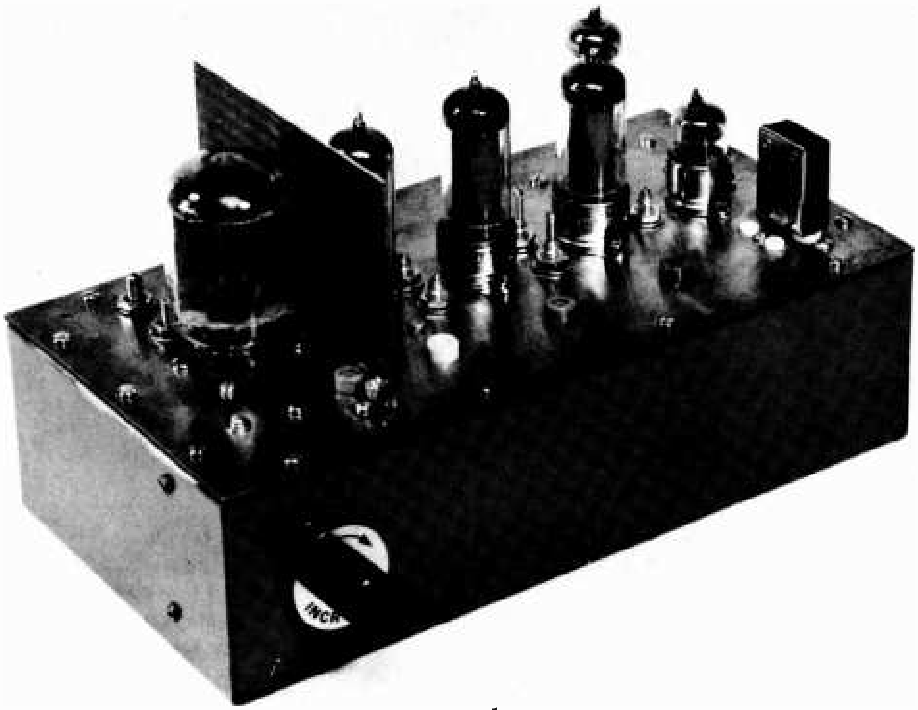


Figure 1

40-WATT, 2-METER BROADBAND EXCITER/TRANSMITTER

Two-meter transmitter employs broadband coupled circuits in driver stages to achieve complete coverage of 144-MHz band without retuning. The unit is designed for continuous service and provides a powerful signal with high attenuation to unwanted spurious emissions. A 7984 "Compactron" tube is used in the final amplifier stage (left). Plate tuning capacitor and output link tuning capacitor are adjacent to 7984. Power output level is adjusted by screen control potentiometer mounted at left-hand end of chassis. Grid current test jacks are seen along the front edge of the chassis. Two crystal sockets are at the front right-hand corner, wired in parallel, to accommodate either FT-243 or HC-6/U style crystals. To the right of 7984 is the interstage shield and immediately adjacent to it is the 5763 driver, followed by the two 6CL6 multiplier tubes. At the extreme right is the 6AS6 crystal oscillator tube. Heat-sink tube shields are used on the 6AS6 and 6CL6 tubes. To the rear of the oscillator tube is the voltage regulator.

jection to 24-MHz, 32-MHz, 40-MHz, and 54-MHz energy, all of which are present to some degree in the plate circuit of the second 6CL6.

A 5763 is used as a tripler from 48 MHz to 144 MHz and, in turn, is inductively coupled to a 7984 amplifier. A small amount of 48-MHz energy is present in the grid circuit of the 7984 stage, but it is effectively suppressed in the high-Q plate and antenna tuned circuits.

The 7984 has good internal shielding and when used with an external shield separating it from the exciter stages proves to be self-neutralized in the 2-meter region. Double screen-terminal bypassing is used on the 7984 socket to provide the proper low-impedance screen-to-ground path necessary at this frequency.

Power output of the exciter is controlled by varying the screen voltage of the 7984 stage by means of the *adjust output* potentiometer (R_1). With a maximum plate potential of 450 volts on the 7984, an input as high as 80 watts may be run. For the unit shown, the usual power input is about

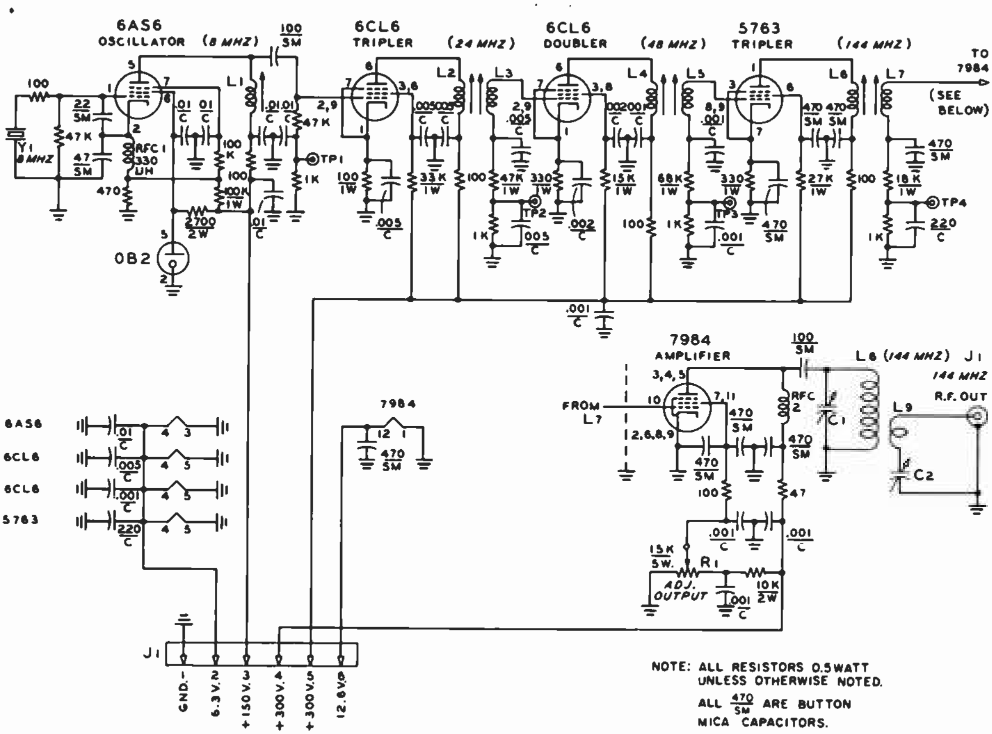


Figure 2

SCHEMATIC, 2-METER TRANSMITTER

C_1, C_2—9 pf, Johnson 160-104
Note: Coils L_1-L_4 are wound on Cambion (CTC) ceramic form 1538-4-3(powdered-iron slug for 50-200 MHz), ½" diam., coded white. Coils L_6 and L_7 are wound on J. W. Miller ceramic form 4500-4 powdered-iron slug for 50-200 MHz ¼" diam., coded white
L_1— 46 turns # 30 enamel wire, closewound
L_2—19 turns #24 enamel, closewound
L_3—15 turns #24 enamel, closewound
L_4—6 turns #24 enamel, closewound. Space last turn or two wire diameter, if necessary to establish proper broadbanding of circuit
L_5— 7 turns #24 enamel, turns spaced wire diameter. Coils L_4 and L_5 are mounted 11/16" apart
L_6— 3 turns #18 enamel, turns spaced wire diameter
L_7—2 turns #18 enamel, turns spaced wire diameter. Coils L_6 and L_7 are mounted 11/16" apart, center to center
L_8—2 turns #12 tinned wire, ½" inside diam., turns spaced wire diameter
L_9—3 turns #18 tinned wire, ½" inside diam., turns close spaced. Optimum spacing between adjacent ends of coils is about 3/16". Adjust spacing for maximum output
RFC_1—330 µH. J. W. Miller 70F-334-A1
RFC_2—2-meter plate choke. Ohmite Z-144, or J. W. Miller X-144 (2µH)

40 watts at about 315 plate volts. Sufficient drive exists, however, to run the higher input level. For continuous RTTY service, the exciter is usually run at about 35 watts input, providing an output of nearly 25 watts. This is more than sufficient to drive a tetrode amplifier using two 4CX250Bs to the 1-kilowatt level. Complete operating data for the exciter is given in Table 1.

The plate circuit of the 7984 is a parallel-tuned configuration having high Q to aid in suppressing undesired subharmonics. The antenna circuit, too, is tuned to resonance by means of a series capacitor in the ground return path. Only the amplifier plate circuit and antenna capacitor need be retuned for frequency excursions within the 2-meter band.

Table 1. Voltage and Power Measurements
(Voltages Measured to Ground. 20,000-ohms/volt Meter used.)
300-Volt Power Supply

Tube	Circuit	Voltage	Current
6AS6 Oscillator	Plate, end of coil L_1	150	
	Screen, pin 6	105	
	Suppressor, pin 7	0.9	
6CL6 Tripler	Plate, end of coil L_2	300	
	Screen, pin 3	150	
	Cathode, pin 1	2.2	
	Cathode		22 ma
	Grid, test point #1		350 μA
6CL6 Doubler	Plate, end of coil L_4	300	
	Screen, pin 3	90	
	Cathode, pin 1	8	
	Cathode		24 ma
	Grid, test point #2		4.3 ma
5763 Tripler	Plate, end of coil L_6	300	
	Screen, pin 6	245	
	Cathode, pin 7	10	
	Cathode		31 ma
	Grid, test point #3		2.5 ma
7984 Amplifier	Plate, end of RFC	300	
	Screen, pin 7	110	
	Screen		1.4 ma
	Plate		100 ma
	Grid, test point #4		3 ma
	Power Input	30 watts	
	Power Output	22 watts	
Filament requirements: 6.3 volts at 2.25 amperes and 12.6 volts at 0.6 ampere			

Transmitter Construction The transmitter is constructed upon a piece of copper-plated (two sides) epoxy circuit board measuring 9½" × 5". It is placed atop an inverted aluminum chassis used as a base and dust cover. The chassis measures 9½" × 5" × 2½". Layouts of the major components are shown in the photographs and in the chassis drawing of figure 4. It must be remembered that a ½-inch border must be left around the circumference of the circuit board to permit the board to sit flush on the chassis lips. The board is cut to size and temporarily placed on the inverted chassis and the border allocated and marked with a pencil. The board is then removed for drilling and cutting to mount the sockets, coil forms, and other major components.

The under-chassis vertical shield is made of copper-coated (one side) circuit board and measures 4" × 2". It is placed across the underside of the 7984 tube socket, carefully notched to fit snugly over the socket and against the board. It is mounted off center so that one edge may be affixed to the aluminum-inclosure by means of a small angle bracket. It is not bolted in place until after the board is wired and attached to the chassis. This shield is located on a line running between pins 8 and 9 and pins 12 and 1 of the 7984 socket.

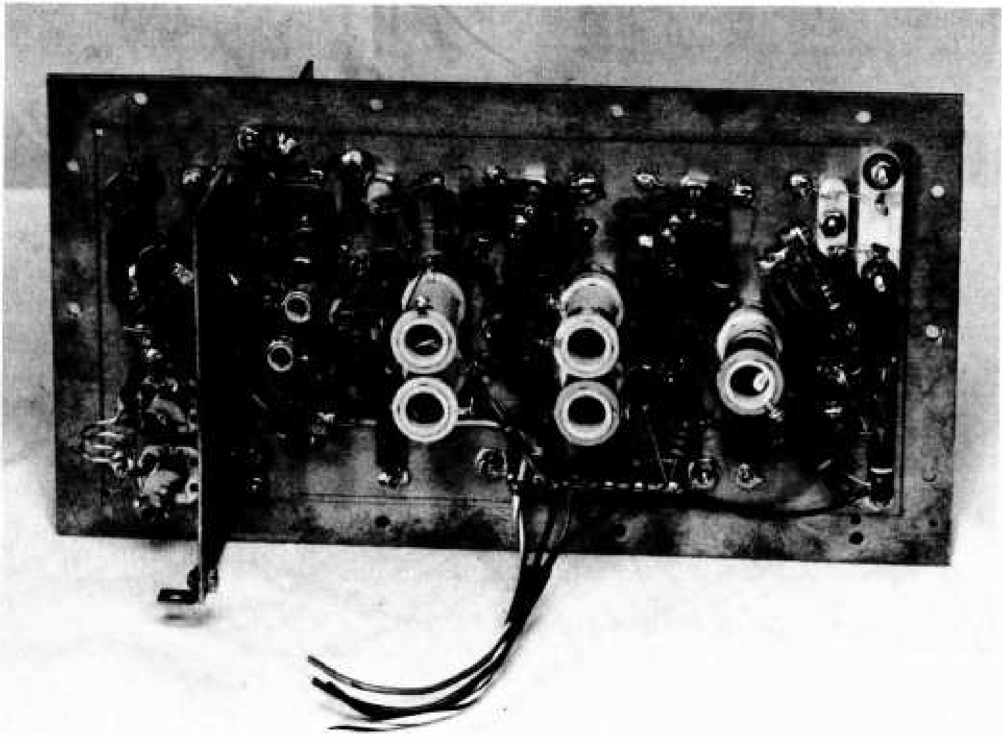


Figure 3

UNDER-CHASSIS PARTS LAYOUT

The 7984 plate tank coil is in the lower left-hand corner of the chassis, to the left of the under-chassis shield, mounted across the center of the 7984 tube socket. One corner of the shield is attached to the chassis base.
At the center of the chassis are the coil forms for the bandpass transformers, with the crystal oscillator stage at the right. A terminal strip at the rear of the chassis is used to terminate the leads to the main power receptacle placed on the back of the aluminum support chassis. Placement of major components is shown in figure 4.

The 7984 cathode pins (2, 6, and 9) are grounded to the circuit board with very short, wide straps of flashing copper strap. Cathode pin 8 and filament pin 1 are soldered to the shield after all circuit-board wiring is completed and the shield is bolted in place. The copper side of the shield faces pins 8 and 1.

All sockets, coil forms, test-point jacks, etc. are mounted on the circuit board before wiring is started. Space is at a premium in some areas and components must be carefully located and wired in proper order. The filament wiring, grounding of socket pins, and socket bypass capacitors should be connected first. Most capacitors are chosen to be series resonant at the operating frequency of the circuit and should be wired in position with very short leads. Filament and d-c voltages are distributed to each stage from the tie-point strip mounted between coil L_2 and the chassis power plug. A tie-point strip placed near the 6AS6 oscillator tube socket serves to terminate power leads to that tube and the OB2 regulator tube associated with it. The +300 volt line for the final amplifier is run in a shielded wire from the power plug to the 10K resistor and *adjust output* potentiometer control (R_1) and from there to the 47-ohm plate circuit resistor for the 7984. The potentiometer and series screen resistor are mounted on the wall of the aluminum chassis, while the power plug and coaxial receptacle for the antenna are mounted to the rear wall of the chassis. Wiring to the power plug, antenna receptacle, and screen potentiometer is done after the circuit board has been wired and placed atop the chassis.

A 2¾" × 4" shield made of plated circuit board is mounted atop the circuit board between the 7984 envelope and coil L_7. The shield is notched to fit snugly over the edge of the mounting nut of coil L_7 and the edge of *test point 4*. Foil side of the board faces the 7984 and almost touches the glass envelope. There is just enough space between the 7984 envelope and coil L_7 tuning screw to mount the shield. Relative spacing between coil L_7 and the 7984 tube socket is critical in this respect.

The 470-pf mica button capacitor that bypasses the 7984 screen terminal to ground (see schematic) is mounted adjacent to pin 7 and connection is made to it by means of a short length of 3/16"-wide copper shim stock. On the opposite side of the shield, screen pin 11 must be bypassed to cathode pin 9 with a second button capacitor, making leads as short as possible. This will stabilize the 7984 by making the screen circuit self-resonant to ground.

Coil spacing is chosen so as to provide adequate intrastage coupling when the circuits are stagger-tuned for 144 MHz and 148 MHz. The end of antenna coil L_9 adjacent to plate coil L_8 connects to the antenna receptacle (J_1) and the opposite end of coil L_9 connects to the antenna tuning capacitor (C_2). This capacitor is fully meshed when working into a nonreactive 50-ohm dummy load.

Transmitter Adjustment and Tuning When the wiring is completed it should be carefully checked against the schematic for possible errors. Before power is applied, the various coils may be roughly grid-dipped to the frequencies indicated on the schematic. Coil dimensions given are based on the use of heat-sink tube shields on the oscillator (6AS6) and 6CL6 multiplier tubes. There is a noticeable detuning effect in the plate circuits of these stages when a tube shield is removed. Heat-sink shields are recommended to reduce bulb temperature and extend tube life.

The 6AS6 and OB2 tubes are plugged in their respective sockets and coil L_1 adjusted for proper oscillation with a crystal in the 8-MHz region. When the 6CL6 tripler is in the socket and L_1 is tuned for reliable crystal oscillation, about 0.35 volt should be measured between *test point 1* and ground.

The 6CL6 tripler is tuned to 24 MHz, and about −4 volts should be observed at *test point 2* in the grid circuit of the 6CL6 doubler stage, whose plate circuit is tuned to 48 MHz. Finally, the 5763 tripler is placed in the socket and its plate circuit is peaked at 144 MHz.

The *adjust output* control (R_1) is set to reduce the 7984 screen voltage to zero (arm at ground end of potentiometer) and the amplifier tube is plugged into its socket. The plate circuit of the 5763 stage is repeaked to provide about 3 volts indication at *test point 4*. Screen voltage may now be in-

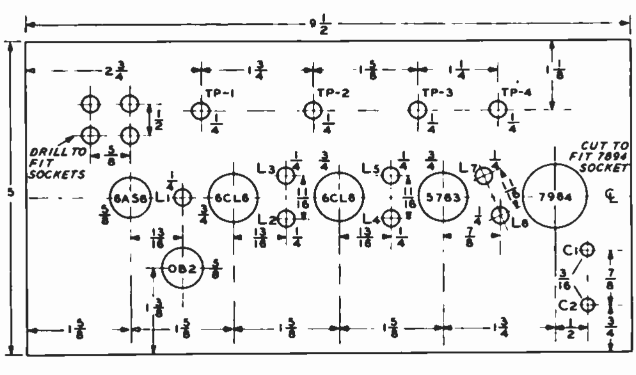


Figure 4

PARTS PLACEMENT FOR TRANSMITTER CHASSIS

creased to provide a plate current reading on a 0 to 200 d-c milliammeter placed in series with the B-plus lead to the 7984 (pin 4 on the power plug). The amplifier plate circuit is resonated, with the transmitter connected to an antenna or dummy load. Adjustment of the coupling between plate coil L_8 and antenna coil L_9 should permit the amplifier to be loaded to 100 milliamperes or more with the *adjust output* control set in an advanced position.

Once the transmitter is ascertained to be operating correctly, the intermediate stages should be adjusted for proper bandpass operation. This requires the use of two crystals, one at about 8.02 MHz and one at about 8.20 MHz. Adjustments are carried out to ensure that the same power output is obtained from the final amplifier stage when either crystal is used, providing the amplifier has been retuned for each frequency in use. Generally speaking, broadbanding may be accomplished by first tuning multiplier plate coils L_2, L_4, and L_6 for maximum final amplifier grid current using the 8.02-MHz crystal and then tuning grid coils L_3, L_5, and L_7 for maximum grid current using the 8.2-MHz crystal. The oscillator coil (L_1) may be adjusted for smooth oscillator operation and used to help to adjust grid drive at either frequency. Once rough alignment has been achieved, the grid and plate coils may be adjusted slightly at either end of the band to equalize the output. When properly adjusted, amplifier grid drive remains practically constant across the full 4-MHz bandwidth. Amplifier plate efficiency runs around 70 percent, and plate voltages between 300 and 500 may be used, provided the maximum plate dissipation figure of 25 watts (intermittent) is not exceeded. Power output remains high, even at excitation levels as low as 1-milliampere grid current to the 7984.

The final check is to test for self-oscillation in the 7984 stage. With the transmitter working into a 50-ohm load, the crystal may be removed from its socket briefly. When this is done, the 7984 output should drop to zero and there should be no grid current (observed as voltage at *test point 4*). This test should be restricted to a second or two, since plate current of the 7984 is excessive when excitation is removed.

21-2 An Advanced Six-Band Solid-State SSB Exciter

The SSB exciter described in this section was designed and built by K9HTK/5. It is a state-of-the-art device capable of exceptionally good efficiency and low intermodulation distortion (IMD) over the range of 3.5 MHz to 54 MHz (figure 5). Power output is in excess of 5 watts PEP on all bands except the 50-MHz band where the output is 1 watt PEP. The IMD is better than —33 decibels below one tone of a two-tone test signal on the lower bands and —45 db on the 50-MHz band. Operating convenience has not been overlooked as provision is made for VOX operation and/or push-to-talk. In addition, a frequency-spotting switch for split operation and a carrier-insertion circuit for linear amplifier tuneup have been incorporated. No tuning of the exciter is required when changing frequency or bands as the output circuits are broadbanded over the full 3.5- to 54-MHz frequency range.

Also incorporated in this exciter is front-panel control of both audio and r-f clipping (variable from zero to 20 db of clipping). This allows the operator to tailor his signal to meet the existing conditions; clipping may be reduced for local ragchews or turned up for more audio punch in DX pileups. An audio speech compressor adjustable from the panel is also incorporated in the design. All of these features add up to provide a very potent SSB exciter for the advanced amateur who has had experience with the sophisticated components and circuitry used in this unit.

Circuit Description The exciter and power supply are completely solid state and wideband circuitry is employed to simplify tuning and adjustment. Special, switchable filters are used in the low-level stages to eliminate unwanted mixing signals, and dual crystal filters are used in the r-f processing circuitry. A phase-locked loop synthesizer is used to generate the conversion signal. This results in an exceptionally clean signal, free of the spurious problems often associated with a premixer and also provides the same tuning rate and degree of fre-

Figure 5

SOLID-STATE SIX-BAND SSB EXCITER

This compact, solid-state SSB exciter delivers over 5 watts PEP output over the range of 3.5 to 29.7 MHz and provides over 1-watt PEP output on the six-meter band. Audio and r-f clipping circuits provide good audio "punch." A phase-locked-loop synthesizer is used for the conversion oscillator and r-f circuits are broadbanded over the full operating range. The main tuning dial is at the right, with the phase-lock light above it. Across the bottom of the panel are (left to right): Audio level, Audio compression (gain and recovery time), carrier insertion, VOX (gain, delay, and antivox gain), and the VOX override switch. The general purpose multimeter and switch are at the upper left of the panel and to the right are the clipping-level controls for audio and r-f clipping. Immediately beneath these controls are the bandswitch and drive-level controls, with the sideband selector switch centered between them. The multimeter has two ranges: 0 to 30 volts and 0 to 900 milliamperes. The +28, +12, and −12 volt supplies are monitored, as well as amplifier current.

quency stability on all bands. The master reference oscillator tunes over the range of 3.21 to 3.71 MHz, providing excellent stability on all bands. Provision is made for coverage of four 500-kHz bands in the 10-meter range and four 500-kHz bands in the six-meter range, although this combination may be changed, if desired. Operation on nonamateur frequencies is also possible (with some exceptions) by the proper choice of crystal and tuned-circuit components. The 3.21- to 3.71-MHz oscillator tuning range was chosen by careful consideration of all mixing products up to the tenth order with the aid of a digital computer. During several months of on-the-air operation no spurious problems have been observed.

For best spurious rejection, the mixing frequency is 9 MHz *above* the desired operating band, which places the mixing frequency quite high for 6-meter operation. However, the use of the frequency synthesizer provides stable frequency control from a low-frequency oscillator of good stability.

If the exciter is used to drive the antenna directly, a half-wave low-pass filter such as described in Chapter 16, Section 3 should be used between the exciter and the antenna to attenuate the harmonics of the fundamental signal. If a linear amplifier with high-Q tuned circuits is used after the exciter, however, the low-pass filter may not be required since the tuned circuits of the linear amplifier will attenuate the harmonics. If desired, an extra switch section could be added on the exciter bandswitch to remotely select the appropriate low-pass filter automatically.

Exciter Circuitry

Audio and VOX Circuits—Shown in figure 6 is a block diagram of the audio and VOX

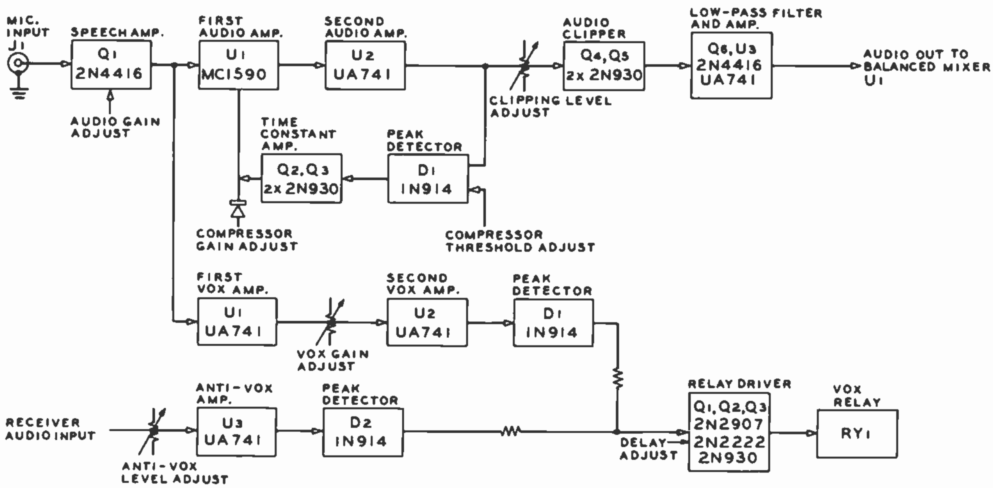


Figure 6

BLOCK DIAGRAM OF AUDIO AND VOX CIRCUITS

Audio clipping and compression are included in the speech amplifier of this versatile exciter. Compression gain and recovery time are adjustable. An audio filter follows the clipper to remove higher order harmonics. Vox gain and delay are adjustable permitting the operating time and hold-in time to be varied at the operator's preference.

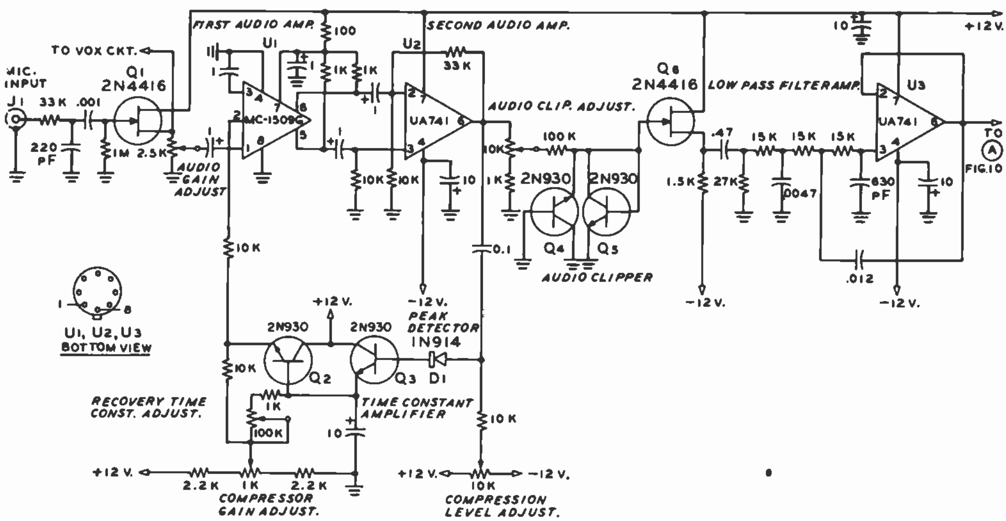


Figure 7

SCHEMATIC, AUDIO CIRCUITRY OF SSB EXCITER

U_1—Motorola MC 1590G
U_2, U_3—Fairchild μA 741
Note: All resistors ¼ watt. All potentiometers audio taper

circuits. The schematic for these circuits is shown in figures 7 and 8. An FET device (Q_1) provides a high input impedance for the microphone and drives the first IC audio amplifier (U_1, figure 7) and the VOX amplifier (U_1, figure 8). The *AUDIO*

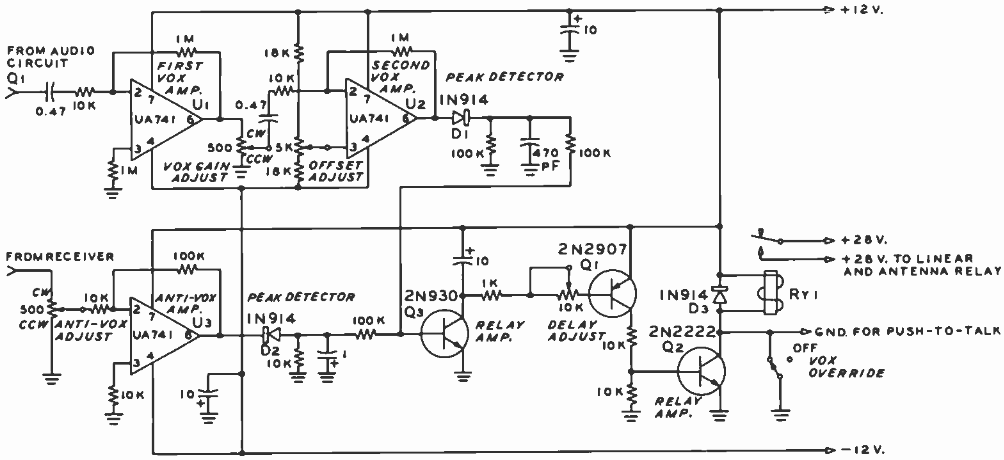


Figure 8

SCHEMATIC, VOX CIRCUITRY

U_1-U_3—Fairchild μA 741
RY_1—Crystal can relay or reed relay. Potter Brumfield JMF 1080-61
Note: All resistors ¼ watt. All potentiometers audio taper

GAIN control in the source circuit of Q_1 allows the drive level of U_1 and U_2 to be set for optimum operation of the compressor circuit consisting of U_1, U_2, Q_2, and Q_3. The *COMPRESSOR GAIN ADJUST* control varies the amount of compression and the *RECOVERY TIME CONSTANT* adjustment varies the time required for the circuit to return to maximum gain after a large signal is removed from the input. The *COMPRESSION LEVEL ADJUST* is an internal control which is set to give 3 volts rms at pin 6 of device U_2 with a large input signal. The *AUDIO CLIPPING ADJUST* control then allows the clipping to be varied from zero to 20 decibels. Transistors Q_1 and Q_5 are used as reverse-connected diodes to provide clipping and they function much better than ordinary diodes in this circuit. Integrated circuit U_3 and associated components form an active three-pole low-pass filter with a 3-kHz cutoff frequency which removes the higher-frequency components generated by the clipping circuit.

Referring to figure 8, integrated circuit U_1 amplifies the signal from the microphone by 40 db and the *VOX GAIN* control varies the signal level applied to U_2. The output signal of U2 is rectified and the positive voltage coupled to the base of transistor Q_3, turning it on when an audio signal is generated by the microphone. This causes Q_1 and Q_2 to turn on and the VOX relay (RY_1) to close. Integrated circuit U_3 amplifies the signal from the receiver output circuit, which is rectified, and the resulting negative voltage also is applied to the base of Q_3. Adjusting the *ANTIVOX* control prevents the speaker output picked up by the microphone from closing the VOX relay. The *DELAY* adjustment allows the hold-in time of the relay to be varied at the operator's preference.

The RF Circuitry—The block diagram of the rest of the exciter is shown in figure 9, including the phase-locked synthesizer. Schematic diagrams of the r-f circuits are shown in figures 10, 11, and 12. The balanced mixer (U_1 in figure 10) generates a DSB signal from the processed audio (Q_3, Q_1). Diode switches are used to remotely select crystals for upper- or lower-sideband operation. The first 9-MHz crystal filter (FL_1) selects one sideband which is amplified by FET device Q_1 and applied to the r-f clipping circuit. Diodes D_1 and D_2 are inexpensive ultrafast switching diodes and perform almost perfect clipping of the signal. The amount of clipping is adjustable by varying the gain of Q_1 over the range

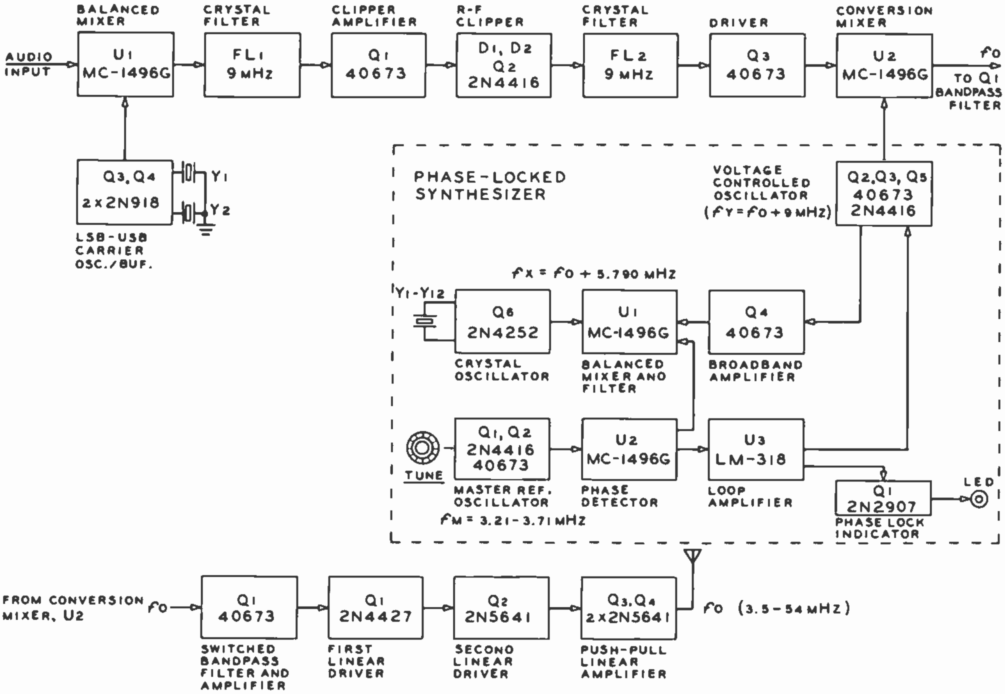


Figure 9

BLOCK DIAGRAM, R-F CIRCUITRY OF SSB EXCITER

The conversion frequency (F_y) is 9 MHz above the signal frequency. The crystal frequency for the phase-locked synthesizer is 5.970 MHz above the signal frequency. The master reference oscillator tunes the range of 3.21 MHz to 3.71 MHz. The SSB signal is passed through a switched bandpass filter (lower left) before being amplified by the three-stage linear amplifier. Operation of phase-locked loop is indicated by light-emitting diode (LED).

of zero to 20 db of r-f clipping. The clipped signal is then passed through a second crystal filter (FL_2, figure 11) to remove high-order products outside the passband of the filter. The clipped signal, now restored to its original bandwidth, is amplified by driver Q_3 and applied to the conversion balanced mixer (U_2). The *DRIVE ADJUST* potentiometer in the #2 gate of Q_3 allows the drive level to the following circuits to be adjusted as required. Drive is not adjusted by the audio circuits as is done in conventional exciters due to the various clipping circuits in this design.

The Conversion Mixer—The conversion mixer (U_2 in figure 11) has three inputs: conversion-oscillator injection from the phase-locked synthesizer; a 9-MHz signal from Q_3; or the carrier-insertion signal from the circuit consisting of diodes D_3, D_4, and associated components. The diodes are long-storage-time PIN devices which act as variable resistors (instead of diodes) at this frequency. This allows a variable amount of carrier signal to be inserted by a front-panel control when the *PUSH TO SPOT* switch is depressed. In normal operation of the exciter these diodes are biased open to prevent the carrier from appearing at the output of U_2. Depressing the switch allows the bias to be adjusted by the *CARRIER INSERTION* potentiometer, causing the diodes to act as a variable attenuator, controlling carrier level as desired.

Transformer T_2 at the output of mixer U_2 is a broadband device (balun) which matches the mixer output impedance to the low-impedance coaxial cable interconnection to Q_1 in figure 12. The output of this device contains a double-tuned filter circuit

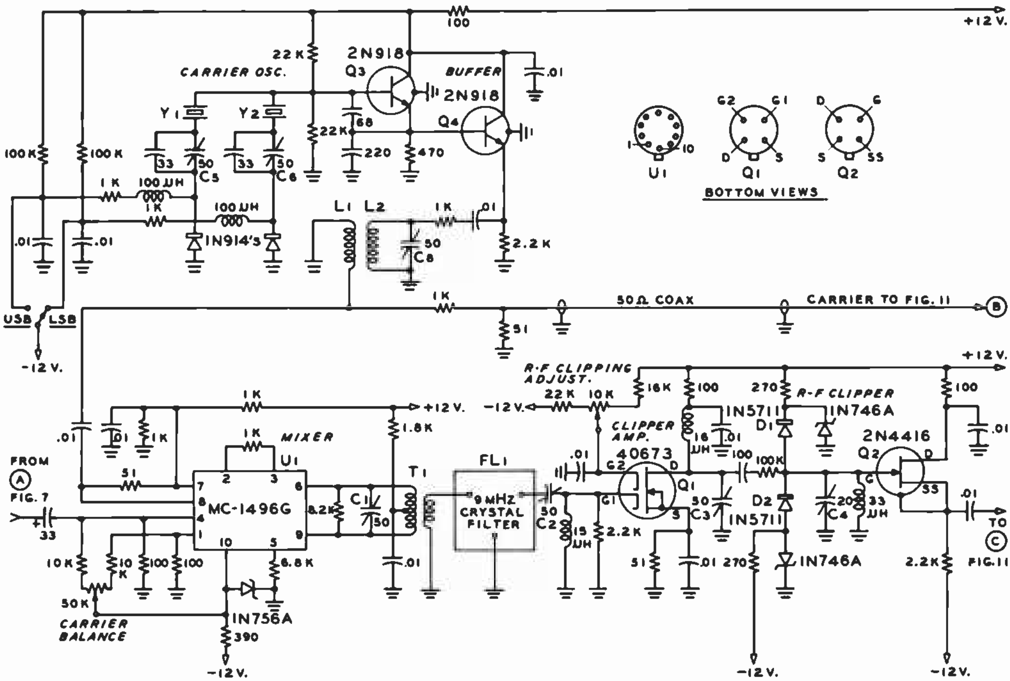


Figure 10

SCHEMATIC, SSB GENERATOR AND R-F CLIPPER

D_1, D_2—Hewlett-Packard HPA 5082-2800 (1N5711)
C_1-C_3—5-50 pf. Johanson 9305
C_4—20 pf. Johanson 9302
C_5, C_6—5-50 pf. Johanson 9305
L_1, L_2—19 turns #28 on CTC 1536-6-2 form. Link is 5 turns #28 closewound on "cold" end of L_2
FL_1—9-MHz filter with 2.4-kHz bandwidth. KVG XF-9A (Spectrum International, Box 1084, Concord, Mass. 01742)
Y_1—8998.5 kHz. KVG 901 (see above)
Y_2—9001.5 kHz. KVG XF 902 (see above)
T_1—Primary: 19 turns #30 bifilar wound, secondary: 8 turns #24 e. Wound on .437 × .250 × .187 Carbonyl SF toroid

Note: All resistors ¼ watt

which passes only the desired mixer product to the high-gain, three-stage linear amplifier. Switch S_1 selects the proper filter for the band in use and may be eliminated if a single-band exciter is desired. In that case, the proper filter is wired directly into the circuit.

The Linear Amplifier—The linear amplifier (figures 13 and 14) consists of two class-A driver stages (Q_1, Q_2) followed by a push-pull class-AB power output stage (Q_3, Q_4). All stages are broadbanded across the 3.5- to 55-MHz range and the power gain is essentially flat to 30 MHz, decreasing to about 6 db at 50 MHz. The resistive attenuator at the input to Q_1 is necessary to ensure stable operation on all bands. Devices Q_2, Q_3, and Q_4 are vhf power transistors with balanced emitter construction; the *Motorola* type 2N5641 was found to combine excellent linearity and ruggedness. Other manufacturer's 2N5641s were not tested and unless linearity testing equipment is available, the *Motorola* devices should be used. A 1N645 diode provides temperature compensation for the bias of Q_3 and Q_4 and should be thermally connected to one of these transistors with heat-sink thermal compound (*Dow Corning* 340 or equivalent). As mentioned previously, a low-pass filter for the band of operation should follow the linear amplifier to suppress the r-f harmonics of the signal if the amplifier is connected directly to an antenna.

The Power Supply—The circuit of the

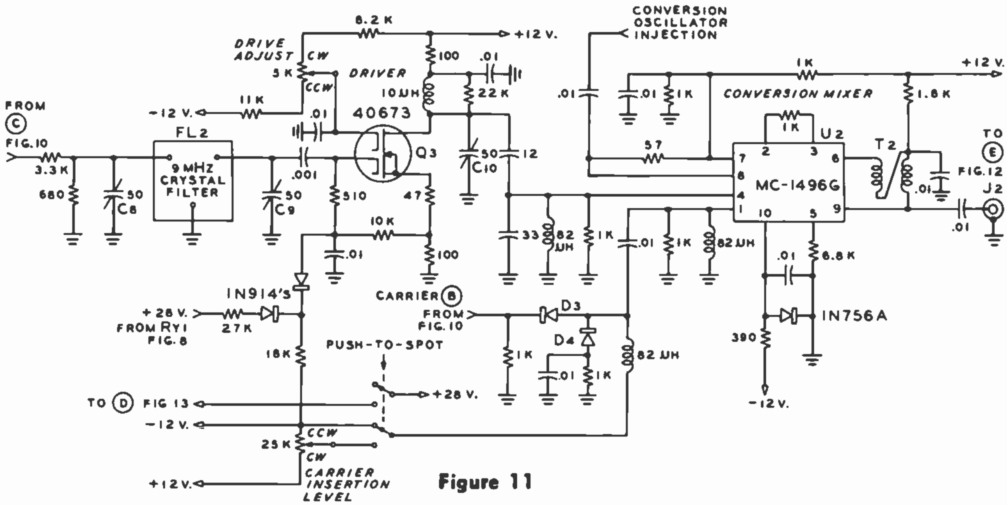


Figure 11

SCHEMATIC, FILTER AND CONVERSION MIXER

D_3, D_4—Hewlett-Packard HPA 5082-3081
C_8-C_{10}—5-50 pf. Johanson 9305
T_2—9 turns #28 insulated, bifilar wound on CF-102 core (Indiana General)
Note: All resistors 1/4 watt. See figure 10 for filter data

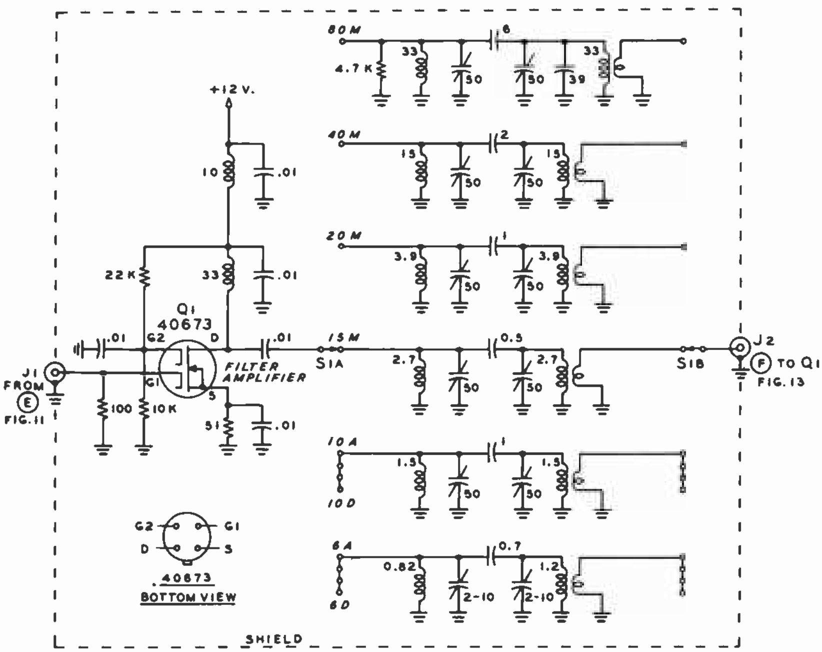


Figure 12

SCHEMATIC, SWITCHED FILTER

J_1, J_2—Subminiature coaxial receptacle
S_1—2-pole, 12-position ceramic switch, 2 decks
Note: All inductance values in microhenries. All inductors are J. W. Miller 9200 series or equivalent with 2 turn link of #28 insulated wire wound on ground end. All variable capacitors are 5-50 pf (Johanson 9305) except for 6 meters. All resistors 1/4 watt

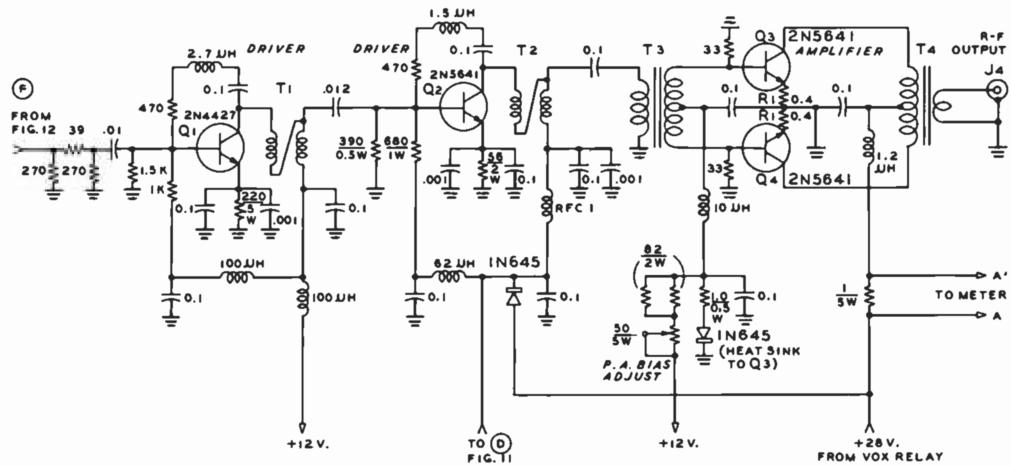


Figure 13

SCHEMATIC, LINEAR AMPLIFIER STAGES

Q_1-Q_4—Motorola transistors (see text)
T_1, T_2—Four-to-one wideband transformer. 8 turns of #28 twisted pair, 8 twists per inch wound on CF-102 core (Indiana General)
T_3, T_4—See figure 14 and text
R_1—Three 1.2-ohm, ½-watt carbon resistors in parallel
RFC_1—Ferroxcube VK200-10/3B
Note: All resistors ¼ watt unless otherwise specified. Dual emitter leads of Q_1 bypassed with .001 µfd and .1 µfd on each lead. All inductance values in microhenrys

power supply is shown in figure 15. It utilizes IC power regulators to provide plus and minus regulated 12 volts. Both positive and negative full-wave rectifier circuits are connected to the secondary of transformer T_1. The +28 volts is used to drive the linear power-amplifier circuits directly and is also connected to regulator U_1, which delivers +12 volts at a maximum current of 500 ma. Regulator U_2 is connected to the negative supply and delivers —12 volts at up to 500 ma. The metering circuit allows the power-supply voltages to be measured as well as allowing the operator to monitor the power amplifier supply current.

The Master Reference Oscillator—Shown in figure 16 is the circuit of the master reference oscillator. The circuit is of the *Seiler* type and gives excellent frequency stability. A box made of ¼" thick aluminum plate is used for the assembly and is mounted on the rear of a *National NPW-0* dial mechanism to achieve the required mechanical rigidity. Drift of the unit shown is less than 50 Hz during the first five minutes of operation at 20°C and less than 10 Hz per hour thereafter at a given temperature. With the temperature compensation shown, frequency change is less than 200 Hz over the range of 0° to 50°C.

A buffer amplifier (Q_2) provides isolation between the oscillator and the load and is partially responsible for the excellent performance of the circuit. Variable capacitor C_3 allows the output level of the unit to be adjusted to 100 millivolts, rms, to drive the following circuit.

The Phase Lock Synthesizer—Shown in figures 17 and 18 are the schematic diagrams of the phase-lock synthesizer. Component values for the oscillators are given in Table 2. Referring to figure 17, integrated circuit U_1 is a balanced mixer with two inputs: one from the crystal oscillator (figure 19) and the other from the voltage-controlled oscillator (VCO) and buffer amplifier shown in figure 18. The crystal frequency is chosen to be below the voltage-controlled oscillator (VCO) frequency so that the difference frequency between the two oscillators falls in the range of 3.21 to 3.71 MHz, determined by the exact frequency the VCO is tuned to. The output of mixer U_1 is filtered by a three-pole bandpass

TABLE 2 OSCILLATOR COMPONENTS										
VCO VALUES ($f_Y = f_O + 9$ MHz)						XTAL OSC. VALUES ($f_X = f_O + 5.790$ MHz)				
BAND (MHz)	f_O	R_4	L_1	C_5 (pF)	C_7 (pF)	C_9 (pF)	C_{10} (pF)	C_{11} (pF)	L_2 (µH)	Y_1 (MHz)
3.5 - 4.0	80 M	4.3 K	19 T. N°28	—	120	10	27	270	6.8	9.290
7.0 - 7.5	40	4.3 K	15T. N°28	1 - 15	82	10	33	250	2.7	12.790
14.0 - 14.5	20	4.3 K	11T. N°28	1 - 15	47	10	200	18	1.5	19.790
21.0 - 21.5	15	4.3 K	8T. N°24	1 - 15	33	22	62	—	1.2	26.790
28.0 - 28.5	10A	6.2 K	6T. N°22	—	30	18	62	—	0.82	33.790
28.5 - 29.0	10B	—	6T. N°22	—	30	18	62	—	0.82	34.290
29.0 - 29.5	10C	—	6T. N°22	—	30	18	62	—	0.82	34.790
29.5 - 30.0	10D	—	6T. N°22	—	30	18	62	—	0.82	35.290
50.0 - 50.5	6A	—	2.5 T. N°18	—	17	12	62	—	0.33	55.790
50.5 - 51.0	6B	—	2.5T. N°18	—	17	12	62	—	0.33	56.290
51.0 - 51.5	6C	—	2.5T. N°18	—	17	12	62	—	0.33	56.790
51.5 - 52.0	6D	—	2.5T. N°18	—	17	12	62	—	0.33	57.290

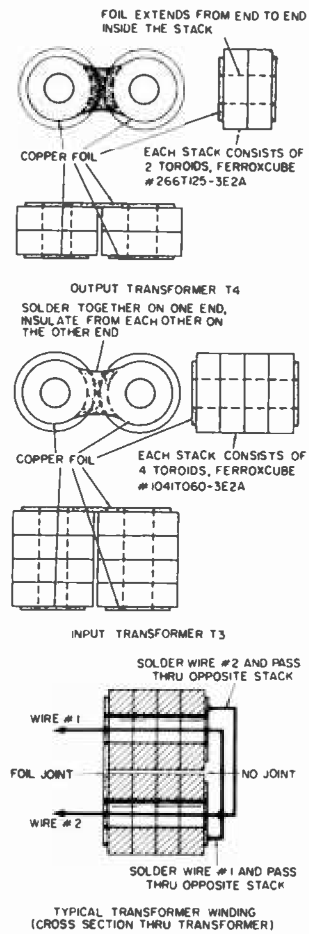


Figure 14

CORE STACK FOR WIDEBAND R-F TRANSFORMERS

Transformers T_3 and T_4 in the linear amplifier are wideband devices made up of stacks of ferrite cores. The stacks are held together by a cylinder of copper foil with adhesive on one side (available Newark Electronics part 38F1301 or 38F1222). Roll the foil around a drill shank of proper size ,adhesive side out ,to form a cylinder. Slide the toroids on the cylinder. Remove the drill and cut the foil so it is 1/8" longer than the stack of toroids on each end. Make 4 to 6 slits in the extended foil and flare out flat against the toroids. Fill in the gaps with small pieces of foil tape and carefully solder in place. Trim even with the edge of the core. Place two stacks side by side and tape together with paper tape. Solder the foil on the end of one stack to the foil on the end of the other stack. This junction forms the center tap of one winding. Solder a short piece of #24 insulated wire to the foil on the other end of each stack and pass the two wires through the adjacent toroid stack. This completes one turn on either side of the center tap. Wind on the remaining turns of the center-tap winding. Finally, wind on the second winding so that the ends of the winding extend from the opposite end of the assembly from the center-tap connection. (Ferroxcube cores available from: Ferroxcube Corporation, 5635 Yale Blvd., Dallas, Texas). See Lowe, QST, December, 1971 for additional transformer data.

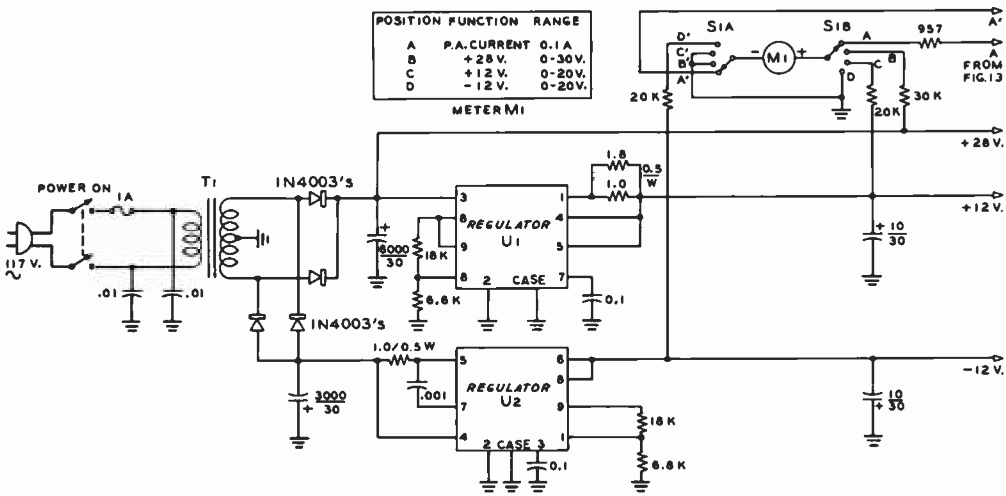


Figure 15

SCHEMATIC, POWER SUPPLY

T_1—Stancor TP-4. Use green, yellow, and red secondary leads
U_1—MC 1461 R
U_2—MC 1463 R
M_1—0-1 ma d-c (Simpson or Weston)

filter (figure 16) which removes unwanted mixer products before they reach the input of the phase detector U_2. The phase detector compares the phase of the signal (and consequently the frequency) to the phase of the master reference oscillator, shown in figure 16, and generates an output signal proportional to the phase difference between the two input signals. This reference signal is d-c coupled to the input of the loop amplifier (U_3, figure 17) after passing through the loop filter (R_1, C_4). This filter shapes the gain-frequency response of the loop and is very important for proper operation of the synthesizer. The values are chosen so that the loop has a 100-kHz pull-in range, that is, if the frequency difference between the master oscillator and the output of U_1 is less than 100 kHz, the loop will lock-up and remain locked. Thus, the VCO will have the same stability as the master reference oscillator.

The output of U_3 is connected to the varicap diode (D_2) in the VCO circuit (figure 18) and also to the lock indicator circuit (Q_1, figure 17). When the loop is locked, only a d-c voltage is present at the output of U_3 and Q_1 is turned off, preventing current from flowing through the light-emitting diode (LED) placed on the exciter panel above the main tuning dial. Should the loop become unlocked, however, a large a-c voltage is developed at the output of U_3, which is rectified by the diodes, thus turning on Q_1. This causes the LED to light, signaling the loop is unlocked. On-the-air opera-

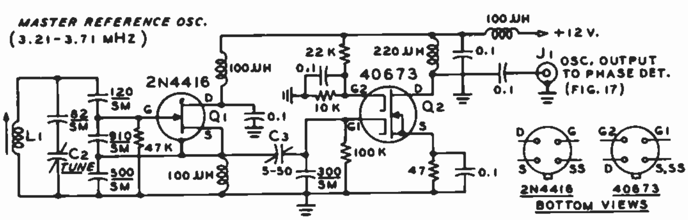


Figure 16

SCHEMATIC, MASTER REFERENCE OSCILLATOR AND BUFFER

C_1—82-pf silver mica with 54-pf, N220 capacitor in parallel
C_2—6 to 78 pf. Polar C341-20/016 (Jackson Bros.)
C_3—5 to 50 pf, Johanson 9305
L_1—51 turns #28 e. on CTC 3354-6 coil form

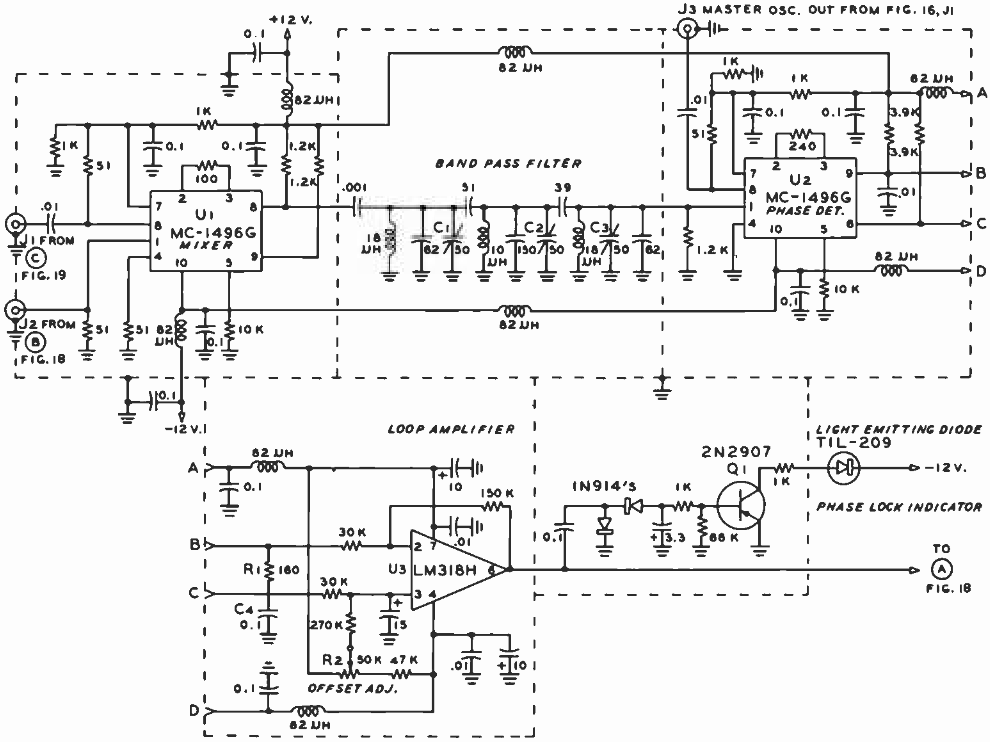


Figure 17

SCHEMATIC, MIXER, BANDPASS FILTER, PHASE DETECTOR, LOOP FILTER, AND LOCK INDICATOR

C_1-C_3—5 to 50 pf. Johanson 9035
Note: All inductance values in microhenries. All inductors J. W. Miller 9200 series. All resistors ¼ watt

tion of the exciter should never be attempted if the loop is unlocked because in this condition the VCO output consists of many frequencies instead of one.

The Voltage-Controlled Oscillator (figure 18)—Another *Seiler* circuit similar to the one used for the master reference oscillator is used as a voltage-controlled oscillator. Two varicap diodes are used to tune the frequency; the first (D_1) is driven from a potentiometer (*coarse tune*) which is mechanically coupled to the dial shaft of the master reference oscillator. This coupling causes the frequency of the VCO to be approximately tuned to the desired frequency selected by the reference oscillator. The second diode (D_2) driven by the loop amplifier, readjusts the frequency slightly so that the loop will lock-up.

Component values for the frequency determining circuit of the VCO (Table 2) are selected to allow the circuit to tune the proper frequency range for the bands shown. Other bands may be covered after considering the mixer products. Devices Q_3, Q_4, and Q_5 (figure 18) are broadband amplifiers which isolate the VCO from the loads and the loads from each other. The output of Q_4 is used to drive the phase-locked loop and the output of Q_5 drives the conversion mixer (U_2, in figure 11).

The Crystal Oscillator (figure 19)—The crystal oscillator consists of a grounded-base Colpitts circuit with the crystal in the feedback path. These crystals have a series-resonant frequency as listed in Table 2. Coil L_2 is a subminiature choke about the size of a ¼-watt resistor. Link L_3 consists

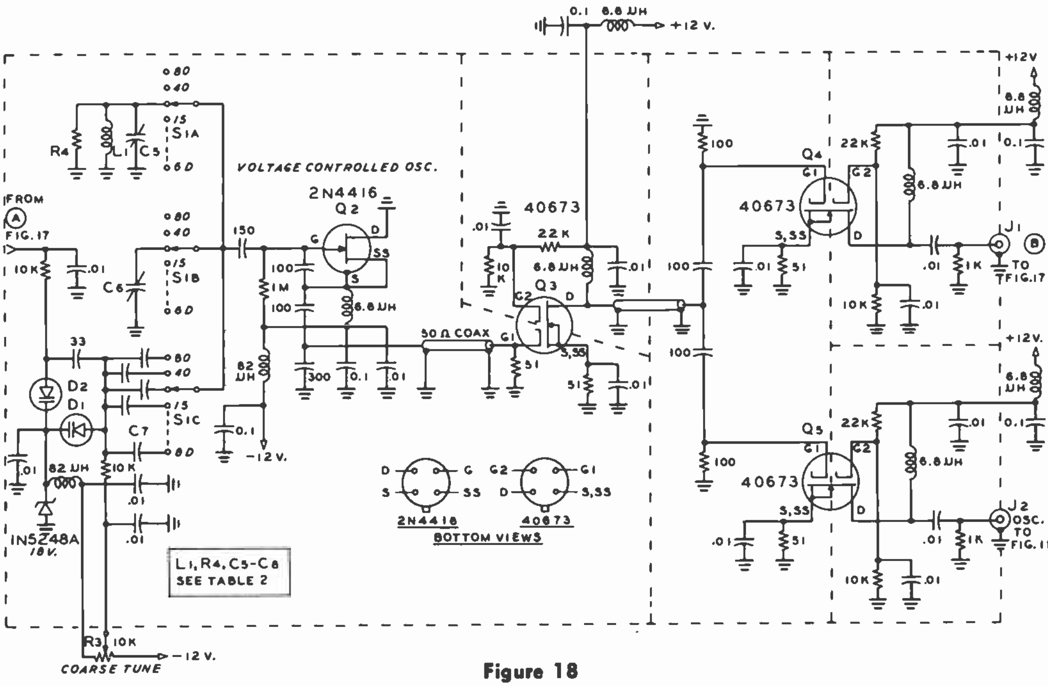


Figure 18

SCHEMATIC, VOLTAGE-CONTROLLED OSCILLATOR AND BUFFERS

D_1, D_2—1N5148A
S_1—6-pole, 12-position switch
Note: For coil and capacitor data, See Table 2. All resistors 1/4 watt. All inductance values in microhenrys

of 1½ turns of #28 insulated wire wound on the ground end of L_2. The output of this oscillator connects to U_1 in figure 17 to complete the phase-lock synthesizer circuit.

Exciter Construction The exciter is built in several modules which are mounted on an aluminum chassis measuring 9" × 14" × 2". The unit is housed within a *Bud Shadow Cabinet* (SB-2142), as shown in figure 5. Placement of the modules is shown in the rear-view photograph (figure 20). The chassis is mounted to the panel with two end brackets. A small gap is left at the rear of the chassis to aid cooling and the rear panel of the cabinet is replaced with a sheet of perforated aluminum. The

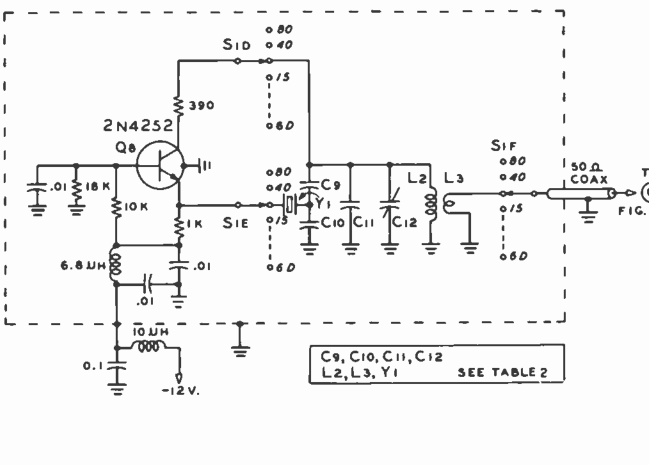


Figure 19

SCHEMATIC CRYSTAL OSCILLATOR

Note: For coil and crystal data, see Table 2. All resistors 1/4 watt

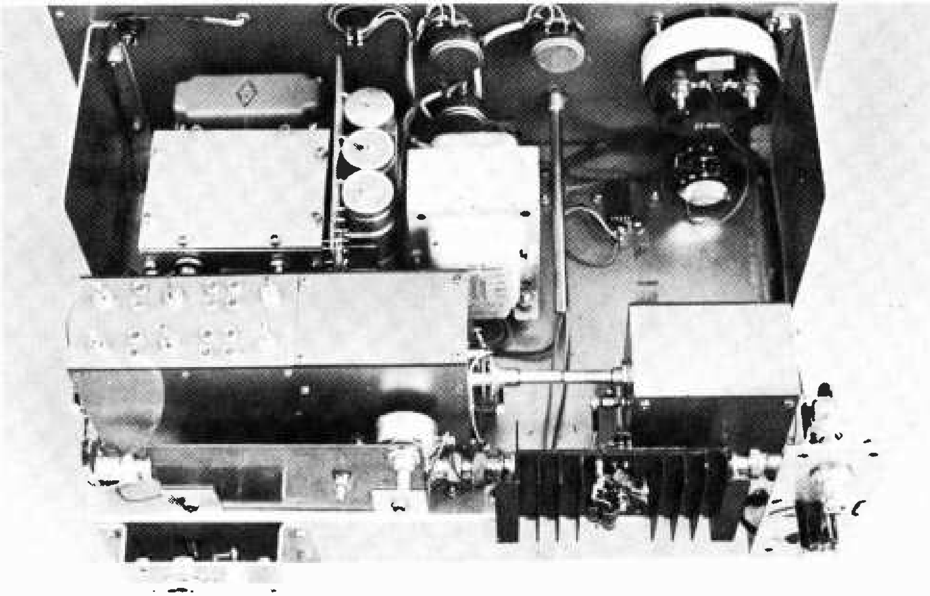


Figure 20

TOP VIEW OF SOLID-STATE SSB EXCITER

Exciter is built in modules that may be tested and aligned one at a time. In the upper left corner is the reduction dial drive and the rugged aluminum box for the master reference oscillator. At center of the chassis are the power transformer and filter capacitors. Directly behind the oscillator is the synthesizer assembly containing the circuits of figures 16, 17, and 18. The bandswitch passes out the end of the module and is ganged with the bandswitch of the switched filter (figure 12). Across the rear of the chassis (foreground of photo) are the driver stages and bias potentiometer (left) and the push-pull linear amplifier (right). Power plug, fuse, and antenna receptacle are on bracket at right. Microphone and VOX inputs are at lower left.

linear-amplifier module is built on a finned heat sink with all but the end fins removed on one side. It is mounted in a vertical position at the rear of the chassis with the remaining fins projecting beyond the chassis into the gap between chassis and cabinet.

Type BNC connectors and miniature coaxial cable are used to interconnect the various modules. Immediately in front of the linear-amplifier module are the switched-filter module (figure 12) and the synthesizer module (figures 16 through 19). Adjacent to the linear amplifier is the driver module (part of figure 11) which includes the *P-A BIAS ADJUST* potentiometer. These modules are tested individually then bolted together and mounted as one unit to the rear of the main chassis.

The aluminum box containing the master reference oscillator is behind the *National* dial drive assembly, with the power supply centered on the chassis.

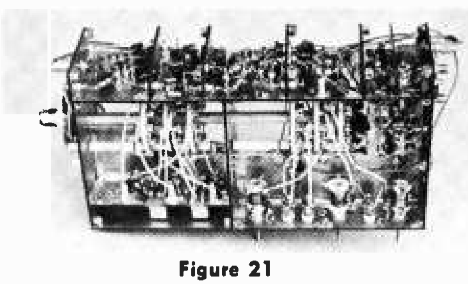


Figure 21

OBLIQUE VIEW OF SYNTHESIZER MODULE

Module is built of double-sided fiberglas printed-circuit board cut into rectangles and soldered together. The bandswitch passes through the two compartments of the assembly. At the left is the crystal oscillator and crystals with the bandpass filter (figure 16) at the upper right. Across the top of the assembly are compartments containing (left to right): mixer, bandpass filter, phase detector, loop filter, and buffer stages (Q_1, Q_2, figure 17). Ten, 15, and 20 meter oscillator coils are at the right.

The enclosures for the switched filter and the synthesizer are built from rectangles cut from double sided 0.06" *fiberglas* p.c. board material and are soldered together. The synthesizer box measures 7" × 4" × 2" and the filter box measures 3" square. A view of the interior of the synthesizer is shown in figure 21. The enclosures are assembled in a similar manner. Threaded brass spacers ¼" long are soldered in the corners to add strength to the box and to secure the covers. The pieces of p.c. board used for the center dividers in the synthesizer box should have the copper soldered together along the exposed edge to provide the best grounds. This was done by wrapping a narrow strip of .001" copper shim stock over the edge of the dividers and soldering on both sides. A good fit between the box panels is obtained by sawing the parts to a slightly large size and then filing the pieces to exact size. After the boxes are soldered together the exposed edges where the covers fit are ground flat with a piece of fine emery cloth placed on a flat surface. This results in a neat enclosure which is strong and compact.

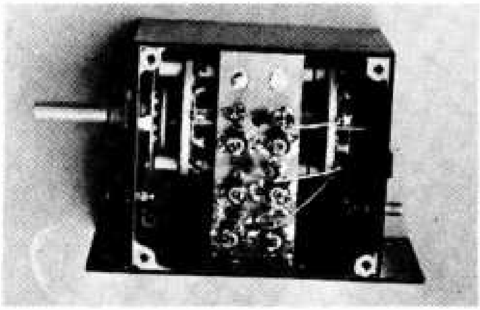


Figure 22

INTERIOR OF SWITCHED FILTER

Filter schematic is shown in figure 12. Filter components are mounted to printed-circuit board placed between the switch decks. Capacitor adjustment screws project from side of box. Input and output coaxial receptacles are on the ends of the box.

The switches for both units are assembled from *Centralab* PA-1 ceramic decks with a PA-302 index assembly used in the synthesizer and a PA-301 assembly used in the switched filter. An interior view of the switched filter is shown in figure 22.

The remainder of the exciter circuitry is mounted on two pieces of p.c. board, each measuring 4" × 10", mounted below the chassis on ¼" spacers (figure 23). The board nearest the front panel contains the audio processing circuits, VOX, and anti-VOX circuitry. The rear board contains the r-f circuitry including the two crystal filters, r-f clipper, and conversion mixer. Input and output terminations are made with BNC coaxial fittings, and each board is tested and aligned before it is placed in the chassis. Small standoff terminals are soldered directly to the copper foil to provide tie points (since no holes are drilled). This is a very fast and convenient method to build the circuits and provides a good ground plane since all grounds may be soldered directly to the copper. Circuit changes or modifications can be done easily and quickly, should the need arise.

Miniature components are used throughout the exciter. The resistors are ¼-watt carbon units, the inductors are approximately the same size (*J W Miller* 9200 series or equivalent), and the bypass capacitors are miniature ceramic units. The small capacitors are *El Menco* DM-5 type mica units. The power-supply components except for the IC regulators are mounted on a vertical p.c. board between the power transformer and the master reference oscillator. A right-angle drive is used to drive the bandswitch from the front panel. When wiring the switches remember that one switch rotates in a direction opposite that of the other when viewed from the front of the switch. The IC regulators are mounted on either end of the chassis to distribute the power dissipation.

Exciter Adjustment Exciter tuneup is not complicated if all modules have been pretested before installation on the chassis. An electronic voltmeter with an r-f probe is required. as well as an audio generator and an oscilloscope. A frequency counter is desirable, but not mandatory.

After checking the units and the power-supply voltage, connect a 5-watt, 50-ohm load to the output terminal. Connect the audio generator through a variable attenuator to the microphone input receptacle and adjust the *COMPRESSION LEVEL ADJUST* control to provide 3 volts rms at pin 6 of U_2 (figure 7) when the *AUDIO GAIN*

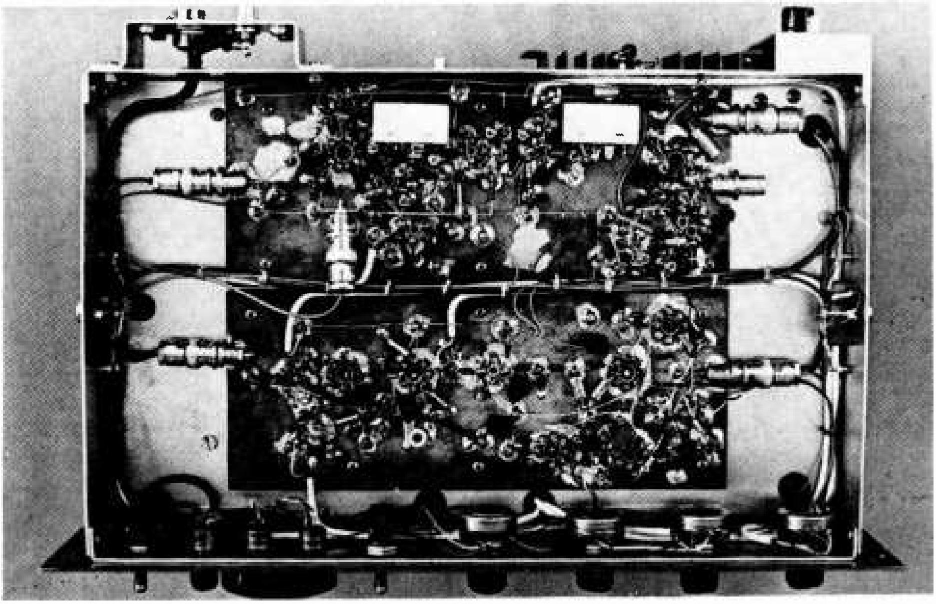


Figure 23

UNDER-CHASSIS VIEW OF SSB EXCITER

The two printed-circuit boards are mounted beneath the aluminum chassis. The board at the rear of the chassis contains the r-f circuitry and the two i-f crystal filters. The board adjacent to the front panel contains the audio circuitry. At the left is the microphone-input coaxial receptacle and at the right is the audio-output coaxial fitting. The r-f output to the switched filter atop the chassis is at the left of the rear circuit board.

and *COMPRESSOR GAIN ADJUST* controls are at mid-setting. The output at pin 6 should remain constant over a signal input range of 40 decibels from the threshold point to the point where waveform distortion becomes visible on the oscilloscope. The *AUDIO GAIN* control can be used to adjust the input level to the compressor to compensate for different microphones and the *COMPRESSOR GAIN* adjustment used to determine how much, if any, compression is used.

To adjust the VOX controls, set the *ANTIVOX* and *VOX GAIN* controls at minimum, turn down the receiver output and turn up the *VOX GAIN* until the VOX relay closes when speaking into the microphone in a normal manner. Now, turn up the speaker to normal output and adjust the *ANTIVOX* control until the relay does not close on loud signals. The *VOX DELAY* control can now be adjusted for proper hold-in time, as desired.

As a final check, connect the oscilloscope to the source of Q_6 (figure 7). No clipping of the waveform should be observed when the *AUDIO CLIP* control is at minimum and clean clipping of the waveform should be visible at maximum clipping setting.

RF Alignment—To align the r-f circuits first adjust capacitors C_5 and C_6 (figure 10) to midrange and peak capacitor C_8 in the emitter circuit of the buffer stage (Q_4) for a maximum reading on the electronic voltmeter with the r-f probe connected to the top of coil L_1. Indicated voltage should be about 100 millivolts rms and may be adjusted, if necessary, by changing the value of the 1K resistor connected to L_2 and C_8.

Now, apply 300 millivolts rms of 1-kHz audio signal to pin 4 of U_1 (figure 10) and peak the r-f output at the source of Q_2, the buffer FET, by tuning capacitors C_1, C_2, C_3, and C_4 in the first crystal filter stage, the clipper amplifier, and the buffer stage. Set the audio frequency to 2.7 kHz and adjust capacitors C_5 (or C_6, depending upon the sideband selected) in the oscillator stage for

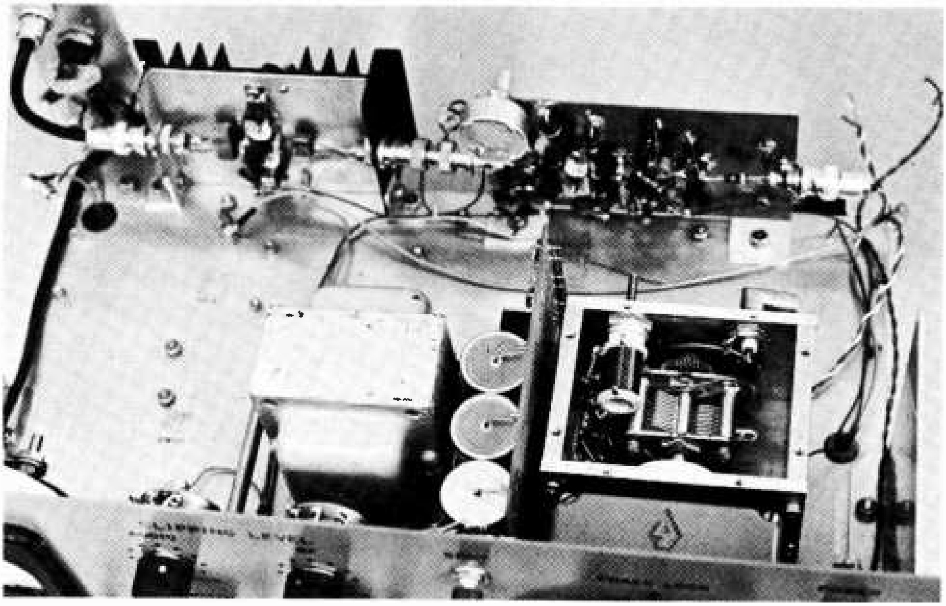


Figure 24

SSB EXCITER CHASSIS AND INTERIOR VIEW OF VFO

The synthesizer module has been removed for this photograph to show the linear amplifiers mounted across the rear of the chassis. The VOX reed relay is directly behind the power-supply filter capacitors. The lid of the vfo module has been removed to show the tuning capacitor and the 1:1 gear drive to the "coarse-tune" potentiometer.

maximum response. Continue tuning the capacitor until the output decreases 3 to 6 decibels. Repeat this procedure with the other capacitor for the opposite sideband. Next, vary the frequency of the audio generator from 300 to 3000 Hz and note the ripple in the filter passband and the upper frequency at which the output has fallen off by 6 decibels. The ripple should be less than plus or minus one decibel across the band. If it is greater than this, adjust capacitors C_1 and C_2 slightly. In an extreme case, it may be necessary to alter the number of secondary turns of transformer T_1.

Next, disconnect the +28-volt line from the VOX relay to the linear amplifier and turn on the *VOX OVERRIDE* switch to remove the cutoff bias applied to Q_3 in the standby mode. Connect the r-f probe to pin 4 of U_2 (figure 11) and adjust capacitors C_8, C_9, and C_{10} (filter FL_2 and the driver transistor) for maximum response with the *DRIVE ADJUST* potentiometer at mid-setting. Again, check the passband ripple and realign capacitors C_8 or C_9, if necessary.

Synthesizer Alignment—Apply +12 volts to the master reference oscillator (figure 24) and adjust coil L_1 and capacitor C_1 so the oscillator tunes the range of 3.185 to 3.735 MHz. Adjust the potentiometer coupled to the shaft of capacitor C_2 so that it is at the clockwise end of its rotation when the oscillator is tuned to 3.735 MHz. Rotate the shaft of the potentiometer back about 10 degrees before locking in place to eliminate the nonlinear portion of rotation next to the stop. Adjust capacitor C_3 to provide 100 millivolts rms output when the oscillator is connected to a 50-ohm load through the subminiature 50-ohm coaxial line and connectors. Apply —12 volts to the crystal oscillator (figure 19) and tune capacitor C_{12} for maximum output on each band. Adjust the coupling between coil L_2 and link L_3 by sliding L_3 up or down the form until the oscillator output is about 100 millivolts on each band, using a 50-ohm load.

Disconnect pin 6 of U_3 (figure 17) from the 10K resistor and varicap diodes (figure

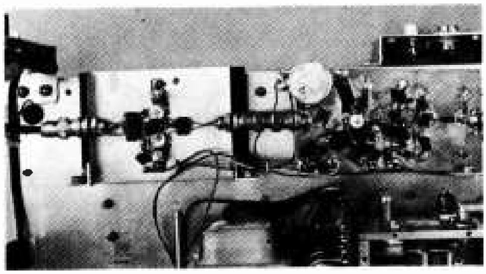


Figure 25

LINEAR AMPLIFIER STAGES

The push-pull amplifier is on the left chassis, with the driver stages on the right chassis. The chassis have been removed from the main chassis and tilted forward for this picture. The "bias adjust" potentiometer is mounted to the plate connecting the two assemblies together.

18) and ground the open end of the resistor. (Do *not* ground pin 6 of U_3). Place all padding capacitors in the VCO (figure 18) to midrange and tune coil L_1 until the output frequency (as measured at the drain terminal of FET Q_4) is nearly correct for each band, starting with 80 meters and working up in frequency. (Remember the frequency you are measuring is 9 MHz higher than the desired band).

With the electronic voltmeter connected to pin 6 of U_3 and power disconnected from the crystal oscillator, adjust the *OFFSET ADJUST* potentiometer (figure 18, loop amplifier U_2) for a reading of zero volts, d-c. Turn off the power and reconnect the 10K resistor to pin 6 of U_3.

Next, tune the master reference oscillator to 3.185 MHz and set the bandswitch to *80 meters*. Connect the electronic voltmeter and oscilloscope to pin 6 of U_3, being careful not to short this point to ground. Turn on power and adjust coil L_1 (80 meters) for zero volts d-c at pin 6. The phase-lock indicator should be off and the oscilloscope should indicate no a-c voltage present. Repeat the tuning of L_1 for each band in sequence, leaving the oscillator at 3.185 MHz. Now set the master reference oscillator to 3.735 MHz and adjust capacitor C_5 or C_6 for the appropriate band for a zero volt d-c reading. Again, start with the 80-meter band and work up in frequency. It probably will be necessary to repeat the procedure twice to get all bands properly tuned. As a final check, tune completely across each band to make sure the loop does not become unlocked at any frequency. If the loop unlocks, the voltage at pin 6 will rise, possibly as high as 10 volts. If this happens, readjust the oscillator capacitor and inductor slightly for a different L/C ratio. For conditions of lock, the voltage at any point in the band should remain between zero and 6 volts.

If a frequency counter is available, the above procedure can be speeded up. Break the line that connects one end of the *coarse tune* potentiometer to the 1N5248A diode and insert a switch in the line. Tune the master reference oscillator to 3.185 MHz with the switch closed and adjust coil L_1. Open the switch and the VCO will be tuned to the high end of the band, even though the master oscillator is still tuned to the low-frequency end. Now capacitors C_5 and C_6 may be aligned as indicated by the counter which is connected to pin 1 of the mixer (U_1, figure 17). Remember that the counter reads a frequency that is 9 MHz above the desired band.

Switched Filter Alignment—A 50-ohm load is connected to the linear amplifier and power is applied to all stages. The *DRIVE ADJUST* potentiometer is set for minimum drive and the *VOX OVERRIDE* switch is turned on. The idling current to the power output stage (as read on the panel meter) is set to 20 ma by adjusting the *PA BIAS ADJUST control.* Connect the audio generator and inject a 1-kHz tone into the exciter, advancing the *DRIVE ADJUST* control (figure 25) to mid-setting. Set the master reference oscillator to a midband frequency and tune the capacitors in each filter section (starting with 80 meters) for a peak current reading on the meter, adjusting the drive control as necessary so as not to exceed 400 ma. The higher bands will require more drive than the lower bands, and output on the 50-MHz band is drive-limited. With 400 ma indicated current, power output will be in excess of 5 watts. If a two-tone source is available, maximum current drain should be limited to 300 ma for 5 watts PEP output.

21-3 A Single-Band 200-Watt PEP SSB Transceiver

Probably the most popular item of equipment for general SSB operation is the transceiver—a complete station in a compact package. Since many of the tubes and components are common to both the transmitting and receiving functions, the transceiver can be built compactly and rather inexpensively, and it is well suited for both fixed-station and mobile operation.

The most economical and least complex transceiver to build is one designed for use on a single amateur band. Multiple mixing schemes and complex coil catacombs are thus eliminated, and the "birdie" problem is greatly simplified. Shown in this section is a 200-watt PEP, single-band transceiver (figure 26) which may be used on any one amateur band from 160 to 20 meters. It is relatively simple in design and is an ideal "first" project for those amateurs interested in building their own sideband gear. While a commercial 9-MHz crystal filter is used, substitution of a homemade crystal filter is practical, further reducing the cost of the transceiver.

The Transceiver Circuit The transceiver circuit is a proven one that has been employed in many com-

Figure 26

FRONT VIEW OF TRANSCEIVER

The transceiver panel measures 12⅛" wide by 6⅝" high. The two large controls at center are for final amplifier tank and vfo tuning. On the left area of the panel are the modulator balance control (top), r-f gain adjustment, receiver volume, and microphone gain control (next to the microphone jack). The lower switch is the main power control (S_2) and the meter switch is at the top, right. Below the plate tuning control are the grid tuning adjustment and the function switch, S_3. On the right of the panel are the carrier level control, R_6, and the antenna loading capacitor, C_4. The cabinet is a wrap-around style made from two pieces of perforated aluminum sheet bent into a U-shape and riveted together at the sides. Panel and cabinet are primed and painted with aerosol (spray) paint.

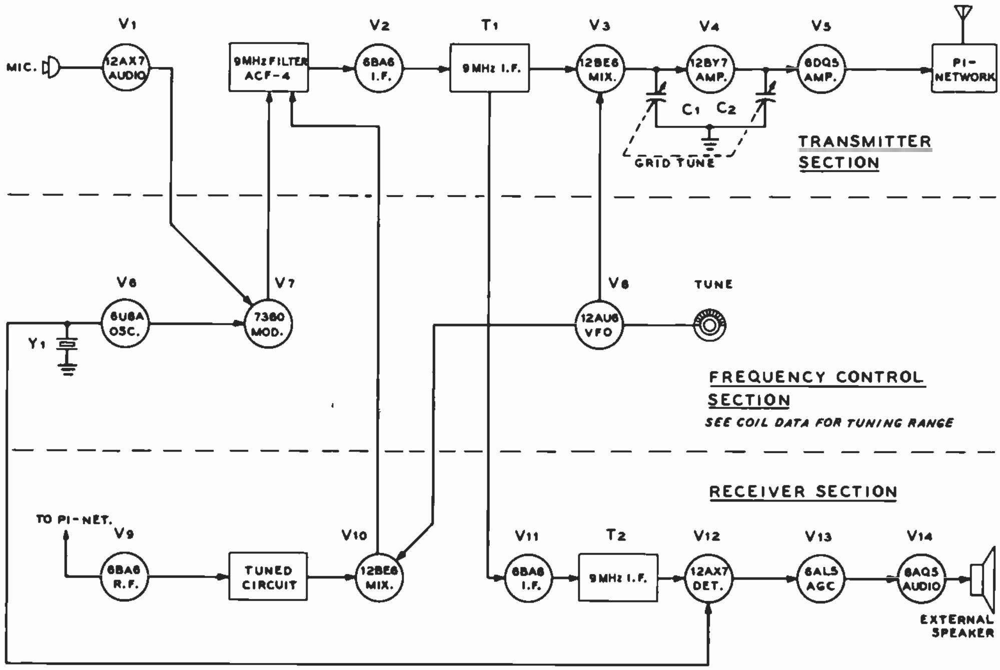


Figure 27

BLOCK DIAGRAM OF SINGLE-BAND SSB TRANSCEIVER

Fifteen tubes are used in a multipurpose circuit. Common r-f tank circuits and i-f filter system simplify construction and reduce cost. A single vfo tunes both receiving and transmitting sections.

mercial units and is a version of the original W6QKI (*Swan*) circuit. Fifteen tubes are used, including a voltage regulator and the unit is designed to be operated from either a 115/230-volt a-c primary supply or a 12-volt transistor power pack (external). Operation of the single-band SSB transceiver and the dual function of some of the tubes and tuned circuits may be seen from an inspection of the block diagram of figure 27.

Reception—In the receiving mode, the circuit takes the form of a single-conversion superheterodyne featuring product detection. The received SSB signal is resonated in the antenna input circuit which, in this case, is the pi-network of the transmitter portion of the unit. The network is capacitively coupled to a 6BA6-remote cutoff r-f amplifier (V_9). The plate circuit (L_1-C_1) of the 6BA6 is common to both receiver and transmitter circuits. A 12BE6 (V_{10}) serves as a receiver mixer, the input signal being mixed with the local vfo signal to produce a 9-MHz intermediate frequency. The vfo stage is common to both transmit and receive circuits and tunes approximately 200 kHz in the region of 5 to 8 MHz, the exact tuning range depending on the band in use. A 12AU6 (operated at slightly reduced filament voltage serves as the oscillator tube (V_8).

The 9-MHz i-f signal passes through the selective crystal lattice filter (ACF-4) and is amplified in a common i-f stage (V_2) which is transformer coupled to a second (receiving) i-f stage (V_{11}) and then fed to a product detector (V_{12}). At this point in the circuit, carrier is injected in the detector from the 6U8A common crystal oscillator (V_6) and the resulting audio product is amplified in one-half of the 12AX7 dual triode (V_{12}) and the 6AQ5A output tube (V_{14}). A portion of the audio signal returns to the 6AL5 automatic gain control rectifier (V_{13}) to provide an audio-

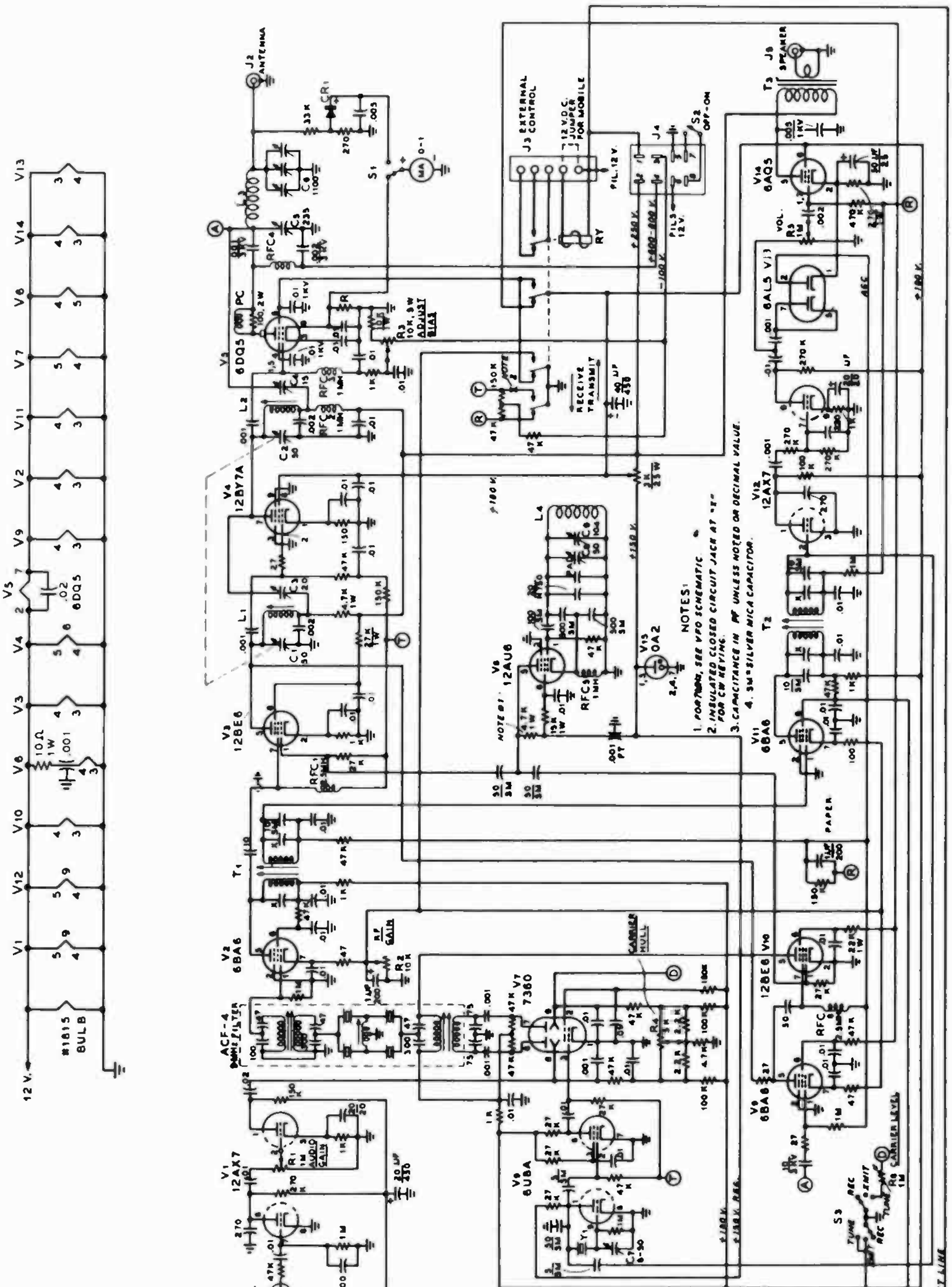


Figure 28

SCHEMATIC, SINGLE-BAND TRANSCEIVER

PARTS LIST FOR FIGURE 28

C_1, C_2—50-pf each; two Hammarlund HF-50 ganged
C_3—20-pf variable mica trimmer
C_4—15-pf, type APC.
C_5—235-pf. Gap .024"; Bud 1859
C_6—365-pf per section; J. W. Miller 2113
C_7—50-pf. Centralab 827
C_8—50-pf, type APC
C_9—104-pf precision capacitor; Miller 2101
CR_1—1N34
J_1—Amphenol 80-PC2F
J_2—Coaxial receptacle. SO-239
J_4—Chassis receptacle; Cinch-Jones P-308AB
MA—Calrad, 0-1 ma d-c, 1¾" meter
PC—4 turns #18 around 100-ohm, 2-watt resistor
R—Meter shunt for 300 ma. Use #30 enamelled, wire wound on 47-ohm, ½-watt resistor
RFC_1 thru RFC_6—2.5 mH subminiature choke; Miller 70F-253-AI
RFC_7, $_8$, $_9$—1 mH choke; Miller 4652
RFC_4—Use Miller RFC-14 for 80-40-20 meters; Use Miller RFC-3.5 for 160 meters
RY—4PDT, 12-volt coil; Potter-Brumfield KHP-17-D11
S_1—Centralab PA-2007
T_1, T_2—10.7-MHz i-f transformer; capacitor X is internal part of unit; Miller 1457
T_3—5000 ohms to 4 ohms; Stancor A-3877
Y_1—International Crystal Co types CY6-9LO (9001.5 kHz) or CY6-9HI (8998.5 kHz) as required
ACF-4—International Crystal Co 9 MHz SSB filter
1—Chassis, 10" x 12" x 3", Bud AC-413
1—Box, 4" x 5" x 3", Bud AU-1028
1—Box, 4" x 4" x 2"; Bud AU-1083
2—Insulated shaft couplers; Johnson 104-264
1—Dial drive; Eddystone 892

derived agc voltage for the receiver section. A fixed positive voltage taken from the cathode of the 6AQ5A stage provides delay voltage for the agc circuit to allow maximum receiver sensitivity to be realized with weak signals. Receiver volume is controlled in the grid of the 6AQ5A stage instead of the low-level audio circuit so that agc action is independent of the audio volume level.

Transmission—In the transmitting mode, the circuit takes the form of a single-conversion, crystal-filter SSB exciter, featuring a 7360 balanced modulator and a 6DQ5 linear amplifier. Switching the circuitry from receive to transmit is accomplished by a single relay (RY) which applies blocking bias (−100 volts) to inactivate tubes used only in the receiving mode. The relay also applies screen voltage to the 6DQ5 r-f amplifier (V_5) and grounds the cathode of the common 6BA6 i-f amplifier stage to nullify the receiving r-f gain control during transmission. The receiver r-f amplifier stage remains connected to the plate circuit of the linear amplifier of the transmitter section, but the 6BA6 amplifier is protected from strong-signal damage by virtue of the high negative bias applied to it in the transmission mode.

When transmitting, the sideband carrier is generated by the common crystal oscillator and buffer stage (V_6). The carrier is coupled into #1 grid of the 7360 balanced modulator (V_7) and the audio signal from the 12AX7 speech amplifier is applied to one deflection plate of the 7360. The resulting double-sideband signal passes into the crystal filter which suppresses the undesired sideband and the carrier, which is already somewhat attenuated by the balanced modulator stage. The desired sideband is amplified in the common 6BA6 i-f stage and passed to the 12BE6 transmitting mixer (V_3) where it is mixed with the vfo signal to produce an SSB signal on the same frequency as the signal being received. The SSB signal is further amplified in the 12BY7A driver stage (V_4) and the 6DQ5 linear amplifier (V_5). When the pi-network plate circuit of the 6DQ5 has been properly tuned for transmission, it is also tuned for optimum reception and requires no further adjustment unless a large frequency excursion is made. The same is true of the 12BY7A tuned circuit (marked *grid tune*).

Transceiver Layout and Assembly The transceiver measures 12⅛" wide by 6⅝" high by 10⅛" deep. A 10" × 12" × 3" aluminum chassis is used for the assembly, with the vfo components mounted in two 4" × 4" × 2" aluminum utility boxes, one atop and one beneath the chassis. The final amplifier plate circuit components are inclosed in a third utility box measuring 4" × 5" × 3" in size. Layout of the major components may be seen in the drawings and photographs (figures 29, 30, 31, and 32). The cabinet is a homemade wrap-around type made of two pieces of perforated aluminum sheet bent into a U-shaped inclosure and riveted together at the sides.

Data is given in the tables for coils, crys-

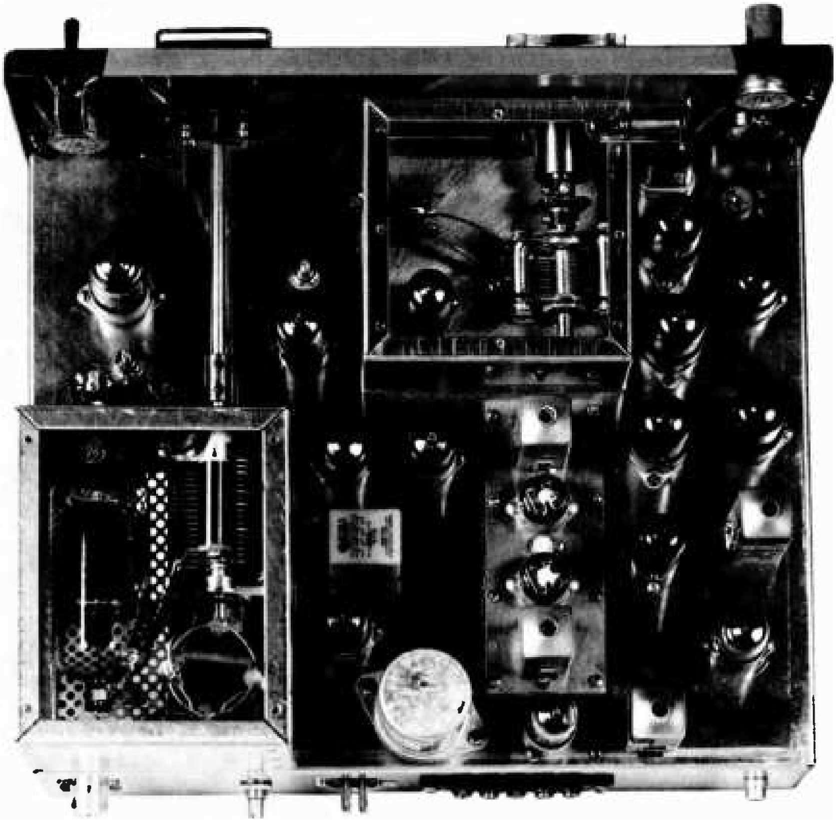


Figure 29

TOP VIEW OF CHASSIS

The SSB transceiver is compact in size, yet not crowded on the 10" x 12" chassis. The use of standard aluminum utility boxes for component inclosures provides excellent shielding at low cost. The box covers have been removed to show interior layout. Ventilation is provided for the horizontally mounted 6DQ5 linear amplifier tube by making a cutout in the chassis below the tube and covering the opening with a sheet of perforated aluminum. A new box cover is made of the same material. The relay to the right of the amplifier box is fully inclosed in a dust cover. Along the rear apron of the chassis are the coaxial antenna receptacle, the bias adjustment potentiometer, the power plug and relay terminal strip, with the speaker jack at the far right.

The 12BY7A driver tube is located between the amplifier box and the front panel, with the 12BE6 transmitter mixer to the right. The 6BA6 receiver r-f stage and 12BE6 mixer are between the relay and the vfo, with the OA2 regulator behind the relay, adjacent to the filter capacitor. The 9-MHz i-f filter strip is at center with the 6BA6 common i-f tube behind it.

At the right, next to the vfo are (going back from the panel): the 9-MHz crystal, the 6U8A oscillator, the 7360, and the 6AQ5A audio amplifier. At the extreme right of the chassis are the 6AL5 agc tube, the 12AX7 speech amplifier stage and the 6BA6 receiver i-f stage.

tals and frequencies to be used to build a transceiver for 160-, 80-, 40-, or 20-meter operation using standard components. The layout has been planned to allow short r-f leads where necessary, and to permit proper circuit isolation. In most cases, resistors and bypass capacitors are mounted directly at the tube-socket pins with liberal use of tie-point terminals to achieve solid construction. The resistor network for balancing the voltage on the deflection plates of the 7360 modulator tube is mounted on a separate

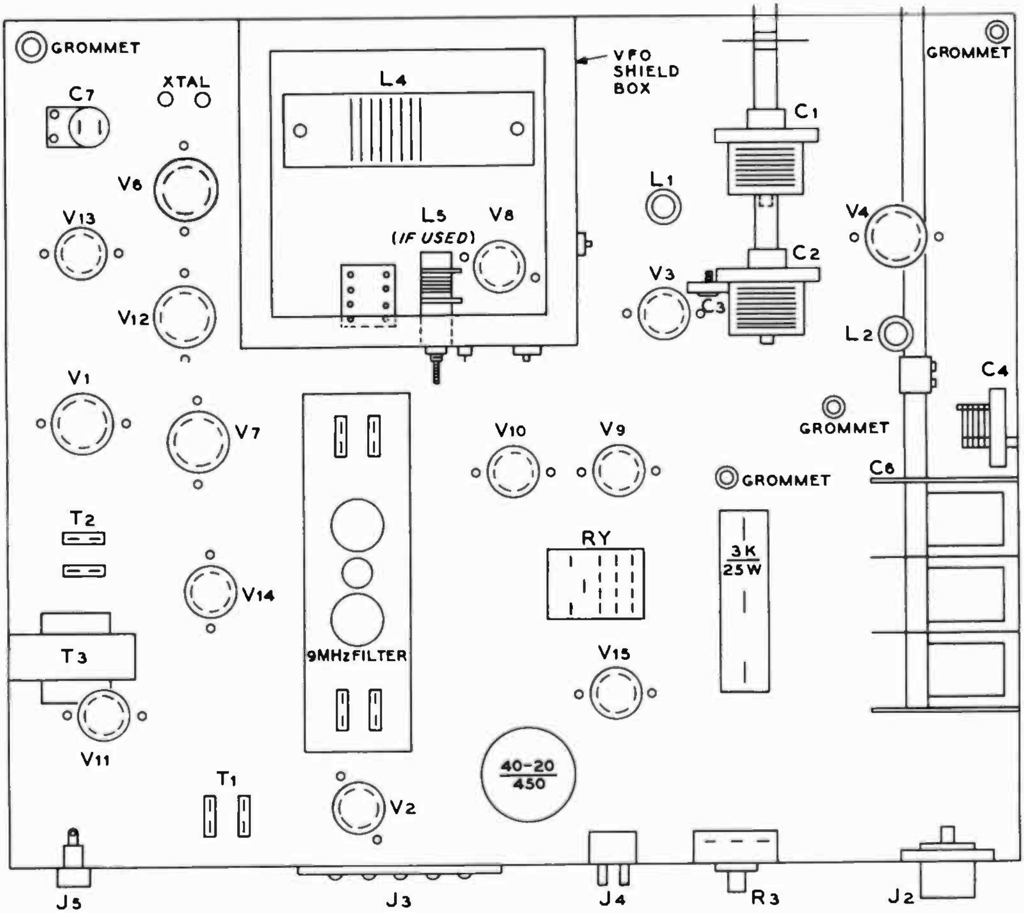


Figure 30

UNDER-CHASSIS LAYOUT OF TRANSCEIVER

terminal board fastened to the side of the chassis, and a second terminal board is used for mounting the r-f choke in the vfo, cathode circuit and the associated capacitors (figure 32). The power plug, relay terminal strip, final amplifier bias potentiometer, and speaker jack are placed on the rear apron of the chassis.

Final amplifier components are placed inside the utility box bolted to the top rear corner of the chassis. The chassis area beneath the 6DQ5 tube is cut out and covered with a perforated aluminum sheet, as are the top and rear of the box, to achieve proper circulation of air around the tube.

The vfo (figure 33) is placed at the front-center of the chassis and is constructed on a ⅛-inch thick plate of aluminum measuring 4″ × 4¼″ in size. The vfo tuning capacitor is fastened to this sturdy base by mounting bolts from the underside of the plate. A precision, silver-plated tuning capacitor having ball bearings and closely controlled torque is used in conjunction with a 10-to-1 ratio epicyclic driving head to achieve a smooth, backlash-free tuning system.

One aluminum utility box is bolted to this mounting plate from the bottom side to serve as a shield compartment for the vfo coil and circuit components. The vfo coil is made from airwound inductor stock (*miniductor*) securely affixed to a ¼-inch

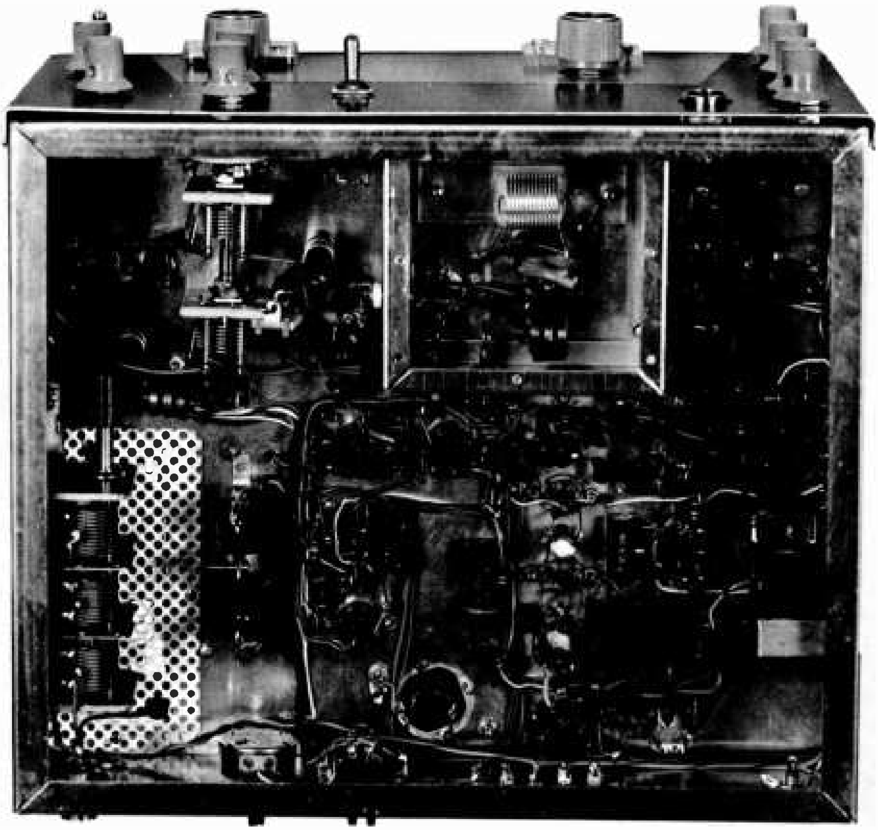


Figure 31

UNDER-CHASSIS VIEW OF THE TRANSCEIVER

The bottom plate has been removed from the vfo compartment to show internal layout. The three-gang antenna loading capacitor, C_6, is bolted to the side apron of the chassis (right) as is the audio output transformer (left). Small components are soldered directly to tube socket terminals and adjacent tie-point strips, leaving the sockets clear for voltage measurements. See Figure 30 for placement of major components.

thick block of *plexiglas* or other insulating material which, in turn, is bolted to the chassis with similar insulating blocks spacing it away from the metal.

Operating voltages are brought into the under-chassis shield box via feedthrough capacitors and the vfo output leads are connected to feedthrough bushings on the sides of the box nearest the transmitting and receiving mixer tubes. A second utility box is bolted to the top of the vfo plate, spaced about ¼ inch back from the front apron of the chassis to permit clearance for the dial and drive mechanism. The drive head is passed through a ¾-inch hole in the front of the utility box and is bolted to the box in line with the capacitor shaft and affixed to it with a flexible coupler. A 4½" diameter circular piece of sheet plastic is attached to the drive head to form the tuning dial. It is spray-painted white and calibration marks are lettered on it with India ink after final calibration is completed. Sufficient clearance is left between the dial and the chassis so the plastic does not rub on the metal.

The front panel is spaced away from the chassis by virtue of the large nuts holding the various controls on the front apron of the chassis and is affixed in place with a

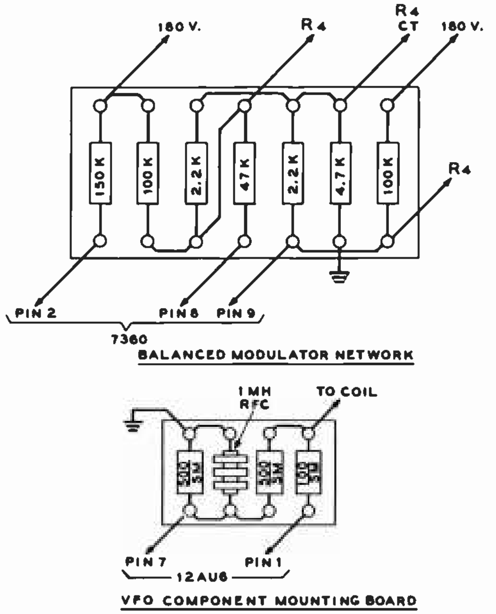


Figure 32

TERMINAL BOARD LAYOUT

second set of nuts on the control bushings. The ⅛-inch space thus created provides room for the dial to rotate freely. A cutout is made in the panel in front of the dial to match the appearance of the meter. The opening is covered with a section of *plexiglas* or *lucite* inscribed with a hairline indicator. A pilot light behind the dial provides proper illumination. The hole in the panel for the tuning shaft should be made sufficiently large so the shaft does not touch the panel, making the tuning mechanism independent of any panel movement.

Transceiver Wiring It is suggested that the receiver portion of the transceiver be wired and tested first. The sideband filter comes as a wired package with matching transformers and requires only a slight modification. The mounting plate is cut down to a width of 1¾″ to conserve space and new mounting holes are drilled along the edges of the plate. The filter assembly is then attached to the transceiver chassis over a slot cut just behind the vfo assembly. The output connection of the filter assembly goes to the grid of the 6BA6 i-f amplifier tube (V_2). The grounded side of the input transformer secondary is lifted from ground, bypassed and connected to the 1000-ohm decoupling resistor in the supply-voltage circuit. The other end of this secondary winding is connected to the plate of the 12BE6 receiver mixer tube. The primary winding is modified for balanced input by grounding the junction of the two 75-pf capacitors and connecting the end of the winding to the plates of the 7360 balanced-modulator tube through the .001-μfd coupling capacitors.

The driver (*grid tune*) capacitors (C_1-C_2) are *Hammarlund HF-50* units ganged together and mounted on the chassis by means of the supplied brackets. A flexible coupling is used to extend the shaft through the front panel. The 12BY7A neutralizing capacitor (C_3) is soldered directly to the stator terminal of the plate-circuit capacitor (C_2) of the amplifier stage. The final amplifier neutralizing capacitor (C_4) is placed on the side apron of the chassis in front of the three-gang antenna loading capacitor (C_6).

Transceiver Coils and Circuits—Coil and tuned-circuit data for the various amateur bands are given in figure 34. For the 160-, 80-, and 20-meter bands, the fundamental frequency of the vfo is employed. For 40-meter operation, the plate circuit of the vfo doubles the oscillator frequency to the 16-MHz range. Lower sideband is used for the 160-, 80-, and 40-meter bands, and upper sideband for the 20-meter band. Substitution of crystal Y_1 will reverse the sidebands, as shown in the table. Additional loading capacitance may be required for proper amplifier operation on 160 meters and may take the form of a 1000-pf (1250-volt) mica capacitor placed in parallel with antenna loading capacitor C_6.

Transceiver Alignment Before starting alignment of the transceiver, it is suggested that a wiring check be made and a voltage check be done with a suitable power supply. No high voltage is required to begin with, and the screen power lead of the 6DQ5 should be temporarily disconnected at the socket pin and taped until preliminary alignment is completed. After the slider on the 300-ohm high-voltage drop-

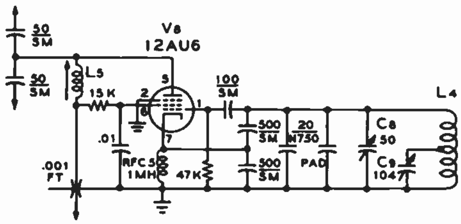


Figure 33

40-METER VFO SCHEMATIC

The 40-meter model of the single-band SSB transceiver employs the second harmonic of the oscillator frequency. A doubler coil, L_5, is placed in the plate circuit of the vfo in place of the 4.7K load resistor. Tuning capacitor C_9 is tapped down the grid coil to cover the tuning range desired. Tap point and padding capacitor data are given in Figure 34.

ping resistor has been adjusted to provide a tap voltage of about 180, tube-socket voltages should be compared to the voltage chart (figure 35). The difference noted in receive and transmit voltage in some cases is due to the cutoff bias being switched in and out of the circuit by the changeover relay. The relay is d-c operated, and for fixed-station service a 12-volt d-c source must be used. When operating mobile this relay terminal is jumpered to the 12-volt d-c filament supply.

The receiver i-f system is aligned first by injecting a 9-MHz modulated test signal at the grid of the receiver i-f amplifier (V_{11}) and tuning the slugs in transformer T_2 for maximum audio signal in the attached speaker. The test generator is then moved back to the input grid of the common i-f amplifier (V_2) and transformer T_1 is adjusted for maximum signal. A vacuum-tube voltmeter on the agc line is helpful in alignment.

When the test signal is injected at the plate terminal of the receiving mixer tube (V_{10}) tuning becomes rather sharp going through the sideband filter. The filter is factory tuned and needs little adjustment other than peaking the top slugs of the two filter transformers. The secondary of the input transformer should be checked, but should not require adjustment more than one-half turn in either direction.

Before an "outside" signal is received, the variable-frequency oscillator must be

COIL DATA

L_1, L_2

3/8" DIA. SLUG TUNED COILS

160 METERS-	220 µH	*MILLER #21A224RBI*
80 METERS-	22 µH	*MILLER #21A225RBI*
40 METERS-	15 µH	*MILLER #21A155RBI*
20 METERS-	3.3 µH	*MILLER #21A336RBI*

L_3

160 METERS -	55 TURNS #20 ENAMEL WIRE CLOSE WOUND. 1 1/4" DIA., 1 3/4" LONG.
80 METERS -	24 TURNS #18 TINNED WIRE. *AIR-DUX #1014A.* 1 1/4" DIA., 1 3/4" LONG.
40 METERS -	14 TURNS #18 TINNED WIRE. *AIR-DUX #1014A.* 1 1/4" DIA., 1" LONG.
20 METERS -	11 TURNS #18 TINNED WIRE. *AIR-DUX #808T.* 1" DIA., 1 1/2" LONG.

L_4 NOTE: C_9 *TAPPED ON* L_4 *FOR VARIOUS RANGES.*

BAND	TUNING RANGE	
160	7200-7000 kHz	9 TURNS #20 TINNED WIRE, 3/4" DIA., 3/4" LONG. TAP 4TH TURN FROM GROUND END. *AIR-DUX #616.* PAD CAPACITOR 51-PF SM.
80	5500-5000 kHz	12 TURNS #20 TINNED WIRE, 3/4" DIA., 3/4" LONG. *AIR-DUX #616.* PADDING CAPACITOR 100-PF SM.
75 PHONE	5200-5000 kHz	SAME COIL AS ABOVE. TAP 8TH TURN FROM GROUND END. PADDING CAPACITOR 180-PF SM.
40	8000-8150 kHz	9 TURNS #20 TINNED WIRE. 3/4" DIA. 3/4" LONG. TAP 3RD TURN FROM GROUND END. *AIR-DUX #616.* NO PADDING CAPACITOR.
40 PHONE	8100-8150 kHz	SAME DATA AS ABOVE, EXCEPT TAP 2ND TURN FROM GND END
20	5000-5500 kHz	SAME DATA AS FOR 80 METERS.
20 PHONE	5200-5350 kHz	SAME DATA AS FOR 75 PHONE. ADJUST TRIMMER C_8 FOR DESIRED RANGE.

L_5

40 ONLY	16000-16300 kHz	3/8" DIA. SLUG-TUNED COIL. 3.3 µH. *MILLER 21A336RBI*

CRYSTAL DATA (Y_1)

160 METERS LOWER SIDEBAND	- USE 9001.5 kHz
80 METERS LOWER SIDEBAND	- USE 9001.5 kHz
40 METERS LOWER SIDEBAND	USE 9001.5 kHz
20 METERS UPPER SIDEBAND	- USE 8998.5 kHz

Figure 34

aligned to cover the desired operating range, as listed in the coil table. The alignment procedure is the same for any band; only the frequency range is different as indicated on the chart. Use of a good frequency meter (such as a BC-221) will be helpful at this point. With the 80-meter unit as an example, the vfo must tune from 5.5 to 5.0-MHz for proper coverage of 3.5 to 4.0 MHz. The carrier crystal is at 9001.5 kHz to properly place the carrier on the slope of the filter for lower sideband output. Coil L_1 of the 6BE6 transmit mixer is tuned to

TUBE-SOCKET VOLTAGE CHART											
TUBE			1	2	3	4	5	6	7	8	9
V_1	12AX7	R-	50	0	.8	0	12	40	0	0	CT
		T-	55	0	1	0	12	45	0	0	CT
V_2	6BA6	R-	0	0	0	6	175	75	.8		
		T-	0	0	0	6	175	70	.5		
V_3	12BE6	R-	-40	0	0	12	220	220	-40		
		T-	-1	6	0	12	210	80	0		
V_4	12BY7	R-	0	-35	0	0	12	CT	250	180	0
		T-	4	-5	0	0	12	CT	250	180	0
V_5	6DQ5	R-	-60	6	0	0	-60	0	6	0	
		T-	-60	6	0	180	-60	0	6	180	
V_6	6U8	R-	75	-40	180	6	0	180	0	0	-2
		T-	75	0	100	6	0	35	0	0	-2
V_7	7360	R-	0	180	-40	6	0	180	180	24	24
		T-	0	75	-1	6	0	140	140	24	24
V_8	12AU6	R-	*	0	0	10	120	115	0		
		T-	*	0	0	10	120	115	0		
V_9	6BA6	R-	0	0	0	6	210	80	.2		
		T-	-70	0	0	6	200	0	0		
V_{10}	12BE6	R-	-.5	0	0	12	180	80	-.2		
		T-	-6	0	0	12	175	0	-107		
V_{11}	6BA6	R-	0	0	0	6	175	80	.5		
		T-	-107	0	0	6	175	140	0		
V_{12}	12AX7	R-	145	0	0	0	12	100	0	.4	CT
		T-	175	-75	0	0	12	100	0	.4	CT
V_{13}	6AL5	R-	10	0	6	0	0	0	0		
		T	0	-140	6	0	0	0	0		
V_{14}	6AQ5	R-	0	10	0	6	225	180	0		
		T-	-60	0	0	6	250	180	-60		
V_{15}	0A2	R-	150	0	0	0	150	0	0		
		T-	150	0	0	0	150	0	0		

NOTE: *MEASUREMENTS MADE WITH A 20,000 OHM-PER-VOLT METER. NO SIGNAL INPUT, R-F GAIN ADVANCED TO MAXIMUM, AUDIO GAIN OFF. FILAMENTS A.C.*

POWER SUPPLY REQUIREMENTS

LOW VOLTAGE -	250 VOLTS AT 110 MA.
BIAS -	110 VOLTS NEGATIVE AT 10 MA.
HIGH VOLTAGE -	600 TO 800 VOLTS AT 300 MA.
FILAMENTS -	12.6 VOLTS A.C. OR D.C. AT 3.7 A.

Figure 35

3.5 MHz with the aid of a grid-dip oscillator, the slug being adjusted with capacitor C_1 set near maximum capacitance. The entire 80-meter band can then be covered by peaking the pi-network and grid-circuit tuning controls.

Alignment of the transmitting circuits is best done with the v.t.v.m. using an r-f probe for signal indication. The function switch is placed in the *tune* position and the carrier-level control (R_6) advanced toward maximum. R-f voltage at the plate of the 6U8A oscillator should measure about 3 or 4 volts, and about the same value should be observed at the plate of the buffer section of this tube. Inasmuch as the filter transformers and transformer T_1 have been adjusted previously, no further adjustment of these circuits is required. The r-f probe can now be placed at the grid of the 6DQ5 amplifier tube socket and the slug in coil L_2 adjusted for maximum r-f voltage reading. This peaks grid tuning so that coil L_2 will track with the previous alignment of coil L_1.

Final Adjustment and Neutralization The 12BY7A stage should now be neutralized. To accomplish this, all power is turned off and the screen lead temporarily removed from the 12BY7A socket. With power again turned on, circuits resonated, and the function switch in the *tune* position, neutralization capacitor C_3 is adjusted with a nonmetallic screwdriver for minimum feedthrough of r-f voltage as measured with the v.t.v.m. probe placed at the #1 grid terminal of the 6DQ5 socket. The screen lead to the 12BY7A socket is replaced when this operation is concluded. The same technique is employed with the 6DQ5 stage as was used with the driver stage. With screen (and plate) voltage removed from the 6DQ5, but with drive applied, the v.t.v.m. is placed on the antenna terminal of the transceiver and neutralizing capacitor C_4 adjusted for minimum voltmeter indication. The pi-network circuit, of course, is in resonance for this operation, as determined by a grid-dip oscillator.

Up to this point, all tuning has been done with carrier injection. For proper sideband operation, the carrier must be removed and the unit excited by an SSB signal. The technique is to position the carrier crystal frequency properly on the filter "slope" and then to balance out the carrier in the 7360 modulator stage. Capacitor C_7 varies the frequency of the crystal oscillator a sufficient amount to find the proper point for the carrier on the passband slope of the filter. The adjustment of this point can best be made by ear, when receiving a sideband signal. Adjust capacitor C_7 until the received audio of an SSB signal sounds natural and pleasing. The crystal should be about 1500 Hz away from the 9-MHz filter center frequency. The frequency displacement, of course, will remain the same while transmitting.

Carrier null is accomplished by adjustment of the balance control (R_4) on the panel. The r-f probe is placed at the grid of the 6DQ5 stage and the function switch turned to *transmit*. No audio signal is desired. The balance potentiometer is adjusted for minimum indicated reading on the v.t.v.m., which should be 1 volt or less. Operation of the audio system and balanced modulator may now be checked by noting the voltage swing while talking into the microphone. A sustained audio tone will swing the meter to 30 or 40 volts peak reading. It is helpful to monitor the signal in a nearby receiver while these adjustments are being made.

Transmit Operation The screen-voltage lead may now be reconnected to the 6DQ5 tube socket and high voltage provided for the plate circuit. Potentials between 400 and 800 volts may be used for the 6DQ5, with proportionately higher output at the higher plate voltages. An antenna or dummy load must be connected to the transceiver to complete the final checkout and bias adjustment. The meter switch is set for *plate current* and the function switch for *transmit*. The bias potentiometer on the rear apron is adjusted for a 6DQ5 resting plate current of 25 milliamperes. Antenna loading is done with the function switch in the *tune* position. As the carrier control is advanced, the final-amplifier plate current will rise in a linear fashion. The amplifier plate circuit is brought into resonance and the grid circuit adjusted for peak plate current reading. Loading control C_6 is adjusted for further increase, reestablishing resonance with the tuning control until the indicated cathode current reaches a value of 275 to 300 milliamperes. Full load current should not be run for more than 20 seconds at a time to achieve maximum amplifier tube life. When the function switch is advanced to *transmit*, amplifier plate current will drop back to the original idling value of 25 ma. As the audio level is raised, speech will kick the indicated current up to values in the vicinity of 125 to 170 milliamperes depending on the individual voice. Too high values of peak current will result in distortion and splatter.

The meter may be switched to read relative power output which, in some cases, will simplify loading the amplifier, especially during mobile operation, as tuning may be done for maximum output reading under a controlled level of excitation.

The 80-meter version of the SSB transceiver is shown in the photographs. The only difference in a unit designed for a different band is modification of the r-f coils and the vfo circuitry. Alignment and tuneup is the same for all bands. The transceiver may be used for c.w. by employing block-grid keying. Operation on c.w. is with carrier control fully advanced and function switch in the *tune* position while transmitting. The switch is manually returned to *receive* for reception.

A discussion of suitable power supplies is given in a later chapter of this Handbook.

21-4 A 200-Watt 3-Band Sideband Transceiver

A mobile SSB transceiver covering three bands can be built utilizing few more parts than a single-band unit, and without requiring any great increase in size over a single-band model. This compact and inexpensive triband transceiver (figure 36) is designed for 80-, 40-, and 20-meter operation at levels up to 200 watts peak envelope power input. Upper sideband, lower sideband, or amplitude modulation may be transmitted on each band. Push-to-talk circuitry is included and the transceiver may be operated from a six- or twelve-volt d-c power source or from a 115-volt a-c supply. Weighing only a few pounds, the transceiver measures only 10" × 12" × 6½" in size — small enough to fit into "compact" cars!

Circuit Description A block diagram of the transceiver is shown in figure 37. Fourteen tubes and two voltage regulators are used. As practically all mobile operation is done on voice, the tuning range of the transceiver can be limited to the phone segments of the bands used. With such a restricted tuning range, bandpass coupling between low-level r-f stages is practical in both the transmitting and receiving sections of the unit, thus eliminating the need of variable tuning controls for several stages. The variable-frequency oscillator is common to both transmitting and receiving sections and tunes only 350 kHz, which is ample range for the 80-meter band and provides full coverage of the 40- and 20-meter bands. Although several of the tubes in the unit are common to both trans-

Figure 36

200 WATT PEP SIDEBAND TRANSCEIVER FOR 80, 40, AND 20 METERS

Less than a cubic foot in volume, this inexpensive transceiver will fit into today's "compact" automobile. Unit may also be used with auxiliary 115-volt a-c supply for the home station. The major controls on the panel are (l. to r.): sideband switch (S_1), SSB/a-m selector switch (S_3), audio volume (R_3), microphone gain (R_1), carrier injection (S_2), band-selector switch (S_{5-8}), microphone jack (J_1), r-f gain (R_4), meter-selector switch (S_9), antenna loading capacitor (C_{15}), and final amplifier tuning (C_{14}). The main frequency-control dial (C_5) is at top center. Wrap-around, perforated cabinet provides ventilation and acts as TVI shield.

Once adjusted for a particular band, the only tuning required is done with the vfo control. Bandpass coupling allows large excursions in frequency. The vfo tuning mechanism with 100:1 ratio makes sideband tuning a pleasure.

mit and receive sections, the receiver r-f section is independent of the transmitter section to make construction easier and to facilitate alignment. The final amplifier tank circuit, however, is used as the antenna input circuit for the receiver to take advantage of the high Q of the circuit and to conserve space. Only two relays are required for receive-transmit changeover and these relays are actuated by the microphone push-to-talk circuit. One miniature relay (RY_2) grounds the grid of the r-f amplifier in the receiver (V_9) for protection during transmissions and a second relay (RY_1) switches various voltages between transmit and receive circuits. Full automatic gain control (agc) is incorporated in the receiver, together with an auxiliary r-f gain control. When transmitting, an automatic level control (alc) system reduces flat-topping and serious overload distortion. The single panel meter may be switched to read cathode current of the linear amplifier stage or relative power output at the antenna receptacle.

The transceiver is designed around the *McCoy* 9-MHz sideband filter, utilizing the sum and difference products created by mixing with a 5-MHz vfo signal to cover the 80- and 20-meter bands. Forty-meter output is obtained by premixing the vfo signal with a 21.5-MHz crystal oscillator to provide a tuneable 16.5-MHz variable-frequency injection signal. This, mixing in turn with the 9-MHz sideband signal, produces a difference frequency in the 7-MHz range.

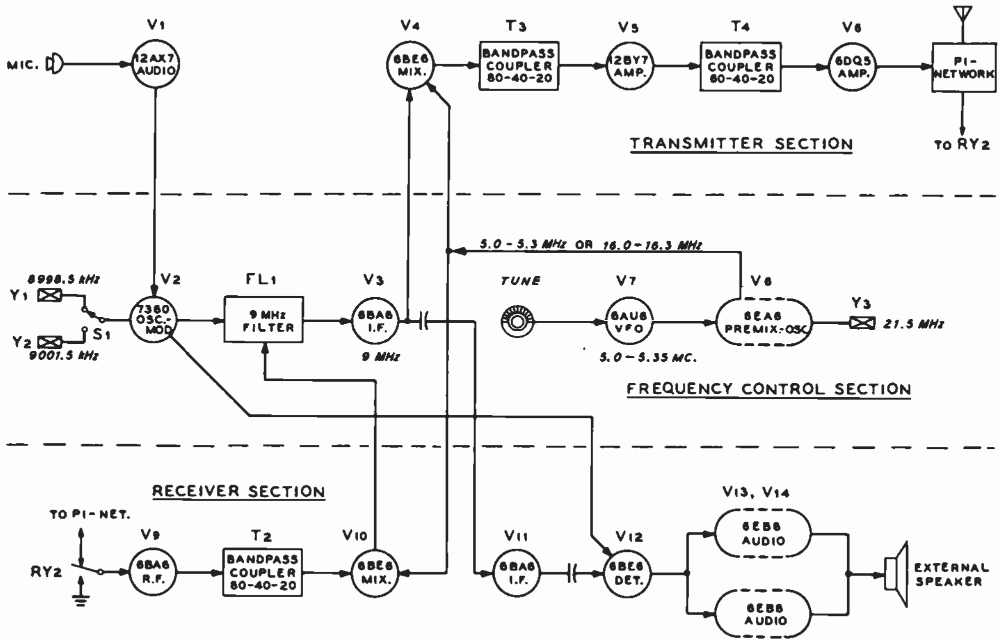


Figure 37

BLOCK DIAGRAM OF TRANSCEIVER

Frequency-control section of unit is common to both receiver and transmitter sections. Beam-deflection type 7360 serves as carrier oscillator and modulator, followed by 9-MHz crystal sideband filter and i-f amplifier stage. Variable-frequency oscillator and mixing oscillator for 7-MHz operation are also common to both sections of transceiver. Transmitter section comprises microphone amplifier and transmitter mixer followed by two linear amplifier stages. Receiver section consists of r-f amplifier and mixer followed by additional i-f stage, product detector, and audio amplifier. Simple push-to-talk circuit switches configuration from transmit to receive.

The Receiver Portion—The receiver portion of the unit starts with a 6BA6 remote-cutoff r-f amplifier (V_9) bandpass-coupled to a 6BE6 mixer (V_{10}) whose injection grid receives mixing voltage from the common 6AU6 vfo (V_7) via the buffer stage (V_8). The 6EA8 buffer functions as a pre-mixer for the vfo on 40 meters when the cathode of the triode section is grounded to activate the 21.5-MHz crystal oscillator.

The intermediate-frequency output of the 6BE6 receiver mixer is 9 MHz and the i-f signal is link-coupled via L_6 to the input of the 9-MHz crystal filter (FL_1). A matching transformer couples the low output impedance of the filter to the grid circuit of the common i-f amplifier (V_3). The received signal is capacitively coupled from this stage to a second 6BA6 receiver i-f amplifier (V_{11}) whose output circuitry is capacitively coupled to a 6BE6 product detector (V_{12}). Oscillator injection for SSB reception is from either of the two sideband crystals in the grid circuit of the 7360 carrier oscillator-balanced modulator (V_2) which is common to receive and transmit sections. Collector plate voltage is removed from the 7360 during reception by relay RY_1C but the oscillator section always functions since deflector and screen voltage is applied in either mode.

The 6BE6 product detector (V_{12}) may be switched to function as a plate detector for reception of a-m signals (S_3ABC). This changeover requires disabling the 7360 carrier oscillator, but since this oscillator is required for transmitting, the a-m changeover switch is routed through the main changeover relay (RY_1B) so voltage is applied to the carrier oscillator when trans-

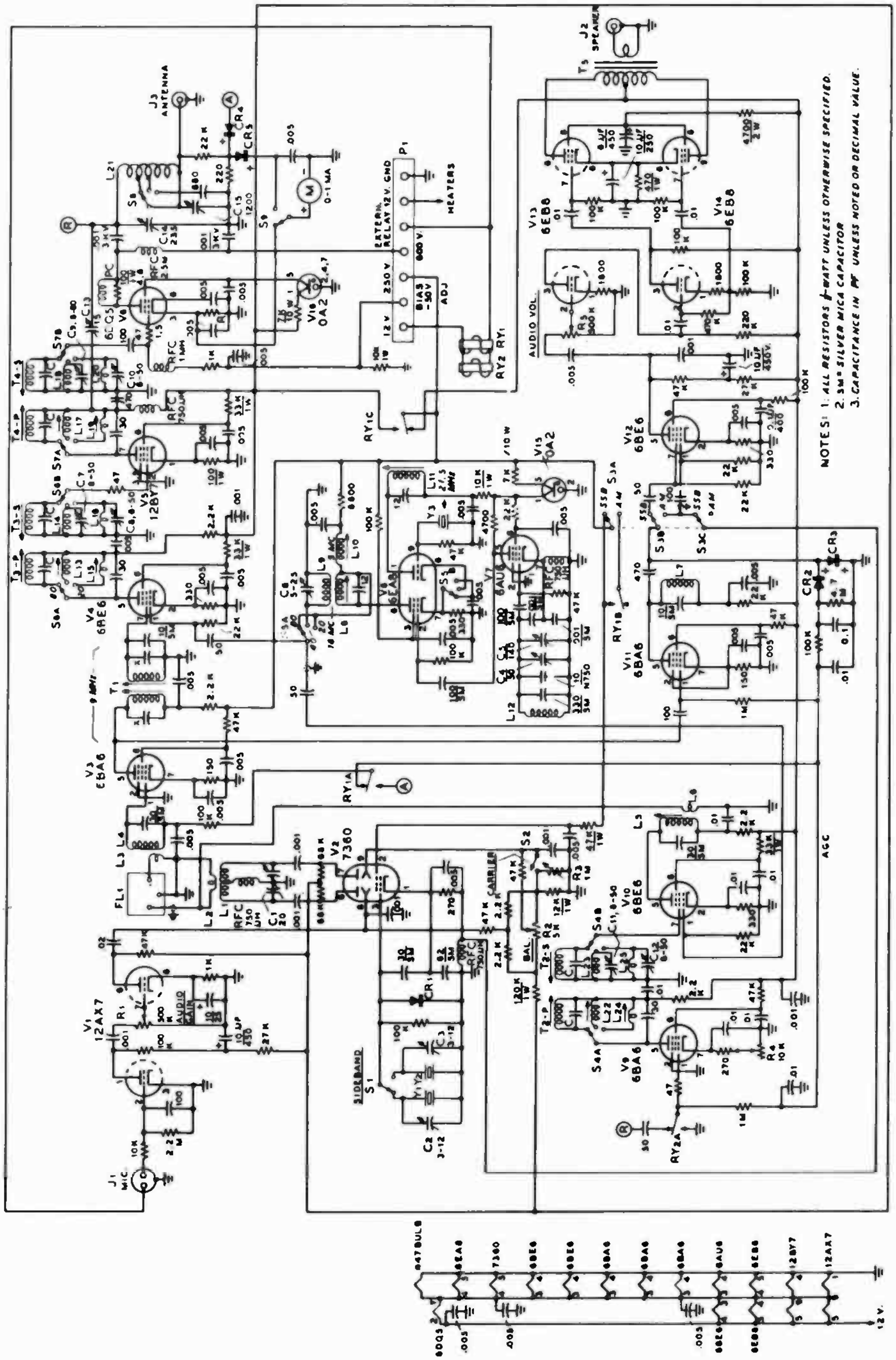


Figure 38

SCHEMATIC OF TRANSCEIVER

PARTS LIST FOR FIGURE 38

C_1—20-pf differential capacitor (E. F. Johnson 160-311)
C_2, C_3—12-pf ceramic variable (Centralab CRL-827)
C_4—50-pf (Hammarlund MAPC)
C_5—140-pf (Hammarlund MC-140M)
C_6—25-pf ceramic variable (Centralab CRL-827)
C_7 thru C_{12}—50-pf ceramic variable (Centralab CRL-827)
C_{13}—15-pf (Hammarlund MAPC).
C_{14}—235-pf (Bud 1859)
C_{15}—1200-pf, 3-gang broadcast-type capacitor (J. W. Miller 2113)
CR_1 thru CR_5—Diode, 1N34 or equivalent
FL_1—9-MHz crystal sideband filter (McCoy SSB-9, McCoy Electronics, Mt. Holly Springs, Pa.)
M—0-1 d-c milliammeter, 1¾" square (Cal-Rad, or equiv.).
PC—Parasitic choke. 7 turns #18e. wire on 100-ohm, 1-watt composition resistor
P_1—8-contact chassis-mounting plug (Cinch-Jones P-308AB)
R_1—1-megohm potentiometer with switch S_2 attached
R—Meter shunt for 300-ma range. Approx. 10" #32 copper wire wound on 47-ohm, ½-watt resistor placed at cathode terminal of 6DQ5
RFC—2.5-mH, 300-ma. (National R-300U)
RY_1—3PDT relay, 12-volt d-c coil (Potter-Brumfield KM-14D or equiv.)
RY_2—DPDT relay, 12-volt d-c coil (Potter-Brumfield KM-11D or equiv.)
S_1 A, B, C—3-pole, 2-position wafer switch (Centralab CRL PA1007)
S_2A, B; S_3A, B; S_4A, B; S_5A, B; S_6—2-pole ceramic wafer sections (Centralab PA-2 each, ganged on Centralab PA-301 index assembly)
T_1—Transformer, 10.7-MHz TV i-f type, (J. W. Miller 1463). (x indicates internal component)
T_2, T_3, T_4—Transformer, 4.5-MHz TV interstage type (J. W. Miller 6270). (c indicates internal component)
T_5—Universal output transformer, 10K plate-to-plate (Stancor A-3823)
Y_1—8898.5-kHz crystal (furnished with FL_1)
Y_2—9001.5-kHz crystal (furnished with FL_1)
Y_3—21.50-MHz crystal (International Crystal Co. FA-5)

mitting, regardless of the setting of the SSB/a-m switch (S_3).

Mobile operation requires a receiver having a reserve of audio power and the audio section is designed to meet this requirement. Two 6EB8 triode-pentode tubes (V_{13}, V_{14}) are employed, with the pentode sections used as a push-pull audio stage. One triode section of the first 6EB8 is used as an audio phase inverter and the second triode is used as the driving amplifier for the phase inverter. The two dual-purpose tubes take up no more space than the usual two-tube amplifier stages but produce nearly 5 watts of high-quality audio. The speaker is not incorporated in the transceiver, since use of the speaker in the auto radio is contemplated. For home use, an auxiliary speaker is incorporated in the 115-volt a-c power unit.

The Transmitter Portion—The transmitter portion of the unit starts with a 12AX7 two-stage speech amplifier (V_1) driving a deflection plate of the 7360 carrier oscillator—balanced modulator (V_2). When transmitting, voltage is applied to the collector plates of the 7360 via relay RY_1C and the carrier is generated by the triode section of the tube functioning as a crystal oscillator. Choice of upper or lower sideband is made by proper crystal selection by means of sideband-selector switch S_1. The balanced-modulator plate circuit of the 7360 is link-coupled to the 9-MHz filter for rejection of the unwanted sideband and passage of the desired sideband to the common 6BA6 i-f amplifier (V_3). The sideband signal is then transformer-coupled to the 6BE6 transmitter mixer (V_4). This mixer stage receives its mixing voltage from the vfo and buffer premixer stages (V_7, V_8) in the same manner as the receiver. Output of the 6BE6 transmitter mixer is at either 80, 40, or 20 meters and is bandpass-coupled on the desired band to a 12BY7 amplifier-driver (V_5). This stage, in turn, is bandpass-coupled to a neutralized 6DQ5 (V_6) serving as a class-AB_1 linear amplifier. The final tank circuit of the amplifier is a pi-network configuration providing good harmonic attenuation and ease of adjustment.

Transceiver Construction Transceiver construction is straightforward and should be no problem for the advanced amateur. The vfo is built as a separate unit and may be tested and aligned before it is installed in the transceiver. The receiver portion of the unit should be wired and tested before the various transmitter stages are completed. The transceiver is constructed on a 10" × 12" × 3" steel chassis. Layout of the major components and shield partitions are observed in the photographs and drawings. The 6DQ5 amplifier tube socket is recessed so that panel height is only 6½". Standard parts are used throughout with the exception of the vfo tuning capacitor. The vfo is built as a unit on the frame of a worm-gear driven capacitor removed from the amplifier stage of a surplus SCR-274N/ARC-5 transmitter. Only the worm gear and frame assembly are used and the original capacitor plates are removed (figure 39). A double bearing 140-pf receiving-type variable capacitor is installed in the frame in place of the original capacitor assembly, slipping the spring-loaded drive gear over the shaft of the new capacitor so that it engages the worm gear as did the rotor of the original capacitor. The free space inside the framework is used to mount the various components of the vfo as shown in the photograph. An aluminum plate is bolted to the back frame to support the tube socket (V_7) and an L-shaped shield is bolted over the top and end of the frame to inclose the assembly.

A circular dial cut from 1/16-inch plastic or *plexiglas* is placed on the large gear in lieu of the original metal dial. The new dial is spray-painted white on the front and calibration marks are lettered with India ink. The complete vfo is bolted to a base plate of ⅛" thick aluminum, slightly larger in area than the capacitor framework. The completed assembly is then bolted to the transceiver chassis with the center of the dial at the center line of the chassis. The plastic dial will extend below the front apron of the chassis, requiring a slight amount of clearance so that it does not rub. The panel is spaced away from the chassis apron by the lock washers and nuts that fasten the various controls, allowing clearance for the dial. The panel is secured in place with a second nut on each control. The upper edge of the panel and the rear lip of the chassis are bolted to the wrap-around cabinet to provide a rigid structure immune to vibration.

Component Layout—Most of the major components are mounted atop the chassis as shown in figures 40 and 41. The antenna receptacle (J_3), power plug (P_1) and jack for the external speaker (J_2) are placed on the rear apron of the chassis and all other major controls are mounted on the front panel with the exception of the phase-balance capacitor (C_1) and voltage-balance potentiometer (R_2) which are placed on the chassis to the rear of the 7360 tube socket. These controls need be adjusted only in the initial alignment and ordinarily require no further attention.

The main bandswitch runs down the center line of the under-chassis area (figure 44) with wafer section S_{1-7} (inclusive) bolted individually to the small partitions that act as interstage shields. Switch wafer S_8 for the 6DQ5 amplifier plate tank coil is mounted in the amplifier compartment on the rear apron of the chassis below tank coil L_{21}, with the connecting wires from the coil brought below deck through an oblong hole in the chassis. The shaft of this switch is ganged to the main bandswitch shaft by means of a link-and-arm arrangement shown in figure 35. Two small lever arms are made by taking apart a flexible shaft coupler. One arm is slipped over the main band-switch shaft at the point where it enters the under-chassis shield plate behind the main panel, and the second arm is attached to the fiber extension shaft driving the amplifier switch wafer (S_8) mounted on the rear apron of the chassis. The two lever arms are interconnected by a narrow strip of aluminum having a hole at each end for small bolts to secure it to the two lever arms. Panel bushings in the shield plate act as bearings for the switch shafts.

The bandpass coils are constructed as indicated in the coil table (figure 43) with the exception of the coils for the 80 meter band. These are ready-made 4.5-MHz TV replacement interstage transformers (T_2, T_3, and T_4). They are used without alteration and provide the desired bandpass effect by virtue of stagger-tuning between 3.8 and 4.0 MHz.

A great deal of the wiring may be done before the shield partitions or switch assemblies are put in place. The switch wafers are installed one at a time, beginning with the receiver segment at the rear of the chassis. The side and front shield plates are made of thin aluminum and are installed last, being bolted to each other, the switch partitions, and the chassis to make a rigid assembly (figure 44).

Terminal boards are used for the small

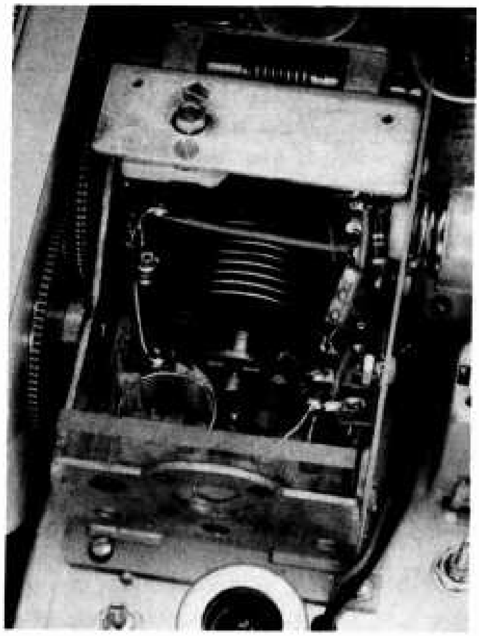


Figure 39

CLOSEUP OF TRANSCEIVER OSCILLATOR

Stable vfo for triband transceiver is made from frame of SCR-274N capacitor. Capacitor plates are removed and 140-pf capacitor substituted. A small bracket bolted to the frame supports padding capacitor C_4. Airwound vfo coil is in foreground, cemented to a ¼-inch thick block of polystyrene which is bolted to capacitor frame. Oscillator tube socket is mounted on side of capacitor and tie point behind it supports cathode r-f choke and various mica capacitors. Connections to vfo unit are terminated at lug strip mounted below the tube socket.

components of the balanced modulator and audio systems. Other small components are mounted to tube-socket terminals and tie-point terminal strips.

Testing and Alignment The transceiver will operate with any power supply capable of delivering between 500 and 800 volts at an intermittent load of 250 milliamperes for the final amplifier, and 250 volts at 125 milliamperes for the receiver and exciter sections. Bias requirement is −50 volts at 5 milliamperes (adjustable).

For fixed-station use and bench alignment, a voltage-doubler power supply using a TV replacement transformer works very well. Two 6.3-volt windings in series will provide filament voltage and this may be rectified to provide direct current to operate the relays. A −50 volt bias supply for the final amplifier stage is also required.

Alignment of VFO and premixer—The first step in the alignment procedure is to adjust the main vfo to tune the range of 5.0 to 5.35 MHz. Since the vfo is made as a separate assembly, it may be aligned and tested before installation on the chassis by applying voltage to the various terminals and monitoring the frequency in a well-calibrated receiver capable of tuning the operating range of the oscillator. A BC-221 frequency meter will aid in this effort. The 21.5-MHz crystal oscillator (V_8) and premixer stage can be adjusted with a vacuum-tube voltmeter and r-f probe placed at the switch arm of S_5 A. With the bandswitch in the 80- or 20-meter position, a voltage will be observed at this point and the slug of coil L_{10} adjusted for maximum indication. This coil is broadly resonant in the 5-MHz region and is tuned for an output reading of not over 2 volts r.m.s.

With the bandswitch in the 40-meter position, the cathode of the triode section of the 6EA8 premixer is connected to the cathode of the pentode section, energizing the crystal-oscillator stage and changing the circuit to a cathode-coupled mixer. The slug in the crystal-oscillator coil (L_{11}) is adjusted for maximum r-f voltage at the grid of the triode section of V_8. The premixer coils (L_8 and L_9) are tuned for maximum r-f voltage at the arm of switch S_5A. The voltage measured at this point is the 16-MHz product of the crystal and vfo frequencies.

Receiver I-F and 80-meter Alignment—The receiver i-f amplifier is aligned by disabling the vfo and injecting a 9-MHz signal at the input grid (pin #7) of the 6BE6 receiver mixer (V_{10}). The i-f coils (L_1, L_5, L_7) and the primary *only* of transformer T_1 are tuned for maximum signal response using avc voltage as indication of resonance. With the bandswitch in the 80-meter position and the vfo functioning, a 4.0-MHz signal is injected at the antenna receptacle and the primary of r-f transformer T_2 tuned for maximum signal. This transformer is stagger-tuned by peaking the secondary

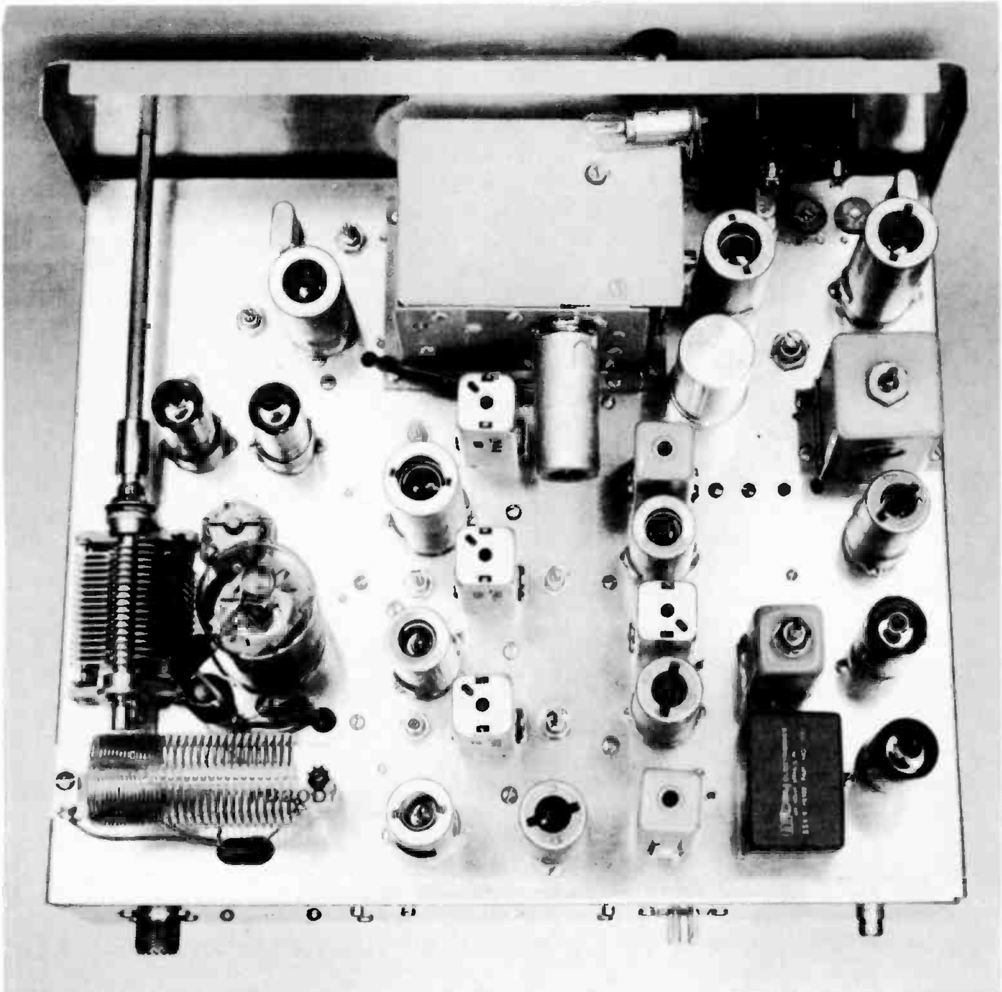


Figure 40

TOP VIEW OF TRIBAND TRANSCEIVER

Identification of various components may be done by comparison to chassis-layout drawing (figure 34). Variable-frequency oscillator is centered behind panel which is spaced away from chassis to allow clearance for circular dial. Pilot lamp is atop oscillator compartment, with oscillator padding capacitor (C_6) adjustable from top of compartment. Carrier crystals and their padding capacitors (C_2-C_3) are visible below panel meter at right. Across the rear chassis apron are (l. to r.): Antenna coaxial receptacle (J_1), power receptacle (P_1) and speaker jack (J_2).

at 3.8 MHz and checking at several points in between where a further slight adjustment of the slugs should result in a fairly flat response over the desired 200-kHz range. It will be noted that the final amplifier tank adjustment (which is the input circuit when receiving) must be peaked slightly when tuning from one end to the other of the 200-kHz range.

Receiver Alignment, 40 and 20 Meters—The tuning of the 40- and 20-meter band-pass r-f coils is done in a different manner. The grid coils (L_{23}, L_{25}) are temporarily unsoldered from the bandswitch (S_1B) to

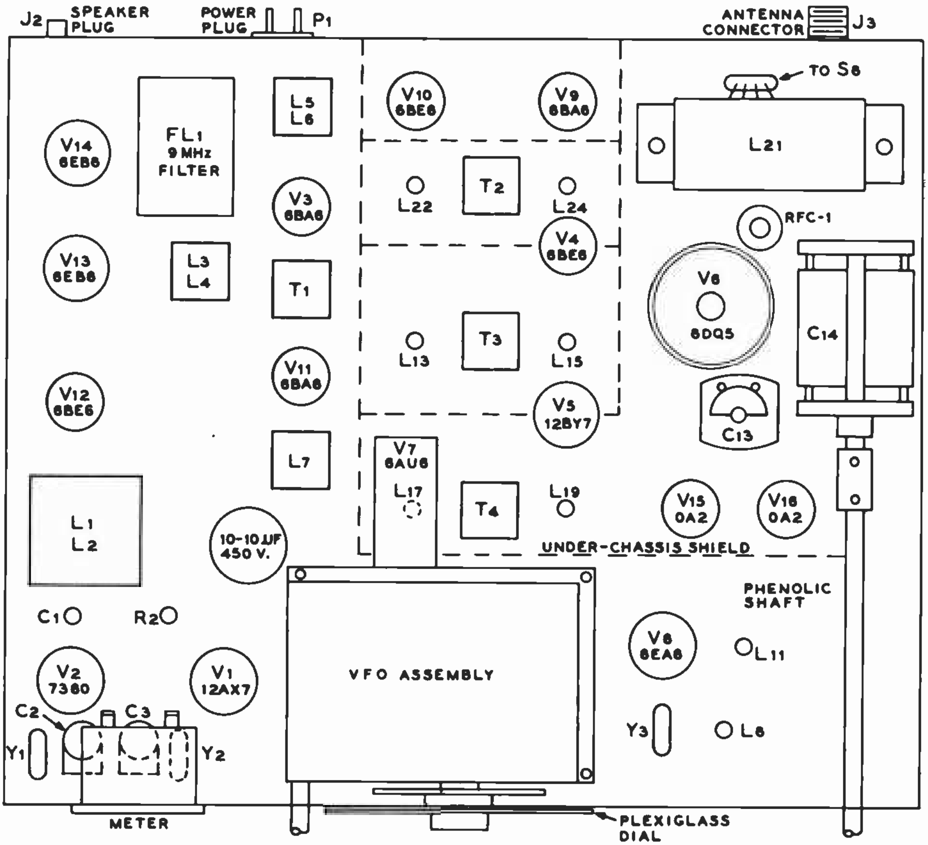


Figure 41

PLACEMENT OF MAJOR COMPONENTS ON TRANSCEIVER CHASSIS

remove them from the active circuit and a grid-dip oscillator is used to set the frequency of the primary circuits (L_{22}, L_{24}) by adjustment of the slugs. The 40-meter plate coil is adjusted to 7.3 MHz and the 20-meter plate coil to 14.35 MHz. The grid coils are then resoldered to the bandswitch terminals and the 6BA6 r-f amplifier tube (V_9) is removed from its socket. This raises the resonant frequency of the primary windings so they will not affect the adjustment of the grid circuitry. The grid coils are then dipped to 7.0 and 14.0 MHz. With the r-f tube back in its socket, the transceiver can be turned on and checked for receiver operation on each band.

Transmitter Alignment Alignment of the transmitter section is done with the high voltage disconnected and with screen voltage removed from the 6DQ5 amplifier. If the OA2 screen-regulator tube is wired so that the dropping resistor goes to pin #1 and the screen lead to pin #5, the screen voltage will be disconnected by removing the OA2 from its socket, since the OA2 has an internal jumper between these pins.

Much of the transmitter alignment is completed once the receiver section has been adjusted. The 7360 balanced-modulator plate coil (L_1) is tuned first, placing the r-f probe of the v.t.v.m. at the grid (pin

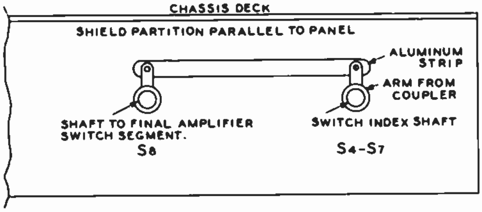


Figure 42

BANDSWITCH ARM DETAIL

#7) of the 6BE6 transmitting mixer (V_4) to obtain an r-f voltage reading. The transmitter circuitry is energized by pressing the push-to-talk switch on the microphone (with the microphone gain control R_1 turned down). The carrier control (R_3) is turned on and advanced to provide carrier injection until a reading is obtained on the v.t.v.m. The slug of coil L_1 is adjusted for maximum r-f indication. The phase-balance capacitor (C_1) should be set for equal capacitance and the voltage-balance potentiometer (R_2) set near the center of rotation. When the carrier control is turned off, the indicated r-f voltage will drop and balance potentiometer R_2 should be adjusted for a minimum r-f reading. This is the adjustment for carrier suppression and at this time the phase-balance capacitor should be adjusted slightly to achieve lowest possible r-f reading. Both controls affect carrier suppression and are slightly interlocking and should be adjusted in sequence for lowest reading on the v.t.v.m. The whole process may be monitored with a receiver used as an r-f probe with the antenna lead placed near the socket of the 6BE6 transmitter mixer tube (V_4).

Carrier Oscillator Adjustment—Capacitors C_1 and C_2 across the upper- and lower-sideband crystals are used to trim the crystal frequencies for proper positioning of the carrier on the slope of the sideband filter. To realize the rated sideband rejection of 40 decibels, the carrier oscillator should be placed 1500 Hz above or below the 9-MHz center frequency of the filter. Carrier suppression is also affected by proper positioning of the carrier frequency on the filter slope. When making the frequency adjustments, carrier suppression should be checked on both upper-and lower-sideband positions. The minimum voltage reading with carrier

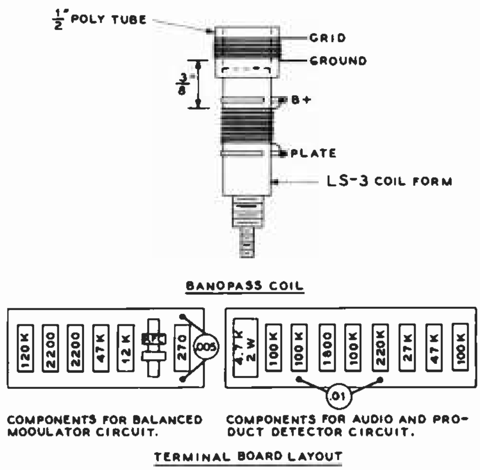


Figure 43

COIL TABLE FOR TRANSCEIVER

L_1—12 bifilar turns (24 in all) #24 enamel wire, closewound on slug-tuned form, ½" diam. (National XR-50). Tune to 9 MHz

L_2—4 turns #24 hookup wire around center of L_1

L_3—4 turns #24 hookup wire on "cold" end of L_4

L_4, L_5, L_7—30 turns #30 enamel closewound on 5/16" diameter form. Tune to 9 MHz

L_6—4 turns #24 hookup wire on "cold" end of L_5

L_8—12 turns #24 enamel closewound on ⅜" diam. slug-tuned form (CTC-LS3 or equiv.). Tune to 16 MHz

L_9—8 turns #24 enamel wire closewound on ⅜" length of ½" diam. polystyrene tubing slipped over top end of coil L_8 to make premixer transformer. Tune to 16 MHz

L_{10}—Ferrite rod loop-antenna coil ("loopstick") with turns removed to resonate to 5MHz (J. W. Miller 6300)

L_{11}—15 turns #24 enamel wire closewound on ⅜" diam. slug-tuned form (CTC-LS3). Tune to 21.5 MHz

L_{12}—7½ turns #20, ¾" diam., ¾" long (B & W 3011). Tunes 5.0 to 5.35 MHz

L_{13}, L_{17}, L_{22}—30 turns #30 enamel wire closewound on ⅜" diam. slug-tuned form (CTC-LS3). Tune to 7 MHz

L_{14}, L_{18}, L_{23}—25 turns #30 enamel wire closewound on ⅜" length of ½" diam. polystyrene tubing cemented to top of L_{13}, L_{17}, L_{22} to make bandpass transformer (see sketch). Tune to 7 MHz

L_{15}, L_{19}, L_{24}—14 turns #28 enamel wire closewound on ⅜" diam. slug-tuned form (CTC-LS3). Tune to 14 MHz

L_{16}, L_{20}, L_{25}—12 turns #28 enamel closewound on ⅜" length of ½" diam. polystyrene tubing cemented to top of L_{15}, L_{19}, L_{24} to make bandpass transformer. Tune to 14 MHz

L_{21}—Final amplifier tank coil. 32 turns #16 wire, with 16 turns spaced twice wire diameter; 16 turns spaced wire diameter Coil is 1" diam., 2½" long, tapped at 10 and 18 turns from plate end. (Air-Dux 820-D10).

Note: L_4, L_5 and L_7 are mounted in ¾" square shield cans similar to transformer T_1.

turned off should be very nearly the same with either crystal. Final adjustment may be made with voice modulation, striving for good audio quality on either sideband as monitored in a nearby receiver.

Bandpass Adjustment—The bandpass circuits in the linear amplifier stages of the transmitter are aligned in the same manner as the receiver circuits using carrier injection from either sideband crystal. The 40- and 20-meter coils are checked with a grid-dip oscillator as before, but the 80-meter transformers (T_3, T_4) as well as the secondary of T_1, are adjusted with voltage applied to the transmitter and the transformer slugs tuned for uniform 6DQ5 drive-voltage reading over the 200-kHz tuning range with the r-f probe placed at the grid of the 6DQ5. A maximum of 15 to 20 volts rms can be obtained with full carrier injection. Under final operating conditions, the 40- and 20-meter coils may require some slight adjustment for uniform drive across these bands.

Amplifier Neutralization—The last step

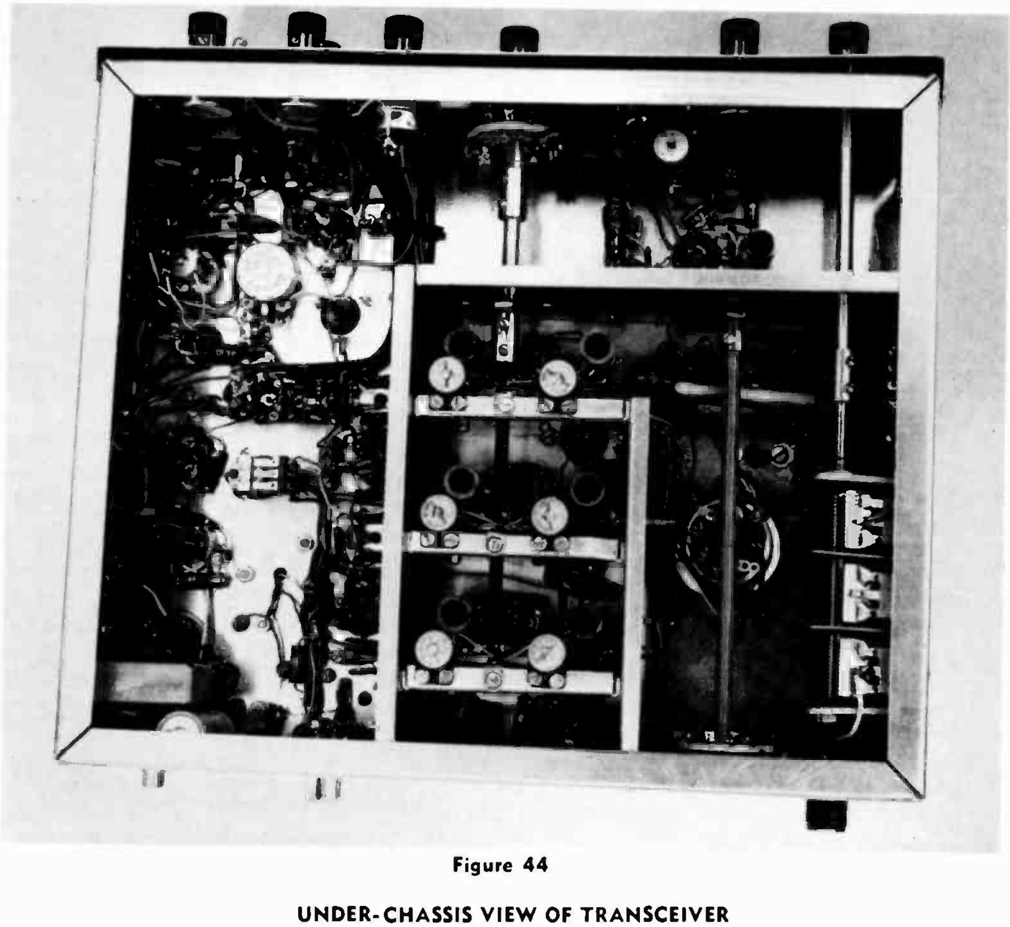


Figure 44

UNDER-CHASSIS VIEW OF TRANSCEIVER

Placement of shield partitions about tuned circuits may be seen at right side of chassis. Trimming capacitors for 40- and 20-meter circuits are mounted on partitions as are switch decks. First switch segment behind front panel is S_5. The three-gang antenna loading capacitor is bolted to the side apron of the chassis near antenna receptacle and tank switch wafer S_6. The opposite side apron is used to mount the audio output transformer (T_5) and two terminal boards that support most of the resistors and capacitors in the audio and balanced modulator circuits. Terminal strips and tie points are used to mount small components securely to resist vibration encountered in mobile work. The antenna relay (RY_2) is mounted on the rear apron above the 6BE6 (V_{10}) socket. The voltage changeover relay RY_1 is mounted in the center of the chassis area between the i-f amplifier tubes and the audio tubes.

is to neutralize the final amplifier stage. With plate and screen voltage removed and grid drive applied to the 6DQ5, neutralization is accomplished by placing the r-f probe at the antenna receptacle and adjusting neutralizing capacitor C_{13} for minimum r-f indication when the 6DQ5 tank circuit is tuned to resonance.

Final Amplifier Adjustment—Amplifier bias is adjusted to provide 50 ma of resting current. The transceiver should be coupled to a dummy load and loading and grid drive (carrier insertion) adjusted to provide the desired input level.

Antenna loading requires that a fixed ratio of grid drive to plate-load impedance be achieved. Maximum drive level is fixed and loading is accomplished at this level and may be increased until flat-topping is first observed on a monitor oscilloscope. Advantage is taken of the high peak-to-average-power ratio in the human voice, and up to 200 watts peak input may be run to the 6DQ5 without overheating the tube. Carrier injection and tune-up conditions, on the other hand, impose maximum dissipation conditions on the tube and tune-up operation at full input should be limited to periods of 20 seconds or less in one minute as tube dissipation runs near 65 watts or so under these conditions. With the average voice, peak plate-current indication on the meter will run below 50 percent of the full carrier injection plate current, even taking into account the alc action of this circuit. Thus, under intermittent carrier tune-up at 800 volts plate potential, maximum plate current may run as high as 275 to 300 milliamperes, with indicated voice peaks running about 125 to 175 milliamperes meter reading. Excessive peak plate current readings under voice conditions indicate flat-topping and consequent distortion of the signal.

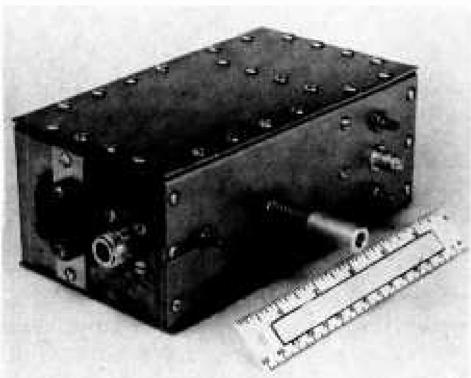


Figure 45

432-MHz TRIPLER/AMPLIFIER USING 4X150A OR 4CX250B

This compact unit functions either as a tripler to 432 MHz, or as an amplifier on that band. It uses an external-anode tetrode in a modified cavity plate circuit. Inclosure is made up of side pieces held together with sheet-metal screws or "pop" rivets. In this oblique view, the B-plus connector is at the left side of the unit, with the coaxial antenna receptacle immediately adjacent to it. The antenna tuning capacitor is mounted to the end piece of the box, which may be removed by loosening the holding screws and the capacitor nut. At the center of the box is the spring-loaded tuning capacitor and at the right end is the coaxial input receptacle and the input tuning capacitors.

Transceiver Antenna The triband transceiver output circuit is designed for a nominal 50-ohm load. Low-frequency whip antennas, particularly 80-meter loaded whips present a low-impedance load which may inhibit proper transceiver loading under certain conditions. If this situation exists, there are several solutions to the problem. The easiest one to apply is to change the length of the coaxial line running from the transceiver to the antenna. By lengthening the line in five- or ten-foot increments, a condition of proper load may be achieved, even though the SWR on the transmission line remains quite high. A better solution is to make use of an auxiliary impedance-matching coil placed at the base of the antenna, as shown in figure 16A. Typically, a matching coil for 80 meters may consist of about fifteen turns of No. 12 wire, 1 inch in diameter and about 2 inches long. The number of active turns in the coil are adjusted, one by one, and the SWR on the transmission line monitored. A proper impedance match will drop the SWR to a value less than 1.2/1 at the resonant frequency of the loaded antenna.

21-5 A Tripler/Amplifier For 432 MHz

An efficient tripler or amplifier for 432-MHz operation may be designed around the

CHART 1

TUBE SOCKET VOLTAGE CHART									
TUBE	1	2	3	4	5	6	7	8	9
V_1	100	−2	0	F	F	180	−0.5	1	—
V_2	3	150	—	F	F	220	220	18	18
V_3	—	0	F	F	220	100	1.8	—	—
V_4	—	2	F	F	240	75	0	—	—
V_5	2	—	0	F	F	F	240	110	0
V_6	−50	F	—	150	−50	—	F	150	—
V_8	100	—	125	F	F	200	3	3	—
V_9	—	0	F	F	220	100	1.8	—	—
V_{10}	—	2.5	F	F	225	80	0	—	—
V_{11}	—	1.8	F	F	220	100	1.8	—	—
V_{12}	—	0.8	F	F	200	35	0.8	—	—
V_{13}	1	—	100	F	F	3	—	180	240
V_{14}	70	—	175	F	F	3	—	180	240

NOTES:

Readings taken with 20,000 ohms-per-volt meter and may vary 10%.
Voltages—0 on pins 6, 7, 8, 9, of V_2 on receive.
Voltage—120 on pin 2 of V_2 on receive.
R-f gain and audio gain fully advanced.

POWER-SUPPLY REQUIREMENTS

Low voltage—250 volts at 115 ma receive
80 ma xmit
Bias— —50 volts d.c. 5 ma

High voltage—600 to 800 volts at 300 ma, xmit only
Filaments—12.6 volts a.c. or d.c. at 4 A
Relay—12 V.D.C. 80 ma, xmit only

4X150A or 4CX250B external-anode tetrode. Rated at 250 watts anode dissipation (the late production 4X150As also have the higher rating) this high-perveance tetrode is one of the few tubes that performs well as a tripler from 144 MHz or as a straight amplifier at 432 MHz. A power output of better than 60 watts may be obtained as a tripler, and over 200 watts output may be achieved in amplifier service.

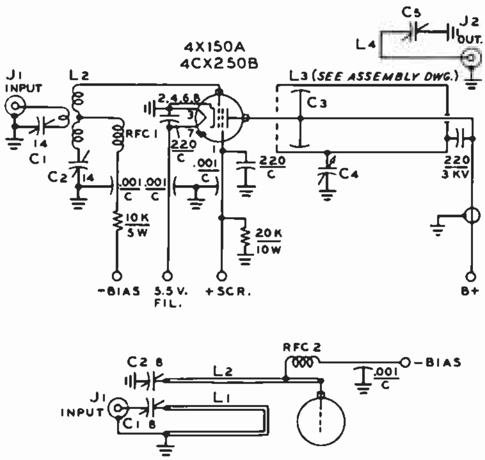
Two units such as described may be built; one acting as a tripler to drive the second one as an amplifier at a power input up to 500 watts.

The Tripler/ Amplifier Circuit The general schematic of the amplifier is shown in figure 46. An easily built coaxial plate-tank circuit provides high efficiency at 432 MHz and the unit operates in the same manner as if it were on the lower-frequency bands. The circuit consists of a short, loaded resonant cylindrical line which uses the amplifier case as the outer conductor. Plate voltage is fed through the line to the anode of the tube, which is insulated from the cylindrical line by means of a thin *teflon* sheet wrapped about the anode.

For tripler service, the grid circuit is tuned to 144 MHz, with the input capacitance of the tube and tuning capacitor C_2 forming a balanced tank circuit. The isolation choke (RFC_1) is at the center, or "cold" point of the grid inductor. A series-tuned link circuit couples the unit to the external exciter.

In amplifier service, the grid circuit is tuned to 432 MHz and takes the form of a half-wavelength line, tuned to resonance by a small capacitor placed at the end of the line opposite the tube.

A special air-system socket designed for the external-anode tetrode must be used. For



ALTERNATIVE GRID CIRCUIT FOR 432 MHZ AMPLIFIER

Figure 46

SCHEMATIC—432-MHz TRIPLER/AMPLIFIER

C_1, C_2, C_5—14 pf, Johnson 160-107
C_3, C_4—See text
Note: Use 8 pf, Johnson 160-104 for 432-MHz alternative grid circuit
L_1—1 turn hookup wire, ¾" diam, inside L_2 (432 MHz)
L_2—3½ turns #14, 1" diameter, ¾" long (432 MHz)
L_3—See text
L_4—⅜"-wide copper strap form inductor ⅜" long × ⅛" high
RFC_1—1.7 μH. J. W. Miller RFC-144, or Ohmite Z-144
RFC_2—0.2 μH. J. W. Miller RFC-420, or Ohmite Z-420
J_1—Coaxial receptable. UG-290A/U
J_2—Coaxial receptacle. UG-58A/U
Blower—6 cfm at 0.4" back pressure. Use #2½ impeller at 6000 r.p.m.

tripler service, the builder has the choice of either the *EIMAC SK-600, SK-610, SK-620,* or *SK-630* socket, together with the appropriate air chimney. The *SK-606* chimney is to be used with the *SK-600* or *SK-610* socket, and the *SK-626* chimney is to be used with the *SK-620* or *SK-630* socket.

For amplifier service at 432 MHz only, the *EIMAC SK-620* or *SK-630* sockets are recommended, as the other versions have screen terminals exposed to the plate-circuit field and exhibit more r-f feedthrough than do the suggested sockets, which have shielded screen terminals. Using the proper sockets, intrastage feedthrough at 432 MHz is sufficiently low so that stage neutralization is not required under normal, loaded operating conditions.

Not shown in the schematic is the fact that an external centrifugal blower is required to adequately cool the filament and plate seals of the external anode tube. Approximately 6.4 cubic feet per minute of cooling air at a pressure drop of 0.82 inches of water is required for full, 250-watt anode dissipation. For operation at reduced voltages and a limitation of 150 watts dissipation, the cooling-air requirement is 3.4 c.f.m. at a pressure drop of 0.15 inches of water.

At a frequency of 432 MHz, cathode backheating is observed in tubes of this type, and to maintain proper cathode temperature, the filament voltage should be reduced to 5.5 volts and held within plus or minus five percent of this value.

Finally, it should be noted that under certain operating conditions, the screen current of a tetrode may become negative. In order to protect the tube from excessively high screen voltage under certain negative-current conditions, it is mandatory to connect a bleeder resistor at the tube that will draw a value of current greater than drawn by the tube under negative-current operating conditions.

Tripler/Amplifier Construction The tripler/amplifier is constructed within a metal box measuring 7" × 2¾" × 2⅝". The top and bottom of the box are flat pieces of aluminum or brass measuring 7" × 2¾". The two side pieces are identical in size with matching holes for sheet-metal screws. Each side has small flanges along the edge which match the sides to the top and bottom pieces. The end section of the box which makes up the plate-circuit assembly is made of brass so that the brass quarter-wavelength plate line may be soldered to it. The opposite end of the box has a hole drilled off center in it to accept a fitting for an air hose or blower orifice (figure 47).

The plate line is made of a 3¾" length of brass tube having a 1¾" outside diameter. The line is soldered to the brass end to accept the anode of the 4X150A or 4CX250B.

An internal partition separates the grid and plate circuits and supports the socket for the tetrode. The socket is bolted atop the partition, as shown in figure 47. Connection is made to the anode for the supply

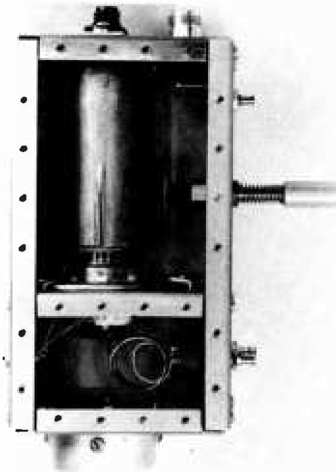


Figure 47

INSIDE VIEW OF TRIPLER/AMPLIFIER

Tetrode tube socket mounts on small partition placed across interior of box. Plate-tuning capacitor and antenna capacitor are at right of assembly. The anode line of the 4X150A (4CX250B) is slotted and slips over the tube, insulated from it by a teflon wraparound insulator. B-plus passes down through the tube to a spring that makes connection to the anode Below the partition are the grid circuit and various bypass capacitors. Power leads pass through feedthrough capacitors mounted in the rear wall of the inclosure (left). Aluminum fitting at the bottom of the box matches air-hose connection to external centrifugal blower. Blower should be turned on when filament voltage is applied to the tube.

voltage by means of an extension shaft run from the high-voltage connector mounted on the top plate of the box. The shaft has a section of spring steel bolted at the end to make a press fit to the top of the anode of the tube.

The plate-blocking capacitor is made of a length of 3-mil *teflon* tape, wrapped twice around the tube anode. The tape is cut to a width of one inch to allow overlap on both sides of the anode. The tape is carefully wrapped around the metal anode before the tube is pressed into the open end of the plate line, as shown in figure 49.

The top plate of the box, in addition to the plate-line and high-voltage connector, supports the antenna receptacle (J_2) and the series antenna-tuning capacitor. The antenna pickup loop (L_4) is soldered between the receptacle and the stator of the capacitor, and is spaced away from the plate line about ⅛ inch.

Plate-tuning capacitor is a 1¾" disc made of brass material soldered to the smooth end of a shaft that is threaded to match a panel bushing. The outer portion of the shaft is ¼-inch diameter to fit the dial drive. Tension is maintained on the shaft and bearing by placing a spring between the shaft extension and the panel bushing, as shown in the side view photograph.

Tripler/Amplifier Operation After the unit has been assembled, it should be tested for operation at reduced voltages. The first step is to grid-dip the input and output circuits to resonance to make sure they tune properly. An r-f output meter or SWR bridge should be used in conjunction with a dummy load for the initial tests. A good dummy load for 432 MHz is 500 feet of RG-58/U coaxial cable. The far end should be shorted and waterproofed and the cable may be coiled up in a tub of water.

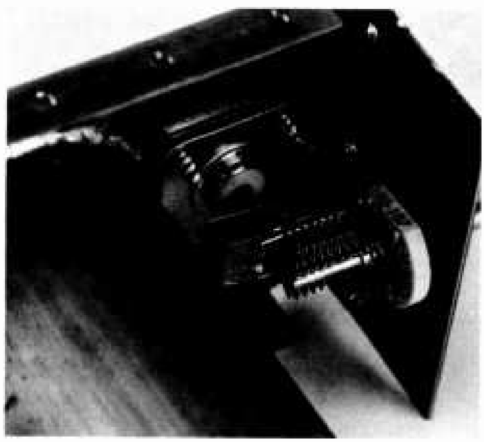


Figure 48

CLOSEUP OF ANTENNA CIRCUITRY

Small series-tuned loop is made of copper strap connected between coaxial output receptacle and stator rods of antenna tuning capacitor. Coupling is adjusted by setting of capacitor, and link is fixed about ⅛-inch away from plate line. Plate line is soldered to brass end plate.

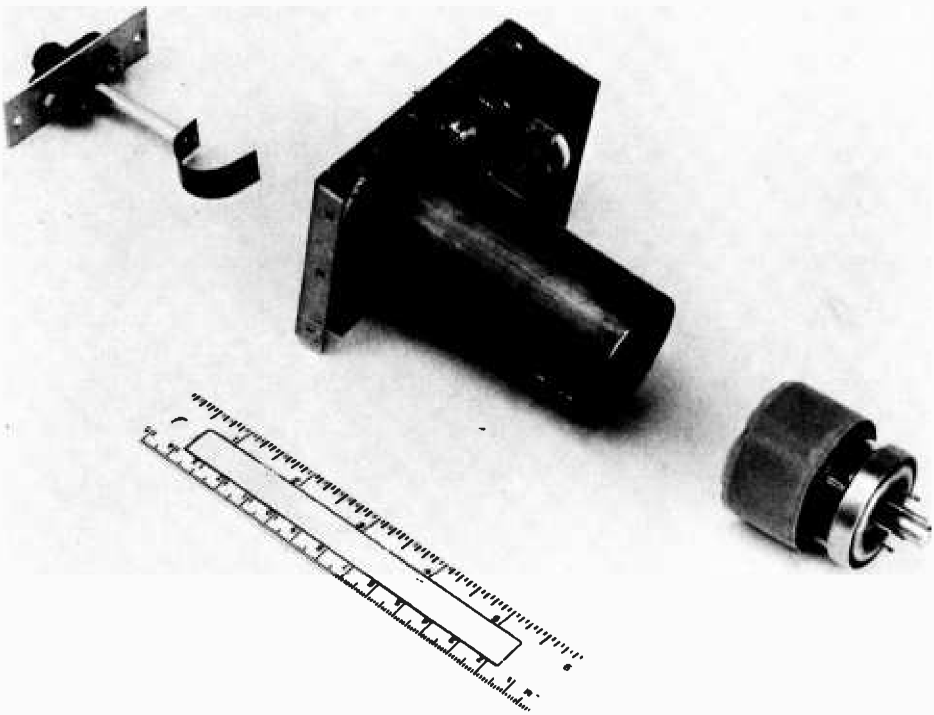


Figure 49

EXPLODED VIEW OF PLATE-LINE ASSEMBLY

The high-voltage receptacle, plate bypass capacitor, and anode connector spring are at left. The brass end plate of the box and plate-circuit assembly are at the center, with the 4X150A tetrode at the right. The tube anode is wrapped with teflon tape to form a bypass capacitor, removing the d-c voltage from the coaxial line. Copper line makes press fit over the anode of the tube.

As with any tetrode, plate current is a function of screen voltage, and screen current is a function of plate loading. Screen voltage, therefore, should never be applied before plate voltage, and screen current should be monitored for proper plate loading. The amplifier should never be tested or operated without a proper dummy load.

To operate as a tripler, the following electrode voltages are suggested: plate voltage, 1000; screen voltage, 250; grid bias, −90 volts. The bias may be obtained from a small voltage-regulator tube or zener diode. Cooling air is applied with filament voltage which should be 5.5 volts. When these voltages are applied to the tube, plate current will be near-zero with no drive, and the screen current will be about 10 milliamperes, or less. The screen current noted will be the sum of the positive current flowing through the bleeder resistor and the negative screen current of the tube.

A small amount of excitation at 144 MHz is applied and the grid circuit resonated, as noted by a small rise in plate current. The plate circuit should be brought into resonance. Excitation is boosted, and the tripler tuned for maximum power into the dummy load. Loading and grid drive may be increased until a plate current of 250 ma is achieved. At this level, total screen current will be about 15 to 20 ma, and grid current will be about 12 ma. Power input is about 250 watts and power output, as measured at the antenna receptacle with a vhf wattmeter is about 70 watts. Over-all tripler efficiency is about 28 percent and plate dissipation is nearly 180 watts.

Screen current is a sensitive indicator of circuit loading. If the screen current falls below 10 to 12 ma (including bleeder current), it is an indication that plate loading is too heavy or grid drive too light. Screen current readings of over 30 ma indicate drive is too heavy or plate loading is too light. A plate voltage as low as 800 volts may be used on the tripler stage, with an output of about 55 watts at a plate current of 250 ma. Plate voltages below this value are not recommended as screen current starts to climb rapidly at low plate potentials. For amplifier service, the alternate grid circuit is employed. The amplifier may be operated either class C or class AB_1. Operating data for both classes of service are given in the 4X150A data sheet.

CHAPTER TWENTY-TWO

H-F and VHF Power Amplifiers

A *power amplifier* is a converter that changes d-c into r-f output. Chapter Seven of this Handbook discussed the various classes of r-f power amplifiers and Chapters Eleven and Twelve covered the calculation of input and output circuit parameters. This chapter covers power-amplifier design, construction, and adjustment.

Modern h-f amateur transmitters are capable of operating on c-w, SSB and often RTTY, on one or more amateur bands between 3.5 MHz and 29.7 MHz. Very few pieces of commercially built amateur equipment have amplitude-modulation capability, other than some gear designed for 6- and 2-meter operation, since the changeover from a.m. to SSB during the past decade is now almost complete. On the other hand, expansion of 160-meter privileges in the past years has not brought about the inclusion of that band in most amateur equipment.

The most popular and flexible amateur h-f transmitting arrangement usually includes a compact bandswitching exciter or transceiver having 100 to 500 watts PEP input on the most commonly used h-f bands, followed by a single linear power-amplifier stage having 1 kW to 2 kW PEP input capacity. In many instances, the exciter is an SSB transceiver unit capable of mobile operation, while the amplifier may be a compact table-top assembly. The amplifier is usually coupled to the exciter by a coaxial cable and changeover relay combination, permitting the exciter to run independently of the amplifier, if desired, or in combination with the amplifier for maximum power output. For c.w. or RTTY, the amplifier is usually operated in the linear mode, since conversion to class-C operation is not required.

These practical designs are a natural outgrowth of the importance of vfo operation and the use of SSB and c-w modes in amateur practice. It is not practical to make a rapid frequency change when a whole succession of stages must be retuned to resonance, or when bandswitching is not employed. Another significant feature in station design is the popularity of 100- to 250-watt output SSB exciter/transceivers. These provide sufficient drive for high-level linear amplifiers without the need for intermediate stages of amplification.

Power-Amplifier Design Power amplifiers are classified according to operating *mode* and *circuitry*. Thus, a particular amplifier mode may be class AB_1, class B, or class C; the circuitry can be either single-ended or push-pull; and the unit may be grid- or cathode-driven. Mode of operation and circuit configuration should not be confused, since they may be mixed in various combinations, according to the desire of the user and the characteristics of the amplifier tube.

High-frequency silicon power transistors are used in some advanced amateur and commercial equipment designs up to the 100-watt PEP power level or so. Undoubtedly solid-state devices will become of increasing importance in h-f power amplifiers in the coming decade.

Either triode or tetrode tubes may be used in the proper circuitry in h-f and vhf power amplifiers. The choice of tube type is often dependent on the amount of drive power available and, in the case of home-made gear, the tube at hand. If an exciter of 100 to 200 watts PEP output capacity is to be used, it is prudent to employ an amplifier whose drive requirement falls in the same power range as the exciter output. Triode or tetrode tubes may be used in cathode-driven (grounded-grid) circuitry which will pass along an excess of exciter power in the form of feedthrough power to the antenna circuit. The tubes may also be grid-driven in combination with a power absorption network that will dissipate excess exciter power not required by the amplifier.

On the other hand, if the power output of the exciter is only a few watts PEP, either low-drive, high-gain tetrodes must be used in grid-driven configuration, or an intermediate amplifier must be used to boost the drive to that level required by triode tubes. Thus, the interface between the exciter and the amplifier in terms of PEP level must be reconciled in the design of the station transmitting equipment.

22-1 Triode Amplifier Design

Triode tubes may be operated in either grid- or cathode-driven configuration, and may be run in class-AB_1, class-AB_2, class-B or class-C mode. Plate dissipation and amplification factor (μ) are two triode characteristics which provide the information necessary to establish proper mode and circuitry and to evaluate the tube for linear-amplifier or class-C service.

Plate dissipation is important in that it determines the ultimate average and peak power capabilities of the tube. Linear amplifiers commonly run between 55- and 65-percent plate efficiency, with the majority of the remainder of the power being lost as plate dissipation. Class-C service often runs at about 70- to 75-percent plate efficiency. Knowing the plate dissipation rating of the tube, the approximate maximum power input and output levels for various modes of service may be determined by the methods outlined in Chapter 7.

Amplification Factor (μ) of a triode expresses the ratio of change of plate voltage for a given change in grid voltage at some fixed value of plate current. Values of μ between 10 and 300 are common for triode transmitting tubes. High-μ tubes (μ greater than about 30) are most suitable in cathode-driven (grounded-grid) circuitry as the cathode-plate shielding of a high-μ tube is superior to that of a comparable low-μ tube, and because a high-μ tube provides more gain and requires less driving power than a low-μ tube in this class of service. Low-μ triodes, on the other hand, are well suited for grid-driven class-AB_1 operation since it is possible to reach a high value of plate current with this type of tube, as opposed to the high-μ equivalent, without driving the grid into the power-consuming, positive region. Even though a large value of driving voltage is required for the low-μ tube, little drive power is required for class-AB_1 service, since the grid always remains negative and never draws current.

As a rule-of-thumb, then, a triode tube to be used for linear r-f service in a power amplifier should have a large plate-dissipation capability, and the output power to be expected from a single tube will run about twice the plate-dissipation rating. High-μ triodes, generally speaking, perform better in cathode-driven, class-B circuitry; whereas medium- and low-μ triodes are to be preferred in grid-driven, class AB_1 circuitry. Circuit neutralization may often be disposed with in the first case (at least in the h-f region), and is always necessary in the second case, otherwise the circuits bear a striking similarity.

Grid-Driven Circuitry Representative grid- and cathode-driven triode circuits are shown in figure 1. The classic grid-driven, grid-neutralized circuit is shown in illustration A. The drive signal is applied to a balanced grid tank circuit (L_1, C_1) with an out-of-phase portion of the exciting voltage fed through capacitor NC to the plate circuit in a bridge neutralization scheme.

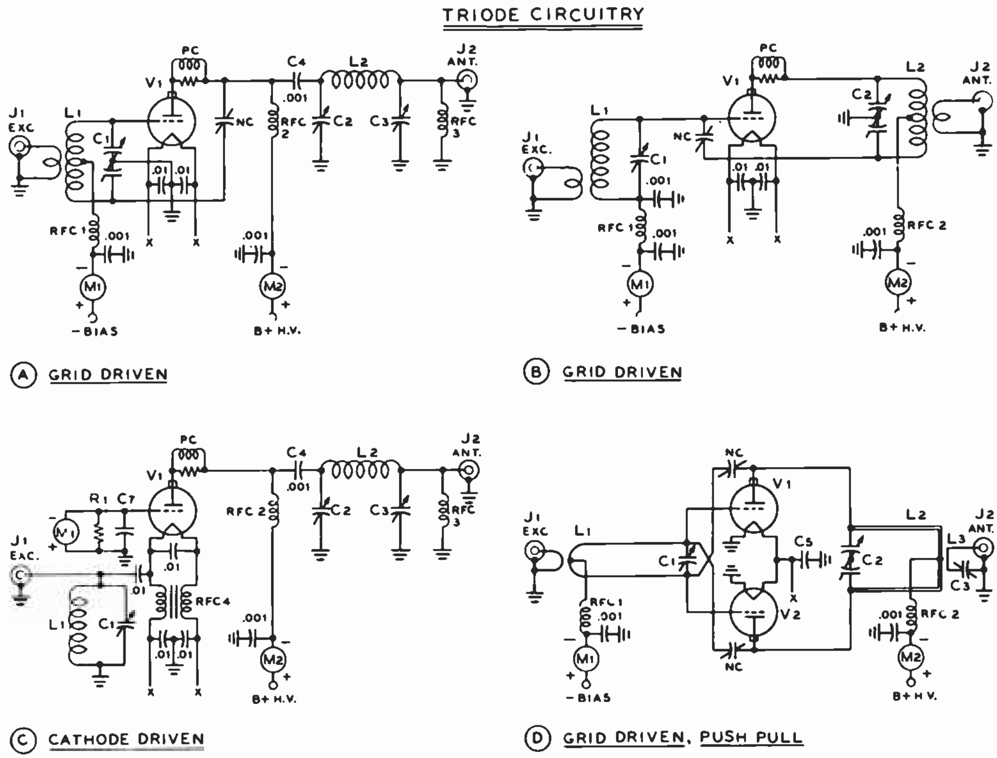


Figure 1

REPRESENTATIVE TRIODE AMPLIFIER CIRCUITS

Circuits A, B and C are for the 3–54 MHz region. Circuit D is intended for the 50–500 MHz region. Note that one filament leg is grounded in circuit D to reduce inductance of filament return circuit.

C_1—Input tuning capacitor. Typically, 3 pf per meter of wavelength. Spacing 0.03″ for power level up to 2 kW, PEP

C_2—Output tuning capacitor. Refer to plate-circuit design data in Chapter 11.

C_3—Loading capacitor. Typically, 20 pf per meter of wavelength. Refer to Chapter 11.

C_4—Plate-blocking capacitor. Typically, 500 pf to 1000 pf, 5 kV

C_5, C_7—Low-inductance mica or ceramic capacitor, series resonant near operating frequency. See Chapter 17

M_1—Grid-current meter

M_2—Plate-current meter

RFC_1—Grid choke, receiving type rated to carry grid current. Typically, 1 to 2.5 mH for 3 to 30-MHz range

RFC_2—Plate choke, transmitting type, solenoid. Rated to carry plate current. Typically, 800 μH. See Chapter 17

RFC_3—Receiving-type choke. 2½ mH for 3- to 30-MHz range.

RFC_4—Bifilar windings, 15 turns each #12 wire on ½-inch diameter ferrite core, 3″ long for 3- to 54-MHz range

PC—Plate parasitic suppressor. Typically, 3 turns #18 enamel, ½-inch diameter, ½-inch long, in parallel with 50-ohm 2-watt composition resistor. See Chapter 17

A pi network is employed for the plate output coupling circuit. The plate inductor (L_2) may be tapped or otherwise variable and is normally adjustable from the amplifier panel, eliminating the necessity of plug-in coils and access openings into the shielded amplifier inclosure. The grid circuit may also be switched or varied in a similar manner.

Neutralization may be accomplished in the plate circuit, as shown in figure 1B. A split plate-tank circuit is required in place of the split grid circuit, making the use of a single-ended pi-network output circuit impractical. Theory and adjustment of grid and plate neutralizing circuits are covered in Chapter 11. In either configuration, care must be taken in construction to make sure that a minimum of stray coupling exists between grid and plate tank circuits. Whenever possible, the grid and plate coils should be mounted at right angles to each other, and should be separated sufficiently to reduce coupling between them to a minimum. Unwanted coupling will tend to make neutralization frequency-sensitive, requiring that the circuit be reneutralized when a major frequency change is made.

Cathode-Driven Circuitry A representative cathode-driven (grounded-grid) triode circuit is shown in figure 1C. A pi-network plate circuit is used, and excitation is applied to the filament (cathode) circuit, the grid being at r-f ground potential. If the amplification factor of the triode is sufficiently high so as to limit the static plate current to a reasonable value, no auxiliary grid bias is required. A parallel-tuned cathode input circuit is shown, although pi-network circuitry may be used in this position. Filament voltage may either be fed via a shunt r-f choke as shown, or applied through a bifilar series-fed cathode tank coil.

While nominally at r-f ground, the grid of the triode may be lifted above ground a sufficient amount so as to insert a monitoring circuit to measure d-c grid current. The grid to ground r-f impedance should remain very low, and proper attention must be paid to the r-f circuit. A considerable amount of r-f current flows through the grid bypass capacitor (C_7) and this component should be rated for r-f service. It should be shunted with a low value of resistance (of the order of 10 ohms or less) and the d-c voltage drop across this resistor is monitored by the grid voltmeter, which is calibrated in terms of grid current. Both resistor and capacitor aid in establishing a low-impedance path from grid to ground and should be mounted directly at the socket of the tube. If multiple grid pins are available, each pin should be individually bypassed to ground. Control of the grid-to-ground impedance in the cathode-driven circuit establishes the degree of intrastage feedback, and an increase in grid impedance may alter stage gain, leading to possible uncontrolled oscillation or perhaps making the stage difficult to drive. At the higher frequencies, stage gain may be controlled by the proper choice of the grid-to-ground impedance.

Push-Pull Circuitry A push-pull triode amplifier configuration is shown in figure 1D. This circuit design is now rarely used in the h-f region because of the mechanical difficulties that ensue when a large frequency change is desired. In the vhf region, on the other hand, where operation of an amplifier is generally restricted to one band of frequencies, linear push-pull tank circuits are often employed. Lumped-inductance tank coils are usually avoided in the vhf region since various forms of parallel-line or strip-line circuitry provide better efficiency, higher Q and better thermal stability than the coil-and-capacitor combination tank assemblies used at the high frequencies. Push-pull operation is of benefit in the vhf region as unavoidable tube capacitances are halved, and circuit impedances are generally higher than in the case of single-ended circuitry. At the higher vhf regions, parallel- and strip-line circuitry give way to coaxial tank circuits in which the tube structure becomes a part of the resonant circuit.

The output coupling circuit may be designed for either balanced or unbalanced connection to coaxial or twin-conductor transmission line. In many cases, a series capacitor (C_3) is placed in one leg of the line at the feed point to compensate for the inductance of the coupling coil.

Common h-f construction technique employs plug-in plate and grid coils which necessitate an opening in the amplifier inclosure for coil-changing purposes. Care must be taken in the construction of the door of the opening to reduce harmonic leakage to a minimum. While variations in layout, construction, and voltage application are found, the following general remarks apply to h-f amplifiers of all classes and types.

Circuit Layout The most important consideration in constructing a push-pull amplifier is to maintain electrical symmetry on both sides of the balanced circuit. Of utmost importance in maintaining electrical balance is the control of stray capacitance between each side of the circuit and ground.

Large masses of metal placed near one side of the grid or plate circuits can cause serious unbalance, especially at the higher frequencies, where the tank capacitance between one side of the tuned circuit and ground is often quite small in itself. Capacitive unbalance most often occurs when a plate or grid coil is located with one of its ends close to a metal panel. The solution to this difficulty is to mount the coil parallel to the panel to make the capacitance to ground equal from each end of the coil, or to place a grounded piece of metal opposite the "free" end of the coil to accomplish a capacity balance.

All r-f leads should be made as short and direct as possible. The leads from the tube grids or plates should be connected directly to their respective tank capacitors, and the leads between the tank capacitors and coils should be as heavy as the wire that is used in the coils themselves. Plate and grid leads to the tubes may be made of flexible tinned braid or flat copper strip. Neutralizing leads should run directly to the tube grids and plates and should be separate from the grid and plate leads to the tank circuits. Having a portion of the plate or grid connections to their tank circuits serve as part of a neutralizing lead can often result in amplifier instability at certain operating frequencies.

Filament Supply The amplifier filament transformer should be placed right on the amplifier chassis in close proximity to the tubes. Short filament leads are necessary to prevent excessive voltage drop in the connecting leads, and also to prevent r-f pickup in the filament circuit. Long filament leads can often induce instability in an otherwise stable amplifier circuit, especially if the leads are exposed to the radiated field of the plate circuit of the amplifier stage. The filament voltage should be the correct value specified by the tube manufacturer when measured *at the tube sockets*. A filament transformer having a tapped primary often will be found useful in adjusting the filament voltage. When there is a choice of having the filament voltage slightly higher or slightly lower than normal, the lower voltage is preferable.

Filament bypass capacitors should be low internal inductance units of approximately .01 μfd. A separate capacitor should be used for each socket terminal. Lower values of capacitance should be avoided to prevent spurious resonances in the internal filament structure of the tube. Use heavy, shielded filament leads for low voltage drop and maximum circuit isolation.

Plate Feed The series plate-voltage feed shown in figure 1D is the most satisfactory method for push-pull stages. This method of feed puts high voltage on the plate tank inductor, but since the r-f voltage on the inductor is in itself sufficient reason for protecting the inductor from accidental bodily contact, no additional protective arrangements are made necessary by the use of series feed.

The insulation in the plate supply circuit should be adequate for the voltages encountered. In general, the insulation should be rated to withstand at least four times the maximum d-c plate voltage. For safety, the plate meter should be placed in the cathode return lead, since there is danger of voltage breakdown between a metal panel and the meter movement at plate voltages much higher than one thousand.

Parallel plate feed, such as shown in figures 1A and 1B, is commonly used for single-ended pi-network amplifier configurations. The plate r-f choke is a critical component in this circuit, and a discussion of choke design is covered in Chapter 17. The plate-blocking capacitor (C_4) should be rated to withstand the peak r-f plate current (usually about three to four times the d-c plate current) and the peak r-f voltage (up to twice the d-c plate voltage.)

In the case of the push-pull stage, the amplifier grid and plate circuits should be symmetrically balanced to ground. In some instances, a small differential capacitor is placed in the grid circuit to effect balance, and the grid current of each tube is monitored individually to ascertain correct balance. The rotor of the split-stator plate-tuning capacitor is usually ungrounded, permitting the plate tank circuit to establish its own r-f balance.

The various filament, grid, and plate bypass capacitors are often vhf coaxial types which have inherently low inductance well into the vhf region. These capacitors should be checked to make sure that their internal self-resonant frequency is well above the operating frequency of the amplifier.

In most cases, the push-pull amplifier may be cross-neutralized in the normal manner. At the higher frequencies (above 150 MHz or so) it is common practice to operate the triode tubes in cathode-driven configuration

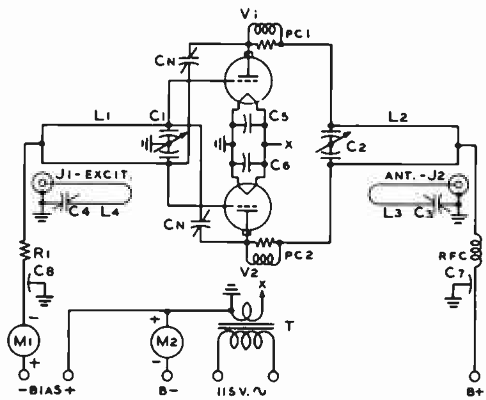


Figure 2

TYPICAL PUSH-PULL VHF TRIODE AMPLIFIER CIRCUIT

C_1, C_2—Low-capacitance, balanced split-stator capacitor. Typically, 10 pf per section for 144 MHz.
C_3, C_4—Loading capacitor. Capacitance chosen to series resonate at operating frequency with coupling loop
C_5, C_6—Low inductance mica or ceramic capacitor, series resonant near operating frequency. See Chapter 17
C_7, C_8—Low-inductance feedthrough capacitor. See Chapter 17
C_N—Neutralizing capacitor. Approximately equal to grid-plate capacitance of triode tube
M_1—Grid-current meter
M_2—Plate-current meter
R_1—Wire resistor (100–500 ohms) to act as low-Q r-f choke
RFC—Vhf choke rated to carry plate current. See text

which usually eliminates the need for neutralization if proper shielding is used.

Plate parasitic suppressors may or may not be necessary depending on the operating frequency of the amplifier and the natural parasitic frequency of the input and output circuits. Both grid- and plate-tuning capacitors should be located close to the tube elements and not tapped down the tuned lines, otherwise unwanted parasitic circuits may be created. If oscillations are encountered, they may possibly be suppressed by placing noninductive carbon resistors across a portion of the plate (and grid) lines as shown in figure 2.

The plate choke (RFC) should be mounted at right angles to the plate line and care should be taken that it is not coupled to the line. In particular, the choke should not be mounted within the line, but rather outside the end of the line, as shown. A resistor (R_1) is used to take the place of a grid choke, thus eliminating any possibility of resonance between the two chokes, with resulting circuit instability.

In order to prevent radiation loss from the grid and plate lines, it is common practice to completely inclose the input and output circuits in "r-f tight" inclosures, suitably ventilated to allow proper cooling of the tubes.

22-2 Tetrode Amplifier Design

As in the case of triode tubes, tetrodes may be operated in either grid- or cathode-driven configuration and may (within certain limits) be run in class-AB_1, -AB_2, -B, or class-C mode. Much of the information on circuit layout and operation previously discussed for triode tubes applies in equal context to tetrodes. Other differences and additional operational data will be discussed in this section.

Tetrode tubes are widely used in h-f and vhf amplifiers because of their high power gain and wide range of simple neutralization. Tetrode circuitry resembles triode circuitry in that comparable modes and circuit configurations may be used. Various popular and proven tetrode circuits are shown in figure 3. Illustration A shows a typical single-ended neutralized tetrode circuit employing a pi-network output circuit and a bridge neutralization scheme. Tetrode neutralization techniques are discussed in detail in Chapter 11.

Tetrode plate current is a direct function of screen voltage and means must be employed to control screen voltage under

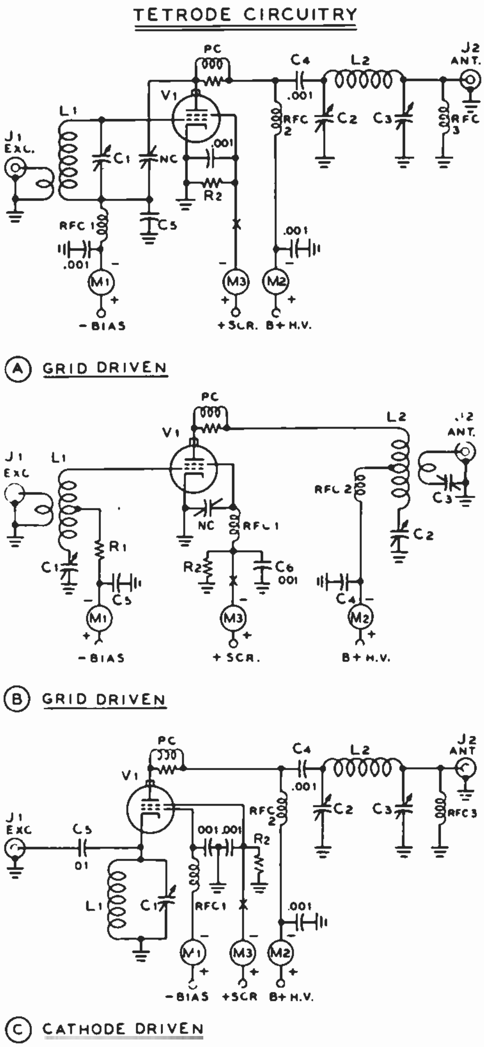


Figure 3

REPRESENTATIVE TETRODE AMPLIFIER CIRCUITS

Circuit B is intended for operation above the self-neutralizing frequency of the tetrode. Above 30 MHz or so, the screen bypass capacitor of circuits A and C is often chosen so as to be self-resonant at the operating frequency of the amplifier.

C_1, L_1—Input tuned circuit. Typically, 3 pf per meter of wavelength for circuits A and B. 20 pf per meter of wavelength for circuit C

C_2, C_3, L_2—Pi-network plate circuit. Refer to plate-circuit design data in Chapter 11

C_4—Plate-blocking capacitor. Typically, 500 pf to 1000 pf at 5 kV

R_1—Wirewound resistor (100-500 ohms) to act as low-Q r-f choke

R_2—Screen resistor to carry negative screen current and complete screen-to-ground circuit. See tube data sheet for details

PC—Plate parasitic suppressor. See Chapter 17 and figure 1 of this chapter. For vhf operation, suppressor may consist of composition resistor shunted across a short portion of the plate lead

RFC_1—Grid choke, receiving type. Typically, 2.5 mH for 3- to 30-MHz range. Vhf-rated choke for 50 MHz and 144 MHz

RFC_2—Plate choke, transmitting type, solenoid. Rated to carry plate current. Typically, 800 µH for 3- to 30-MHz range. Vhf-rated choke for 50 MHz and 144 MHz.

RFC_3—Receiving-type choke. 2.5 mH for 3- to 30-MHz range

M_1—Grid-current meter

M_2—Plate-current meter

M_3—Screen-current meter

all conditions of operation of the tetrode. In particular, if the *d-c screen-to-ground* path is broken, the screen voltage may rise to equal the plate potential, thus damaging the tube and rupturing the screen bypass capacitor. It is dangerous, therefore, to reduce screen voltage for tuneup purposes by simply breaking the screen power lead unless a protective screen bleeder resistor (R_2) is placed directly at the tube socket, as shown in the illustrations of figure 3. If this resistor is used, the screen supply may be safely broken at point X for tuneup purposes, or for reduced-power operation. The value of screen bleeder resistance will vary depending on tube characteristics, and a typical value is generally specified in the tube data sheet. For tubes of the 4CX250B family, the value of resistance is chosen to draw about 15 to 20 ma from the screen power supply. The 4CX1000A, on the other hand, requires a screen bleeder current of about 70 ma.

In any case, regardless of whether the screen circuit is broken or not, the use of a screen bleeder resistor in the circuit at all times is mandatory for those tetrodes which produce reverse screen current under certain operating conditions. This is a normal characteristic of most modern, high-gain tetrodes and the screen power supply should be designed with this characteristic in mind so that correct operating voltages will be maintained on the screen at all times.

With the use of a screen bleeder resistor, full protection for the screen may be provided by an overcurrent relay and by interlocking the screen supply so that the plate voltage must be applied before screen voltage can be applied.

Power output from a tetrode is very sensitive to screen voltage, and for linear service a well-regulated screen power supply is required. Voltage-regulator tubes or a series-regulated power supply are often used in high-power tetrode linear-amplifier stages.

A tetrode neutralizing circuit suitable for the lower portion of the vhf region is shown in figure 3B. When the operating frequency of the tetrode is higher than the self-neutralizing frequency, the r-f voltage developed in the screen circuit is too great to provide proper voltage division between the internal capacitances of the tube (see Chapter 11). One method of reducing the voltage across the screen lead inductance and thus achieving neutralization is to adjust the inductive reactance of the screen-to-ground path so as to lower the total reactance. This reactance adjustment may take the form of a variable series capacitor as shown in illustration B. This circuit is frequency sensitive and must be readjusted for major changes in the frequency of operation of the amplifier.

Balanced input and output tuned circuits are used in the configuration of figure 3B. In the grid circuit, the split capacitance is composed of variable capacitor C_1 and the grid-cathode input capacitance of the tube. The coil (L_1) is chosen so that C_1 approximates the input capacitance. The same technique is employed in the plate circuit, where a split tank is achieved by virtue of capacitance C_2 and the output capacitance of the tetrode tube.

A cathode-driven tetrode amplifier is shown in illustration C. Many tetrodes do not perform well when connected in class-B grounded-grid configuration (screen and grid both at ground potential). These tubes are characterized by high perveance, together with extremely small spacing between the grid bars, and between the grid structure and the cathode. Tubes of the 4-65A, 4X150A/4CX250B, and 4CX1000A family are in this class. For proper operation of these high-gain tubes, the screen requires much larger voltage than the control grid. When the electrodes of these tubes are tied together, the control grid tends to draw heavy current and there is risk of damaging the tube. Lower-gain tetrodes, such as the 813, 4-400A, and 4-1000A have a more balanced ratio of grid to screen current and may be operated in zero-bias, grounded-grid mode. The best way to employ the higher-gain tetrode tubes in cathode-driven service is to ground the grid and screen through bypass capacitors and to operate the elements at their rated class AB_1 d-c voltages. In all cases, grid and screen current should be monitored so as to keep maximum currents within ratings.

Tetrode Amplifier Circuitry The most widely used tetrode circuitry for h-f use is the single-ended pi-network configuration, variations of which are shown in figure 4.

A common form of pi-network amplifier is shown in figure 4A. The *pi* circuit forms the matching system between the plate of the amplifier tube and the low-impedance, unbalanced, antenna circuit. The coil and input capacitor of the *pi* may be varied to tune the circuit over a 10 to 1 frequency range (usually 3.0 to 30 MHz). Operation over the 20- to 30-MHz range takes place when the variable slider on coil L_2 is adjusted to short this coil out of the circuit. Coil L_1 therefore comprises the tank inductance for the highest portion of the operating range. This coil has no taps or sliders and is constructed for the highest possible Q at the high-frequency end of the range. The adjustable coil (because of the variable tap and physical construction) usually has a lower Q than that of the fixed coil.

The degree of loading is controlled by capacitors C_4 and C_5. The amount of circuit capacitance required at this point is inversely proportional to the operating frequency and to the impedance of the antenna circuit. A loading capacitor range of 100 to 2500 pf is normally ample to cover the 3.5- to 30-MHz range.

The *pi* circuit is usually shunt-fed to remove the d-c plate voltage from the coils and capacitors. The components are held at ground potential by completing the circuit to ground through the choke (RFC_1). Great stress is placed on the plate-circuit choke (RFC_2). This component must be specially designed for this mode of operation, having low interturn capacitance and no spurious

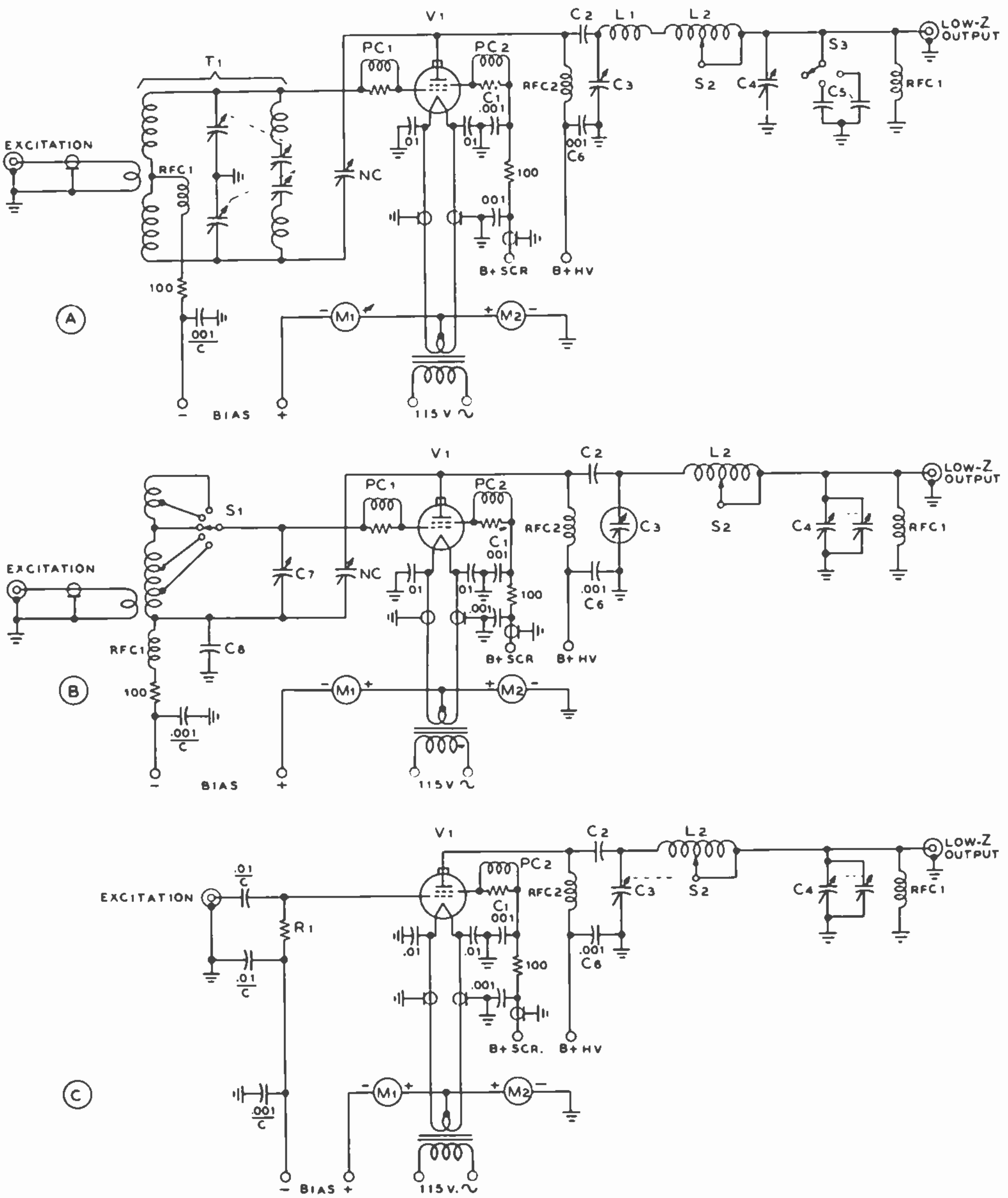


Figure 4

TYPICAL PI-NETWORK CONFIGURATIONS

A—Split grid circuit provides out-of-phase voltage for grid neutralization of tetrode tube. Rotary coil is employed in plate circuit, with small, fixed auxiliary coil for 28 MHz. Multiple tuning grid tank T_1 covers 3.5-30 MHz without switching

B—Tapped grid and plate inductors are used with "bridge-type" neutralizing circuit for tetrode amplifier stage. Vacuum tuning capacitor is used in input section of pi-network

C—Untuned input circuit (resistance loaded) and plate inductor ganged with tuning capacitor comprise simple amplifier configuration. R_1 is usually 50-ohm, 100-watt carbon resistor.

PC_1, PC_2—57-ohm, 2-watt composition resistor, wound with 3 turns #12 enam. wire

Note: Alternatively, PC_2 may be placed in the plate lead.

internal resonances throughout the operating range of the amplifier.

Parasitic suppression is accomplished by means of chokes PC_1 and PC_2 in the screen, grid, or plate leads of the tetrode. Suitable values for these chokes are given in the parts list of figure 4. Effective parasitic suppression is dependent to a large degree on the choice of screen bypass capacitor C_1. This component must have extremely low inductance throughout the operating range of the amplifier and well up into the vhf parasitic range. The capacitor must have a voltage rating equal to at least twice the screen potential (four times the screen potential for plate modulation). There are practically no capacitors available that will perform this difficult task. One satisfactory solution is to allow the amplifier chassis to form one plate of the screen capacitor. A "sandwich" is built on the chassis with a sheet of insulating material of high dielectric constant and a matching metal sheet which forms the screen side of the capacitance. A capacitor of this type has very low internal inductance but is very bulky and takes up valuable space beneath the chassis. One suitable capacitor for this position is the *Centralab type 858S-1000*, rated at 1000 pf at 5000 volts. This compact ceramic capacitor has relatively low internal inductance and may be mounted to the chassis by a 6-32 bolt. Further screen isolation may be provided by a shielded power lead, isolated from the screen by a .001-μfd ceramic capacitor and a 100-ohm carbon resistor.

Various forms of the basic pi-network amplifier are shown in figure 4. The *A* configuration employs the so-called "all-band" grid-tank circuit and a rotary pi-network coil in the plate circuit. The *B* circuit uses coil switching in the grid circuit, bridge neutralization, and a tapped pi-network coil with a vacuum tuning capacitor. Figure 4C shows an interesting circuit that is becoming more popular for class-AB_1 linear operation. A tetrode tube operating under class-AB_1 conditions draws no grid current and requires no grid-driving power. Only r-f voltage is required for proper operation. It is possible therefore to dispense with the usual tuned grid circuit and neutralizing capacitor and in their place employ a noninductive load resistor in the grid circuit across which the required excitation voltage may be developed. This resistor can be of the order of 50 to 300 ohms, depending on circuit requirements. Considerable power must be dissipated in the resistor to develop sufficient grid swing, but driving power is often cheaper to obtain than the cost of the

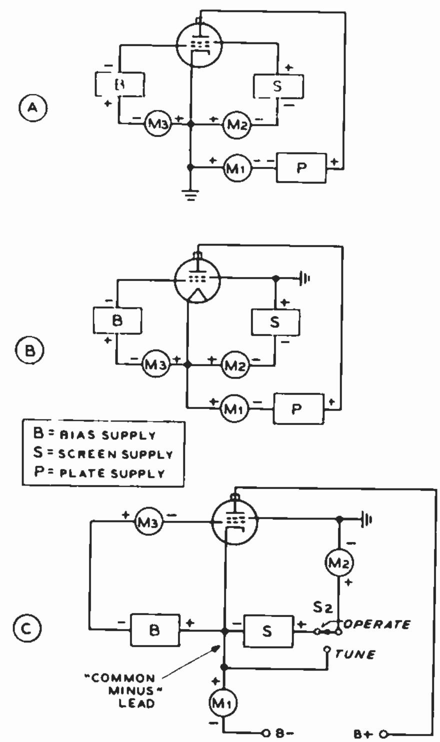


Figure 5

GROUNDED-SCREEN-GRID CONFIGURATION PROVIDES HIGH ORDER OF ISOLATION IN TETRODE AMPLIFIER STAGE

A—Typical amplifier circuit has cathode return at ground potential. All circuits return to cathode.

B—All circuits return to cathode, but ground point has been shifted to screen terminal of tube. Operation of the circuit remains the same, as potential differences between elements of the tube are the same as in circuit A.

C—Practical grounded-screen circuit. "Common minus" lead returns to negative of plate supply, which cannot be grounded. Switch S_2 removes screen voltage for tune-up purposes.

usual grid-circuit components. In addition, the low-impedance grid return removes the tendency toward instability that is often common to the circuits of figures 4A and 4B. Neutralization is not required of the circuit of figure 4C, and in many cases parasitic suppression may be omitted. The price that must be paid is the additional excitation that is required to develop operating voltage across grid resistor R_1.

The pi-network circuit of figure 4C is interesting in that the rotary coil (L_2) and the plate tuning capacitor (C_3) are ganged together by a gear train, enabling the circuit to be tuned to resonance with one panel control instead of the two required by the circuit of figure 4A. Careful design of the rotary inductor will permit the elimination of the auxiliary high-frequency coil (L_1), thus reducing the cost and complexity of the circuit.

The Grounded-Screen Configuration For maximum shielding, it is necessary to operate the tetrode tube with the screen at r-f ground potential. As the screen has a d-c potential applied to it (in grid-driven circuits), it must be bypassed to ground to provide the necessary r-f return. The bypass capacitor employed must perform efficiently over a vast frequency spectrum that includes the operating range plus the region of possible vhf parasitic oscillations. This is a large order, and the usual bypass capacitors possess sufficient inductance to introduce regeneration into the screen circuit, degrading the grid-plate shielding to a marked degree. Nonlinearity and self-oscillation can be the result of this loss of circuit isolation. A solution to this problem is to eliminate the screen bypass capacitor, by grounding the screen terminals of the tube by means of a low-inductance strap. Screen voltage is then applied to the tube by grounding the positive terminal of the screen supply, and "floating" the negative of the screen and bias supplies below ground potential as shown in figure 5. Meters are placed in the separate-circuit cathode return leads, and each meter reads only the current flowing in that particular circuit. Operation of this grounded-screen circuit is normal in all respects, and it may be applied to any form of grid-driven tetrode amplifier with good results.

The Inductively Tuned Tank Circuit The output capacitance of large transmitting tubes and the residual circuit capacitance are often sufficiently great to prevent the plate tank circuit from having the desired value of Q, especially in the upper reaches of the h-f range (28- to 54-MHz). Where tank capacitance values are small, it is possible for the output capacitance of the tube to be greater than the maximum desired value of tank capacitance. In some cases, it is possible to permit the circuit to operate with higher-than-normal Q, however this expedient is unsatisfactory when circulating tank current is high, as it usually is in high-frequency amplifiers.

A practical alternative is to employ *inductive tuning* and to dispense entirely with the input tuning capacitor which usually has a high minimum value of capacitance (figure 6). The input capacitance of the circuit is thus reduced to that of the output capacitance of the tube which may be more nearly the desired value. Circuit resonance is established by varying the inductance of the tank coil with a movable, shorted turn, or loop, which may be made of a short length of copper water pipe of the proper diameter. The shorted turn is inserted within the tank coil by a lead-screw mechanism, or it may be mounted at an angle within the coil and rotated so that its plane travels from a parallel to an oblique position with respect to the coil. The shorted turn should be silver plated and have no joints to hold r-f losses to a minimum. Due attention should be given to the driving mechanism so that unwanted, parasitic shorted turns do not exist in this device.

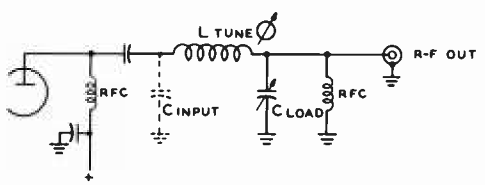


Figure 6
INDUCTIVE TUNING ELIMINATES INPUT TUNING CAPACITOR

Push-Pull Tetrode Circuitry Tetrode tubes may be employed in push-pull amplifiers, although the modern trend is to parallel operation of these tubes. A typical circuit for push-pull operation is shown in figure 7. The remarks concerning the filament supply, plate feed, and grid bias in Section 22-1 apply equally to tetrode stages. Because of the high circuit gain of the tetrode amplifier, extreme care must be taken to limit intrastage feedback to an absolute minimum. It must be remembered with high-gain tubes of this type that almost full output can be obtained with practically

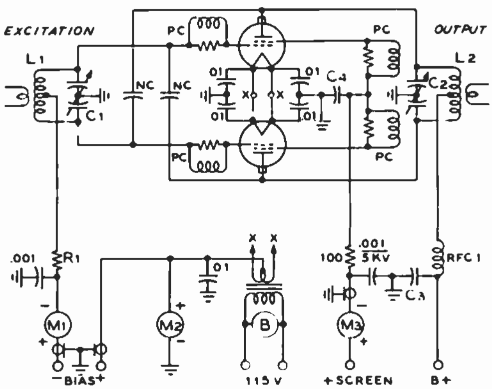


Figure 7

REPRESENTATIVE PUSH-PULL TETRODE AMPLIFIER CIRCUIT

The push-pull tetrode amplifier uses many of the same components required by the triode amplifier of figure 2. Parasitic suppressors may be placed in grid, screen, or plate leads. A low-inductance screen capacitor is required for proper amplifier operation. Capacitor C_t may be .001 µfd, 5 kV. Centralab type 858S-1000. Strap multiple screen terminals together at socket with ⅛-inch copper strap for operation below 30 MHz and attach PC to center of strap. Blower required for many medium- and high-power tetrode tubes to cool filament and plate seals.

zero grid excitation. Any minute amount of energy fed back from the plate circuit to the grid circuit can cause instability or oscillation. *Unless suitable precautions are incorporated in the electrical and mechanical design of the amplifier, this energy feedback will inevitably occur.*

Fortunately these precautions are simple. The grid and filament circuits must be isolated from the plate circuit. This is done by placing these circuits in an "electrically tight" box. All leads departing from this box are bypassed and filtered so that no r-f energy can pass along the leads into the box. This restricts the energy leakage path between the plate and grid circuits to the residual plate-to-grid capacity of the tetrode tubes. This capacity is of the order of 0.25 pf per tube, and under normal conditions is sufficient to produce a highly regenerative condition in the amplifier. Whether or not the amplifier will actually break into oscillation is dependent upon circuit loading and residual lead inductance of the stage. Suffice to say that unless the tubes are actually neutralized a condition exists that will lead to circuit instability and oscillation under certain operating conditions.

Parasitic suppression is required with most modern high-gain tetrodes and may take place in either the plate or screen circuit. In some instances, suppressors are required in the grid circuit as well. Design of the suppressor is a cut-and-try process: if the inductor of the suppressor has too few turns,

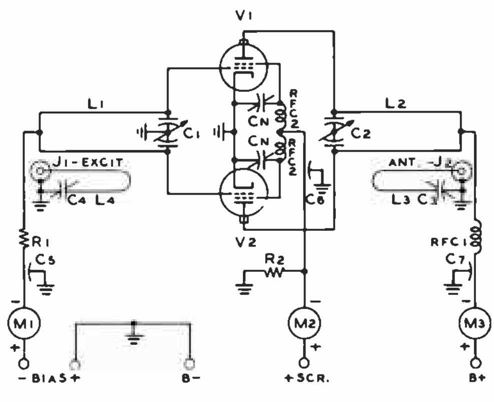


Figure 8

REPRESENTATIVE VHF PUSH-PULL TETRODE AMPLIFIER CIRCUIT

Tuned lines are used in grid and plate tank circuits in place of lumped inductances. Each screen circuit is series resonated to ground by neutralizing capacitor C_N. Wirewound resistor (R_1) is used in the grid-return circuit and frequency-rated r-f chokes in the plate and screen power leads. Screen resistor is included to complete screen-to-ground circuit, as discussed in text. Vhf type feedthrough capacitors are used for maximum suppression of r-f currents in power leads.

the parasitic oscillation will not be adequately suppressed. Too many turns on the suppressor will allow too great an amount of fundamental frequency power to be absorbed by the suppressor and it will overheat and be destroyed. From 3 to 5 turns of #12 wire in parallel with a 50-ohm, 2-watt composition resistor will usually suffice for operation in the h-f region. At 50 MHz, the suppressor inductor may take the form of a length of copper strap (often a section of the plate lead) shunted by the suppressor resistor.

VHF Push-Pull Tetrode Amplifiers The circuit considerations for the vhf triode amplifier configuration apply equally well to the push-pull tetrode circuit shown in figure 8. The neutralization techniques applied to the tetrode tube however, may vary as the frequency of operation of the amplifier varies about the *self-neutralizing* frequency of the tetrode tube. At or near the upper frequency limit of operation, the inductance of the screen-grid lead of the tetrode cannot be ignored as it becomes of importance. Passage of r-f current through the screen lead produces a potential drop in the lead which may or may not be in phase with the grid voltage impressed on the tube. At the self-neutralizing frequency of the tube, the tube is inherently neutralized due to the voltage and current divisions within the tube which place the grid at the filament potential as far as plate-circuit action is concerned (see Chapter 11, Section 6). When the tetrode tube is operated below this frequency, normal neutralizing circuits apply; operation at the self-neutralizing frequency normally does not require neutralization, provided the input and output circuits are well shielded. Operation above the self-neutralizing frequency (in the range of 25 MHz to 100 MHz for large glass tubes, and in the range of 120 MHz to 600 MHz for ceramic, vhf tubes) requires neutralization, which may take the form of a series screen-tuning capacitor, such as shown in the illustration.

Neutralization is frequency sensitive and the amplifier should be neutralized at the operating frequency. Adjustment is conducted so as to reduce the power fed from the grid to the plate circuit. The amplifier may be driven with a test signal (filament and d-c voltages removed) and the signal in the plate tank circuit measured with an r-f voltmeter. The neutralizing capacitors are adjusted in unison until a minimum of fed-through voltage is measured. A good null will be obtained provided that intrastage feedback is reduced to a minimum by proper shielding and lead-bypassing techniques.

Sweep Tubes in Linear Service Listed in figure 9 are intermittent voice operation ratings for various TV sweep tubes when used for linear operation in the amateur service. While the plate dissipation of these tubes is of the order of 30 to 35 watts, the intermittent nature of amateur transmission and the high ratio of peak to average power in the human voice allow a good balance between peak power input, tube life, and tube cost to be achieved. For lower levels of intermodulation distortion, the user must shift to transmitting-type tubes rated for linear service, and which are designed to have low intermodulation distortion characteristics.

22-3 Cathode-Driven Amplifier Design

The *cathode-driven,* or *grounded-grid* amplifier has achieved astounding popularity in recent years as a high-power linear stage for sideband application. Various versions of this circuit are illustrated in figure 10. In the basic circuit the control grid of the tube is at r-f ground potential and the exciting signal is applied to the cathode by means of a tuned circuit. Since the grid of the tube is grounded, it serves as a shield between the input and output circuits, making neutralization unnecessary in many instances. The very small plate-to-cathode capacitance of most tubes permits a minimum of intrastage coupling below 30 MHz. In addition, when zero-bias triodes or tetrodes are used, screen or bias supplies are not usually required.

Feedthrough Power A portion of the exciting power appears in the plate circuit of the grounded-grid (cathode-driven) amplifier and is termed *feedthrough*

R-F LINEAR AMPLIFIER SERVICE FOR SSB AND CW GRID DRIVEN, CLASS AB_1 MODE													
TUBE	FIL V/A	BASE	PLATE VOLTAGE Eb	SCREEN VOLTAGE Ec2	GRID VOLTAGE Ec1	ZERO SIG. PLATE CUR. Ibo	MAX. SIG. PLATE CUR. Ib	MAX SIG. SCREEN CUR. Ic2	PL LOAD IMPEDAN Rp-Ω	PLATE INPUT PWR. W.	USEFUL POWER OUT. Po	AVERAGE PLATE DISSIP. Pd	3d ORDER IMD Db
6146 6146B	6.3/1.2	7CK	600	200	-46	25	103	9	3570	61	41	16	-25
			750	200	-51	25	118	7	2825	88	55	28	-22
			800	290	-69	30	125	10	3620	100	59	35	-24
			800	290	-77	25	180	13	2300	145	91	45	-19
807	6.3/0.9	5AW	600	300	-34	18	70	8	4300	42	28	12	-23
			750	300	-35	15	70	8	5200	53	36	14	-23
6550	6.3/1.6	7S	680	340	-39	48	140	20	3010	95	67	28	-32
			800	290	-33	45	127	15	3920	102	70	29	-30
6DQ5	6.3/2.5	8JC	500	150	-46	48	170	17	1800	85	54	27	-28
			500	150	-48	48	182	13	1625	91	56	29	-26
			700	150	-49	35	182	11	2210	127	78	41	-23
			800	180	-67	30	250	13	1710	200	121	70	-19
6GB5	6.3/1.38	9NH	600	200	-41	23	192	14	1900	115	80	30	-18
6GE5	6.3/1.2	12BJ	600	200	-45	30	132	15	2500	79	51	23	-22
			800	250	-61	25	172	18	2750	138	90	39	-19
6HF5	6.3/2.25	12FB	500	140	-46	40	133	5	1900	67	35	29	-27
			800	125	-45	30	197	7	2170	158	100	48	-21
6JE6A	6.3/2.5	9QL	500	125	-44	40	110	4	2300	55	30	24	-28
			750	175	-63	27	218	15	1850	163	102	51	-20
6LQ6	6.3/2.5	9QL	750	175	-60	25	215	9	1850	161	102	49	-18
			800	200	-69	25	242	13	1850	197	124	60	-18

Figure 9

SWEEP TUBE DATA FOR CLASS AB_1 LINEAR AMPLIFIER SERVICE

power. In any amplifier of this type, whether it be triode or tetrode, it is desirable to have a large ratio of feedthrough power to peak grid-driving power. The feedthrough power acts as a swamping resistor across the driving circuit to stabilize the effects of grid loading. The ratio of feedthrough power to driving power should be about 10 to 1 for best stage linearity. The feedthrough power provides the user with added output power he would not obtain from a more conventional circuit. The driver stage for the grounded-grid amplifier must, of course, supply the normal excitation power plus the feedthrough power. Many commercial sideband exciters have power output capabilities of the order of 70 to 100 watts and are thus well suited to drive high-power grounded-grid linear amplifier stages whose total excitation requirements fall within this range.

Distortion Products Laboratory measurements made on various tubes in the circuit of figure 10A show that a distortion reduction of the order of 5 to 10 decibels in odd-order products can be obtained by operating the tube in cathode-driven service as opposed to grid-driven service. The improvement in distortion varies from tube type to tube type, but some order of improvement is noted for all tube types tested. Most amateur-type transmitting tubes provide signal-to-distortion ratios of −20 to −30 decibels at full output in class-AB_1 grid-driven operation. The ratio increases to approximately −25 to −40 decibels for class-B grounded-grid operation. Distortion improvement is substantial, but not as great as might otherwise be assumed from the large amount of feedback inherent in the grounded-grid circuit.

A simplified version of the grounded-grid amplifier is shown in figure 10B. This configuration utilizes an untuned input circuit, circuit of figure 10A. It has inherent limitations, however, that should be recognized. In general, slightly less power output and efficiency is observed with the untuned-cathode circuit, odd-order distortion products run 4 to 6 decibels higher, and the circuit is harder to drive and match to the exciter than is the tuned-cathode circuit of figure 10A. Best results are obtained when the

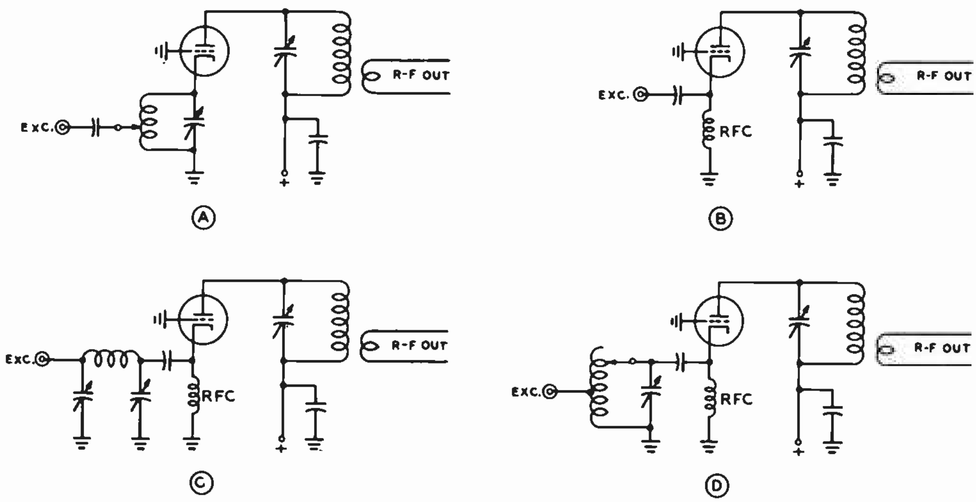


Figure 10

THE CATHODE-DRIVEN AMPLIFIER

Widely used as a linear amplifier for sideband service, the cathode-driven (g-g) circuit provides economy and simplicity, in addition to a worthwhile reduction in intermodulation distortion. A—The basic g-g amplifier employs tuned input circuit. B—A simplified circuit employs untuned r-f choke in cathode in place of the tuned circuit. Linearity and power output are inferior compared to circuit of figure A. C—Simple high-C pi-network may be used to match output impedance of sideband exciter to input impedance of grounded-grid stage. D—Parallel-tuned, high-C circuit may be employed for bandswitching amplifier. Excitation tap is adjusted to provide low value of SWR on exciter coaxial line.

coaxial line of the driver stage is very short —a few feet or so. Optimum linearity requires cathode-circuit Q that can only be supplied by a high-C tank circuit.

Since the single-ended class-B grounded-grid linear amplifier draws grid current on only one-half (or less) of the operating cycle, the sideband exciter "sees" a low-impedance load during this time, and a very high-impedance load over the balance of the cycle. Linearity of the exciter is thereby affected and the distortion products of the exciter are enhanced. Thus, the *driving signal* is degraded in the cathode circuit of the grounded-grid stage unless the unbalanced input impedance can be modified in some fashion. A high-C tuned circuit, stores enough energy over the operating r-f cycle so that the exciter "sees" a relatively constant load at all times. In addition, the tuned circuit may be tapped or otherwise adjusted so that the SWR on the coaxial line coupling the exciter to the amplifier is relatively low. This is a great advantage, particularly in the case of those exciters having fixed-ratio pi-network output circuits designed expressly for a 50-ohm termination.

Finally, it must be noted that removal of the tuned cathode circuit breaks the amplifier plate-circuit return to the cathode, and r-f plate-current pulses must return to the cathode via the outer shield of the driver coaxial line and back via the center conductor! Extreme fluctuations in exciter loading, intermodulation distortion, and TVI can be noticed by changing the length of the cable between the exciter and the grounded-grid amplifier when an untuned-cathode input circuit and a long interconnecting coaxial line are used.

Cathode-Driven Amplifier Construction Design features of the single-ended and push-pull amplifiers discussed previously apply equally well to the grounded-grid stage. The g-g linear amplifier may have either configuration, although the majority of the g-g stages are single ended, as push-pull offers no distinct advantages and adds greatly to circuit complexity.

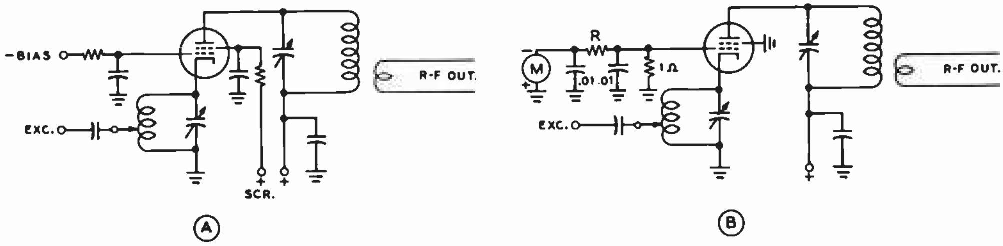


Figure 11

TETRODE TUBES MAY BE USED IN CATHODE-DRIVEN AMPLIFIERS

A—Tetrode tube may be used in cathode-driven configuration, with bias and screen voltages applied to elements which are at r-f ground potential. B—Grid current of grounded-grid tube is easily monitored by RC network which lifts grid above ground sufficiently to permit a millivoltmeter to indicate voltage drop across 1-ohm resistor. Meter is a 0-1 d-c milliammeter in series with appropriate multiplier resistor.

The *cathode circuit* of the amplifier is resonated to the operating frequency by means of a high-C tank (figure 10A). Resonance is indicated by maximum grid current of the stage. A low value of SWR on the driver coaxial line may be achieved by adjusting the tap on the tuned circuit, or by varying the capacitors of the pi-network (figure 10C). Correct adjustments will produce minimum SWR and maximum amplifier grid current at the same settings. The cathode tank should have a Q of 2 or more.

The cathode circuit should be completely shielded from the plate circuit. It is common practice to mount the cathode components in an "r-f tight" box below the chassis of the amplifier, and to place the plate circuit components in a screened box above the chassis.

The *grid (or screen) circuit* of the tube is operated at r-f ground potential, or may have d-c voltage applied to it to determine the operating parameters of the stage (figure 11A). In either case, the r-f path to ground must be short, and have extremely low inductance, otherwise the screening action of the element will be impaired. The grid (and screen) therefore, must be bypassed to ground over a frequency range that includes the operating spectrum as well as the region of possible vhf parasitic oscillations. This is quite a large order. The inherent inductance of the usual bypass capacitor plus the length of element lead within the tube is often sufficient to introduce enough regeneration into the circuit to degrade the linearity of the amplifier at high signal levels even though the instability is not great enough to cause parasitic oscillation. In addition, it is often desired to "unground" the grounded screen or grid sufficiently to permit a metering circuit to be inserted.

One practical solution to these problems is to shunt the tube element to ground by means of a 1-ohm composition resistor, bypassed with a .01-μfd ceramic disc capacitor. The voltage drop caused by the flow of grid (or screen) current through the resistor can easily be measured by a millivoltmeter whose scale is calibrated in terms of element current (figure 11B).

The *plate circuit* of the grounded-grid amplifier is conventional, and either pi-network or inductive coupling to the load may be used. There is some evidence to support the belief that intermodulation distortion products are reduced by employing plate circuit Q's somewhat higher than normally used in class-C amplifier design. A circuit Q of 10 or greater is thus recommended for ground-grid amplifier plate circuits.

Tuning the Grounded-Grid Amplifier Since the input and output circuits of the grounded-grid amplifier are in series, a certain proportion of driving power appears in the output circuit. If full excitation is applied to the stage and the output circuit is opened, or the plate voltage removed from the tube, practically all of the driving power will be dissipated by the grid of the tube. Overheating of this element will quickly occur under these circumstances, followed by damage to the tube. Full excitation should therefore never be applied to a grounded-grid stage unless plate voltage is applied beforehand, and the stage is loaded to the antenna.

Tuneup for sideband operation consists of applying full plate voltage and sufficient excitation (carrier injection) so that a small rise in resting plate (cathode) current is noted. The plate loading capacitor is set near full capacitance and the plate tank capacitor is adjusted for resonance (minimum plate current). Drive is advanced until grid current is noted and the plate circuit is loaded by decreasing the capacitance of the plate loading capacitor. The drive is increased until about one-half normal grid current flows, and loading is continued (re-resonating the plate tank capacitor as required) until loading is near normal. Finally, grid drive and loading are adjusted until PEP-condition plate and grid currents are normal. The values of plate and grid current should be logged for future reference. At this point, the amplifier is loaded to the maximum PEP input condition. In most cases, the amplifier and power supply are capable of operation at this power level for only a short period of time, and it is not recommended that this condition be permitted for more than a minute or two.

The exciter is now switched to the SSB mode and, with speech excitation, the grid and plate currents of the cathode-driven stage should rise to approximately 40 to 50 percent of the previously logged PEP readings. The exact amount of meter movement with speech is variable and depends on meter damping and the peak to average ratio of the particular voice. Under no circumstances, however, should the voice meter readings exceed 50 percent of the PEP adjustment readings unless some form of speech compression is in use.

To properly load a linear amplifier for the so-called "two-kilowatt PEP" condition, *it is necessary for the amplifier to be tuned and loaded at the two-kilowatt level,* albeit briefly. It is necessary to use a dummy load to comply with the FCC regulations, or else a two-tone test signal should be used, as discussed in Chapter 9.

For best linearity, the output circuit of the grounded-grid stage should be overcoupled so that power output drops about 2-percent from maximum value. A simple output r-f voltmeter is indispensable for proper circuit adjustment. Excessive grid current is a sign of antenna undercoupling, and overcoupling is indicated by a rapid drop in output power. Proper grounded-grid stage operation can be determined by finding the optimum ratio between grid and plate current and by adjusting the drive level and loading to maintain this ratio. Many manufacturers now provide grounded-grid operation data for their tubes, and the ratio of grid to plate current can be determined from the data for each particular tube.

Choice of Tubes for G-G Service Not all tubes are suitable for grounded-grid service. In addition, the signal-to-distortion ratio of the suitable tubes varies over a wide range. Some of the best g-g performers are the 811A, 813, 4-400A, and 4-1000A. In addition, the 3-400Z, 3-500Z, 8873, 8877 and 3-1000Z triodes are specifically designed for low distortion, grounded-grid amplifier service.

Certain types of tetrodes, exemplified by the 4-65A, 4X150A, 4CX300A, and 4CX-1000A should not be used as grounded-grid amplifiers unless grid bias and screen voltage are applied to the elements of the tube (figure 11A). The internal structure of these tubes permits unusually high values of grid current to flow when true grounded-grid circuitry is used, and the tube may be easily damaged by this mode of operation.

The efficiency of a typical cathode-driven amplifier runs between 55- and 65-percent, indicating that the tube employed should have plenty of plate dissipation. In general, the PEP input in watts to a tube operating in grounded-grid configuration can safely be about 2.5 to 3 times the rated plate dissipation. Because of the relatively low average-to-peak power of the human voice it is tempting to push this ratio to a higher figure in order to obtain more output from a given tube. This action is unwise in that the odd-order distortion products rise rapidly when the tube is overloaded, and because

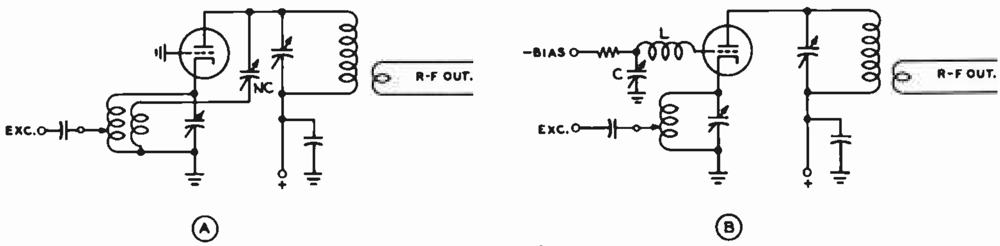


Figure 12

NEUTRALIZING CIRCUITS FOR CATHODE-DRIVEN STAGES

Neutralization of the g-g stage may be necessary at the higher frequencies. Energy fed back in proper phase from plate to cathode is used to neutralize the unwanted energy fed through the tube (A). Reactance placed in series with the grid return lead (B) will accomplish the same result. The inductance (L) usually consists of the internal grid lead of the tube, and capacitor C may be the grid bypass capacitor. A series-resonant circuit at the operating frequency is thus formed.

no safety margin is left for tuning errors or circuit adjustments.

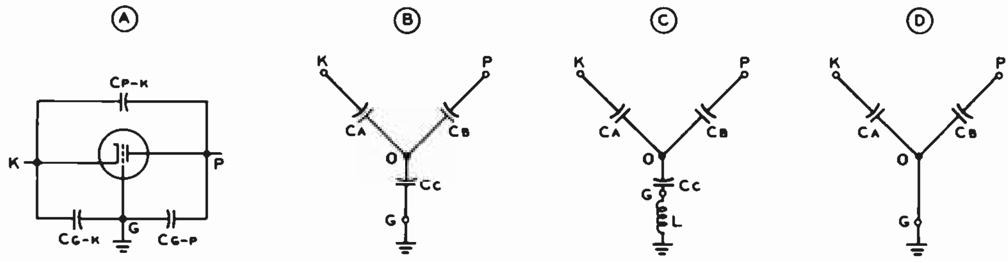
Neutralization of the G-G Stage At some high frequency the shielding action of the grid of the g-g amplifier deteriorates. Neutralization may be necessary at higher frequencies either because of the presence of inductance between the active grid element and the common returns of the input and output circuit, or because of excessive plate-cathode capacitance.

Neutralization, where required, may be accomplished by feeding out-of-phase energy from the plate circuit to the filament circuit (figure 12A) or by inserting a reactance in series with the grid (figure 12B). For values of plate-cathode capacitance normally encountered in tubes usable in g-g service, the residual inductance in the grid-ground path provides sufficient reactance, and in some cases even series capacitance will be required. Typical tube electrode capacitances are shown in figure 13A. These can be represented by an equivalent star connection of three capacitors (figure 13B.). If an inductance (*L*) is placed in series with C_C so that a resonant circuit is formed (figure 13C), point O will be at ground potential (13D). This prevents the transfer of energy from point *P* to point *K*, since there now exists no common coupling impedance. The determination of value C_C and L are shown in figure 13.

It is apparent that when the plate-cathode capacitance of the tube is small as compared to the plate-grid and the grid-cathode capacitance, C_C is a large value and the required value of inductance L is small. In practical cases the value of L is supplied by the tube and lead inductance, and the grid-to-ground impedance can be closely adjusted by proper choice of the bias bypass capacitor (figure 12B). Below a certain frequency determined by the physical geometry of the tube, neutralization may be accomplished by adding inductance to the grid-return lead; above this frequency it may be necessary to series tune the circuit for minimum energy feedthrough from cathode to plate. Most tubes are sufficiently well screened so that series inductive neutralization at the lower frequencies is unnecessary, but series capacitance tuning of the grid-return lead may be required to prevent oscillation at some parasitic frequency in the vhf range.

22-4 Two Solid-State Broadband Linear Amplifiers for SSB

Described in this section are two transistorized, broadband, class-B linear amplifiers covering the 1.5- to 30-MHz range. They are designed by the *Semiconductor Division of TRW, Inc.* The amplifiers are untuned, operate from a nominal 12.5 d-c volt source and provide outputs of 25 watts PEP and 100 watts PEP, respectively. They exhibit intermodulation distortion product



$$C_C = \frac{(C_{P-G} \times C_{P-K}) \pm (C_{P-K} \times C_{G-K}) \pm (C_{G-K} \times C_{P-G})}{C_{P-K}}$$

$$L = \frac{1}{(2\pi f)^2 \times C_C}$$

Figure 13

Tube electrode capacitances can be represented by an equivalent star connection of three capacitors. If inductance is placed in series with C_c so that a resonant circuit is formed (drawing C), point 0 will be at ground potential.

levels of better than —30 db below one tone of a two-tone test signal at full output level.

The amplifiers combine small size, good efficiency, and wide instantaneous bandwidth with high stability and excellent tolerance to various operating conditions. In particular, these units are designed to withstand a wide range of operating temperatures (such as encountered in portable or mobile work), bias variation, extremes of load SWR, and overdrive condition. The amplifiers are assembled on circuit boards which are mounted on aluminum heat sinks to achieve proper temperature control. The units may be placed in a cabinet or case at the builder's choice.

The 25-Watt Amplifier The 25-watt PEP output amplifier is shown in figures 14 through 18. It requires only 0.4 watt PEP drive at 30 MHz for full output, having a power gain of about 18 decibels. Amplifier efficiency is about 55 percent under c-w (carrier) conditions. Even-order harmonics are better than —35 decibels below peak power output. The level of the odd-order harmonics is such that a harmonic filter should be incorporated after the amplifier to suppress the 3rd, 5th, and 7th order harmonics which are attenuated less than —30 decibels below peak power output in the amplifier.

Amplifier Circuitry—The schematic of the 25-watt amplifier is shown in figure 15. Two *TRW* type PT5740 epitaxial silicon NPN power transistors, specially designed for h-f SSB service are used (Q_1, Q_2). The transistors incorporate temperature-compensating emitter resistors on the chip and are designed to work into an infinite SWR load without damage at a maximum collector potential of 16 volts.

The PT5740 devices are connected in a push-pull configuration with broadband, ferrite-loaded transformers used in the input and output circuits to match unbalanced terminations. A simple RLC compensation network is placed across the input winding of transformer T_1 to equalize amplifier gain across the operating range.

The input impedance of a PT 5740 power transistor is below 5.5 ohms and is capacitive over the operating range of the amplifier. The output impedance is of the order of 4 ohms. As a result, special r-f transformers must be built to match these very low impedance levels to 50 ohms.

The push-pull collectors of the transistors are connected to a balanced feed transformer (T_2) and to a matching output transformer (T_3) to provide single-ended output at a nominal impedance value of 50 ohms. The push-pull configuration is used since the amplifier covers five octaves of bandwidth, and suppression of even har-

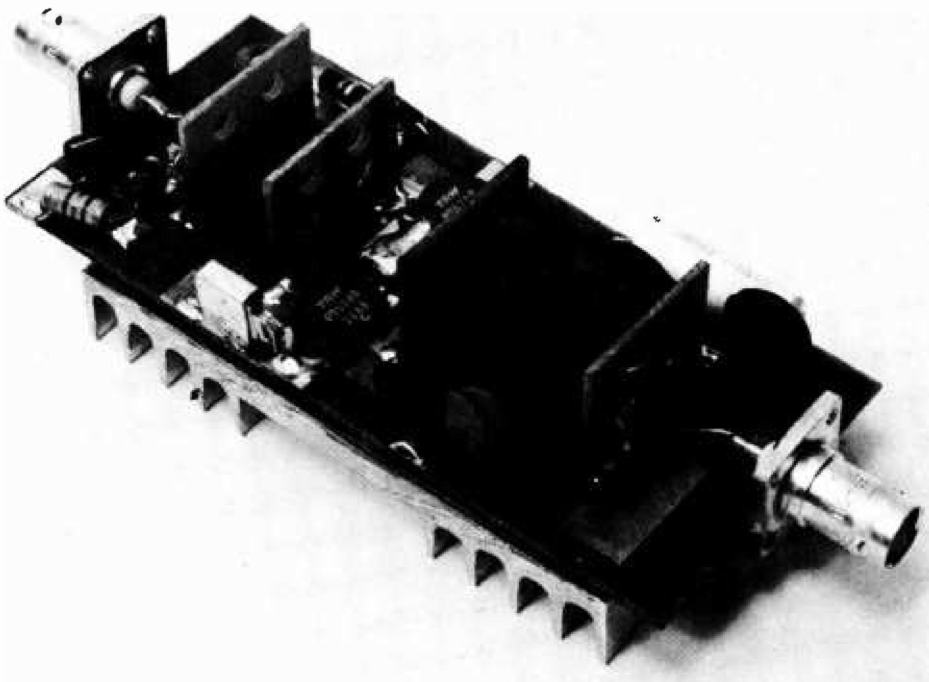


Figure 14

25-WATT PEP OUTPUT SOLID-STATE H-F LINEAR AMPLIFIER

Two TRW type PT 5740 transistors in a broadband circuit provide high performance over the 1.5- to 30-MHz range. The amplifier is built on an etched-circuit board with ferrite-loaded input and output transformers. The input transformer is at the left with the two NPN power transistors at center. The multiple output transformer and feed transformer are at the right of the assembly. Transistors are heat-sinked to an aluminum radiator beneath the circuit board. Ground points atop the board are jumpered to the copper foil on the underside of the printed-circuit board.

monics is of major importance since the harmonics are a function of the ratio of the cutoff frequency to the operating frequency and the selectivity of the output matching network.

Bias Stability—One of the most demanding aspects of solid-state linear amplifier design is the bias network and the associated temperature stability of the transistors. Factors influencing the bias value and network include: (1) Large signal r-f amplifiers generally rectify a portion of the input signal and if the base-emitter resistance is high the amplifier will be biased class AB for small signals, but will self-bias to class-C operation under large signal conditions. This shift in operating point seriously increases intermodulation distortion. The bias source resistance, therefore, must be held to a low value, typically 0.5 to 1 ohm. (2) Intermodulation distortion is usually minimum over a relatively narrow range of resting collector current. The devices used in this amplifier have a large safe operating current range and the resting collector current may be set high enough to achieve the lowest value of intermodulation. (3) Under small-signal conditions transistor dissipation is low and junction temperature is low. However, under conditions of peak power disspation the junction temperature rises. Using a constant-voltage bias source with a device having a negative temperature coefficient for emitter-base voltage change can lead to thermal destruction of the chip unless thermal equilibrium is established by proper transistor design and use of the proper heat sink.

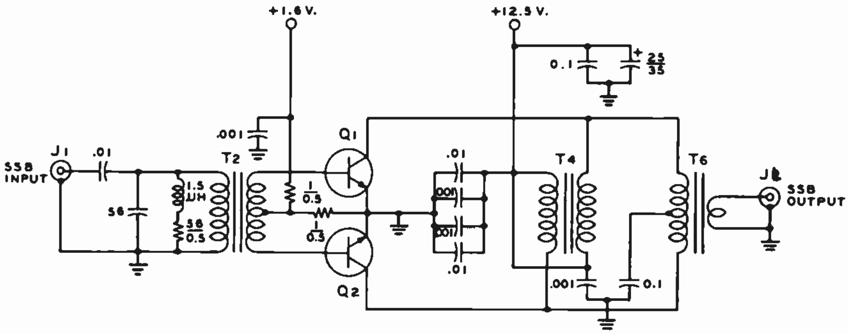


Figure 15

SCHEMATIC, 25-WATT AMPLIFIER

T_2, T_4, T_6—See text and figures 16-18
Q_1, Q_2—TRW type PT 5740 r-f power transistors
Circuit-board material—Glass-filled epoxy board, G-10, 0.060" thick
Heat sink—Wakefield 620 or equivalent
NPO chip capacitors—UNELCO (Underwood Electronics)

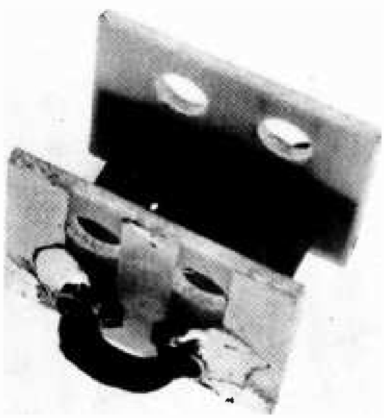


Figure 16

FERRITE-LOADED INPUT TRANSFORMERS T_1 and T_3

Each transformer consists of six ferrite beads in two stacks of three each, epoxied between end plates made of p.c. board material. Each transformer consists of a single-turn winding of two pieces of 0.190" diameter brass rod, each 0.80" long. The pieces are connected together at one end by the foil of one p.c. end board, thus forming a one turn loop. For the 25-watt amplifier, the primary consists of 4 turns #20 e. wire. For the 100-watt amplifier, the primary consists of 5 turns #18 e. wire.

In both of these amplifiers, the design of the PT 5740 power transistor and the accompanying circuitry solves the important bias, temperature and collector current problems.

Impedance Matching—Broadband, ferrite-loaded input and output transformers are used in this amplifier to achieve the required frequency response. The ferrite material used has an initial permeability of 800 which remains above 200 at 30 MHz. Losses in the ferrite material are quite low and ferrite temperature rise is less than 20°C in any transformer at full power output at any frequency in the operating range.

Input transformer T_1 is shown in figures 16 and 17. The unit consists of a very low impedance, split secondary made of two short brass tubes mounted between end plates made of printed-circuit board (foil on one side). One end board serves as the terminal end connections for the tubes and the other acts as a connecting strap and center-tap point between the tubes. Two stacks of three ferrite cores are slid over the tubes which are then soldered in position between the boards and the assembly is epoxied for rigidity. The high impedance wire winding (50 ohms) is threaded in continuous fashion through the tubes.

The d-c feed transformer (T_2) and the output matching transformer (T_3) are mounted side by side between two printed-circuit board end plates, in the manner de-

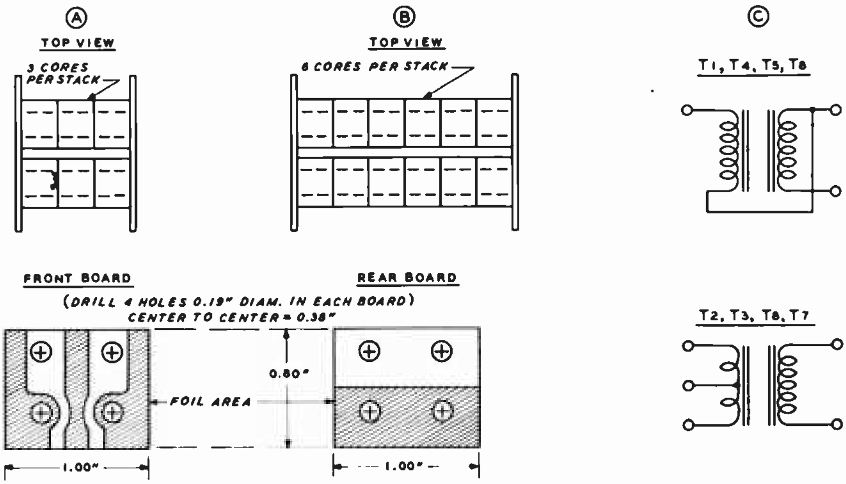


Figure 17

DETAILS OF FERRITE-LOADED TRANSFORMERS

(A)—Top view of input transformer stack of 3 ferrite cores showing assembly and view of front p.c. board. Foil areas provide terminations for brass tubes and connections to main circuit board.
(B)—Top view of transformer assembly of output and feed transformers. Each transformer is made up of two stacks of six cores each. Brass tubes are connected to foil on p.c. board at front and rear.
(C)—Schematic of ferrite transformers. Transformers T_1 and T_8 are identical to T_4 and T_5 but are not mounted on p.c. board frame.

scribed for the input transformer. Each transformer consists of two stacks of six ferrite cores. The assembly is shown in figure 18. The end plates are soldered to the two brass tubes that make up the low impedance winding of transformer T_3 and the whole assembly of ferrites and end plates is epoxied for rigidity. The secondary winding of transformer T_3 and the twisted-pair dual winding of T_2 are then wound back and forth through the ferrite stacks as shown in the photograph. When completed, the transformers are soldered to the copper foil of the circuit board. The low impedance (brass tube) winding ends are soldered directly to the foil of the end boards and the foil to that of the master board.

Amplifier Assembly and Testing—The amplifier is assembled on an etched-circuit board measuring 4½" x 2" and mounted to an aluminum heat sink. The sink ends are trimmed to fit the board.

Upon completion, the amplifier is connected to an exciter, a dummy load, and a metered 12.5-volt source capable of supplying 5 amperes. Base bias is supplied from a well-regulated source and is adjusted for a resting collector current of 150 ma. With

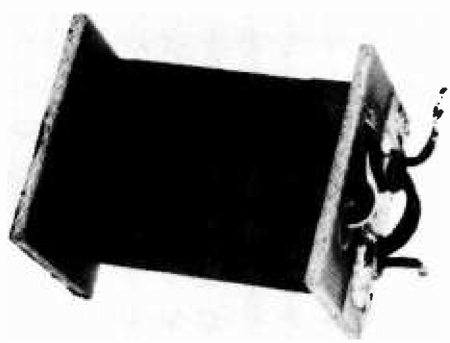


Figure 18

OUTPUT AND FEED TRANSFORMER ASSEMBLY T_4, T_6 and T_5, T_7

The feed transformer is shown atop the output transformer in a four-stack assembly. Twenty-four ferrite cores are used, stacked between two p.c. end plates. The feed transformer has two, one-turn windings made of #18 enamel wire, twisted at 5 twists per inch. Hybrid transformers T_1 and T_8 are identical but are not mounted on p.c. end boards (see figure 19). Transformers T_6 and T_7 are composed of copper-tube windings, similar to T_2 and T_3 except that six stack cores are used and the tubes are 1.375" long. The output winding consists of 4 turns #18 enamel copper wire.

full carrier insertion, the collector current will rise to nearly 5 amperes, and will approximate 2.5 ampere peaks under voice modulation. The third harmonic is —13 db below the fundamental signal level and a suitable harmonic filter should be used before the antenna to reduce this emission. (Note: the unfiltered waveform is essentially a square wave. Output power measurement should be made with a calorimetric power meter or other thermal sensing instrument. Power meters using a diode detector will read low by a factor of 0.785).

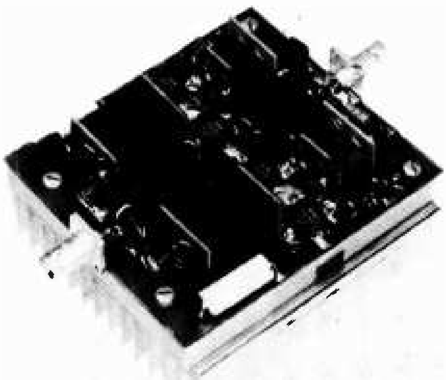


Figure 19

100-WATT PEP OUTPUT SOLID-STATE H-F LINEAR AMPLIFIER

Four TRW type PT5741 transistors are used in a combined, push-pull configuration to cover the 1.5- to 30-MHz range. The four transistors are in line across the middle of the printed-circuit board. At the right are the two input transformers T_2 and T_3 with the hybrid transformer T_1 between them. At the left are the two output-feed transformer assemblies with the hybrid transformer T_8 between them. Ground points atop the board are jumpered to the copper-foil ground plane on the underside of the p.c. board.

The 100-Watt Amplifier The 100-watt PEP output amplifier is shown in figures 19 and 20. The unit requires 3 watts PEP drive power at 30 MHz for full output, having a power gain of about 15 decibels. It may be easily driven by the amplifier described in the previous section.

Amplifier Circuitry—The schematic of the amplifier is shown in figure 20. Two pairs of *TRW* type PT 5741 transistors are operated push-pull and then combined with zero-degree hybrid transformers (T_1 and T_8) which convert the nominal 50-ohm source and load impedances to two 100-ohm ports which are in phase. Any amplitude or phase unbalance causes power to be dissipated in resistors R_1 and R_2. As in the smaller amplifier previously described, an RLC compensation network is placed across the input winding of transformer T_1 to equalize amplifier gain across the operating range.

The collector-feed transformers (T_4, T_5) combine with the output matching transformers (T_6, T_7) to form a modified 180° hybrid combiner. Difference in phase or amplitude that would otherwise exist at the collectors are minimized by allowing the difference current to be bypassed to ground. The resulting output currents in the two transformers are highly balanced and provide good second harmonic rejection. Any minor amplitude or phase unbalance is dissipated in resistor R_2. The port impedance is transformed to an unbalanced value of about 50 ohms by transformer T_8.

Amplifier Assembly and Testing—Data for the various ferrite transformers is given in figures 16, 17, and 18 and the amplifier layout is shown in figure 19. The unit is assembled on an etched-circuit board measuring 4½" x 4" in size. Placement of the four output transistors is critical in that the connection between the collectors and the brass-tubing winding of the output transformers should be extremely short, being composed of the copper foil on the mating circuit boards. Multiple bypass capacitors at the "cold" end of the windings contribute to the low impedance collector path to ground.

Using a 12.5-volt source capable of supplying 16 amperes, the amplifier is adjusted to draw a resting collector current of 0.5 ampere by varying the base bias potential. With full carrier insertion, the collector current will rise to nearly 16 amperes, and will approximate 7-ampere peaks under voice modulation. As in the case of the smaller amplifier, a suitable harmonic filter should be used between the amplifier and the antenna to suppress odd-order harmonics.

(Note additional information on the am-

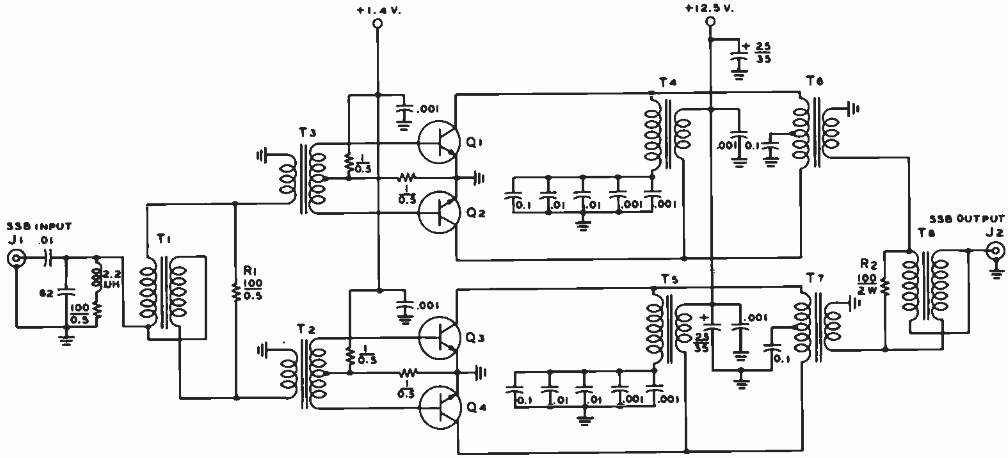


Figure 20

SCHEMATIC, 100-WATT AMPLIFIER

T_1-T_8—See text and figures 16-18
Q_1-Q_4—TRW type PT 5741 r-f power transistors
Circuit-board material—Glass-filled epoxy board, type G-10, 0.060" thick
Heat sink—Wakefield, or equivalent
NPO chip capacitors—UNELCO (Underwood Electronics)

plifiers, a circuit-board template, and data on the TRW transistors may be obtained by requesting Application Notes CT-122-71 and CT-113-71 from the *Semiconductor Division, TRW Inc.*, 14520 Aviation Blvd., Lawndale, CA, 90260.)

22-5 The KW-1 Mark III Linear Amplifier Using the 8875

This compact desktop linear amplifier, designed and built by W3WSQ, is a third generation descendant of the popular 1000-watt PEP amplifier featured in various forms in the last three editions of this Handbook. This new version operates on all amateur bands between 3.5 MHz and 29.7 MHz with good efficiency. The KW-1 amplifier features a single 8875 ceramic high-μ power triode with a 300-watt anode dissipation rating operating in a class-B, cathode-driven configuration. Peak power input is 1000 watts for SSB voice operation, 800 watts for intermittent c-w operation, and 500 watts for continuous RTTY service.

The 8875 anode has a transverse cooler requiring forced-air cooling directed crossways. Maximum dissipation is realized with ducted air to the cooler from a small, low noise blower mounted near the tube.

The 8875 is rated for 250 ma d-c of continuous anode current. In intermittent voice service or keyed c-w operation where short term duty does not exceed 50%, the d-c anode current may be 500 ma during the "on" time. During very short test periods, the tube may be operated at the full 500 ma value but care must be taken to keep the "on" time as short as possible, with sufficient "off" time to allow for tube cooling.

The KW-1, Mark III linear amplifier is small enough to be placed on the operating table next to an SSB transceiver or exciter (figure 21). At 2500 volts anode potential, third-order products are better than —30 decibels below one tone of a two-tone test signal.

The Amplifier Circuit The schematic of the KW-1 amplifier is shown in figure 22. The 8875 is operated in a cathode-driven mode using switchable cathode input transformers for each band (secTable) and a tapped pi-network output circuit. A small degree of r-f feedback is

Figure 21

THE KW-1, MARK III LINEAR AMPLIFIER

This amplifier covers all h-f amateur bands between 80 and 10 meters using an 8875 ceramic, high-μ power triode. A cathode-driven circuit is employed and the amplifier is capable of 1000 watts PEP input for SSB. The unit is housed in an aluminum cabinet and is self-contained except for the power supply. At the top of the panel are the multimeter and plate meter, with the plate tuning control at the center left and the loading control at the right. The plate bandswitch is at center, with the cathode bandswitch at the lower right.

The amplifier cabinet is light gray with a dark gray panel. After the lettering is applied, the panel is sprayed with clear Krylon enamel to protect the lettering.

incorporated in the design by the choice of the 200-pfd grid bypass capacitors on the tube, placing the grid above r-f ground by the small voltage drop created across a divider formed by the plate-grid and grid-ground capacitances.

The power gain of the 8875 is quite high and—even with the r-f feedback—only 25 watts PEP drive power is required. A resistive T-pad is included in the input circuit which raises the drive level to about 100 watts PEP to accommodate some of the higher power SSB exciters. The pad may be omitted if a lower driving level is desired.

Because the grid of the 8875 is not at ground potential, a safety gap (surge arrestor) is placed from grid to ground (SG_1), which will ionize and "fire" when the grid potential exceeds the breakdown voltage of the gap. This protects the grid and cathode of the tube from transient voltages that may develop in the circuit.

Since the 8875 has a separate cathode, the filament may be isolated from the input circuit. It is not necessary in the h-f region, but a special trifilar filament choke is used to permit the cathode to be returned to d-c ground, as shown in the schematic.

Resting plate current of the 8875 is set by the *Adjust Bias* potentiometer. A built-in bias supply also provides control voltage for the transmit relay, RY_1. A series connected diode in the control circuit serves to keep the relay transient voltage from upsetting the bias circuit. A separate filament transformer is used for the 8875 and a primary potentiometer allows the voltage to be set at 6.3 volts at the socket of the tube.

The control circuit is designed to prevent application of r-f drive without plate volt-

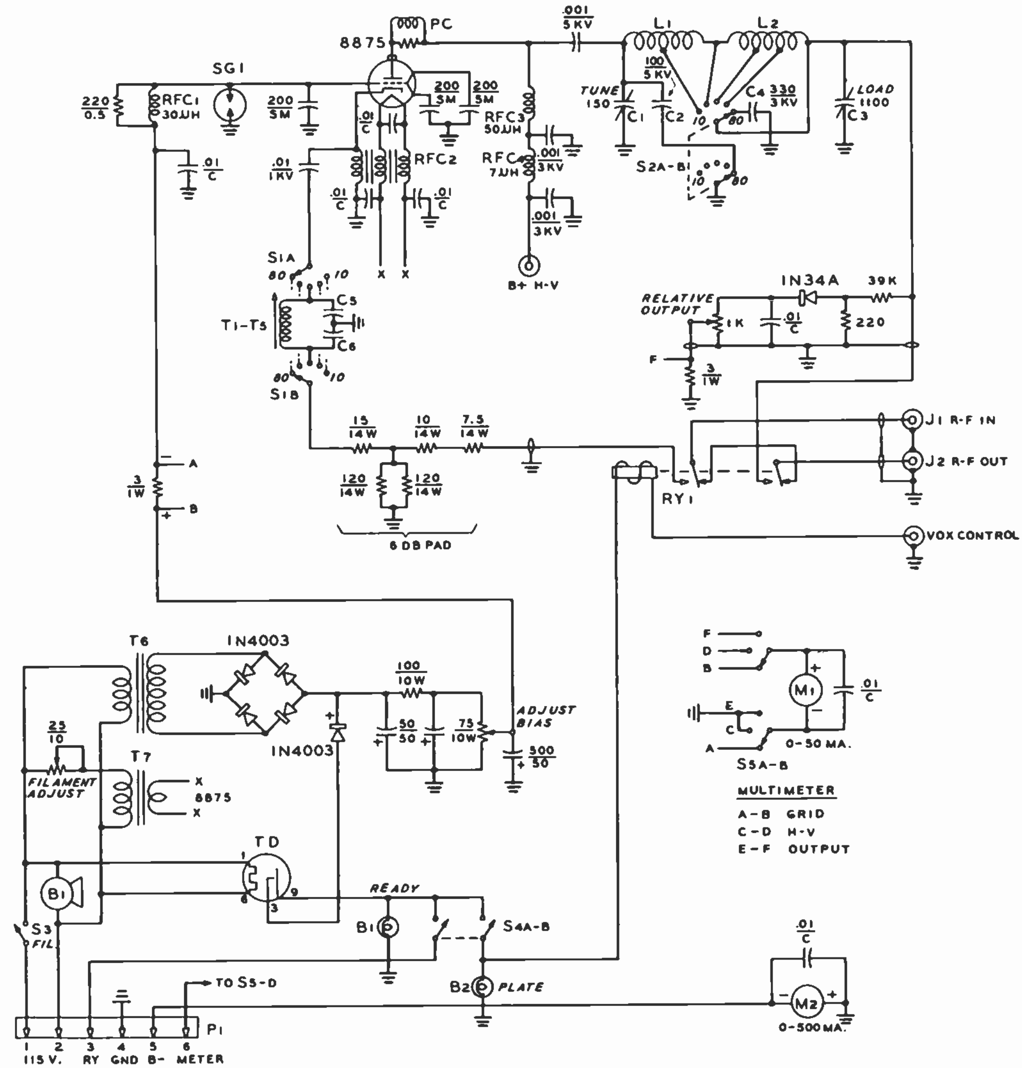


Figure 22

SCHEMATIC, KW-1 LINEAR AMPLIFIER

C_1—150 pf, 3 kV. Johnson 154-8 (26-96)
C_2—Centralab type 858S (21-109)
C_3—1100 pf, (26-97, 1700 pf may be used)
L_1—10-15 meter coil. 9 turns #10 wire, 1⅛" inside diameter, 2¼" long. 10-meter tap 4 turns from plate end (40-596)
L_2—20-40-80 meter coil. 1-13/16" diameter form, 4" long. Wound with #16 wire at 9 turns per inch. 20-meter section, 4 turns; 40-meter section, 7 turns; 80-meter section, 10 turns. Space between sections is ¼".
RFC_1—30 μH (45-18)
RFC_2—Trifilar choke. 20 turns #14 e. on ½" diameter ferrite core, 2¼" long (Indiana General). Interwind with third winding of #22 insulated wire. (45-60 with interwound winding of #22 insulated wire)
RFC_3—50 μH (45-61), or Ohmite Z-14
SG_1—Surge arrestor, 230-volt peak. Signalite CG-230L, Siemens B1-A230 or Reliable Electric SR-P17170
T_6—24 volts, 1 ampere
T_7—6.3 volts, 5 amperes
TD—Time delay relay, 60 seconds. Amperite115C60T
B_1—Dayton 2C782. 3160 rpm, 2¼" wheel
PC—4½ turns #16 around 50-ohm, 1-watt composition resistor
Note: Heath part numbers given in parenthesis.

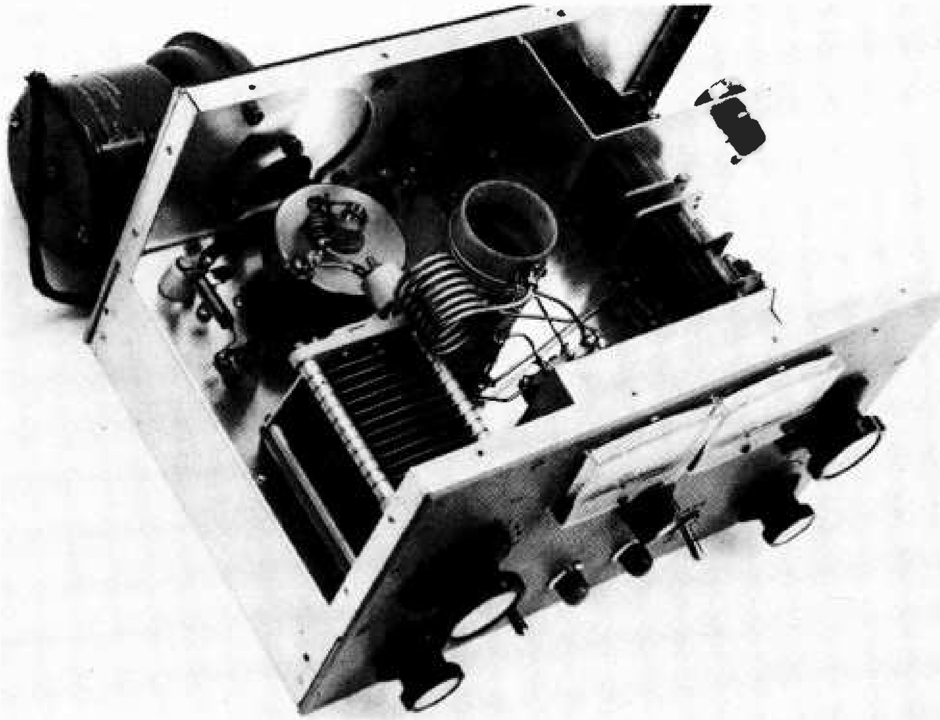


Figure 23

INSIDE VIEW OF THE KW-1 AMPLIFIER

The 8875 tube is at the left with the blower positioned to force air across the anode cooler. Six holes are drilled in the chassis under the 8875 to allow air to escape from under the chassis by convection, thus cooling the tube base. The 80-40-20 meter plate coil is bolted vertically to the chassis at center with the high-frequency air-wound coil supported between the tuning capacitor and the bandswitch. The bias-control potentiometer is mounted on the shield plate behind the loading capacitor.

Cathode Transformers, T_1-T_5 Wound on 3/8" diameter forms, slug-tuned.	
T_1-(80 Meters)	24 turns #16e. C6 (omitted) C5 = 470pf.
T_2-(40 Meters)	17 turns #16e C6 = 510pf. C5 = 310pf.
T_3-(20 Meters)	9 1/2 turns #18e. C6 = 360pf C5 = 200pf
T_4-(15 Meters)	4 1/2 turns #18e C6 (omitted). C5 = 75pf
T_5-(10 Meters)	3 1/2 turns #18e C6 (omitted). C5 = 68pf

age and a 60-second time delay unit (TD) prevents plate voltage from being applied before the cathode of the tube reaches operating potential.

A single 50-μa d-c meter is used to monitor grid current, high voltage, and relative power output. Grid current is read across a 5-percent resistor in the grid-bias return lead. Plate voltage is read indirectly across the last resistor in the power-supply bleeder string. The full-scale meter readings are 50 ma and 5000 volts for grid current and plate voltage respectively.

The KW-1 Mark III amplifier plate circuit is a conventional pi-network arrangement with additional plate tuning capacitance (C_2) added to the circuit on the 80-meter band by means of switch S_2. The plate coil is divided into two sections; the smaller, air-wound coil being used for 10 and 15 meters and the larger coil for 20, 40, and 80 meters. The network is designed to match a nominal 50-ohm load having an SWR of 3 or less. An additional loading capacitance (C_4) is automatically switched into the circuit for 80-meter operation.

Amplifier Construction The amplifier is built on an aluminum chassis measuring 12″ × 8″ × 2½″. Enclosure height is 7″. Front and back panels of the box are cut from ⅛″ aluminum and the U-shaped cover is made of thin aluminum sheet. A 6″ × 3″ perforated aluminum plate is riveted in a cutout in the top of the cover to allow cooling air to escape from the enclosure. Angle stock is bolted around the top and side edges of the front and rear panels as a mounting surface for the cover.

The two meters are enclosed in a cut down minibox which serves as an r-f shield and an L-shaped bracket shields the filament transformer and antenna relay from the amplifier output circuitry.

Placement of the major components may be seen in figure 23. The 8875 is positioned carefully in front of the orifice of the blower and about one inch away. Six quarter-inch holes are drilled in the chassis around the tube socket to allow under-chassis air to be drawn up by convection to cool the base of the tube.

The cathode tuned circuits (T_1, T_5) and the time delay relay are mounted on an under-chassis shield plate, as seen in figure 24. The resistors making up the input attenuator are mounted immediately to the rear of this plate on two phenolic terminal strips.

Many of the components used in this amplifier are replacement parts for the *Heath SB-200* linear amplifier and were ordered directly from the *Service Department, Heath Co.*, Benton Harbor, Michigan 49022 under the identification number given in the parts list. Other similar components will work as well as the particular parts used in this amplifier.

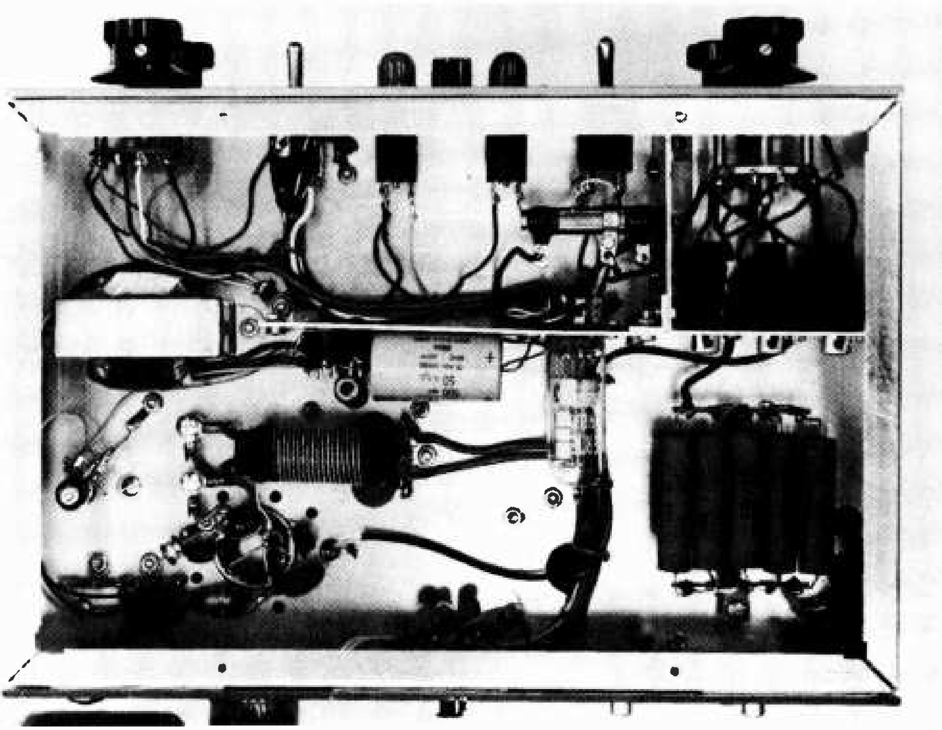


Figure 24

UNDER-CHASSIS VIEW OF AMPLIFIER

The tuned cathode circuits are in the partitioned area at the upper right with the input attenuator pad directly behind it. At center are the glass encapsulated time-delay relay and the bias power supply. The 8875 socket and filament choke are at lower left.

Figure 25

CLOSEUP OF 8875 SOCKET WIRING

To the right of the socket is the small glass-encapsulated spark gap connected between grid terminals and the chassis. The trifilar filament choke is in the foreground.

Transmitter Power Supply The schematic of the KW-1 Mark III power supply is shown in figure 26. A multi-conductor cable connects the supply to the amplifier along with the high voltage lead, which is run in RG-59/U coax. The *filament* switch on the panel of the amplifier controls the primary power circuit and the time delay relay and *plate* switch activate the *transmit* relay control circuitry. The power supply is energized by grounding the VOX *control* terminal on the rear of the amplifier chassis. The power supply provides approximately 2500 volts under no-signal conditions and 2100 volts at a peak plate current of 450 milliamperes. The dynamic characteristics of the power supply allow the amplifier to develop about 20% greater peak SSB envelope power for a given level of c-w input. The power supply utilizes a voltage doubler circuit and incorporates high voltage metering. Supply voltage is checked with a meter of known accuracy and the *meter calibrate* potentiometer is adjusted to provide the same reading on the panel meter of the amplifier.

Amplifier Tuning and Adjustment Wiring should be completely checked before power is applied. The approximate settings of the plate tank circuit should be determined for each band with the aid of a grid-dip oscillator. The slug cores of the cathode transformers are adjusted to mid-band resonance for each position of the bandchange switch.

The *adjust bias* potentiometer is set for maximum grid bias and filament voltage is applied to the 8875 and checked at the socket. *Caution:* The cabinet cover should

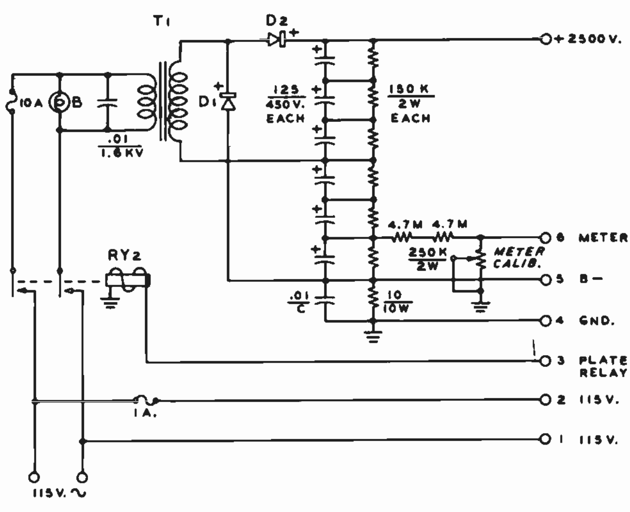


Figure 26

POWER SUPPLY, KW-1 AMPLIFIER

T_1—117-volt primary. 820-volt, 0.5-ampere secondary (54-151)
D_1, D_2—Each leg: Five 1N4005 diodes. Place .01 µf, 1.6-kV disc capacitor and 100K, 1-watt resistor across each divide
RY_2—24-volt d-c coil, DPDT

now be bolted in place as high voltage points are exposed in the amplifier.

An exciter and dummy load are attached to the amplifier and high voltage applied. The VOX circuit should be energized by grounding the VOX terminal. The amplifier is now ready to be tuned up. After the time-delay relay has closed, the bias potentiometer is adjusted for a resting plate current of about 25 ma. A small amount of carrier is applied to the amplifier as a tuning signal until about 150 ma of plate current is indicated. The amplifier is tuned to resonance and peaked for maximum reading on the output meter. Once resonance is established, the tuning and loading controls are adjusted for maximum output as the driving signal is gradually increased. The loading capacitor should be near full capacitance for 80 and 40 meters, about 60 percent meshed for 20 meters and slightly less for 15 and 10 meters. Maximum carrier signal plate current is 450 ma and corresponding grid current is 30 ma.

The last step is to peak the input transformers for maximum grid current on each band, retarding the excitation so as not to overdrive the amplifier.

Carrier is now removed and voice modulation applied. A maximum of 1000 watts PEP input is achieved with peak voice current of about 210 milliamperes. For c-w operation, carrier insertion is used and the amplifier is loaded to a plate current of 400 ma.

22-6 The 500Z 2-kW PEP Linear Amplifier for 10 thru 80 Meters

Two 3-400Z or 3-500Z high-μ triode tubes form the basis for this compact, multiband, high-power desk-top linear amplifier. Heavy-duty design combined with rugged components permit the amplifier to be run at full legal power level for SSB or c-w service. Measuring only 16″ × 8″ × 13″ deep the amplifier is small enough to be placed on the operating table adjacent to the SSB transceiver or exciter.

Auxiliary circuitry permits the exciter to bypass the amplifier, if desired, for low-power operation, and the unit incorporates automatic load control (ALC) for optimum voice efficiency in SSB operation. At maximum input level, the third-order intermodulation products are better than −33 decibels below one tone of a two-tone test signal, attesting to the high degree of linearity attained without the use of auxiliary feedback circuitry. Peak drive power is of the order of 90 watts, and the amplifier may be driven by any SSB exciter capable of this power output.

The Amplifier Circuit This 2000-watt PEP linear amplifier employs two zero-bias triode tubes connected in cathode-driven, grounded-grid configuration. A pi-network output circuit is used, capable of matching 50-ohm or 70-ohm coaxial antenna circuits. For improved linearity and ease of drive, a simple tuned-cathode input circuit is ganged to the pi-network amplifier bandswitch. Separate grid and plate meters are used and a variable ALC circuit is provided for connection to the exciter. The amplifier is designed for operation over a plate voltage range of 2000 to 2700 volts and a plate potential of 2500 volts is recommended.

Amplifier Circuitry—The schematic of the linear amplifier is shown in figure 28. Two 3-400Z or 3-500Z tubes are connected in parallel. Each of the three grid pins of the tube sockets is grounded, and the driving signal is applied to the filament circuit of the tubes, which is isolated from ground by a bifilar r-f choke. Neutralization is not required because of the excellent circuit isolation provided by the tubes and by the circuit layout.

The driving signal is fed in a balanced manner to the filament circuit of the two tubes. Mica capacitors suitable for r-f service are used to properly distribute the driving signal to the tuned-cathode circuit and the filaments of the tubes. Ceramic-disc capacitors are not recommended for use in this portion of the circuit because the peak r-f current under full amplifier input may be as high as 6 amperes or so. The tuned-cathode circuits (L_1-L_5) are fixed-tuned to the center of each amateur band and may be forgotten.

The Plate Circuit—Plate voltage is applied to the tubes through a heavy duty r-f choke bypassed at the B-plus end by a low-

Figure 27

TWO-KILOWATT PEP INPUT IS FEATURED IN THIS COMPACT AMPLIFIER USING ZERO BIAS TRIODES

This desk-top amplifier allows maximum PEP input on all high-frequency amateur bands. Two zero-bias 3-500Z triodes are used in a cathode-driven, grounded-grid circuit. ALC is included as well as a high-efficiency, low-noise fan cooling system.

The amplifier is housed in a perforated aluminum case and is entirely self-contained, except for the power supply. At the top of the front panel are the grid and plate meters. The antenna loading control (C_7) is at the left and the plate tuning control (C_6) at the right. Both capacitors are driven through small precision planetary vernier drives. The bandswitch is centered at the lower portion of the panel.

The amplifier cabinet is gray, with light-green panel. After the lettering is applied, the panel is sprayed with clear Krylon enamel to protect the lettering. The unit is elevated above the desk top on rubber feet to permit good movement of air about the under-chassis area.

inductance, ceramic capacitor. In addition, the high voltage passes through a length of shielded cable to the high-voltage connector at the back of the chassis, and is further bypassed to ground at that point. A single .001-μfd, 5-kV ceramic capacitor is used for the high-voltage plate-blocking capacitor and is mounted atop the plate r-f choke. The pi-network coil is divided into two parts for highest efficiency and ease in assembly. The first portion covers 10, 15, and 20 meters, and an additional section is added to the first to cover operation on 40 and 80 meters. Both coils are homemade and air wound at a minimum cost. The bandswitch is a *Radio Switch Corp.* high-voltage ceramic-insulated unit mounted to the front panel of the amplifier.

A typical circuit Q of 10 was chosen to permit a reasonable value of capacitance to be used at 80 meters and the number of turns in the plate coils was adjusted to maintain this value of Q up through 15 meters. At 10 meters, the Q rises to about 18 and is largely determined by the minimum circuit capacitance achieved at this frequency. The pi-network output capacitor is a three-section, ceramic insulated 1100-pf unit. It is sufficiently large for proper operation of the amplifier on all bands through 40 meters. For 80-meter operation, an additional 500-pf heavy duty mica capacitor is switched in

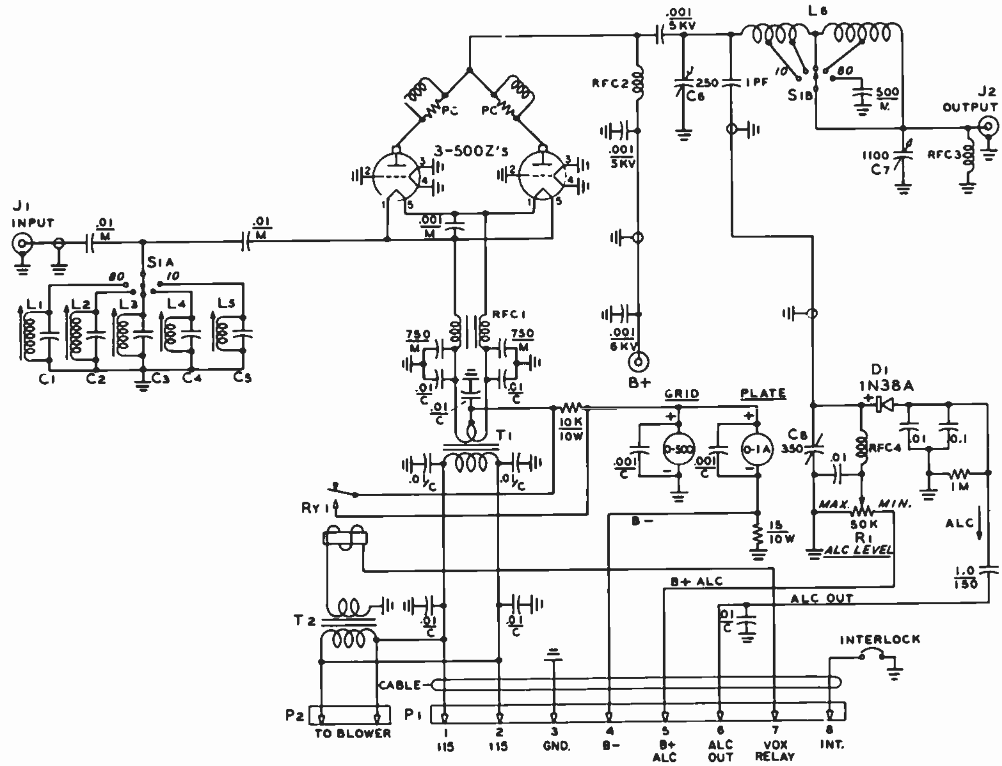


Figure 28

SCHEMATIC OF 500Z LINEAR AMPLIFIER

C_1—200 pf, 1-kV mica.
C_2, C_3—470 pf, 1-kV mica.
C_4, C_5—1000 pf, 1-kV mica.
C_6—250 pf, 3-kV, .075" spacing. Johnson 154-9
C_7—1100 pf, 3-section. Jackson Bros. LE3-4595-380 (M. Swedgal, 258 Broadway, N. Y. 10007)
C_8—350-pf mica compression capacitor
L_1, L_2—(0.15 μH) 4 turns #16 e. on ½" diam., form, powdered-iron core. National XR-50 or equiv. form. (Slug removed from both coils)
L_3, L_4—(0.31 μH) 6 turns #14 e. on National XR-50 or equiv. form. (Slug removed from L_3).
L_5—(1.3 μH) 13 turns # 18 e. on National XR-50 or equiv. form
L_6—(10-15-20 meters) 10½ turns #8 wire, 2" diam., 3½" long. 10-meter tap is 5¼ turns from plate end; 15-meter tap, 7¼ turns. (40-80 meters) 16 turns # 10 wire, 2½" diameter, 4" long. 40-meter tap is 8 turns from "hot" end. Coil wound on lucite plate with edges grooved for proper spacing of turns
RFC_1—Bifilar winding. Each coil is 14 turns #10 e., on ferrite core, 5" long ½" diam. (Indiana General CF-503) (Newark Electronics catalog #59F-1521) Manufactured choke: Jennings Industries J1-FC32
RFC_2—(Approx. 60 μH) 90 turns #26 e., spacewound wire diameter, 3⅜" long, ¾" diam. on ceramic or Teflon form. Series resonant at 26 MHz. Jennings Industries J1-PC800
RFC_3, RFC_4—2.5 mH. National R-100
T_1—5 volts, 30 amp. Stancor P-6492
T_2—6.3 volts, 1 amp. Stancor P-8389
RY_1—Spst, 6.3-volt coil
Fan—Ripley SK-4125 or equivalent.
Meters—Calectro
PC_1—Three 100-ohm, 2-watt resistors in parallel. 3½ turns #18 spacewound about one resistor
S_{1A}—Single-pole, 11-position ceramic switch, 30° index. Centralab PA-6001.
S_{1B}—Single-pole, 11-position ceramic switch, 30° index. Radio Switch Corp. Model 86-A
Sockets—Johnson 122-275-1
Dials—General Radio with Jackson Bros. 4511-DAF Planetary Ball Drive Unit

parallel with the variable unit to provide good operation into low-impedance antenna systems commonly found on this band. The capacitor is connected to the unused 80-meter position of the bandswitch.

The instantaneous r-f plate voltage is sampled by a capacitive voltage divider and applied to a reverse-biased rectifier (D_1). Bias level is set by means of an adjustable potentiometer (*ALC Level*). When the r-f voltage exceeds the bias level, an ALC pulse is applied to the ALC control circuit of the

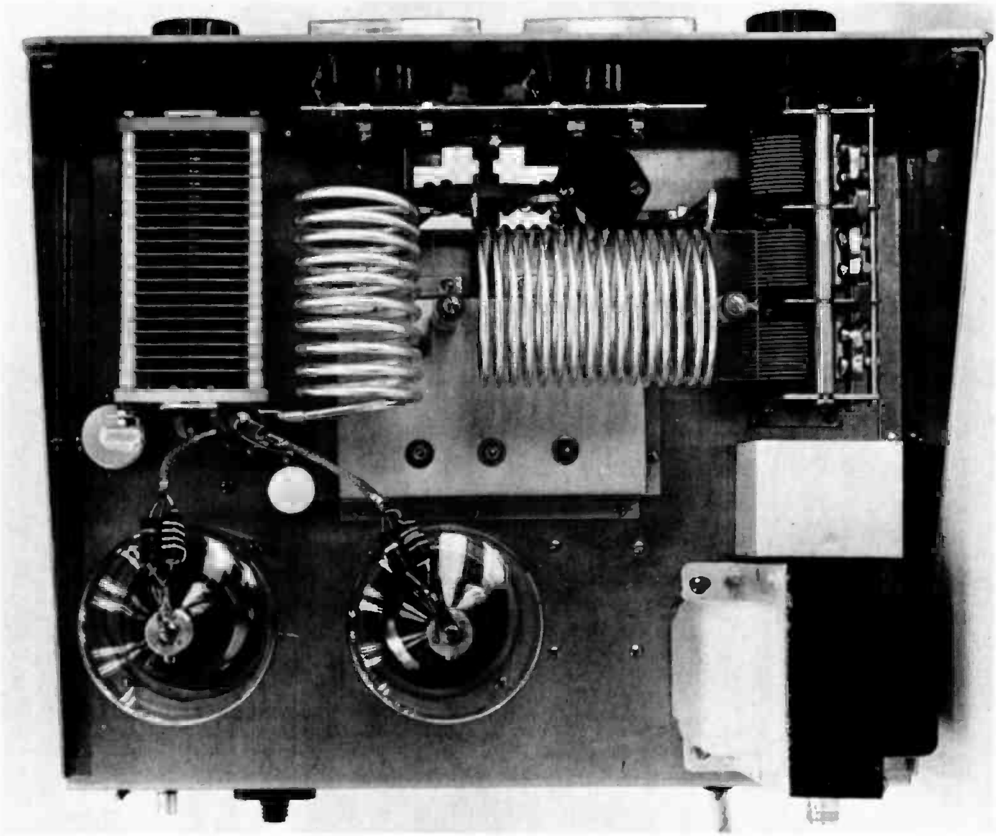


Figure 29

TOP VIEW OF LINEAR AMPLIFIER

Two 3-500Z tubes are placed at the rear corner of the amplifier chassis. The spacing of sockets and blower are shown in figure 32. The plate loading and tuning capacitors are mounted to each side of the pi-network coil assembly. The three stator sections of the output capacitor are connected in parallel by short lengths of copper strap. Directly below the plate coils is the aluminum box containing the cathode tuned circuits, with the adjustment slugs of the coils projecting through the top of the box.

The 500-pf auxiliary 80-meter loading capacitor is placed above the bandswitch, directly in front of the 80-40 meter coil. At the left, the 1-pf coupling capacitor is attached directly to the rotor of the main tuning capacitor (see figure 31).

The filament transformer for the two 3-500Z tubes is at the rear, right corner of the chassis. The portion of the transformer facing the tubes is painted white to reflect the infra-red radiation from the tubes, which run a cherry-red color at full plate dissipation level. The cooling fan is mounted to the rear of the cabinet, and is not seen in this view.

exciter. The r-f level applied to the control circuit is set by adjustment of capacitor C_8 and the voltage is determined by the ratio of this capacitor to the 1-pf capacitor coupling the ALC circuit to the plates of the amplifier tubes. At a plate potential of 2500 or so, the nominal value of r-f plate voltage swing is about 1800 volts. If the ratio of the capacitive divider is 1:200, then about 90 volts of peak pulse is applied to the diode. Under normal operation, the diode is biased to about +30 volts and ALC pulses of about one-half this value are normal. Thus, the r-f voltage at the diode should be not more than 45 volts or so, calling for a capacitance ratio of about 1:300. This ratio is well within the range of the mica compression capacitor used for C_8.

The Metering Circuit—It is dangerous practice to place the plate-current meter in the B-plus lead to the amplifier unless the meter is suitably insulated from ground

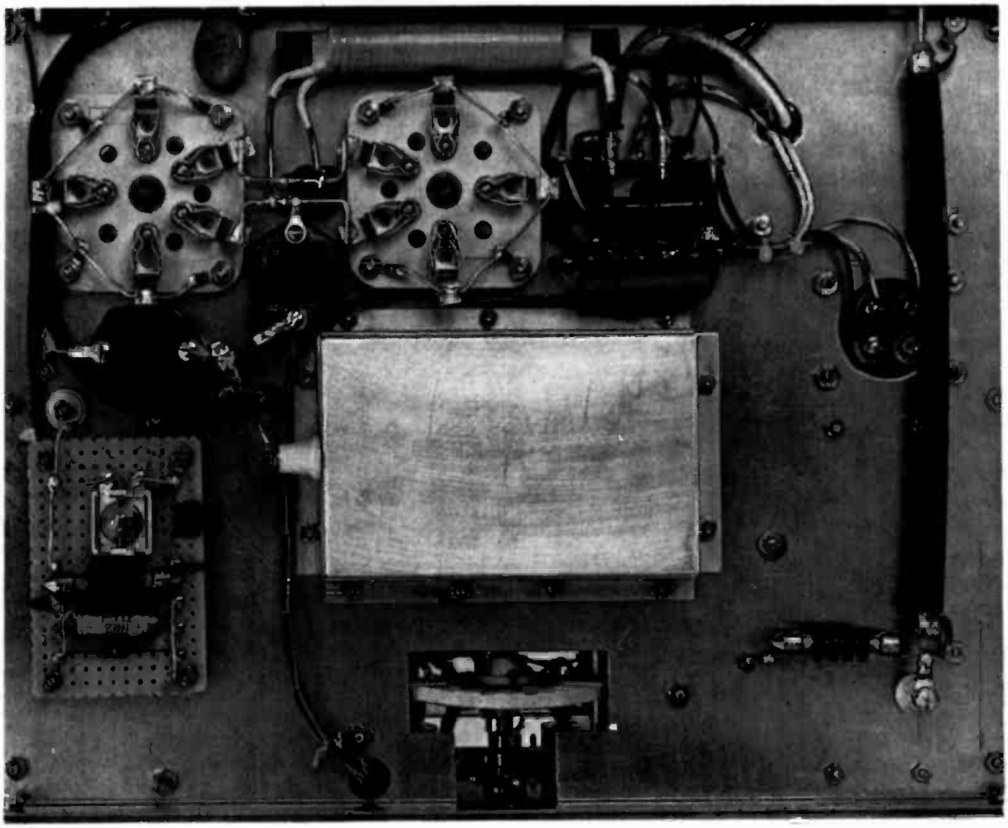


Figure 30

UNDER-CHASSIS VIEW OF AMPLIFIER

The cathode circuit box is at the center of the chassis, with the connecting load passing through a feedthrough insulator at the left. The shaft of switch S_{1A} passes through the wall of the upper section of the box, only about 1/16-inch above the level of the chassis and is joined to the plate bandswitch (S_{1B}) with a brass coupling.

The three grid pins of the tube sockets are grounded to the mounting bolts. The sockets are lowered below the chassis by means of spacers to permit cooling air to flow about the base of the tubes. The two .01-μfd mica coupling capacitors are placed adjacent to the left-hand tube socket, with the ferrite-core filament choke running parallel to the rear of the chassis. Directly to the right of the sockets are placed two phenolic terminal strips which support the filament wiring, the 10K VOX resistor and the 15-ohm meter safety resistor. The bypass capacitors for the "cold" end of the filament choke are also located on one terminal strip.

At the right end of the chassis is a small phenolic board that holds mica compression capacitor C_1 and the components associated with the ALC circuit. The ALC level potentiometer is a small ¾-inch diameter control mounted on the rear lip of the chassis, adjacent to input receptacle J_1. To the right of J_1 is the high-voltage connector, with the .001-μfd, 6-kV disc capacitor mounted behind it. The antenna output circuitry is at the right end of the chassis. The connection from the plate-loading capacitor passes through a ceramic feedthrough insulator near the panel, and the connection to the coaxial receptacle (J_2) at the rear of the amplifier is made via a short length of RG-8/U coaxial line. The outer braid of the line is grounded to the the chassis at each end.

and isolated behind a protective panel so that the operator cannot accidentally receive a shock from the zero-adjustment fixture. If the meter is placed in the cathode return circuit, it will read the cathode current which is the sum of the grid and plate currents. A better idea is to place the plate meter in the B-minus lead between the cathode return circuit and the negative terminal of the power supply. The negative of

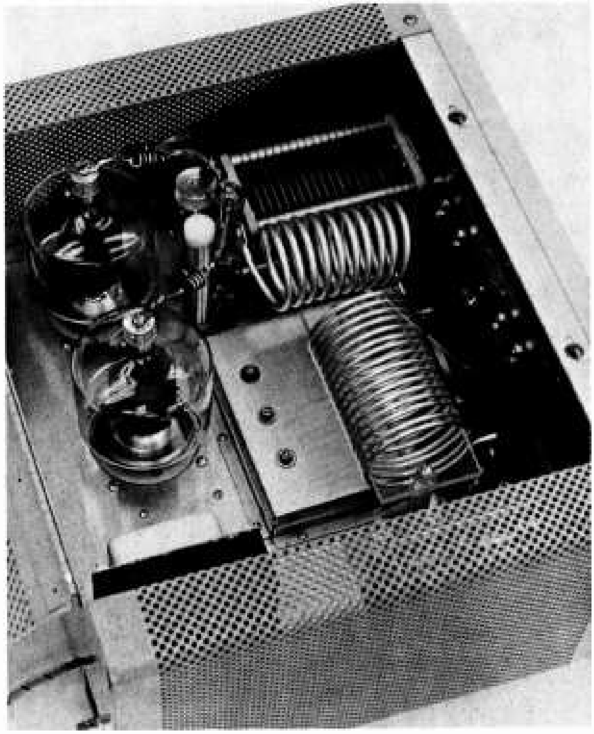


Figure 31

OBLIQUE VIEW OF PLATE CIRCUIT

The Eimac HR-6 anode connectors are used on the 3-500Z tubes, with the parasitic suppressor mounted close to the connector. The plate leads are made of lengths of flexible copper braid. Both leads terminate at the plate-blocking capacitor which is mounted to a small bracket bolted to the stator terminal of the plate-tuning capacitor. At the far side of the tuning capacitor is the 1-pf ALC coupling capacitor, made of two 1-inch diameter copper discs, spaced about ¼-inch apart. The upper disc is affixed to the stator terminal of the capacitor and the lower disc is supported by the feedthrough insulator mounted directly beneath it on the chassis deck.

the supply must thus be left "floating" above ground, or the meter will not read properly (figure 28). A protective resistor is placed across the meter circuit to ensure that the negative side of the power supply remains close to ground potential. A separate ground lead is then run between the chassis ground of the amplifier and that of the supply. Grid current is measured between grid and cathode return as shown in the simplified schematic, with the grid pins of the tubes directly connected to chassis ground.

The Cooling System—It is necessary to provide cooling air about the plate seal and filament seals of either the 3-400Z or 3-500Z tubes. Sufficient air is required to maintain the plate seal at a temperature below 225°C and the filament seals at a temperature below 200°C. Common practice calls for the use of special air-system sockets and chimneys, in conjunction with a centrifugal blower to maintain air flow requirements to meet these temperature limitations. Considerable difficulty with conventional cooling techniques has arisen, caused by the noise created by the blower motor and the movement of air through the cooling system. Extensive tests have shown that for c-w and SSB operation at the legal power limits (1-kW c-w input and 2-kW PEP voice input on SSB) either the 3-400Z or 3-500Z may be adequately cooled by a lateral air blast blown against the tube by a small rotary fan, properly spaced from the tube. A drawing of such an installation is shown in figure 32.

The Johnson 122-275-1 ceramic tube socket is used, which permits a minimum amount of lateral pressure to be exerted on the glass base of the tube. The socket is mounted below the chassis deck about 1/16" to provide an air path around the base of the tube through which under-chassis air is drawn by convection. The rotary fan is mounted between the tubes, in line with the center of the glass envelope and blows cooling air across the envelope and plate caps. Under these conditions, maximum plate dissipation of about 350 watts per tube is achieved for the 3-400Z and 450 watts per tube for the 3-500Z. While maximum dissipation rating is not achieved with either tube, the allowable dissipation is suf-

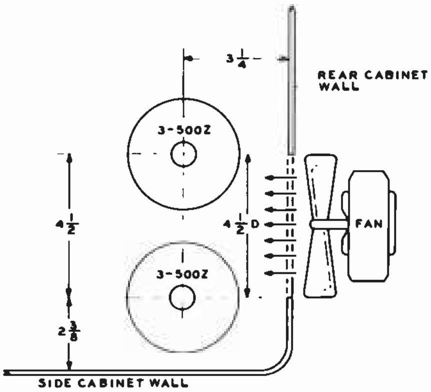


Figure 32

AIR-SYSTEM LAYOUT

The Ripley fan (Ripley Co., Inc., Middletown, Conn.) is bolted to the rear of the cabinet behind a 4½-inch diameter hole, covered with ¼-inch mesh wire screen. The air blast passes across the tube envelopes and the warm air is exhausted out the perforated top of the amplifier cabinet. The tube sockets are located with respect to the fan to permit maximum cooling air to envelop the tubes.

ficiently high so that the maximum amateur power input may be run in either case with adequate safety factor. If it is desired to operate the amplifier under steady-state conditions (RTTY, for example), the power input will have to be reduced to about 850 watts in the case of the 3-500Z's or 750 watts for the 3-400Z's. The alternative is to install a forced-air cooling system to boost the plate dissipation capability to the maximum limit specified in the instruction sheet for the tube type in question. The air cooling system shown, however, is entirely adequate for c-w and SSB operation under normal operating conditions for extended periods of time.

The perforated metal cabinet provides maximum ventilation and, when the lid is closed, provides good r-f inclosure. In order to permit the air to be drawn into the bottom of the amplifier chassis, rubber "feet" are placed at each corner of the cabinet, raising it about one inch above the surface on which it sits. The top surface of the cabinet should be kept clear to permit the heat to freely escape from the amplifier when it is in use.

Amplifier Construction The over-all dimensions of the perforated, wraparound cabinet housing the amplifier are 16″ wide, 8″ high, and 13″ deep. The amplifier is built on a shallow chassis bent from a single sheet of aluminum and measures 15¼″ wide, 12½″ deep and has a 1″ lip at the rear. Clearance under the chassis is 1¼″ to the bottom of the cabinet. An oblique view of the chassis and cabinet, including the placement of the major components is shown in figure 31. The cooling fan is mounted to the rear of the cabinet and forces air against the two transmitting tubes through a 4″ diameter hole cut in the rear panel of the cabinet. The hole is covered with a piece of wire mesh having ¼″ squares.

Placement of the major components may be seen in the photographs. Because of the small depth of the chassis, placement of the bandswitch and tuned cathode assembly is critical. The various cathode tuned circuits and bandchange switch are mounted in an inclosed box placed near the center of the chassis, in line with the main band change switch. The cathode inclosure box is made up of two small aluminum chassis (5″ × 3½″ × 1″) placed back to back, one atop and the other underneath the chassis. The flanges of the chassis are cut off, and substitute flanges are attached to the outside of the chassis lips, permitting the two units to be bolted together, as shown. The various coils and bandswitch are mounted to the top chassis box, in line with the main switch and connected to it with a shaft coupler. The cathode coils and capacitors are assembled and mounted in a vertical position within the box. The cathode tank-coil assembly may be wired and the tuned circuits grid-dipped to the center of each amateur band before the chassis box is bolted to the corresponding cutout in the chassis.

The pi-network coil assembly is seen in the top view photographs. The 10-15-20 meter coil is wound of No. 8 solid copper wire. Ordinary plastic-covered house wire is used, the plastic coating stripped off before the coil is close wound on a suitable form. Once the winding is completed, the coil is spaced and the taps are soldered in place. Thin, ⅛″ wide copper strap is used for the tap leads. Each lead is pretinned at the end and wrapped around the proper coil turn and soldered in place with a large iron. A good connection is important at this point as the r-f current flowing through the joint is high. Once the coil is cut to size, and the tap leads soldered in place, the coil end connections are trimmed to length and adjusted to the proper position. The coil lead to the tuning capacitor terminates in a copper soldering lug and the opposite end is flattened in a vise to make a glove fit with the proper 20-meter tap point on the bandswitch. Once all leads are properly trimmed, the coil is removed and silver plated.

The 40-80 meter coil is wound and tapped in the same fashion. Once completed, it is threaded on a strip of lucite or plastic material that has been grooved along both edges to fit the spaced winding of the coil. The grooves may be easily cut with a small triangular file. The lucite plate is supported by two plastic posts, cut to size and mounted to the chassis behind the bandswitch.

The plate parasitic suppressors for each tube are made of three composition resistors wired in parallel, with a small inductor wound around one resistor. The suppressors are placed immediately adjacent to the anode connectors of each tube, and flexible leads made of copper braid are run from the suppressors to a common terminal of the plate coupling capacitor mounted atop the plate r-f choke.

The placement of the major components beneath the chassis is shown in figure 30. A T-shaped opening is cut in the forward area of the chassis to clear the plate bandswitch, and an opening is cut in the center of the chassis for the cathode tank assembly. The tube sockets are mounted beneath the chassis by 6-32 hardware, with several washers placed on each mounting bolt beneath the chassis to lower the socket about $\frac{1}{16}$ inch, providing additional air passage around the base seal of the tube. The grid pins are grounded to the adjacent socket bolts. The large filament choke is mounted from a phenolic terminal strip to the parallel-connected filament pins of the tubes. The mica coupling capacitors are placed in close position to the filament wiring and the ceramic feedthrough insulator mounted in the side wall of the input coil compartment.

At the side of the under-chassis area a small perforated circuit board supports the various components of the ALC circuit and the connecting lead to the 1-pf air capacitor mounted on the main tuning capacitor passes through a ceramic feedthrough insulator in the chassis deck.

The connection from the pi-network output capacitor to the coaxial receptacle mounted on the rear lip of the chassis is made via a short length of 50-ohm coaxial cable, the outer shield of the cable being grounded at both ends to nearby chassis points.

The filament transformer is mounted atop the chassis in a rear corner as seen in the photographs. The bottom area of the transformer is cleaned of paint so that the end bells make a good ground connection to the chassis to partially shield the windings from the r-f field atop the chassis. The end bell of the transformer nearest the tubes is painted white to reflect the infrared radiation emitted from the tubes, permitting the transformer to run much cooler than otherwise would be the case if the end bell was left black. The remainder of the transformer is left black so as to radiate the heat generated within the transformer.

The VOX relay and auxiliary transformer are mounted in a small shield box placed in front of the filament transformer. Sufficient room exists in this area so the box may be enlarged to also hold a rectifier and filter capacitor should it be desired to substitute a d-c relay for the a-c unit specified.

A shield plate measuring 6" × 2" is affixed to the rear of the meters to shield the movements from the intense r-f field surrounding the plate coils. The shield is held in position by the meter studs, each stud passing through a rubber grommet mounted in the shield plate. The plate is grounded in each corner by a short, direct lead to the meter mounting bolts.

Amplifier Adjustment Before the tubes are inserted in the amplifier, the main bandswitch should be set to the various bands and the plate tank assembly tuned for resonance on each band when the loading capacitor is set to about ⅔ maximum value. The approximate settings should be logged for future reference. The two tubes are now inserted in their sockets and filament voltage applied to the amplifier. Voltage at the tube sockets should run between 4.8 and 5.1 volts, as measured with an accurate meter. The amplifier is now placed in the cabinet and the cooling fan connected so that it runs whenever the filament circuit is energized. An interlock switch atop the cabinet should be immediately wired so that it opens the high-voltage control relay in the power supply. In addition, a high-voltage shorting switch, such as shown in the illustration (figure 33), is suggested as an integral part of the amplifier, since lethal voltages are exposed when the lid of the cabinet is raised unless precautions are taken.

Typical operating voltages and currents for the 3-500Z tube are tabulated in Table 1. An operating plate potential of 2500 is recommended with an intermittent-service power supply capability of 800 milliamperes.

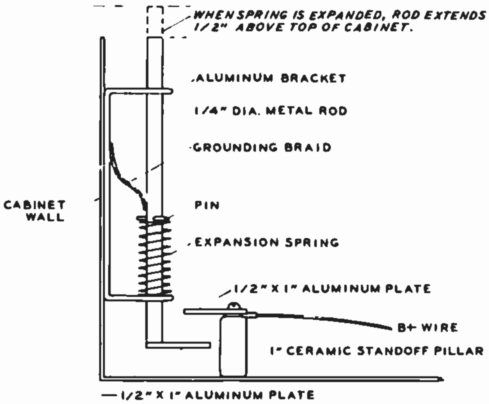


Figure 33

HOME-MADE HIGH-VOLTAGE SAFETY SHORTING SWITCH

This switch is actuated when the cabinet lid is raised, shorting the B-plus lead to ground. The switch is made up of a short section of ¼-inch shaft extension that is spring-loaded in the up position. Closing the lid forces the shaft down about ½ inch, breaking the short connection. When the lid of the cabinet is raised, the expansion spring pushes the rod upward, engaging the B-plus terminal. A grounding braid is used to make good ground connection to the shaft of the switch. The power supply should be fused or otherwise protected against the dead short.

Initial adjustment is greatly facilitated with the aid of an SWR meter or other output indicating device. Plate voltage is applied to the amplifier and the resting plate current is noted. A small amount of grid drive is introduced into the amplifier and resonance established in the plate circuit. Drive and loading are gradually increased, holding a ratio of about 3:1 between indicated plate and grid current. In the case of the 3-500Zs, maximum indicated grid current should be about 240 ma for a plate current of 800 ma. This ratio should be achieved with the minimum drive level and maximum antenna load level possible.

Under voice modulation, the plate current will kick to about 440 ma and grid current will kick to about 130 ma. For c-w operation at 2500 volts, plate loading and grid drive are decreased until 400 ma plate current and 125 ma grid current are noted on the meters. As with all grounded-grid amplifiers, grid drive should never be applied before plate voltage, or damage to the tubes may result.

Table 1.
Typical Operating Data, 3-500Z
R-F Linear Amplifier Service, Class-B
(one tube)

D-C Plate Voltage	1500	2000	2500
Zero signal Plate Current (ma)	65	95	130
Single Tone DC Plate Current (ma)	400	400	400
Single Tone DC Grid Current (ma)	130	130	120
Two Tone DC Plate Current (ma)	260	270	280
Two Tone DC Grid Current (ma)	80	80	70
PEP Useful Output Power (watts)	330	500	600
Resonant Load Impedance (ohms)	1600	2750	3450
Intermodulation Distortion Products (db)	−46	−38	−33

22-7 A 2-KW Linear Amplifier for 6 Meters

This rugged and reliable amplifier, designed and built by W6UOV, is designed for the serious 50-MHz experimenter. It uses an 8877 ceramic, high-μ triode and operates at 1-kW input for continuous c-w or RTTY service and at 2-kw PEP power input for SSB service (figure 34). The amplifier is well shielded and all leads are filtered so that the unit has minimum harmonic radiation. A driver capable of 40

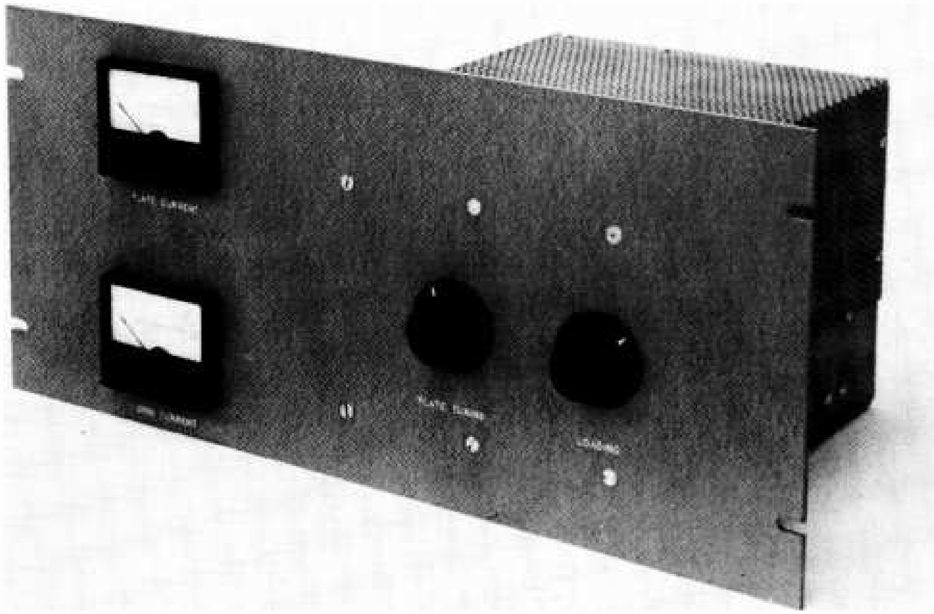


Figure 34

TWO-KILOWATT LINEAR AMPLIFIER FOR 6-METERS

This powerful amplifier features the 8877 high-μ ceramic triode in a cathode-driven circuit. At the left are the grid and plate current meters, with plate tuning and loading controls at the right. Amplifier requires about 40 watts peak drive for full output.

watts PEP power output is required for maximum amplifier input at a recommended anode potential of 2600 volts. Amplifier efficiency is 61 percent and the power gain is about 15 decibels.

Typically, at a potential of 2600 volts and a plate current of 750 milliamperes (2-kW PEP input) the third-order intermodulation products are better than —40 decibels below one tone of a two-tone test signal. This is an order of magnitude better than the majority of SSB exciters on the amateur market. Under these conditions, useful power output is more than 1200 watts, over and above tank circuit losses.

The Amplifier Circuit The 8877 is used in a cathode driven circuit, as shown in figure 35. The control grid is operated at d-c ground with a minimum of inductance between the tube and the chassis. Plate and grid currents are measured in the cathode return circuit. A 12-volt, 50-watt zener diode is placed in series with the cathode return lead to set the desired resting plate current.

Standby current is reduced by means of a 10K, 25-watt cathode resistor which is shorted out by the VOX relay, causing the tube to operate at its normal resting plate current. The 200-ohm resistor from the negative terminal of the plate supply to ground makes certain the negative supply terminal does not soar to the value of the plate voltage if the positive side of the supply is accidentally shorted to ground. Two reverse-connected diodes across the safety resistor limit any transient surges under a shorted condition which might cause insulation breakdown. In addition, the diodes protect the two panel meters from transient currents. A 200-ohm resistor across the zener diode provides a load for it and prevents the cathode voltage from soaring if the zener should burn open.

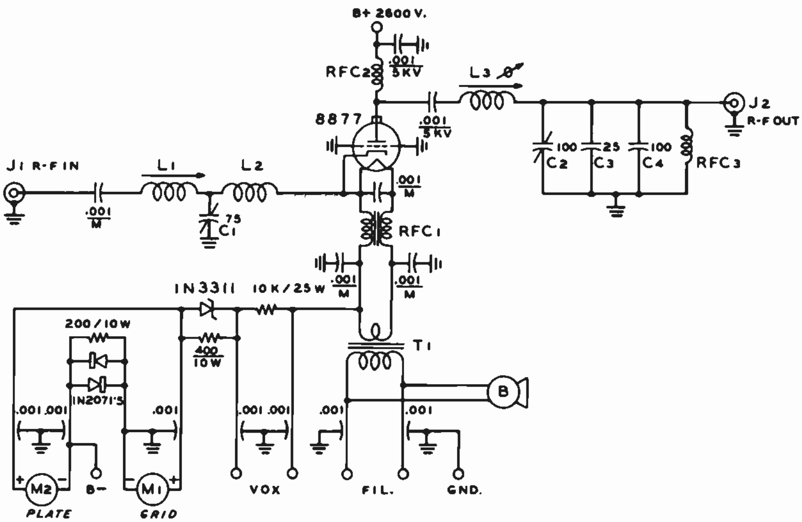


Figure 35

SCHEMATIC, 6-METER AMPLIFIER

B—Dayton 1C-180
C_1—75 pf. Hammarlund APC-75
C_2—100 pf. Johnson 155-10
C_3, C_4—Centralab type CRL 8588
Chimney—SK-2216 (Eimac)
L_1—6 turns #18 e. on CTC-1538-4-3 form, winding length 7/8"-inch
L_2—6 turns #18 e., 1/2-inch diameter, 5/8-inch long, self-supporting
L_3—3 turns 3/8-inch diameter copper tubing, inside diameter 1 7/8-inch; coil length 2 3/8-inch. Shorted turn 2 1/4-inch diameter copper tubing, 1/4 inch from main coil
M_1—100 ma d-c. Weston
M_2—1 ampere d-c. Weston
RFC_1—Bifilar wound choke, 1/2-inch diameter ferrite core (Indiana General CF-503). Three windings of 12 turns #10 formvar
RFC_2—54 turns #20 e. on 1/2-inch diameter Teflon rod; winding length 1-13/16 inches
RFC_3—Ohmite Z-50
Socket—Johnson 122-247-202
T_1—5 volts, 10 amperes. Stancor P-6135

The cathode input matching circuit consists of a simple T-network to transform the nominal 50-ohm input to the cathode impedance of the 8877 which is 54 ohms in parallel with 26-pfd input capacitance.

One coil (L_1) and the shunt capacitor are variable. With these two adjustments it is possible to cover a wide range of impedance transformations. The controls for the variable elements are brought out the left rear side of the chassis. Once the adjustments have been made, no tuning is required over the first megahertz of the band.

The socket for the 8877 is mounted one-half inch below the chassis using threaded brass spacers. Four pieces of brass shim stock, or beryllium copper, are formed into L-shaped contacts placed between the spacers and the chassis to make contact to the control-grid ring (figure 36).

The plate circuit is a standard pi-network (figure 37) with tube output and stray capacitances forming the input capacitance of the network (about 30 pfd). The output loading capacitor is an air variable unit, shunted by two fixed ceramic capacitors. Amplifier tuning is accomplished by varying the inductance of the coil by adjusting the coupling between the coil and a shorted turn.

Amplifier Construction The amplifier is built on an aluminum chassis box which is shielded by a perforated aluminum cover plate and a solid bottom plate. Air is blown into the under-chassis area, drawn up through the anode cooler of the 8877, and exhausted through the perforated cover. Placement of the major components may be seen in the photographs.

The amplifier plate tank coil is supported on two short teflon insulators. The closed ring near the front panel is the shorted turn

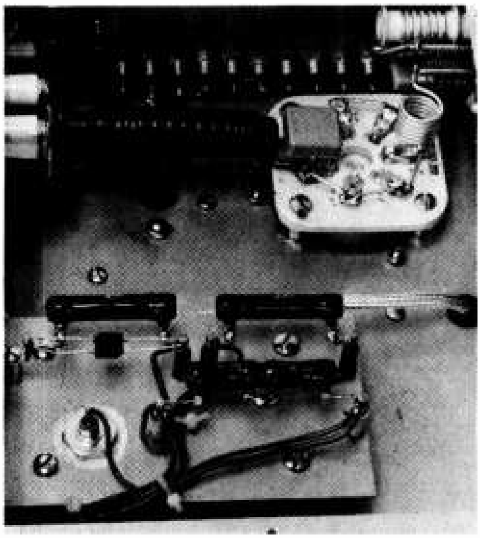


Figure 36

UNDER-CHASSIS VIEW OF AMPLIFIER COMPARTMENT

The input circuit is at the upper right with the filament choke at center. The filament by-pass capacitor is placed across the socket terminals. The grounding clips for the grid ring of the 8877 are next to the chassis, held in position by the socket-mounting studs and bolts. The socket is mounted below chassis level to allow passage of the cooling air. In the foreground are the zener diode, metering resistors, and reverse-connected meter diodes. The various filter capacitors for the power leads are mounted on the rear apron of the chassis, with the Millen high-voltage terminal at the left.

used for tuning; it is made of 3/8" diameter copper tubing, hard-soldered to a brass shaft coupler with copper-silver solder. Soft solder is not advisable in this application since the circulating current in the shorted turn is quite high.

The amplifier coil is adjustable in that the flexible strap connecting the blocking capacitor to the plate r-f choke may be moved about to subtract inductance from the main coil which is slightly oversize (figure 38). The position of the blocking capacitor (supported on a short bracket from the mounting insulator) is variable so that the strap can be flexed and set to the proper position. Note that the strap runs in the reverse direction to the winding direction of the main coil.

To adjust the amplifier for operation at the low-frequency end of the 6-meter band, the tube is placed in the socket, the shorted turn completely decoupled and the position of the blocking capacitor and anode strap adjusted to resonate the plate circuit to 50 MHz with the loading capacitor fully meshed. As the shorted turn is coupled tighter, total tank inductance will be reduced, causing the resonant frequency to rise. When the shorted turn is fully coupled, the resonant frequency of the plate tank circuit will be about 51 MHz.

Amplifier loading is accomplished in the same manner as in a typical pi-network circuit, with the shorted turn taking the place of the plate tuning capacitor.

A homemade plate r-f choke is used, since no commercial chokes were capable of providing proper service at this frequency. The choke is wound on a 1/2-inch diameter teflon rod and mounted atop the ceramic capacitor which bypasses the B-plus end of the choke.

Visible on the back of the front panel are the vernier ball-drive assemblies used on the plate tank controls. These allow the operator good control over the tuning and loading adjustments necessary for proper amplifier operation.

Amplifier Adjustment The output circuit is grid-dipped to frequency with the 8877 in the socket and the output loading capacitor fully meshed. Filament voltage is adjusted to 5.0 volts at the socket and the top and bottom shields are bolted in position. The cooling blower should be checked for proper operation. Amplifier operation is completely stable and tuning and loading follow the same sequence as with any standard grounded-grid amplifier. Grid excitation, of course, should never be applied when plate voltage is removed from the amplifier.

For initial tuneup, an SWR meter should be placed in series with the input line so that the input network may be adjusted for lowest value of SWR. A second SWR meter may be placed in the output line to serve as a power output indicator.

Drive is applied to provide about 20 ma of grid current and the plate circuit is tuned to resonance, drive level is raised in small increments along with output coupling until the desired power level is reached.

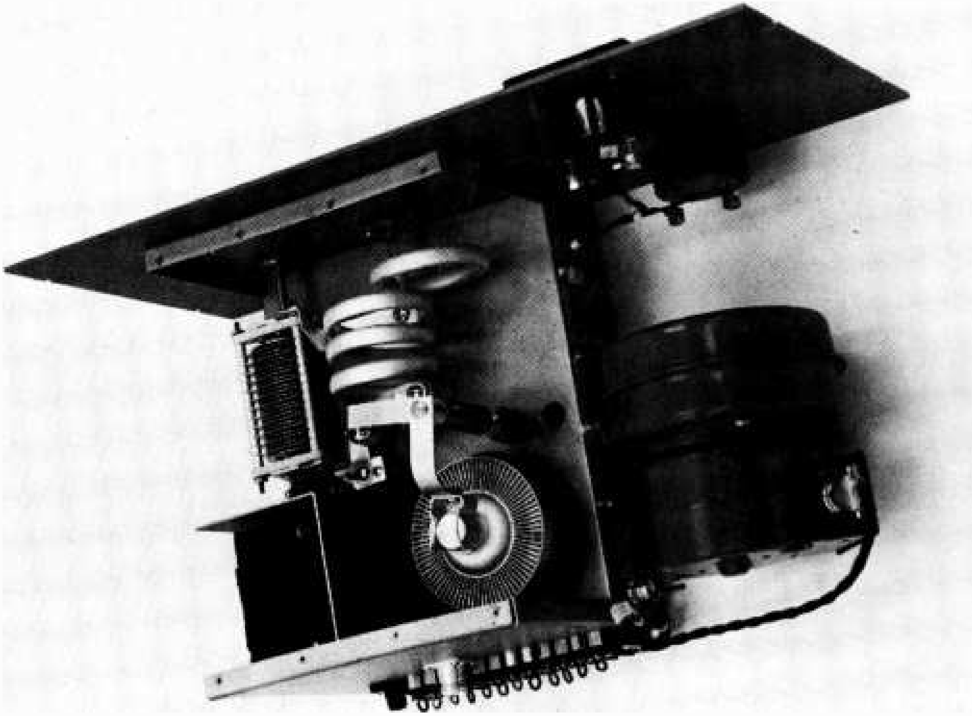


Figure 37

TOP VIEW OF AMPLIFIER

The variable capacitor across the top edge of the chassis is the adjustable portion of the loading circuit (C_t). Two ceramic transmitting capacitors are mounted in parallel with the air capacitor and can be seen at the rear of C_t near the filament transformer shield. The variable, shorted turn is supported from the panel drive. Amplifier shield has been removed for this photograph.

Always tune for maximum power at minimum input power.

When the amplifier is properly loaded, the input circuit may be adjusted for minimum SWR on the coaxial line to the exciter. Once the adjustment has been made, no tuning of the input circuit is required over the first 1000 kHz of the band.

With a plate potential of 2600 volts, the amplifier is loaded to a plate current of 750 ma, with a grid current of 58 ma. This represents 2000 watts input. Under voice waveforms, the grid and plate current will be approximately one-half the above values. For c-w operation, the amplifier may be run at 400 ma plate current and 28 ma grid current, for a power output of about 640 watts.

22-8 The KW-2 Heavy Duty Linear Amplifier

This rugged h-f linear amplifier is designed for operation at the 2-kW power input level for continuous service under the most exacting operating conditions (figure 39). Using a plate potential of 2700 volts, the power output on all bands is better than 1 KW. Peak driving power is about 50 watts and third-order intermodulation distortion products at maximum power output are —40 decibels below one tone of a two-tone test signal. The unit is built by W6HRB.

The KW-2 amplifier operates on any amateur band between 3.5 MHz and 29.7 MHz.

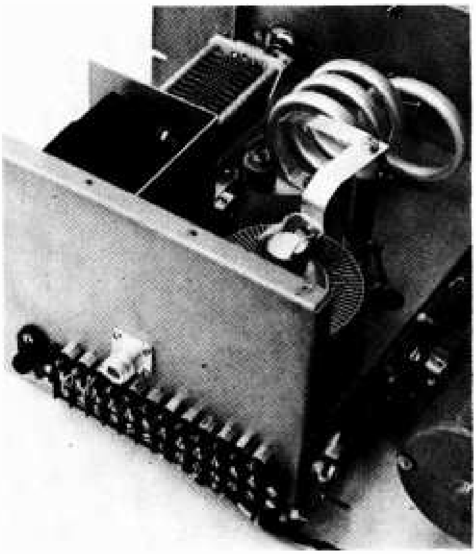


Figure 38

CLOSEUP OF AMPLIFIER PLATE CIRCUIT

The shorted turn is hard-soldered to the shaft coupler to allow front-panel loading and tuning. The "anti-inductance" plate-circuit strap can be seen connecting the top of the plate choke to the plate-blocking capacitor. Note that the position of the plate-blocking capacitor can be changed by loosening one screw and rotating the capacitor around the screw.

Harmonic output is very low, as a result of complete lead filtering and the use of a pi-L network in the output circuit. A single 8877 high-μ, ceramic power triode is used in a cathode-driven circuit to provide good efficiency, high power gain, and excellent stability.

Amplifier Circuitry The complete circuit of the KW-2 linear amplifier is shown in figure 40. The 8877 is operated class AB_2, with the grid at ground potential.

The Input Circuit—The drive signal is fed into the cathode through a pi-network (T_1-T_5) which matches the nominal 50-ohm amplifier input impedance to the 54-ohm cathode impedance of the 8877. The input circuit Q is unity, which is sufficient to preserve the waveform of the input signal. The cathode circuits are broadband and once adjusted to frequency may be forgotten.

The filament of the 8877 is internally insulated from the cathode and can be placed at ground potential, thus eliminating the expensive bifilar filament choke; only a small cathode r-f choke is required.

Grid- and plate-current metering is done in the cathode return circuit with the meters protected by reverse-connected diodes. A fused zener diode sets the resting plate current of the 8877 to the proper value for lowest distortion. The plate meter is placed between the 8877 cathode circuit and the negative terminal of the power supply, which is raised above ground. A protective resistor is placed across the meters to ensure that the negative side of the power supply remains close to ground potential. A separate ground connection is run between chassis ground and the chassis of the external power supply.

A dual purpose meter measures high voltage and amplifier grid current, which is monitored in the ground return circuit to the cathode.

The Plate Circuit—A pi-L network is used for maximum harmonic attenuation. In order to achieve good circuit Q at the upper and lower frequency limits of the operation, the plate circuit tuning capacitor (C_2) is divided into two sections, the smaller section being used for 10, 15, and 20 meters and the larger section added in parallel for 40- and 80-meter operation (figure 41). The network is designed for a plate load impedance of about 1900 ohms, and an image impedance of 220 ohms. Operating Q of the network is 10.

A four-section, two-deck heavy duty ceramic switch switches the plate circuit components. Circuit Q rises at 10 meters due to the output capacitance of the tube and the stray circuit capacitances (a total of about 25 pfd) and drops off slightly at the low frequency end of the 80-meter c-w band. A summation of the pi-L network components for each band is tabulated in Table 1.

Amplifier Cooling—The 8877 requires 20 cfm of air at a pressure drop of 0.23″ for 1000 watts anode dissipation at sea level. A squirrel cage blower provides proper ventilation. For full 1500 watts dissipation, 38 cfm of air is required at a pressure drop of 0.60″. In any case, sufficient cooling air must be supplied to hold tube temperature

Figure 39

THE KW-2 HEAVY DUTY LINEAR AMPLIFIER

This rugged h-f amplifier is designed for continuous duty operation at the 2-kW PEP input level. The amplifier operates at a plate potential of 2700 volts from an external power supply. Output is better than 1 kW on all bands at an intermodulation distortion level of −40 decibels below one tone of a two-tone test signal. At the left are the plate and grid/high-voltage meters. Directly below the meters are (left to right) the primary power switch, the meter-selection switch, and the high-voltage switch. The toggle switches are internally illuminated. The plate tuning control is near the center of the panel, with the loading control at the right. The bandswitch is near the bottom of the panel. The amplifier is enclosed in a perforated metal cabinet, suitable for placing next to your exciter.

below 225°C with 50°C ambient temperature at sea level.

Amplifier Construction A perforated wraparound cabinet measuring 16″ wide, 8″ high, and 13″ deep houses the amplifier. The r-f components are housed in an r-f tight enclosure measuring 10½″ wide, 12″ deep, and 7⅛″ high. The bottom plate of the box is solid and the top is perforated to allow cooling air to escape.

Placement of the major components may be seen in the photographs. The 8877 tube socket is mounted off-center on a small subchassis placed in the corner of the enclosure. Chassis size is 7½″ wide, 6¼″ deep, and 3″ high. The squirrel cage blower is mounted to the side of the enclosure and propels cooling air into the subchassis.

The main tuning capacitor, the loading capacitor, and the bandswitch are mounted to the front panel and to the enclosure. The large, copper-tubing plate inductor (L_2) is supported at one end from the panel and by a short ceramic insulator from the subchassis at the opposite end. The plate r-f choke is mounted to the side wall of the enclosure, as seen in figure 41.

The filament transformer and some small components are mounted on a second subchassis at the side of the main enclosure. This is braced to the panel with a small end bracket.

The under-chassis view (figure 42) shows placement of the various bulkhead mounted r-f filter capacitors and the cathode tuned circuits. The main bandswitch has an extended shaft which drives the cathode-circuit switch and the L-section switch by means of a right angle coupler placed under

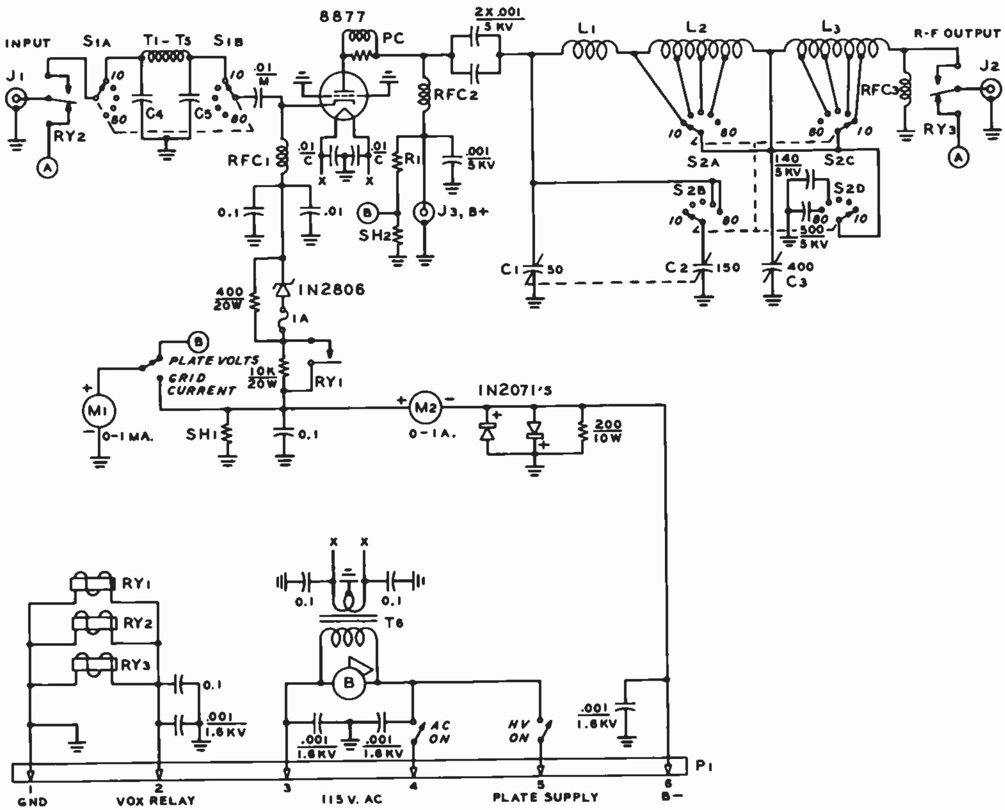


Figure 40

SCHEMATIC, KW-2 HEAVY DUTY LINEAR AMPLIFIER

Blower—Dayton 4C-012
C_1-C_2—Two section capacitor (50 pf + 150 pf, 4.5 kV). Johnson 154-16, modified as described in text
C_3—400 pf, 2 kV. Johnson 154-3 or equivalent
C_4, C_5—(See Table 2), 1-kV silver mica
Feedthrough capacitors—0.1, 600 volt, Sprague Hy-Pass
J_1—Type BNC, UG-185/U
J_2—Type N, UG-23/BU
J_3—Type HN high voltage, UG-496/U
L_1, L_2, L_3—(See Table 1)
Meters—Calectro
P_1—6-contact recessed receptacle. Cinch-Jones P-306RP
PC—2½ turns of ⅛" copper strap, ⅞" diameter around three 150-ohm, 2-watt composition resistors in parallel
R_1—Multiplier for 5000 volts to match meter
RFC_1—2" winding of #22 enamel wire on plastic rod, ⅜" diameter
RFC_2—3½" winding of #22 formvar insulated wire, ¾" diameter on Teflon rod
RFC_3—2.5 mH. National R-100
RY_1—Dpdt with coil to match VOX circuit
RY_2, RY_3—Spdt vacuum relay, with coil to match VOX circuit. Torr Laboratories, Inc. Type TF1 or TCR-1
SH_1—Shunt for 500 ma
SH_2—5 ohms, 10 watts
T_1-T_5—(See Table 2)
T_6—5.0 volt, 10 ampere. Hill Magnetics, Menlo Park, Ca. type HMP-1837

the 8877 chassis. A small shield plate covers the terminals of the bulkhead capacitors that are in the field of the final amplifier tank coils.

The pi-network assembly is made up of three inductors. The 10-meter coil (L_1) is the smallest and may be seen in the under-chassis photograph. It is held in position by leads made of ¼-inch wide copper strap. The coil is placed next to the main band-switch and positioned to have the shortest possible leads. The main coil (L_2) is wound of 3/16" copper tubing and is held in position by a sheet of *Rexolite*, grooved at the

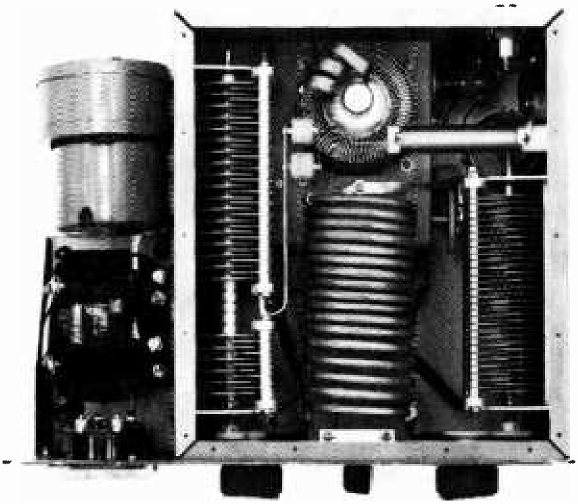


Figure 41

TOP VIEW OF KW-2 AMPLIFIER

The enclosed r-f compartment is at the right with blower, filament transformer, and small components mounted on a small chassis at the left. The main tank coil is wound on a Rexolite plate and supported from the front panel and the subchassis for the 8877 tube. A home-made two-section plate tuning capacitor (left) is used. To the right of the 8877 is the plate r-f choke, mounted in a horizontal position from the side of the enclosure. The L-section coil and antenna relays are just below the choke. The antenna loading capacitor is at the far right.

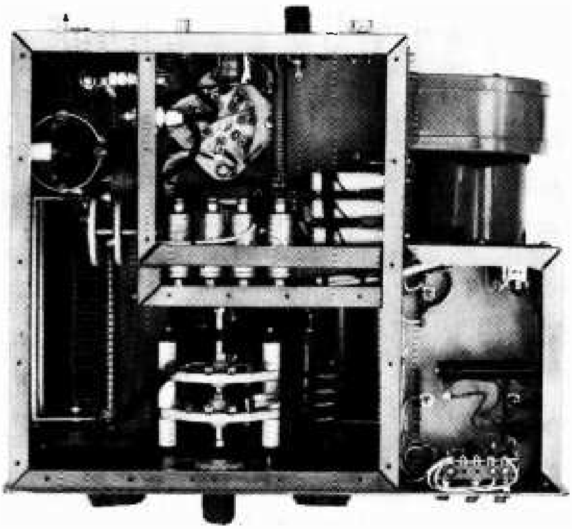


Figure 42

UNDER-CHASSIS VIEW OF KW-2 POWER AMPLIFIER

The bottom plate has been removed for this photograph. The main bandswitch is at lower center, with the 10-meter main tank coil beside it (right). To the left is the loading capacitor, the L-section bandswitch and the L-section coil. Top, center is the 8877 subchassis, showing the cathode inductors and the feedthrough capacitors. The VOX relay is just behind the ceramic tube socket. To the right is the blower and filament-transformer chassis. The main band-switch drives the L-section switch through a right-angle drive unit and at the same time drives the cathode switch, by means of the same unit.

outer edges to accept the coil windings. The L-section coil (L_3) is a small air-wound inductor placed in a vertical position near the L-section switch, which is mounted to the side of the 8877 subchassis. All wiring between plate coils, capacitors, and the main bandswitch is done with 1/4-inch wide copper strap. The ends of the straps are pre-tinned and wrapped around the proper coil turn and soldered in place with a large iron.

The Tuning Capacitor—The tuning capacitor is divided into two sections. It is made from a *Johnson 154-16* unit which has a total capacitance of 240 pfd with an air gap of 0.125″ (4.5 kV rating). The capacitor is divided into two sections by removing 4 rotor plates, leaving two rotor sections of 6 and 18 plates. The stator assembly is carefully removed and cut into two sections of 5 and 17 plates. Extra ceramic insulators from a similar capacitor are used to complete the stator assemblies which are bolted together with metal spacers and replaced in the capacitor end frames. This makes two capacitors having 50 pfd and 150 pfd capacitances, respec-

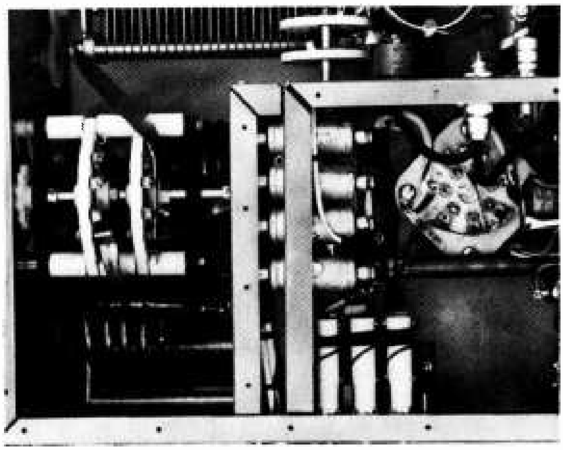


Figure 43

CLOSE UP OF 8877 SUBCHASSIS ENCLOSURE

The ceramic socket is submounted below the chassis deck to allow cooling air to pass up through the anode of the tube. To the right of the socket is the cathode r-f choke. R-f connections are made through BNC-type fittings mounted in the wall of the subchassis. The two antenna relays are visible in the upper-left corner of the chassis.

TABLE 1. Target Values for Pi-L Network

f (MHz)	$C_1 + C_2$ (Input)	C_3 (Loading)	$L_1 + L_2$ (μH)	L_3 (μH)
3.5	182	894	14.7	4.2
4.0	159	782	12.8	3.7
7.0	91	447	7.3	2.1
14.0	45	223	3.7	1.0
21.0	30	149	2.4	0.7
28.0	23	112	1.8	0.5

L_1—(10 Meters): 4½ turns of 3/16" tubing, 1⅛" i.d., 2" long plus leads. Resonate in circuit to 29.7 MHz with 40 pf input capacitance and 112 pf output capacitance (see text).

L_2—17½ turns ¼" tubing, 2⅝" i.d., 6" long. 12 turns spaced to 3¾"; 5½ turns spaced to 2¼". Adjust as follows:

(15 Meters): L_1 plus 2¾ turns of L_2. Resonate to 21.45 MHz with 40 pf input capacitance and 150 pf output capacitance.

(20 Meters): L_1 plus 5¾ turns of L_2. Resonate to 14.35 MHz with 45 pf input capacitance and 225 pf output capacitance.

(40 Meters): L_1 plus 10¾ turns of L_2. Resonate to 7.3 MHz with 90 pf input capacitance and 450 pf output capacitance.

(80 Meters): L_1 plus L_2. Resonate to 3.8 MHz 180 pf input capacitance and 900 pf output capacitance.

L_3—6 turns per inch of #18, 2" diam. I-CORE Air Dux 1606.

(80 Meters): 3.7 μH, 10 turns; (40 Meters): 2.1 μH, 7 turns; (20 Meters): 1.0 μH; 4 turns; (15 Meters): 0.7 μH, 3 turns; (10 Meters): 0.5 μH, 2 turns.

tively. The stator sections are connected into the circuit with lengths of ¼-inch wide copper strap.

Network Adjustment—The values of capacitance and inductance for the plate-circuit network components are listed in Table 1. These are "target" values and may be closely approximated by adjusting the various inductors once they are placed in the circuit. Lead length enters the picture, as well as stray capacitance of the coil to the surrounding environment and the builder must adjust the coil taps when all nearby components are mounted in place. The connecting leads represent circuit inductance that is not wound into the coils and this must be accounted for when the final connections are made.

The network is adjusted on each band by setting the capacitors to the approximate values given in Table 1, taking into account that the residual circuit capacitance is of the order of 25 pfd. The 10-meter coil is adjusted first. Since the target capacitance for C_1 is of this order, it indicates that circuit Q must be considered to be raised in order to allow for a tuning capacitance sufficient for proper adjustment. In this case, the tuning capacitor is set to provide a total circuit capacitance of about 40 pfd and the loading capacitor is set to about 112 pfd. With the aid of a grid-dip oscillator, coil L_1 is adjusted to provide resonance at 29.7 MHz, with C_1 near minimum capacitance. The coil turns are spread apart, or otherwise trimmed until resonance is achieved. It may be necessary to shorten interconnecting leads to the bandswitch to achieve proper results. Once the coil has been properly trimmed, it may be soldered in position.

This operation is repeated with the bandswitch set to the 15-meter position. The tuning and loading capacitors are set as indicated in the table and the 15-meter tap on coil L_2 adjusted for resonance at 21.45 MHz. This process is repeated for the 20-meter tap and also the 40-meter tap. The end of the main coil may now be trimmed for 80-meter operation. Use a frequency of 3.5 MHz if c-w operation is planned, or 3.8 MHz if phone only operation is contemplated. Adjust the capacitors to the correct value indicated in Table 1 for each band. Be sure to recheck each band when all of the taps are in place.

The L-section coil (L_3) is tapped along with the main coil, however, these taps are not so critical and may be set according to the data given in figure 40. The whole pi-L network configuration is grid-dipped as a unit, of course, which includes the L-section coil in the resonant circuit.

Amplifier Tuning and Adjustment After the amplifier has been wired and inspected, it is ready for initial checks. Optimum performance may be obtained with a plate potential of 2700 to 3000 volts. Before plate voltage is applied, filament voltage should be checked at the tube socket and the bottom placed on the amplifier enclosure to force the cooling air into the tube socket. The plate circuit controls should be set to the values indicated during the grid-dip oscillator tests.

When plate voltage is applied, the resting plate current should be about 90 to 100 ma. A small amount of drive is applied through an SWR meter and the amplifier is tuned and loaded into a dummy antenna, adjusting the plate circuit controls for maximum power output and minimum plate current. Grid current should run about 15 percent of the plate current value, as drive level and loading are increased until peak input power is attained. For 2 kW input, plate current should run about 740 ma with 110 ma of grid current. Maximum grid current, minimum plate current, and maximum power output should all occur at a single setting of the amplifier plate circuit controls.

For operation at the 1-kW power level, grid drive should be reduced and anode voltage should be reduced to about 2200.

The final step is to repeak the cathode circuit inductors for minimum SWR on the coaxial cable between the amplifier and the exciter. This should be done at maximum power input to the amplifier.

TABLE 2.
Cathode Transformers, T_1—T_5
Wound on ½" diam. Millen forms, slug-tuned

T_1—(10 Meters):	Approx. 0.34 μH. 4 turns #16e. Each capacitor: 100pf
T_2—(15 Meters):	Approx. 0.45 μH. 5 turns #16e. Each capacitor: 150pf
T_3—(20 Meters):	Approx. 0.65 μH. 6 turns #16e. Each capacitor: 220pf
T_4—(40 Meters):	Approx. 0.9 μH. 10 turns #16e. Each capacitor: 430pf
T_5—(80 Meters):	Approx. 2.8 μH. 13 turns #20e. Each capacitor: 820pf

22-9 A High Performance 2-Meter Power Amplifier

This compact, high performance amplifier is rated for continuous duty at the 2-kW peak power level. It combines reliable service with good linearity and efficiency. Designed and built by W6PO, the amplifier has been used for moonbounce communication with Europe on many occasions.

The amplifier uses an 8877 high-μ ceramic power triode in a cathode-driven circuit. A half-wave plate line is employed, along with a lumped-constant T-network input circuit. The amplifier is fully shielded and built to fit on a standard 19-inch relay rack panel (figure 44). The amplifier requires no neutralization, is completely stable and free of parasitics, and very easy to tune and operate.

The amplifier is designated for continuous duty at the 1-kW input level as well as at the 2-kW level for SSB operation. For the high power operation, plate voltage should be between 2500 and 3000 volts; under this condition the amplifier will deliver 1240 watts output. Stage gain is about 13.8 decibels and amplifier efficiency is 62 percent.

The Amplifier Circuit A schematic of the amplifier circuit is shown in figure 45. The 8877 is operated with the grid at d-c and r-f ground potential. The grid ring at the base of the tube provides a low inductance path between the grid element and the chassis. Plate and grid currents are measured in the cathode-return lead and a 12-volt, 50-watt zener diode is placed in series with the negative return to set the

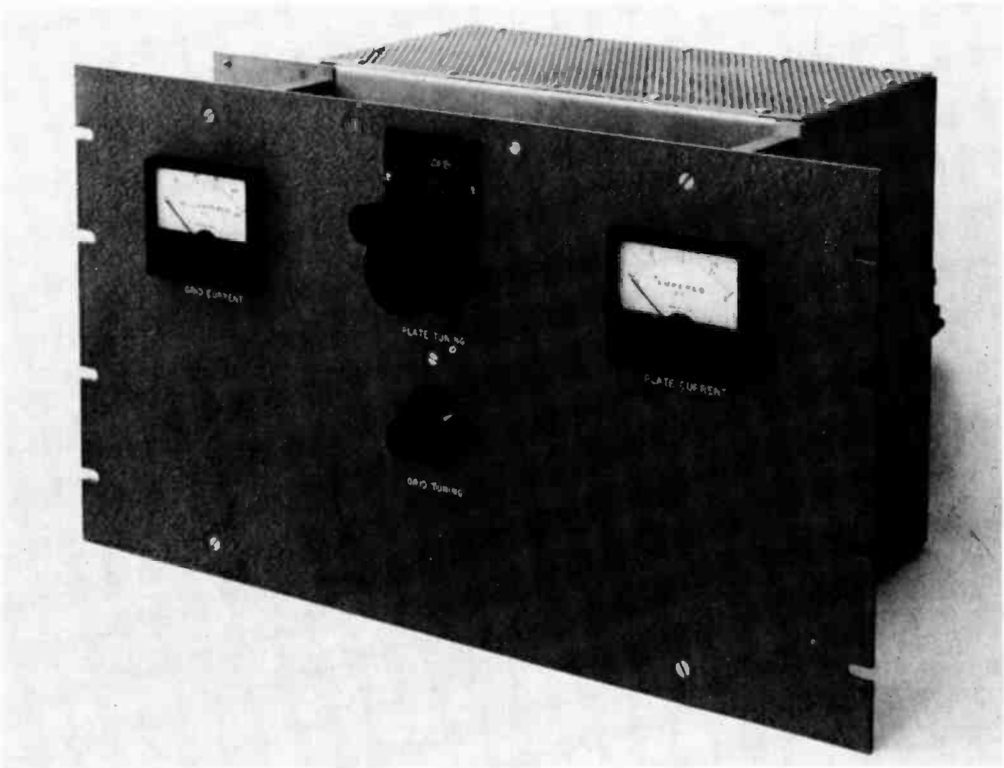


Figure 44

THE HIGH-PERFORMANCE 2-METER POWER AMPLIFIER

This amplifier will operate at the 2-kW PEP input level for heavy-duty performance. The amplifier is built upon a 10½" relay rack panel. The counter dial for the plate-tuning capacitor is at the center of the panel with the grid-tuning control directly beneath it. Grid and plate meters are at the left and right of the panel. The top of the r-f enclosure is covered with perforated aluminum sheet to allow the cooling air to escape from around the tube.

proper value of zero-signal plate current. Two diodes are reverse-connected across the instrument circuit to protect the meters.

Standby plate current of the 8877 is reduced to a very low value by the 10K cathode resistor which is shorted out when the VOX relay is activated, permitting the tube to operate in normal fashion.

A 200-ohm safety resistor ensures that the negative power lead of the amplifier does not rise above ground potential if the positive side of the high-voltage supply is accidentally grounded. A second safety resistor across the zener diode prevents the cathode potential from soaring if the zener should accidentally burn open.

The Input Circuit—The cathode input matching circuit is a T-network which matches the 50-ohm nominal input impedance of the amplifier to the input impedance of the 8877 which is about 54 ohms in parallel with 26 pfd. The network consists of two series-connected inductors and a shunt capacitor. One inductor and the capacitor are variable so the network is able to cover a wide range of impedance transformation. The variable inductor (L_1) is mounted to the rear wall of the chassis and may be adjusted from the rear of the amplifier. The input tuning capacitor (C_2) is adjustable from the front panel. When the network has been properly tuned, no adjustment is then required over the 4-MHz range of the 2-meter band.

The Plate Circuit—The amplifier plate circuit is a transmission-line type resonator.

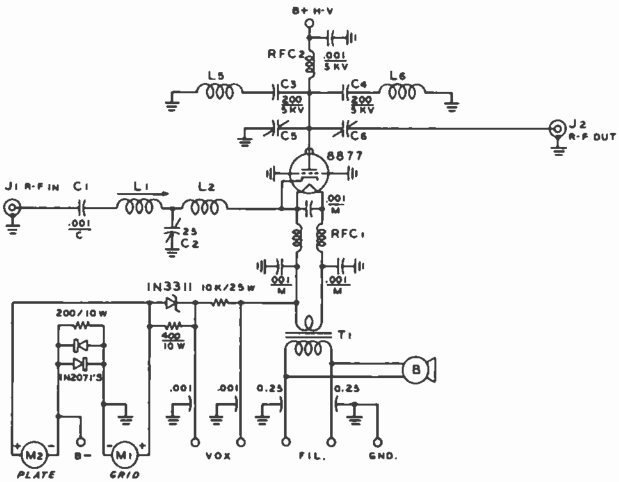


Figure 45

SCHEMATIC, 2-METER POWER AMPLIFIER

C_1—Centralab 858S-1000
C_2—25 pf Hammarlund HFA-25B
C_3, C_4—Each made up of two parallel connected 100 pf, 5 kV ceramic capacitors. Centralab 850S-100
C_5—Plate tuning (see text)
C_6—Plate loading (see text)
L_1—5 turns #14, ¾" long on ½" diameter form (white slug). CTC 1538-4-3
L_2—4 turns #14, ¾" diam., ¾" long
L_3, L_4—(RFC_1)—Two windings; 10 turns #12 enamel each, bifilar wound, ⅝" diameter
L_5, L_6—Plate lines (see text)
RFC_2—7 turns #14, ⅝" diameter, 1⅜" long
T_1—5 volts, 10 amperes. Chicago-Standard
M_1—0-100 ma d-c
M_2—0-1 amp d-c
Socket—Eimac SK-2210
Chimney—Eimac SK-2216

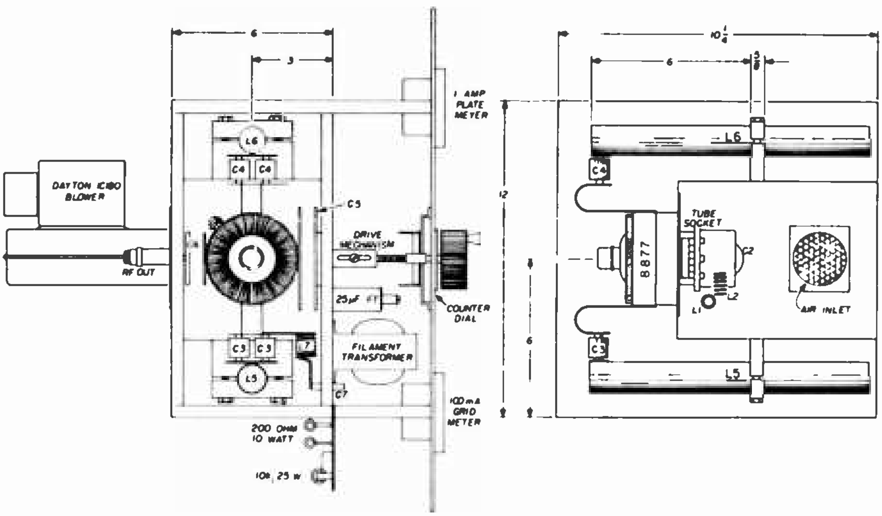


Figure 46

ASSEMBLY OF 2-METER AMPLIFIER

Structural details of the amplifier show relative size and position of the various components. Enclosure is made of aluminum panels. Bottom panel is solid and top panel is perforated to allow cooling air to escape.

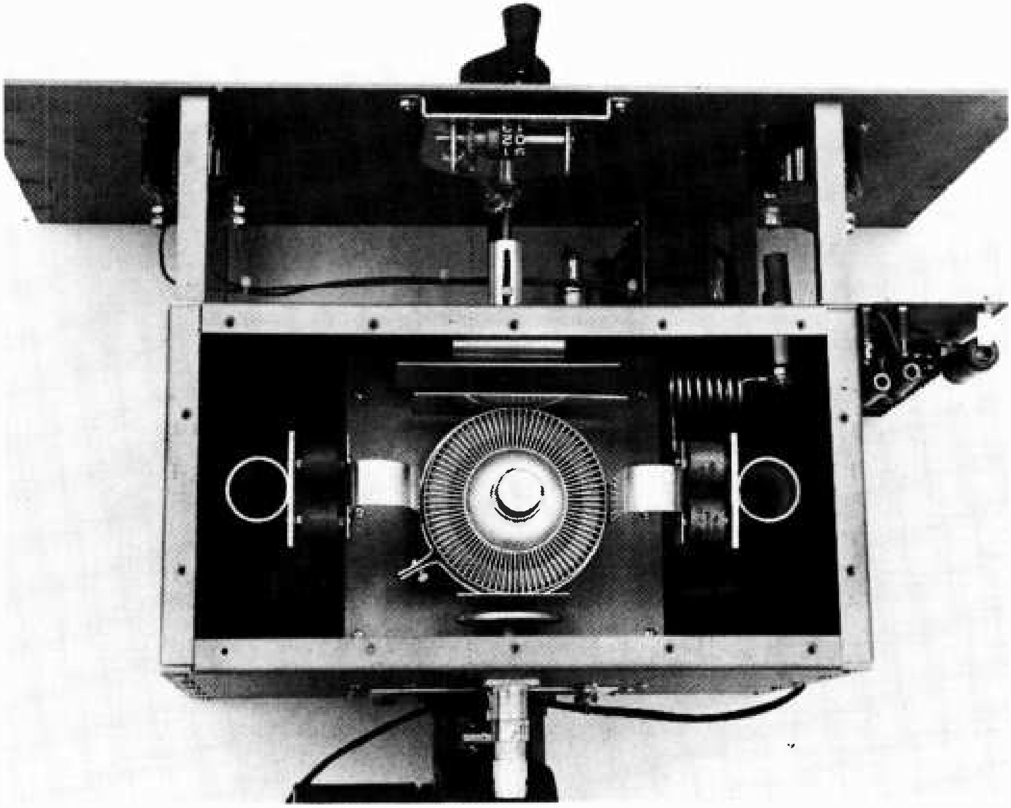


Figure 47

TOP VIEW OF 2-METER AMPLIFIER

The perforated plate is removed from the plate compartment showing the 8877 tube at center. Plate-blocking capacitors and plate lines are at either side of the tube, with the plate r-f choke in the upper right corner of the enclosure. The two-plate tuning capacitor is shown just above the the tube, with one of the plates attached to the anode strap of the tube. The other plate is driven in and out by means of a simple rotary mechanism driven by the counter dial. At the bottom (rear) of the amplifier the variable output coupling capacitor is seen just above the blower motor. The filament transformer and filament feedthrough capacitors are mounted to the front of the enclosure and a small plate at the right holds the various power resistors, diodes, etc.

The line (L_5 plus L_6) is a half wavelength long with the tube placed at the center (figure 46). This circuit, while having less operational bandwidth than an equivalent quarter-wavelength line, is chosen because standard water pipe can be used as the center conductor of the line and the over-all length of the line is long enough to be practical. In addition, the heavy r-f current that flows on the tube seals and control grid would, in the process of charging up the output capacitance to the peak plate voltage swing, tend to concentrate on one side of the tube if a single-ended, quarter-wavelength line were used. This current concentration would cause localized heating of the tube. The best tuned-circuit configuration to minimize this effect is a symmetrical, cylindrical coaxial cavity with the tube at the center. That arrangement is complex and difficult to build. A practical compromise is to use two quarter-wavelength lines connected to opposite sides of the tube. Note that each of the two quarter-wavelength lines used in this design are physically longer than if only one quarter-wavelength line were used.

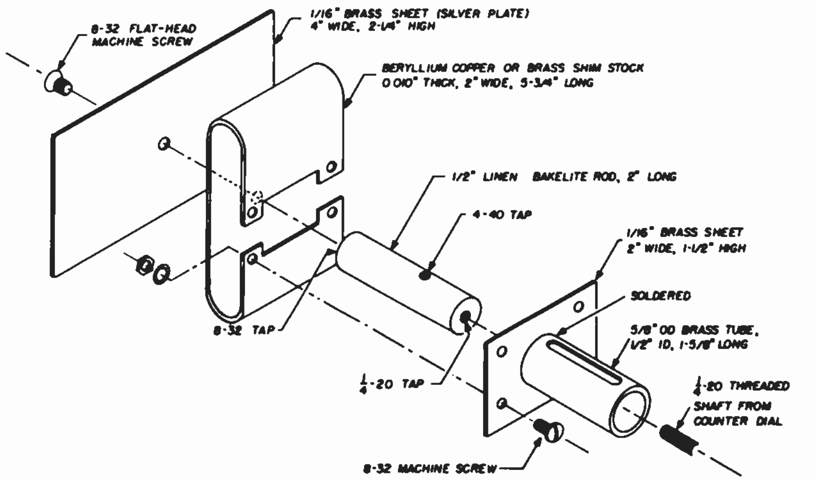


Figure 48

PLATE-TUNING CAPACITOR ASSEMBLY

The variable plate portion of the plate-tuning capacitor. This assembly permits the capacitor to be adjusted under full power since there are no moving or sliding contacts which carry heavy r-f current.

This is because only one-half of the tube output capacitance loads each of the two lines.

Resonance is established by a moving plate capacitor (C_5) and antenna loading is accomplished by a second capacitor (C_6) placed at the anode of the 8877. Output power is coupled through the series capacitor into a 50-ohm output circuit. In the top-view photograph (figure 47) tuning capacitor C_5 is at the front of the compartment; variable loading capacitor C_6 is at the rear. The plate r-f choke is visible in the front corner.

Amplifier Construction The 2-meter power amplifier is built in an enclosure measuring 10¼″ × 12″ × 6¼″. The 8877 socket is centered on a 6″ × 6″ subchassis plate. A squirrel-cage blower forces cooling air into the under-chassis area and the air escapes through the 2⅝″ diameter socket hole.

The plate-tuning mechanism is shown in figure 48. This simple apparatus will operate with any variable plate capacitor, providing a back-and-forth movement of about one inch. It is driven by a counter dial and provides a quick, inexpensive and easy means of driving a vhf capacitor. The ground-return path for the grounded plate is through a wide, low-inductance beryllium-copper or brass strip which provides spring tension for the drive mechanism.

The variable output coupling capacitor is located at the side of the 8877 anode. The type-N coaxial fitting is connected to the moveable plate of the coupling capacitor. The fitting is centered in a tubular assembly which allows the whole connector to slide in

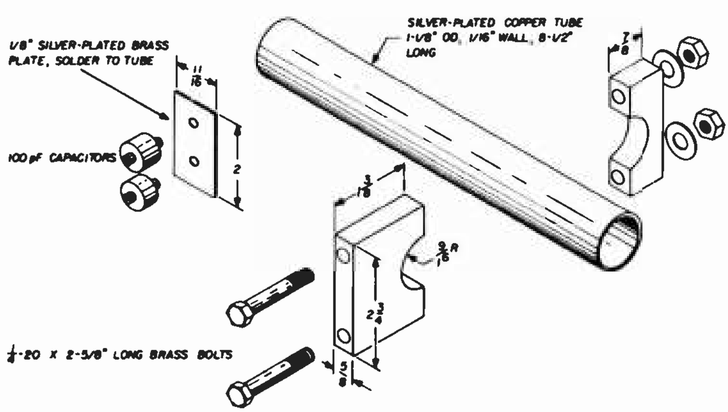


Figure 49

PLATE-LINE ASSEMBLY

Detail of plate lines L_5 and L_6. Copper tubes are standard water pipe.

and out of the chassis, permitting the variable plate of the coupling capacitor to move with respect to the fixed plate mounted on the tube anode clamp (figure 46). When the final loading adjustment has been set, the sliding fitting is clamped by means of a small cable clamp passed around the tubular assembly, as shown in figure 47.

The length of the plate-line inductors (L_5, L_6) is adjusted by means of two dural blocks placed at the shorted ends of the lines (figure 49). The position of the blocks is determined by setting plate-tuning capacitor C_5 at its lowest value and adjusting line lengths so that the plate circuit resonates at 148 MHz with the 8877 tube in the socket.

The plate r-f choke is mounted between the junction of one plate strap and a pair of the dual blocking capacitors and the high-voltage feedthrough capacitor is mounted to the front wall of the plate circuit compartment. The r-f blocking capacitors are rated for r-f service and the substitution of TV-type capacitors at this point is not recommended.

Not observable in the photographs is a short chimney to direct cooling air from the socket through the anode of the 8877. It is made from thin, sheet *Teflon* and is clamped in place between the chassis and the anode strap.

Under-chassis layout is shown in figure 50. The cathode input circuit is in the center compartment. The slug-tuned coil (L_1) is mounted on the rear wall. Air-wound filament chokes are placed in front of the socket. The cathode-heater choke coils are near the top edge of the enclosure. All of the cathode leads of the socket, plus one heater pin (pin 5) are connected in parallel and driven by the input matching network.

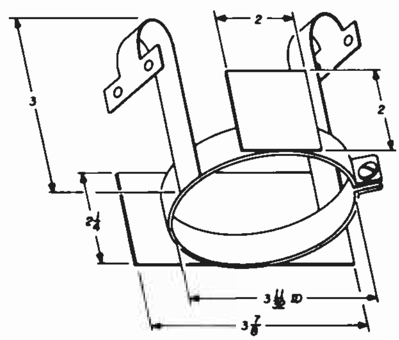


Figure 51

ANODE CLAMP ASSEMBLY

The ceramic socket for the 8877 is mounted one-half inch below chassis level by spacers to permit passage of cooling air to the anode. Four pieces of brass shim stock (or beryllium copper) are formed into gounding clips to make contact to the control-grid ring. The clips are mounted be-

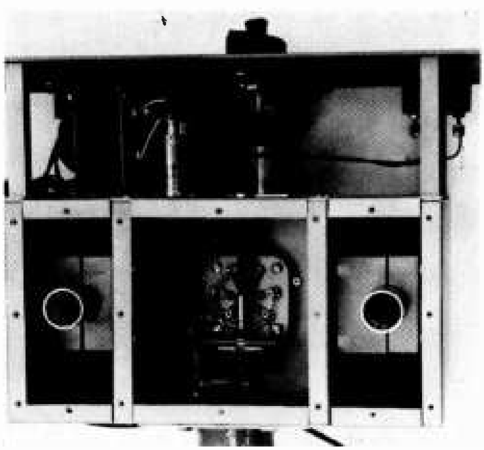


Figure 50

UNDER-CHASSIS VIEW OF 2-METER AMPLIFIER

The cathode input circuit is in the center compartment, with the filament choke just above the tube socket. The socket is mounted below the chassis deck to permit cooling air to escape up around the tube anode. The dural blocks holding the ends of the plate lines are bolted to the side walls of the inner chassis. The walls are slotted to permit the blocks to be moved up and down the lines to establish resonance.

TABLE 3. Operating Data for 8877 for 2-kW PEP and 1-kW Conditions

Plate Voltage	3000	2500	2500 V
Plate Current (peak) (single tone)	667	800	400 ma
Plate Current (no-signal)	54	44	44 ma
Grid Voltage	−12	−12	−12 V
Grid Current (single tone)	46	50	28 ma
Power Input	2000	2000	1000 W
Power Output	1240	1230	680 W
Drive Power	47	67	19 W

tween the spacers and the chassis. The aluminum clamps holding the ends of the plate lines are visible in the side compartments. The filament transformer and dial mechanism are placed in the area between the main enclosure and the panel.

Amplifier Tuning and Adjustment As with all grounded-grid amplifiers, excitation should never be applied when plate voltage is removed from the amplifier.

The first step is to grid-dip the input and output circuits to near resonance with the 8877 in the socket. An SWR meter should be placed in the input line so the input network may be adjusted for lowest SWR.

Tuning and loading follow the same sequence as with any lower-frequency grounded-grid amplifier. Connect an SWR meter and dummy load to the output circuit. Plate voltage is applied, along with a very low drive level. The plate circuit is tuned for resonance and the cathode circuit is peaked for maximum grid current. Final adjustment of the cathode circuit should be done at full power input because the input impedance of a cathode-driven amplifier is a function of the plate current of the tube.

R-f drive is increased in small increments along with output coupling until the desired power level is reached. By adjusting drive and loading together it is possible to attain the operating conditions given in Table 3. Always tune for maximum plate efficiency; that is, maximum output power for minimum input power. Do not overload and underdrive as plate efficiency will drop drastically under these conditions.

CHAPTER TWENTY-THREE

Power Supplies

Vacuum tubes and solid-state devices require an essentially pure direct current power supply for proper operation. Primary power is usually taken from the home electrical system which, in the United States, is nominally 117/240 volts, 60 hertz, in a 3-wire, grounded-neutral circuit. For mobile or portable operation, the primary power source is often a 6- or 12-volt automotive system.

In the common case of the home electrical system, the various d-c voltages required for communication equipment are supplied by a transformer, rectifier and filter network used in conjunction with a control and overload protection device.

In view of the high cost of iron-core components which make up the bulk of a power supply, it is well to carefully consider the design of a power supply in terms of minimum requirements which will permit desired performance to be obtained from the supply. Thus, the a-c primary power must be economically converted to high and lower d-c voltages for the proper operation of the various circuits in the equipment. In addition, certain circuits require voltage control or voltage regulation for proper operation. This Chapter covers the design and assembly of suitable power supplies and control circuits for amateur communication equipment.

23-1 Power-Supply Requirements

A power supply for a transmitter or for a unit of station equipment should be designed in such a manner that it is capable of delivering the required current at a specified voltage, that it has a degree of regulation consistent with the requirements of the application, that its ripple level at full current is sufficiently low for the load which will be fed, that its internal impedance is sufficiently low for the job, and that none of the components shall be overloaded with the type of operation contemplated.

The meeting of all the requirements of the previous paragraph is not always a straightforward and simple problem. In many cases compromises will be involved, particularly when the power supply is for an amateur station and a number of components already on hand must be fitted into the plan.

The power-supply requirements needed to establish the design of a satisfactory unit include the full-load output voltage; minimum, normal and peak current drain; the required voltage regulation; ripple voltage limit, and type of rectifier circuit to be used.

Once these requirements have been ascertained, the actual components for the supply may be selected. It is prudent, however, to design a supply in such a manner that it will have the greatest degree of flexibility; this will allow the supply to be used without change as a portion of new station equipment or as a bench supply to run experimental equipment.

Current-Rating Considerations The *minimum current drain* which will be taken from a power supply will be, in most cases, merely the bleeder current. There are many cases where a particular power supply will always be used with a moderate or heavy load on it, but when the supply is a portion of a transmitter it is best to consider the minimum drain as that of the bleeder. The minimum current drain from a power supply is of importance since it, in conjunction with the nominal voltage of the supply, determines the minimum value of inductance which the input choke must have to keep the voltage from soaring when the external load is removed.

The *normal current rating* of a power supply usually is a round-number value chosen on the basis of the transformers and chokes on hand or available from the catalog of a reliable manufacturer. The current rating of a supply to feed a steady load such as a receiver, a speech amplifier, or a continuously operating r-f stage should be at least equal to the steady drain of the load. However, other considerations come into play in choosing the current rating for a keyed amplifier, an amplifier of SSB signals, or a class-B modulator. In the case of a supply which will feed an intermittent load such as these, the current ratings of the transformers and chokes may be *less* than the maximum current which will be taken; but the current ratings of the rectifier system to be used should be at least equal to the maximum current which will be taken. That is to say that 300-ma transformers and chokes may be used in the supply for a modulator whose resting current is 100 ma but whose maximum current at peak signal will rise to 500 ma. However, the rectifier system should be capable of handling the full 500 ma.

The iron-core components of a power supply which feeds an intermittent load (such as demanded by an SSB transmitter) may be chosen on the basis of the current averaged over a period of several minutes, since it is the heating effect of the current which is of greatest importance in establishing the rating of such components. Since iron-core components have a relatively large amount of thermal inertia, the effect of an intermittent heavy current is offset to an extent by a resting period between words and syllables, or by key-up periods in the case of c-w transmission. However, the current rating of a rectifier tube is established by the magnitude of emission available from the filament of the tube, and the rating of a semiconductor rectifier is established by the maximum temperature limit of the rectifier element, both of which cannot be exceeded even for a short period of time or the rectifier will be damaged.

The above considerations are predicated, however, on the assumption that none of the iron-core components will become saturated due to the high level of intermittent current drain.

Voltage Regulation Since the current drain of a power supply can vary over a large magnitude, it is important to determine what happens to the output voltage of the supply with regard to change in current. Power-supply regulation may be expressed in terms of *static* and *dynamic* regulation. Static regulation relates to the regulation under long-term conditions of change in load whereas dynamic regulation relates to short-term changes in load conditions. Regulation is expressed as a change in output voltage with respect to load:

$$\text{Percent Regulation} = \frac{(E_1 - E_2) \times 100}{E_2}$$

where,

E_1 is no-load voltage,
E_2 is full-load voltage.

Thus static regulation concerns itself with the "on" and "off" voltages of the power

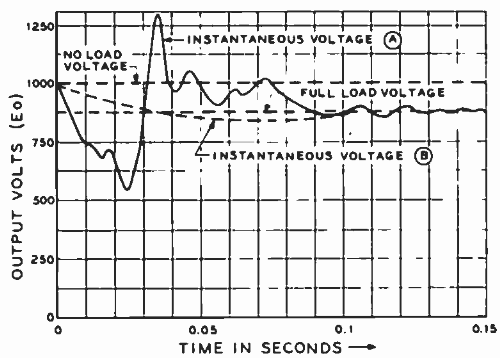


Figure 1

STATIC AND DYNAMIC REGULATION

A—Dynamic regulation illustrates voltage peaks caused by transient oscillations in filter network.
B—Static regulation is expressed in terms of no-load and full-load currents and voltages.

supply and dynamic regulation concerns itself with syllabic or keyed fluctuations in load. Static regulation is expressed in terms of average voltages and currents, whereas dynamic regulation takes into account instantaneous voltage variations caused by peak currents, or currents caused by undesired transient oscillations in the filter section of the power supply. In particular, c-w and SSB transmissions having a high peak-to-quiescent ratio of current drain are affected by poor dynamic regulation in the power system.

Examples of static and dynamic regulation are shown in figure 1. In example A, the no-load power-supply voltage is 1000 and the full-load voltage is 875. Static regulation is therefore 14.3 percent. If an oscilloscope is used to examine the supply voltage during the first fractions of a second when the full load is applied, the instantaneous voltage follows the erratic plot shown in curve A of figure 1. The complex pattern of voltage fluctuations, or transients, are related to resonant frequencies present in the power-supply filter network and are of sufficient magnitude to distort the waveform of c-w signals, or to appreciably increase intermodulation distortion and alter the first syllable of speech in an SSB system. Proper design of the filter system can reduce dynamic voltage fluctuations to a minimum and, at the same time, greatly improve the static regulation of the power supply.

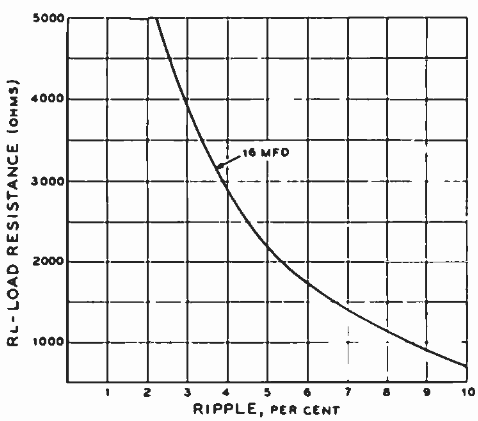


Figure 2

120-Hz RIPPLE ON 16-μFD CAPACITOR AS FUNCTION OF LOAD RESISTANCE

Static and dynamic regulation values of about 10 percent or so are considered to be limits of good design practice in amateur transmitting equipment, as illustrated by voltage curve B in figure 1.

Ripple Voltage The alternating component of the output voltage of a d-c power supply is termed the *ripple voltage.* It is superimposed on the d-c voltage, and the effectiveness of the filter system can be expressed in terms of the ratio of the rms value of the ripple voltage to the d-c output voltage of the supply. Good design practice calls for a ripple voltage of less than 5 percent of the supply voltage for SSB and c-w amplifier service, and less than 0.01 percent of the supply voltage for oscillators and low-level speech amplifier stages.

Ripple frequency is related to the number of pulsations per second in the output of the filter system. A full-wave rectifier, having two pulses of 60 Hz, for example, produces a 120-Hz ripple wave. A simple capacitive filter will reduce 120-Hz ripple as shown in figure 2. Ripple is an inverse ratio with capacitance, so doubling the capacitance will halve the ripple.

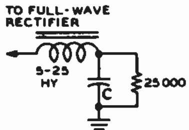
Ripple Filter Circuits The percentage of ripple found in representative LC filter circuits is shown in figure 3. The approximate ripple percentage for filter components may be calculated with the aid of the following formulas, assuming the power line frequency to be 60 Hz and the use of a full-wave or full-wave-bridge rectifier circuit. The ripple at the output of the first section of a two-section choke input filter is:

$$\text{Percent Ripple} = \frac{118}{(L \times C) - 1}$$

where,

L is the input choke inductance in henrys (at the operating current to be used),
C is the capacitance which follows the choke, expressed in microfarads.

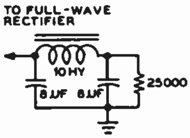
In the case of a two-section filter, the percent ripple at the output of the first section is determined by the foregoing formula.



RIPPLE IN TERMS OF C AT FULL LOAD

CAPACITANCE, C	PERCENT RIPPLE
2 μF	13.1
3 μF	8.5
4 μF	6.2
6 μF	4.0

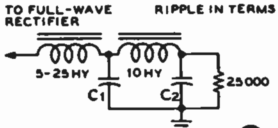
(A)



RIPPLE IN TERMS OF LOAD RESISTANCE

LOAD, OHMS	PERCENT RIPPLE
25000 (BLEEDER ONLY)	0.02
15000	0.04
10000	0.06
5000	0.1
3000	0.17
2000	0.25

(B)



RIPPLE IN TERMS OF C1 AND C2 AT FULL LOAD

C1	C2	PERCENT RIPPLE
2	2	1.2
3	2	0.7
4	4	0.25
8	8	0.06

(C)

Figure 3

VALUES OF RIPPLE VOLTAGE FOR STANDARD POWER-SUPPLY CIRCUITS

This percentage is multiplied by the filter reduction factor of the following section of filter. This reduction factor is determined through the use of the following formula:

$$\text{Filter reduction factor} = \frac{1.76}{LC-1}$$

where LC again is the product of the inductance and capacitance of the filter section. The reduction factor will turn out to be a decimal value, which is then multiplied by the percentage ripple obtained from the use of the preceding formula.

As an example, take the case of the filter diagramed in figure 4. The LC product of the first section is 16. So the ripple to be expected at the output of the first section will be: $118/(16-1)$ or $118/15$, which

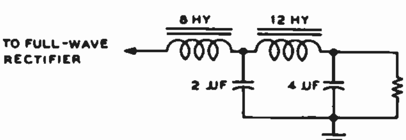


Figure 4

SAMPLE FILTER FOR CALCULATION OF RIPPLE

gives 7.87 percent. Then the second section, with an LC product of 48, will give a reduction factor of: $1.76/(48-1)$ or $1.76/47$ or 0.037. Then the ripple percentage at the output of the total filter will be: 7.87 times 0.037 or slightly greater than 0.29 percent ripple.

Resistance-Capacitance Filters In many applications where current drain is relatively small, so that the voltage drop across the series resistor would not be excessive, a filter system made up of resistors and capacitors only may be used to advantage. In the normal case, where the reactance of the shunting capacitor is very much smaller than the resistance of the load fed by the filter system, the ripple reduction per section is equal to $1/(2\pi RC)$. In terms of the 120-Hz ripple from a full-wave rectifier the ripple-reduction factor becomes: $1.33/RC$ where R is expressed in thousands of ohms and C in microfarads. For 60-Hz ripple the expression is: $2.66/RC$ with R and C in the same quantities as above.

Filter System Resonance The inductance of the filter choke in an LC filter network is dependent to an extent on the current drawn through it. At some values of inductance, it is possible for a 60-Hz or 120-Hz resonant circuit to be set up if the filter capacitance value is low. Filter resonance imposes a heavy peak load on the rectifier system and diodes or mercury-vapor rectifiers can be damaged by such undesired currents.

A 120-Hz resonance is achieved when the product of inductance and capacitance is 1.77. Thus, a 1-μfd capacitor and a 1.77-henry choke will resonate at 120 Hz. The LC product for resonance at 60 Hz is about 7.1. This latter value may occur when a 2-μfd capacitor is used with a 3.55-henry choke, for example. The LC products of 1.77 and 7.1 should be avoided to prevent resonance effects, which can result in destructive transient voltages in the power-supply system. In particular, the use of a swinging choke (one whose inductance varies with current) may lead to resonance effects, because the inductance of such a device may vary over a 5-to-1 range. It is possible for 60-Hz resonance to occur at a low value of current drain, and then for 120-Hz resonance to occur at near-full load current. When a swinging-type input choke is used in the filter system, the *LC* product must be greater than 7.1 at maximum current drain to eliminate unwanted supply resonances.

Back EMF It is possible to place the filter choke in the B-minus lead of the power supply, reducing the voltage potential appearing from choke winding to ground. However, the *back-emf* of a good choke is quite high and can develop a dangerous potential from center tap to ground on the secondary winding of the plate transformer. If the transformer is not designed to withstand this potential, it is possible to break down the insulation at this point.

23-2 Power-Supply Components

The usual components which make up a power supply, in addition to rectifiers which have already been discussed, are filter capacitors, bleeder resistors, transformers, and chokes. These components normally will be purchased especially for the intended application, taking into consideration the factors discussed earlier in this chapter.

Filter Capacitors There are two types of filter capacitors: (1) paper-dielectric type, (2) electrolytic type.

Paper capacitors consist of two strips of metal foil separated by several layers of special paper. Some types of paper capacitors are wax-impregnated, but the better ones, especially the high-voltage types, are oil-impregnated and oil-filled. Some capacitors are rated both for *flash* test and normal operating voltages; the latter is the important rating and is the maximum voltage which the capacitor should be required to withstand in service.

The capacitor across the rectifier circuit in a capacitor-input filter should have a working-voltage rating equal at *least* to 1.41 times the rms voltage output of the rectifier. The remaining capacitors may be rated more nearly in accordance with the d-c voltage.

The *electrolytic capacitor* consists of two aluminum electrodes in contact with a conducting film which acts as an *electrolyte.* A very thin film of oxide is formed on the surface of one electrode, called the *anode.* This film of oxide acts as the dielectric. The electrolytic capacitor must be correctly connected in the circuit so that the anode always is at a positive potential with respect to the electrolyte, the latter actually serving as the other electrode (plate) of the capacitor. A reversal of the polarity for any length of time will ruin the capacitor.

The high capacitance of electrolytic capacitors results from the thinness of the film which is formed on the plates. The maximum voltage that can be safely impressed across the average electrolytic filter capacitor is between 450 and 600 volts; the working voltage is usually rated at 450. When electrolytic capacitors are used in filter circuits of high-voltage supplies, the capacitors should be connected in series. The positive terminal of one capacitor must connect to the negative terminal of the other, in the same manner as dry batteries are connected in series.

Electrolytic capacitors can be greatly reduced in size by the use of etched aluminum foil for the anode. This greatly increases the surface area, and the dielectric film covering it, but raises the power factor slightly. For this reason, ultramidget electrolytic capacitors ordinarily should not be used at full rated d-c voltage when a high a-c component is present as would be the case for the input capacitor in capacitor-input filter.

Bleeder Resistors A heavy-duty resistor should be connected across the output of a filter in order to draw some load current at all time. This resistor avoids soaring of the voltage at no load when swinging-choke input is used, and also provides a means for discharging the filter capacitors when no external vacuum-tube circuit load is connected to the filter. This *bleeder* resistor should normally draw approximately 10 percent of the full load current.

The power dissipated in the bleeder resistor can be calculated by dividing the square of the d-c voltage by the resistance. This power is dissipated in the form of heat, and, if the resistor is not in a well-ventilated position, the wattage rating should be higher than the actual wattage being dissipated. High-voltage, high-capacitance filter capacitors can hold a dangerous charge if not bled off, and wirewound resistors occasionally open up without warning. Hence it is wise to place carbon resistors in series across the regular wirewound bleeder.

Several small resistors may be connected in series, if desired, to obtain the required wattage and voltage rating.

Transformers Power transformers and filament transformers normally will give no trouble over a period of many years if purchased from a reputable manufacturer, and if given a reasonable amount of care. Transformers must be kept dry; even a small amount of moisture in a high-voltage unit will cause quick failure. A transformer which is operated continuously, within its ratings, seldom will give trouble from moisture, since an economically designed transformer operates at a moderate temperature rise above the temperature of the surrounding air. But an unsealed transformer which is inactive for an appreciable period of time in a highly humid location can absorb enough moisture to cause early failure.

Filter Choke Coils Filter inductors consist of a coil of wire wound on a laminated iron core. The size of wire is determined by the amount of direct current which is to flow through the choke coil. This direct current magnetizes the core and reduces the inductance of the choke coil; therefore, filter choke coils of the *smoothing* type are built with an air gap of a small fraction of an inch in the iron core, for the purpose of preventing saturation when maximum current flows through the coil winding. The "air gap" is usually in the form of a piece of fiber inserted between the ends of the laminations. The air gap reduces the initial inductance of the choke coil, but keeps it at a higher value under maximum load conditions. The coil must have a great many more turns for the same initial inductance when an air gap is used.

The d-c resistance of any filter choke should be as low as practical for a specified value of inductance. Smaller filter chokes, such as those used in radio receivers, usually have an inductance of from 6 to 15 henrys, and a d-c resistance of from 200 to 400 ohms. A high d-c resistance will reduce the output voltage, due to the voltage drop across each choke coil. Large filter choke coils for radio transmitters and class-B amplifiers usually have less than 100 ohms d-c resistance.

23-3 Rectification Circuits

There are a large variety of rectifier circuits suitable for use in power supplies. Figure 5 shows the three most common circuits used in supplies for amateur equipment.

Half-Wave Rectifier A *half-wave rectifier* (figure 5A) passes current in one direction but not in the other. Dur-

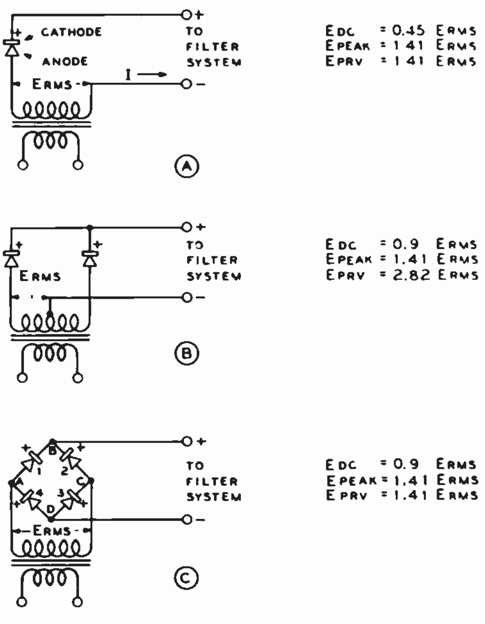


Figure 5

COMMON RECTIFIER CIRCUITS

A—Half-wave rectifier. Ripple is 121%.
B—Full-wave rectifier. Ripple is 48%.
C—Bridge rectifier. Ripple is 48%.

ing one half of an applied a-c cycle when the anode of the rectifier is positive with respect to the cathode the rectifier is in a state of conduction and current flows through the rectifier. During the other half of the cycle, when the anode is negative with respect to the cathode, the rectifier does not conduct and no current flows in the circuit. The output current, therefore, is of a pulsating nature which can be smoothed into direct current by means of an appropriate filter circuit. The output of a half-wave rectifier is zero during one-half of each a-c cycle; this makes it difficult to filter the output properly and also to secure good voltage regulation for varying loads. The *peak inverse voltage* with a resistive or inductive load is equal to the peak a-c voltage of the transformer (1.41 × E_{rms}) and is equal to twice the peak a-c voltage with a capacitive load.

Full-Wave Rectifier A *full wave* rectifier (figure 5B) consists of a pair of half-wave rectifiers working on opposite halves of the a-c cycle, connected in such a manner that each portion of the rectified wave is combined in the output circuit, as shown in figure 6. A transformer with a center-tapped secondary is required. The transformer delivers a.c. to each anode of each rectifier element; one anode being positive at any instant during which the other anode is negative. The center point of the high-voltage winding of the transformer is taken as the negative (B-minus) connection.

The cathodes of the rectifier units are always positive in polarity with respect to the anode of this type of circuit, and the output current pulsates 120 times per second for a 60-Hz supply. The peak output voltage is 1.4 times the rms transformer voltage and the inverse voltage across each rectifier unit is 2.8 times the rms voltage of the transformer (as measured across one half of the secondary winding). For a given value of ripple, the amount of filter required for a full-wave rectifier is half that required for a half-wave rectifier, since the ripple frequency of the former is twice that of the latter.

Bridge Rectifier A *bridge rectifier* (figure 5C) has four rectifier elements operated from a single a-c source. During one half-cycle of the applied a-c voltage, *point A* becomes positive with respect to *point C* and conduction takes place through rectifiers 4 and 2. During the other half of the cycle, conduction takes place through rectifiers 3 and 1 when *point C* is positive with respect to *point A*. On one half of the cycle, therefore, rectifiers 4 and 2 are in series with the output circuit and on the other half-cycle, rectifiers 3 and 1 are in series with the circuit. The bridge circuit is a full-wave system since current flows during both halves of a cycle of the alternating current.

One advantage of a bridge-rectifier connection over a full-wave, two-rectifier system is that with a given transformer voltage the bridge circuit produces a voltage output nearly twice that of the conventional full-wave circuit. In addition, the peak inverse voltage across any rectifier unit is only 1.4 times the rms transformer voltage. Maximum output voltage into an inductive or resistive load is about 0.9 times the rms transformer voltage.

The center point of the high-voltage winding of the bridge transformer is not at ground potential. Many transformers having a center-tapped high voltage winding are not designed for bridge service and insulation between this point and the transformer core is inadequate. Lack of insulation at this point does no harm in a full-wave circuit when the center tap is grounded, but may cause breakdown when the transformer is used in bridge configuration.

Rectifier Circuits *Choke input* is used in many filter systems because it gives good utilization of both rectifier and power-transformer capability (figure 6A). In addition, it provides much better voltage regulation than does a *capacitor input* system. A minimum value of choke inductance exists, and this critical value is equal to $R_L/1000$, where R_L is the load resistance. Inductance above the critical value will limit the no-load output voltage to about the average value ($E_{d.c.}$) in contrast to the capacitor-input filter circuit (figure 6B) wherein the no-load output voltage may rise as high as the peak value of the transformer voltage. The capacitor-

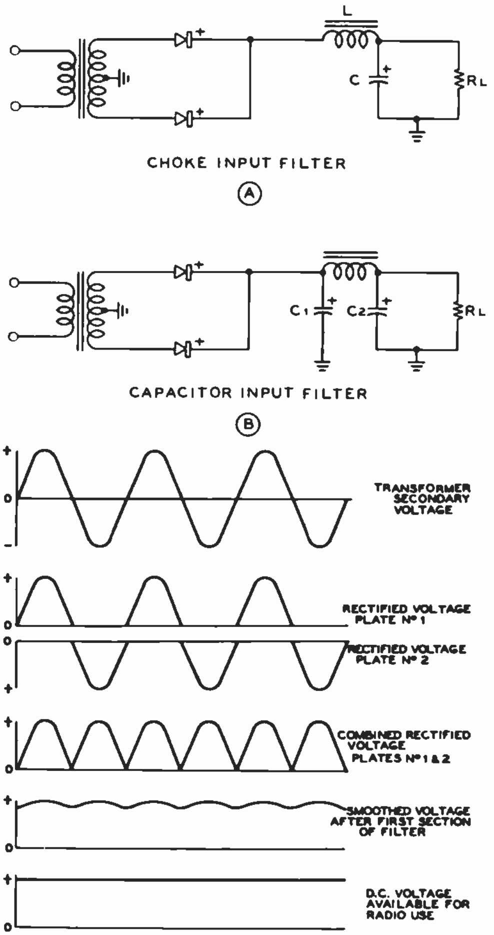


Figure 6

RECTIFICATION AND FILTER ACTION

Showing transformer secondary voltage, the rectified output of each diode, the combined output of the rectifiers, the smoothed voltage after the choke-input filter, and the d-c output voltage of the capacitor input filter.

input filter, at full load, provides a d-c output voltage that is usually slightly above the rms voltage of the transformer.

When capacitor input is used, consideration must be given to the peak value of the a-c voltage impressed on the filter capacitor, which usually runs equal to the peak transformer voltage (1.41 E_{rms}). The input capacitor, therefore, must have a voltage rating high enough to withstand the peak voltage if breakdown is to be avoided. A complete discussion of capacitor- and choke-input filter systems is contained in the *Radiotron Designer's Handbook,* published by the Radio Corporation of America, Harrison, New Jersey.

Special Single-Phase Rectification Circuits Figure 7 shows six circuits which may prove valuable when it is desired to obtain more than one output voltage from one plate transformer or where some special combination of voltages is required. Figure 7A shows a more or less common method for obtaining full voltage and half voltage from a bridge rectification circuit. With this type of circuit, separate input chokes and filter systems are used on both output voltages. If a transformer designed for use with a full-wave rectifier is used in this circuit, the current drain from the full-voltage tap is doubled and added to the drain from the half-voltage tap to determine whether the rating of the transformer is being exceeded.

Figure 7B shows a system which may be convenient for obtaining two voltages which are not in a ratio of 2 to 1 from a bridge-type rectifier; a transformer with taps along the winding is required for the circuit however. With the circuit arrangement shown, the voltage from the tap will be greater than one-half the voltage at the top.

An interesting variable-voltage circuit is shown in figure 7C. The arrangement may be used to increase or decrease the output voltage of a conventional power supply, as represented by transformer T_1, by adding another filament transformer to isolate the filament circuits of the two rectifier tubes and adding another plate transformer between the filaments of the two tubes. The voltage contribution of the added transformer T_2 may be subtracted from or added to the voltage produced by T_1 simply by reversing the double-pole double-throw switch (S). A serious disadvantage of this circuit is the fact that the entire secondary winding of transformer T_2 must be insulated for the total output voltage of the power supply.

An arrangement for operating a full-wave rectifier from a plate transformer not

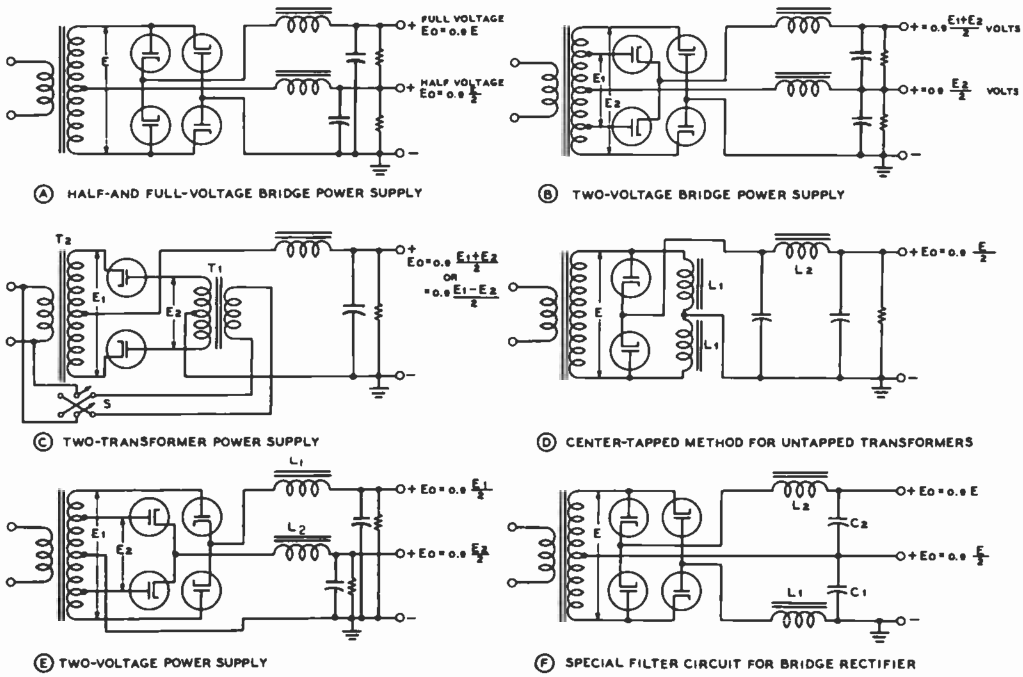


Figure 7

SPECIAL SINGLE-PHASE RECTIFICATION CIRCUITS

A description of the application and operation of each of these special circuits is given in the accompanying text.

equipped with a center tap is shown in figure 7D. The two chokes (L_1) must have high inductance ratings at the operating current of the plate supply to hold down the alternating current load on the secondary of the transformer since the *total* peak voltage output of the plate transformer is impressed across the chokes alternately. However, the chokes need only have half the current rating of the filter choke (L_2) for a certain current drain from the power supply since only half the current passes through each choke. Also, the two chokes (L_1) act as input chokes so that an additional swinging choke is not required for such a power supply.

A conventional two-voltage power supply with grounded transformer center tap is shown in figure 7E. The output voltages from this circuit are separate and not additive as in the circuit of figure 7B.

A special bridge rectifier is shown in figure 7F. Both L_1 and L_2 should be swinging chokes but the total drain from the power supply passes through L_1 while only the drain of the final amplifier passes through L_2. Capacitors C_1 and C_2 need be rated only half the maximum output voltage of the power supply, plus the usual safety factor. This arrangement is also of advantage in holding down the "key-up" voltage of a c-w transmitter since both L_1 and L_2 are in series, and their inductances are additive, insofar as the "critical inductance" of a choke-input filter is concerned. If 20-μfd capacitors are used at both C_1 and C_2 the dynamic regulation of the supply will be adequate for SSB operation.

Polyphase Rectification Circuits It is usual practice in commercial equipment installations when the power drain from a plate supply is to be greater than about one kilowatt to use a

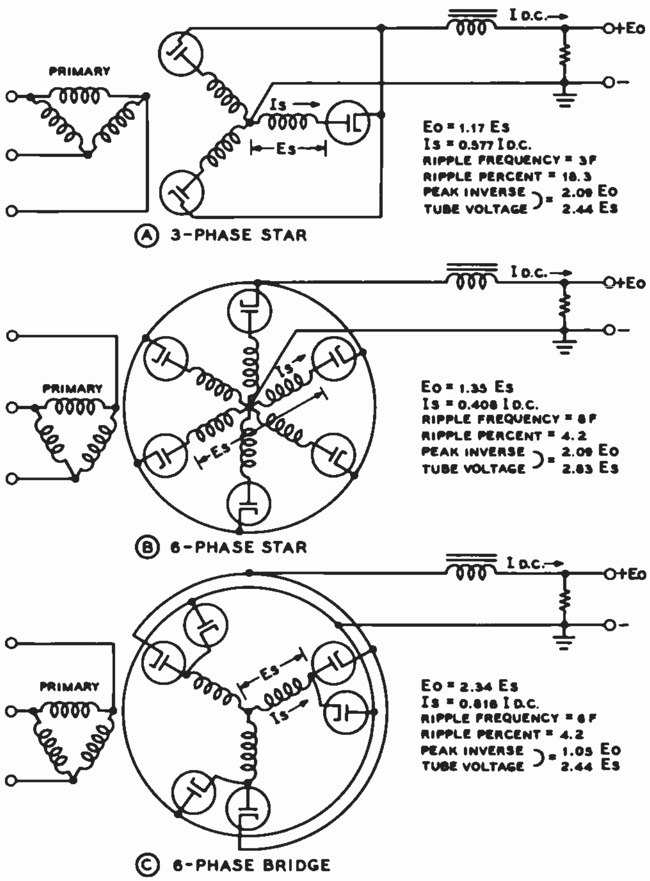


Figure 8

COMMON POLYPHASE-RECTIFICATION CIRCUITS

These circuits are used when polyphase power is available for the plate supply of a high-power transmitter. The circuit at B is also called a three-phase full-wave rectification system. The circuits are described in the accompanying text.

polyphase rectification system. Such power supplies offer better transformer utilization, less ripple output and better power factor in the load placed on the a-c line. However, such systems require a source of three-phase (or two-phase with Scott connection) energy. Several of the more common polyphase rectification circuits with their significant characteristics are shown in figure 8. The increase in ripple frequency and decrease in percentage of ripple is apparent from the figures given in figure 8. The circuit of figure 8C gives the best transformer utilization as does the bridge circuit in the single-phase connection. The circuit has the further advantage that there is no average d-c flow in the transformer, so that three single-phase transformers may be used. A tap at half-voltage may be taken at the junction of the star transformers, but there will be d-c flow in the transformer secondaries with the power-supply center tap in use. The circuit of figure 8A has the disadvantage that there is an average d-c flow in each of the windings.

Rectifiers Rectifying elements in high-voltage plate supplies are usually electron tubes of either the high-vacuum or mercury-vapor type, although silicon rectifier stacks containing a large number of elements are often used. Low-voltage high-current supplies may use argon gas rectifiers (Tungar tubes), silicon rectifiers, or other types of solid-state rectification elements.

Peak Inverse Plate Voltage and Peak Plate Current In an a-c circuit, the maximum peak voltage or current is $\sqrt{2}$, or 1.41 times that indicated by the a-c meters in the circuit. The meters read the *root mean square* (rms) values, which are the peak values divided by 1.41 for a sine wave.

If a potential of 1000 rms volts is obtained from a high voltage secondary winding of a transformer, there will be 1410 volts peak potential from the rectifier plate to ground. In a single-phase supply the rectifier tube has this voltage impressed on it, either positively when the current flows or "*inverse*" when the current is blocked on the other half-cycle. The *inverse peak voltage* which the tube will stand safely is used as a rating for rectifier tubes. At higher voltages the tube is liable to arc back, thereby destroying or damaging it. The relations between peak inverse voltage, total transformer voltage, and filter output voltage depend on the characteristics of the filter and rectifier circuits (whether full- or halfwave, bridge, single-phase or polyphase, etc.).

Rectifier tubes are also rated in terms of *peak plate current*. The actual direct load current which can be drawn from a given rectifier tube or tubes depends on the type of filter circuit. A full-wave rectifier with capacitor input passes a peak current several times the direct load current.

In a filter with choke input, the peak current is not much greater than the load current if the inductance of the choke is fairly high (assuming full-wave rectification).

Mercury-Vapor Rectifier Tubes The inexpensive *mercury-vapor* type of rectifier tube is often used in the high-voltage plate supplies of amateur and commercial transmitters. When new or long-unused tubes are first placed in service, the filaments should be operated at normal temperature for approximately twenty minutes before plate voltage is applied, in order to remove all traces of mercury from the cathode and to clear any mercury deposits from the top of the envelope. After this preliminary warmup with a new tube, plate voltage may be applied within 20 to 30 seconds after the time the filaments are turned on, each time the power supply is used. If plate voltage should be applied before the filament is brought to full temperature, active material may be knocked from the oxide-coated filament and the life of the tube will be greatly shortened.

Small r-f chokes must sometimes be connected in series with the plate leads of mercury-vapor rectifier tubes in order to prevent the generation of radio-frequency hash. These r-f chokes must be wound with sufficiently heavy wire to carry the load current and must have enough inductance to attenuate the r-f parasitic noise current to prevent it from flowing in the filter supply leads and then being radiated into nearby receivers. Manufactured mercury-vapor rectifier hash chokes are available in various current ratings from various manufacturers.

When mercury-vapor rectifier tubes are operated in parallel in a power supply, small resistors or small iron-core choke coils should be connected in series with the plate lead of each tube. These resistors or inductors tend to create an equal division of plate current between parallel tubes and prevent one tube from carrying the major portion of the current. When high-vacuum rectifiers are operated in parallel, these chokes or resistors are not required.

Voltage-Multiplying Circuits Practical voltage-multiplying circuits can be built up using silicon diode rectifiers or vacuum diodes as shown in figure 9. The basic "building block" is the half-wave rectifier shown in illustration A. The rectifier element simply rectifies the transformer voltage and delivers the alternate half-cycles of energy to the filter capacitor. The output voltage will be close to the peak voltage of the secondary winding of the transformer.

Figures 9B and C illustrate two voltage-doubler circuits which will deliver a peak d-c output voltage approximately equal to twice the rms value of the applied voltage. The no load d-c output voltage is equal to 2.82 times the rms input voltage. The full wave circuit is of advantage when the lowest level of ripple is required from the supply, since the ripple frequency is equal to twice the line frequency. The circuit of illustration C is of advantage when it is desired to ground one side of the transformer secondary winding, however the ripple frequency is the same as the a-c line frequency.

The circuit of figure 9D is a quadrupler and, in effect, is two voltage doublers of the type shown in 9C with their outputs con-

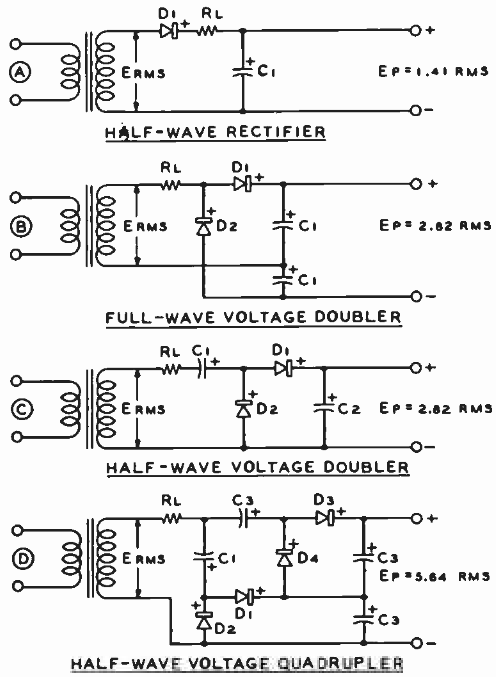


Figure 9

VOLTAGE-MULTIPLYING CIRCUITS

Voltage-multiplying circuits can be built up using silicon diode rectifiers or vacuum diodes. The basic "building block" is the half-wave rectifier (A). Capacitor C_1 is rated for twice the rms voltage of the transformer, and for a receiver supply, should be about 150 μfd. Capacitor C_2 in the voltage doubler circuit of (C) is rated for four times the rms voltage of the transformer. Capacitor C_2 in the quadrupler circuit of (D) is rated for three times the rms voltage of the transformer.

nected in series. The circuit delivers a d-c output voltage under load approximately equal to four times the rms value of the transformer voltage. The no-load d-c output voltage is equal to 5.64 times the rms input voltage.

All of these circuits consist of capacitors which are charged on halves of the voltage cycle through series-connected diodes and, in the case of circuits B and D, the charged capacitors are discharged in series through the secondary load circuit.

Diode Noise The silicon diode which is widely used in these circuits does not conduct until the applied forward potential exceeds the threshold voltage, which is about 0.5 volt. At this voltage the diode conducts abruptly, creating a steep wavefront, capable of generating radio-frequency interference. The interference is often eliminated if a transient suppression capacitor is placed across the diode (figure 10). In some cases, especially with the use of controlled-avalanche diodes, the capacitor is omitted and the *white-noise* interference generated by the diode may be found as an annoying "rush" on the sidebands of the transmitted signal, or as an annoying noise in the receiver. Suppression capacitors and additional lead filtering in the power supply may be required to eliminate the interference created by the abrupt conduction characteristic of the diode rectifier.

23-4 Series Diode Operation

Series diode operation is commonly used when the peak-inverse voltage of the source is greater than the maximum PIV rating of a single diode. For proper series operation, it is important that the PIV be equally divided among the individual diodes. If it is not, one or more of the diodes in the *stack* will be subjected to a PIV greater than its maximum rating and, as a result, may be destroyed. As most failures of this type result in a shorted junction, the PIV on the remaining diodes in the stack is raised, making each diode subject to a greater value of PIV. Failure of a single diode in a stack can lead to a "domino effect" which will destroy the remaining diodes if care is not taken to prevent this disaster. Forced voltage distribution in a stack is necessary when the individual diodes vary appreciably in reverse characteristics. To equalize the steady-state voltage division, shunt resistors may be placed across the diodes in a stack (figure 10A). The maximum value of the shunt resistor to achieve a 10-percent voltage balance, or better is:

$$\text{Shunt resistance} = \frac{PIV}{2 \times \textit{Max. Reverse Current}}$$

Six-hundred-volt PIV diodes, for example, having a reverse current of 0.3 ma at the maximum PIV require a shunt resistance of 1 megohm, or less.

Transient Protection Diodes must be protected from voltage transients which often are many times greater than the permissible peak-inverse voltage. Transients can be caused by d-c switching at the load, by transformer switching, or by shock excitation of LC circuits in the power supply or load. Shunt capacitors placed across the diodes will equalize and absorb the transients uniformly along the stack (figure 12B). The shunt capacitor should have at least 100 times the capacitance of the diode junction, and capacitance values of 0.01 μfd or greater are commonly found in diode stacks used in equipment designed for amateur service.

Controlled avalanche diodes having matched zener characteristics at the avalanche point usually do not require RC shunt suppressors, reducing power-supply cost and increasing over-all reliability of the rectifier circuit.

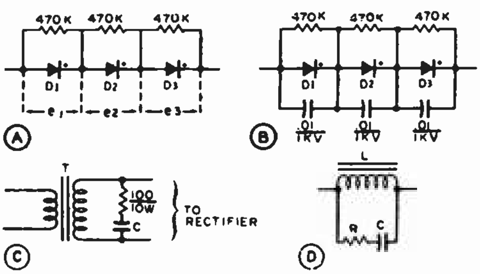


Figure 10

PROTECTION CIRCUITS FOR SEMICONDUCTOR POWER SUPPLIES

A—Peak inverse voltage should be distributed equally between series-connected diodes. If diodes do not have matched reverse characteristics, shunt resistors should be placed across the diodes.

B—Series-connected diodes are protected against high-voltage switching transients by shunt capacitors which equalize and absorb the transients uniformly along the stack.

C—Transient suppressor placed across the secondary of the high-voltage transformer protects diode stack from transients often found on the a-c power line or created by abrupt change in the magnetizing current of the power transformer.

D—Suppressor network across series filter choke absorbs portion of energy released when magnetic field of choke collapses, thus preventing the surge current from destroying the diode stack.

In high-voltage stacks, it is prudent to provide transient protection in the form of an RC suppressor placed across the secondary of the power transformer (figure 10C). The suppresser provides a low-impedance path for high-voltage transients often found on a-c power lines, or generated by an abrupt change in the magnetizing current of the power transformer as a result of switching primary voltage or the load. The approximate value of the surge capacitor in such a network is:

$$\textit{Capacitance } (\mu\text{fd}) = \frac{15 \times E \times I}{e^2}$$

where,

- E is the d-c supply voltage,
- I is the maximum output current of the supply in amperes,
- e is the rms voltage of the transformer secondary winding.

High-voltage transients can also be caused by series filter chokes subject to abrupt load changes. An RC suppressor network placed across the winding of the choke can absorb a portion of the energy released when the magnetic field of the choke collapses, thus preventing the current surge from destroying the diode stack (figure 10D). The approximate value of the transient capacitor is:

$$\textit{Capacitance } (\mu\text{fd}) = \frac{L \times I^2}{10 \times E^2}$$

where,

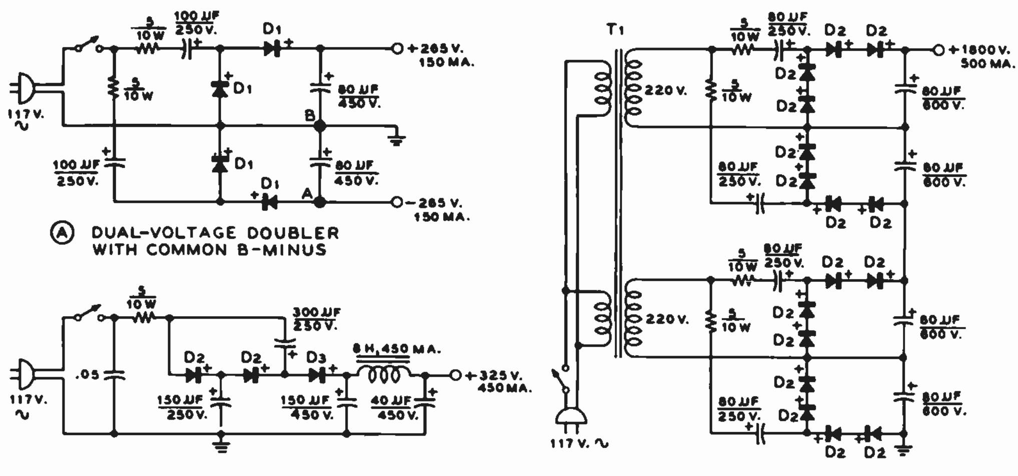
- L is the maximum choke inductance (henrys),
- I is the maximum current passing through the choke (amperes),
- E is the maximum d-c supply voltage (KV)

The resistance in series with the capacitor should equal the load impedance placed across the supply.

23-5 Silicon Supplies for SSB

Shown in figure 11 are three semiconductor power supplies. *Circuit A* provides 500 volts (balanced to ground) at 0.5 ampere. If the supply is isolated from ground by a 1:1 transformer of 250 watts capacity *point A* may be grounded and *point B* will provide half-voltage. *Circuit B* is a half-wave tripler that delivers 440 volts at 0.5 ampere. In this circuit, one side of the power line is common to the negative side of the output. *Circuit C* is a 900-watt, 0.5 ampere supply composed of two voltage doublers supplied from a "distribution" transformer having dual 117/240-volt windings.

Power Supply Rating for SSB Service The *duty cycle* (ratio of duration of maximum power output to total "on" time) of a power supply in SSB and c-w service is much smaller than that of a supply used for a-m equipment. While the power supply must be capable of supplying peak power equal to the PEP input of the SSB equipment for a short duration, the average power demanded by SSB voice gear over a period of time usually runs about one-half or less of the PEP requirement. Then, too, the intervals between words in SSB operation provide periods of low duty, just as the spaces in c-w transmission allow the power supply to "rest" during a transmission. Generally speaking, the average power capability of a power supply designed for *intermittent voice service* (IVS) can be as low as 25 percent of the PEP level. C-w requirements run somewhat higher than this, the average c-w power level running close to 50 percent of the peak level for short transmissions. Relatively small power transformers of modest capability may be used for intermittent voice and c-w service at a worthwhile saving in weight and cost. The power capability of a transformer may be judged by its weight, as shown in the graph of figure 16. It must be remembered that the use of alc or voice compression in SSB service raises the duty, thus reducing the advantage of the IVS power rating. The IVS rating is difficult to apply to very small power transformers,



(A) DUAL-VOLTAGE DOUBLER WITH COMMON B-MINUS

(B) HIGH-CURRENT SUPPLY

(C) 900-WATT HIGH-VOLTAGE SUPPLY

Figure 11

SEMICONDUCTOR POWER SUPPLIES

A—Voltage-quadrupler circuit. If point "A" is taken as ground instead of point "B," supply will deliver 530 volts at 150 ma from 117-volt a-c line. Supply is "hot" to line.

B—Voltage tripler delivers 325 volts at 450 ma. Supply is "hot" to line.

C—900-watt supply for sideband service may be made from two voltage quadruplers working in series from inexpensive "distribution-type" transformer. Supply features good dynamic voltage regulation.

D_1, D_2, D_3—1N4005. Use .01-μfd capacitor and 100K resistor across each diode.

T_1—Power distribution transformer, used backwards. 240/460 primary, 117/240 secondary, 0.75 KVA. Chicago PCB-24750.

since the d-c resistance of the transformer windings tends to degrade the voltage regulation to a point where the IVS rating is meaningless. Intelligent use of the IVS rating in choosing a power transformer, stacked silicon rectifiers, and "computer" type electrolytic capacitors can permit the design and construction of inexpensive, lightweight high-voltage power supplies suitable for SSB and c-w service.

The Design of IVS Power Supplies The low duty of SSB and c-w modes can be used to advantage in the design of high-voltage power supplies for these services.

The Power Transformer — Relatively low-voltage transformers may be used in voltage-doubler service to provide a kilowatt or two of peak power at potentials ranging from one to three thousand volts. Most suitable power transformers are rated for commercial service and the IVS rating must be determined by experiment. Figure 12 shows a relationship between various services as determined by extensive tests performed on typical transformers. The data illustrates the relationship between transformer weight and power capability. Transformer weight excludes weight of the case and mounting fixtures. Thus, a plate transformer weighing about 17 pounds that is rated for 400 watts commercial or industrial service should have an 800-watt peak capacity for c-w service and a 950-watt peak capacity for intermittent SSB service. A transformer having a so-called "two-kilowatt PEP" rating for sideband may weigh as little as 22 pounds, according to this graph.

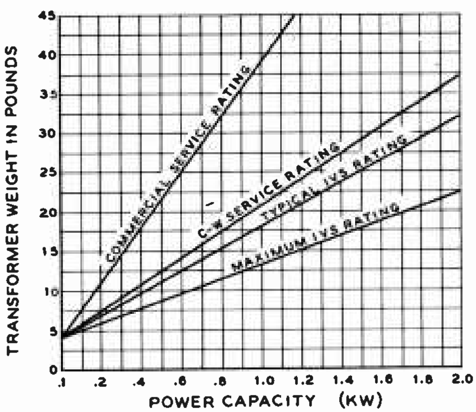


Figure 12

INTERMITTENT VOICE SERVICE IN SSB PERMITS LARGE PEAK POWER TO BE DRAWN FROM POWER TRANSFORMER. PEAK-TO-AVERAGE RATIO OF NEARLY FOUR TO ONE MAY BE ACHIEVED WITH MAXIMUM IVS RATING. POWER CAPACITY OF TRANSFORMER MAY BE DETERMINED FROM WEIGHT

Not shown in the graph is the effect of amplifier idling (standby) current taken from the supply, or the effect of bleeder current. Both currents impose an extra, continuous drain on the power transformer and quickly degrade the IVS rating of the transformer. Accordingly, the IVS curves of figure 12 are limited to the bleeder current required by the equalizing resistors for a series capacitor filter and assume that the idling plate current of the amplifier is cut to only a few milliamperes by the use of a VOX-controlled cathode bias system. If the idling plate current of the amplifier assumes an appreciable fraction of the peak plate current, the power capability of the supply decreases to that given for c-w service.

Most small power transformers work reliably with the center tap of the secondary winding above ground potential. Some of the larger transformers, however, are designed to have the center tap grounded and lack sufficient insulation at this point to permit their use in either a bridge or voltage doubling configuration. The only way of determining if the center-tap insulation is sufficient is to use the transformer and see if the insulation breaks down at this point! It is wise to ground the frame of the transformer so that if breakdown occurs, the frame of the tranformer does not assume the potential of the secondary winding and thus present a shock hazard to the operator.

The Silicon Rectifier—A bewildering variety of "TV-type" silicon rectifiers exists and new types are being added daily. Generally speaking, 600-volt PIV rectifiers, having an average rectified current rating of 1 ampere at an ambient temperature of 75°C with a maximum single-cycle surge-current rating of 15 amperes or better are suitable for use in the power supplies described in this section. Typical rectifiers are packaged in the *top-hat* configuration as well as the epoxy-encapsulated assembly and either type costs less than a dollar per unit. In addition, *potted* stacks utilizing controlled-avalanche rectifiers are available at a cost less than that of building a complete RC stack of diodes. The silicon rectifier, if properly used, is rarely the limiting factor in the design of steady-state IVS power supplies, provided proper transient protection is incorporated in the supply.

The Filter Capacitor—Recently developed "computer"-type aluminum-foil electrolytic capacitors combine high capacitance per unit of volume with moderate working voltage at a low price. Capacitors of this type can withstand short-interval voltage surges of 15 percent over their d-c working voltage. In a stack, the capacitors should be protected by voltage-equalizing resistors, as shown in the power supplies in this section. The capacitors are sheathed in a *Mylar* jacket and may be mounted on the chassis or adjacent to each other without additional insulation between the units. The stack may be taped and mounted to a metal chassis with a metal clamp, as is done in some of the units described here.

Inrush Current Protection — When the power supply is first turned on, the filter capacitors are discharged and present a near short circuit to the power transformer and rectifier stack. The charging current of a high-capacitance stack may exceed the maximum peak-recurrent current rating of the rectifiers for several cycles, thus damaging the diodes. Charging current is limited only by the series impedance of the power-supply circuit which consists mainly of the d-c circuit resistance (primarily the resistance of the secondary winding of the power transformer) plus the leakage reactance of the transformer. Transformers having high secondary resistance and sufficient leakage reactance usually limit the inrush current so that additional inrush protection is unnecessary. This is not the case with larger transformers having low secondary resistance and low leakage reactance. To be on the safe side, in any case, it is good practice to limit inrush current to well within the capability of the diode stack. A current-limiting circuit is shown in figure 13 which can be added at little expense to any power supply. The current-limiting resistor (R) is initially in the circuit when the power supply is turned on, but is shorted out by relay RY after a sufficient time has elapsed to partially charge the filter capacitors of the power supply. The relay coil is in a simple time-delay circuit composed of R_1-C_1. The delay may be adjusted by varying the capacitance value, and need only be about one-half second or so. Surplus 24-volt d-c relays used in dynamotor starting circuits work well in this device, as they have large low-resistance contacts and reasonable coil resistance (250 ohms or so).

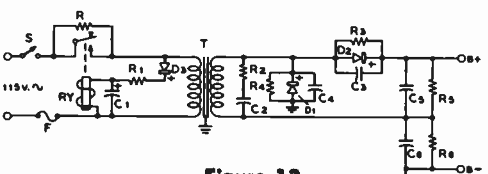


Figure 13

INRUSH CURRENT PROTECTION FOR POWER SUPPLY

Charging current of capacitor filter may be limited by series impedance of the power supply. In voltage-doubler circuit shown here, primary resistor R limits inrush current to within the capability of the diodes. Limiting resistor is shorted out after sufficient time has elapsed to partially charge the filter capacitors. Delay time of 0.5 second is usually sufficient. R_1-C_1 combination determines time delay. Secondary surge suppression (R_2-C_2) is used, and shunt RC equalizing networks are employed across each diode stack. Filter capacitors (C_5, C_6) are "computer-grade" electrolytic capacitors in series with 10K, 10-watt wirewound resistor placed across each capacitor.

Practical IVS Supplies An IVS voltage-doubler power supply may be designed with the aid of figures 12 and 14. A typical doubler circuit, such as shown in figure 13 is to be used. The full-wave voltage doubler is preferred over the half-wave type, as the former charges the filter capacitors in parallel and discharges them in series to obtain a higher d-c voltage than the peak voltage of the secondary winding of the power transformer. This saves transformer weight and expense.

Referring to figure 13, filter capacitors C_5 and C_6 are charged on alternate half cycles, but since the capacitors are in series across the load, the ripple frequency has twice the line frequency.

A second advantage of the full-wave doubler over the half-wave type is that the former tends to be self-protecting against switching transients. One diode stack is always in a conducting mode, regardless of the polarity of a transient, and the transient is therefore discharged into the filter-capacitor stack.

The filter-capacitor stack is rated for the peak no-load voltage (plus a safety factor), while the diode rectifiers must be able to withstand twice the peak no-load voltage (plus a safety factor). Good engineering practice calls for the *d-c working voltage of each portion* of the capacitor stack to be equal to the peak a-c voltage of the power transformer (1.41 × rms secondary voltage) plus 15 percent safety factor.

The R′ Factor—The a-c secondary voltage, secondary resistance, circuit reactance, and IVS capability of a transformer will determine its excellence in voltage-doubler service. The end effect of these parameters may be expressed by an empirical *R′ factor* as shown in figure 14. As an example, assume a power transformer is at hand weighing 25 pounds, with a secondary winding of 840 volts (rms) and a d-c secondary resistance of 8 ohms. The IVS rating of this transformer (from figure 12) is about 1.5 kW, PEP, or more. The appropriate d-c no-load voltage of an IVS supply making use of this unit in voltage-doubler service, such as the circuit of figure 13, is:

$$E_{\text{NO LOAD}} = 2.81 \times e$$

where,

e is the rms secondary voltage.

For this transformer, then, the no-load d-c supply voltage is about 2360 volts. The full load voltage will be somewhat less than this value. For a maximum power capability of 1.5 kW, a full-load current of about 0.75 ampere is required if the full load d-c voltage is in the vicinity of 2000. This is a realistic figure, so a "target" full-load voltage of 2000 is hopefully chosen.

The projected full-load voltage for a doubler-type supply may be determined with the aid of the R' factor and is calculated from:

$$E_{\text{LOAD}} = E_{\text{NO LOAD}} - R' \ (I \times R)$$

where,

R' is determined from figure 16,

I is the full load current in amperes,

R is the secondary resistance of the transformer.

For this example, R' is about 60 for the secondary resistance of 8 ohms, and the full-load d-c voltage of the supply is found to be just about 2000.

The peak rectified voltage across the complete filter-capacitor stack is equal to the no-load d-c voltage and is 2360 volts. Six 450-volt "computer"-type 240-μfd elec-

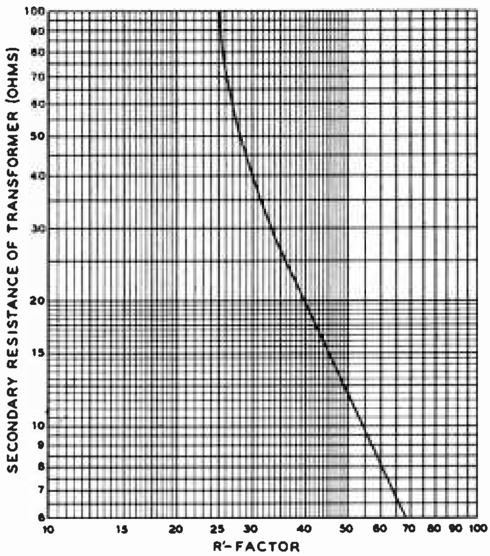


Figure 14

R′ FACTOR GRAPH FOR IVS POWER SUPPLIES

The full load d-c voltage of an IVS-rated voltage-doubler power supply may be determined with the aid of this graph. The secondary resistance of the transformer is measured and the R′ Factor is found. For example, a transformer having a secondary resistance of 20 ohms has an R′ Factor of about 40. The factor is used in the formula to calculate the full load d-c voltage of the power supply. For use with bridge circuits, the R′ Factor derived here should be divided by 2.5 before being used in the formula.

trolytic capacitors in series provide a 40-μfd effective capacitor, with a working voltage of 2700 (peak voltage rating of 3000), a sufficient margin for safety. Each capacitor is shunted with two 100K, 2-watt resistors in parallel.

The total PIV for the diode stack is twice the peak rectified voltage and is 4720 volts. A 100-percent safety factor is recommended for the complete stack, whose PIV should thus be about 9440 volts. The number of individual diodes in a suitable stack is:

$$\textit{Number of diodes} = \frac{11.2 \times \textit{rms voltage}}{\textit{Diode PIV}}$$

For this example, 600-volt PIV rectifiers are chosen and 16 are required, eight in each half of the stack.

The charging current of the capacitor stack may be safely ignored if the power supply is energized through a series primary resistor (R) such as shown in figure 13. One-ampere diodes having a single-cycle surge-current rating of 15 to 30 amperes are recommended for general use. The diffused silicon rectifiers (1N3195 and 1N-4005, for example) have a single-cycle surge-current rating of 30 amperes.

Capacitor Filters Power supplies for SSB service whose current requirements have a large peak-to-average ratio often make use of capacitor filters (figure 15). This simple circuit eliminates the resonant transients that are often found in LC filter systems and, if the capacitance is sufficiently large, provides adequate voltage regulation. In the case of a 2-kW PEP supply (2500 volts at 0.8 ampere) the load resistance is 3100 ohms and the required capacitance for 5-percent regulation is 55 μfd. Dynamic regulation of this degree is satisfactory for SSB and c-w service, as well as for amplitude modulation. As discussed earlier, the rectifier and power transformer must be protected from the inrush charging current of the filter capacitor.

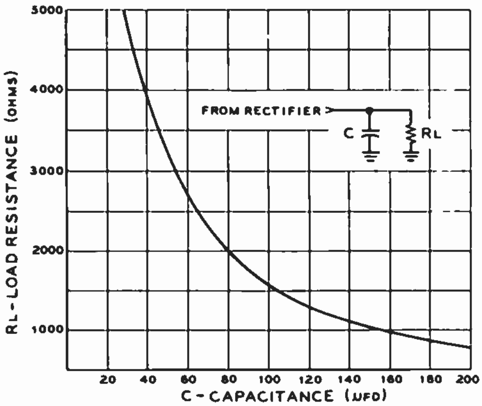


Figure 15

CAPACITOR FILTER

Capacitor filter is often used for SSB linear-amplifier power supplies. For 5-percent regulation, chart shows capacitance required for various values of d-c plate-load resistance.

23-6 A 1-Kilowatt IVS Power Supply

Shown in figures 16 and 17 is a typical 1-kilowatt IVS power supply designed from the above data. This supply is based on a 40-percent duty cycle and may be used for c-w service at 1-kilowatt level, or up to 1200 watts PEP or so for SSB service. The regulation of the supply is shown in the graph (figure 17), and the unit is capable of delivering 2300 volts at 0.5 ampere in IVS operation. The no-load voltage rises to 2750. The power supply is suitable for running a single 3-400Z at maximum rating, or it may be used for a pair of 813, 4CX250B, or 4CX-300A tubes at the kilowatt level. A transformer having less secondary resistance and slightly less secondary voltage would provide improved voltage regulation. The 840-volt transformer having an 8-ohm secondary winding discussed earlier would be ideal in this application.

The power supply is constructed on a steel amplifier foundation chassis and dust cover. The diode stack is mounted on a perforated phenolic board under the chassis. The electrolytic capacitors are taped together and held in position atop the chassis by a clamp cut from an aluminum sheet. The interior of the clamp is lined with a piece of plastic material salvaged from a package of frozen vegetables. The voltage-equalizing resistors are wired across the terminals of the capacitors. Normally, it

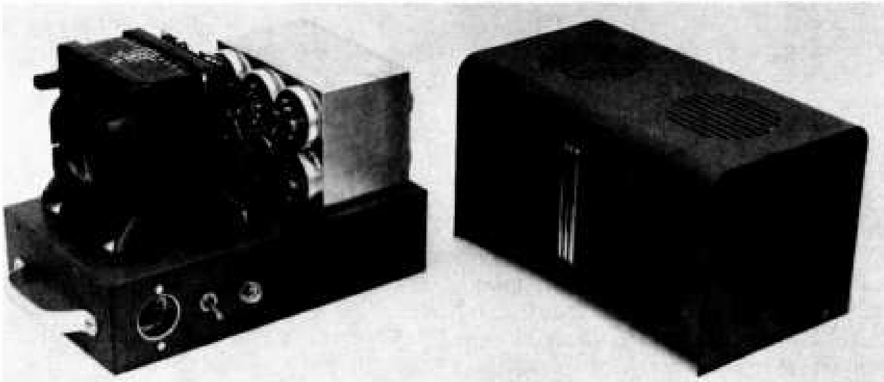


Figure 16

COMPACT ONE-KILOWATT IVS SUPPLY FOR SSB AND C-W SERVICE

This power supply delivers 2250 volts at 500 ma for SSB operation and 2400 volts at 400 ma for c-w operation. The supply is constructed on a covered foundation unit measuring 12" × 7" × 9" high (Bud CA-1751). The electrolytic capacitors are held in position by a bracket cut from aluminum sheet. Primary power receptacle, power switch, and neon pilot light are on the front apron of the chassis, with primary fuse and Miller high-voltage connector on the rear apron. High-voltage diode stack is mounted beneath the chassis on a phenolic board.

takes 10 seconds or so to fully discharge the filter capacitors when no external load is connected to the supply. It is recommended that the supply be discharged with a 1000-ohm, 100-watt resistor before any work is done on the unit. Power-supply components and all terminals should be well protected against accidental contact. The voltage delivered by this supply is lethal and the filter capacitors hold a considerable charge for a surprising length of time. This is the price one pays for an intermittent-duty design, and care should be exercised in the use of this equipment

To reduce the standby current and power consumption, it is recommended that cathode bias be applied to the linear amplifier stage shown in various designs in this Handbook. During transmission, the cathode resistor may be shorted out by contacts of the VOX relay, restoring the stage to proper operation.

Using the alternative 1100-volt transformer, the supply delivers 2600 volts at a c-w rating of 380 ma. Peak IVS voice rating is 500 ma (1.25 KW, PEP). No-load voltage is about 3100, and eight electrolytic capacitors are required in the stack instead of six.

23-7 A 2-Kilowatt PEP Supply for SSB

The power supply described in this section is designed for the maximum power rating for amateur service. It is capable of 1.2 kilowatts power for c-w (50 percent duty cycle) and 2 kilowatts IVS for SSB service. The supply is ideally suited for a grounded-grid amplifier using a single 3-1000Z, 4-1000A, or a pair of 3-400Z's. Regulation of the supply is shown in figure 18. A voltmeter is incorporated in the supply to monitor the plate voltage at all times. The supply makes use of the circuit of figure 13. Twenty 600-volt PIV diodes are used in the rectifier stack to provide a total PIV of 12 KV, which allows an ample safety factor. Eight 240-μfd, 450-volt capacitors are used in the filter stack to provide 30-μfd effective capacitance at 3600 volts working voltage. The voltage across the "bottom" capacitor in the stack is monitored by a 0-to-1 d-c milliammeter recalibrated 0 to 4 KV and which is used with a series multiplier to provide a 0 to 5000-volt full-scale indication. A 0-to-1 d-c ammeter is placed in series with the negative lead to the high-voltage terminal strip.

The supply is built on a steel amplifier foundation chassis in the same style as the 1-kW supply described previously. All safety precautions outlined earlier should be observed with this supply.

23-8 IVS Bridge-Rectifier Supplies

The bridge-rectifier circuit is somewhat more efficient than the full-wave circuit in that the former provides more direct current per unit of rms transformer current for a given load than does the full-wave circuit. Since there are two rectifiers in opposite arms of the bridge in the conducting mode when the a-c voltage is at its peak value, the remaining two rectifiers are back-biased to the peak value of the a-c voltage. Thus the bridge-rectifier circuit requires only half the PIV rating for the rectifiers as compared to a center-tap full-wave rectifier. The latter circuit applies the sum of the peak a-c voltage plus the stored capacitor voltage to one rectifier arm in the maximum inverse-voltage condition.

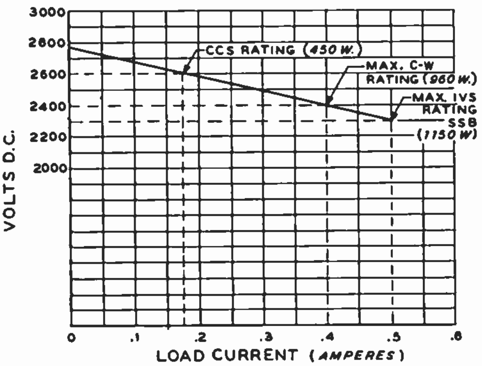


Figure 17

REGULATION CURVE OF ONE KILOWATT IVS SUPPLY

The power supply uses the circuit of figure 13. Primary surge resistor (R) is 5 ohms, 50 watts. Secondary surge-voltage resistor (R_s) is 200 ohms, 10 watts. Surge capacitor (C_s) is .02 µfd, 3 KV (Aerovox P89-M). Sixteen type 1N2071 (600-volt PIV) diodes are used in an assembly such as shown in figures 25 and 26. The diode shunt capacitors are .01-µfd, 600-volt ceramic discs, and the shunt resistors are 470K, ½-watt units. Six 450-volt (working), 240-µfd filter capacitors are used in series, each capacitor shunted with two 100K, 2-watt resistors in parallel. The time delay relay (RY) has a 24-volt d-c coil with a resistance of about 280 ohms (Potter-Brumfield PR5-DY). Contacts are rated at 25 amperes. Delay time is about 0.5 second and is determined primarily by the time constant of R_1-C_1. Suggested values are 800µfd (50 working volts) for C_1 and 600 ohms, 10 watts for R_1. Diode D_1 may be a 1N2070. The power transformer shown is a surplus unit having a 115/230-volt primary and a 960-volt secondary. The transformer weight is 18 pounds and it has an IVS rating of 1.2 KW. (A commercial alternative is Hill Magnetics Co., 2201 Bay Road, Redwood City, Calif. #HMP-1939A. This compact, 825-volt, wound-core transformer has improved regulation and is rated at 1 KW continuous duty [2 KW IVS rating] and provides 2000 volts at a continuous load of 500 ma.)

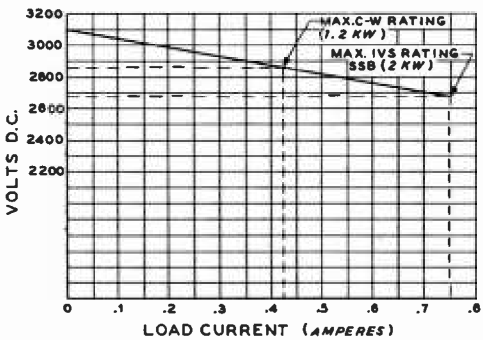


Figure 18

The power supply uses the circuit of figure 13. Surge components are as given in figure 16, except that the surge capacitor (C_s) has a rating of 5 kV. Twenty type-IN2071 (600-volt PIV) diodes are used in an assembly similar to that shown in figures 21 and 22. Eight 240 µfd, 450-working-volt (500-volt peak) capacitors are used to provide 30 µfd effective capacitance. Two 100K, 2-watt resistors are shunted across each capacitor. Time-delay circuit components are as suggested in figure 13. The transformer used has a 117/240-volt primary and an 1100-volt secondary, with an ICAS rating of 1.2 kW. (Berkshire Transformer Corp., Kent, Conn. #BTC-4905B).

A 500-Watt IVS Bridge Power Supply Shown in figure 19 is a 500-watt bridge power supply designed around an inexpensive "TV-replacement" type power transformer. The secondary winding is 1200 volts center-tapped at a current rating of 200 ma. The weight of the transformer is 8 pounds, and the maximum IVS rating is about 500 watts or so. Secondary resistance is 100 ohms. Used in bridge service, the transformer makes practical an inexpensive power supply providing

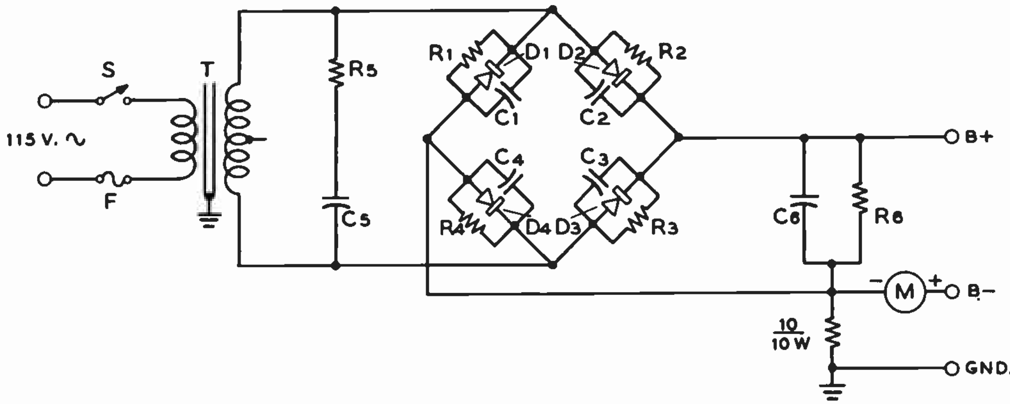


Figure 19

SCHEMATIC OF 500-WATT IVS BRIDGE POWER SUPPLY

Diode package (C_1-D_1-R_1, etc.) is composed of one each: 1N2071 diode in parallel with .01 μfd, 600-volt ceramic capacitor and a 470K, ½-watt resistor. Each bridge arm requires six packages, made as shown in figures 24 and 25. The secondary voltage-surge network (C_5-R_5) is a 100-ohm, 10-watt resistor in series with a .02 μfd, 3 kV capacitor (Aerovox P89-M). The power transformer has a 1200-volt center tapped 200-ma rating. (Stancor PC-8414 or Thordarson 22R36). The filter stack uses four 120-μfd, 450-volt electrolytic capacitors in series, with 10K, 10-watt resistors across each capacitor. Meter (M) is a 0-500 d-c milliammeter. A 10-ampere fuse (F) is used. Transformer core is grounded as a safety measure.

about 1250 volts at an IVS peak current rating of 380 ma. The no-load voltage is about 1600. For c-w use, the current rating is 225 ma at 1400 volts (about 300 watts). Maximum PIV is nearly 1700 volts so each arm of the bridge must withstand this value. Allowing a 100-percent safety factor requires 3400 volts PIV per arm, which may be made up of six 600-volt PIV diodes in series with an appropriate RC network across each diode. The diode assembly is constructed on two phenolic boards, one of which is shown in figures 21 and 22. A total of 24 rectifiers are required. Four 120-μfd, 450-volt electrolytic capacitors in series

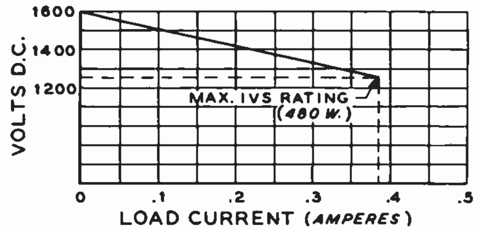


Figure 20

VOLTAGE-REGULATION CURVE OF 500-WATT BRIDGE POWER SUPPLY

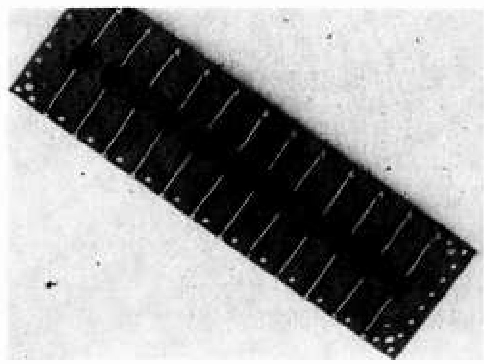


Figure 21

ASSEMBLY OF HIGH-VOLTAGE DIODE STACK

Inexpensive "TV-type" diodes may be connected in series to provide a high value of peak-inverse voltage. Shown here are twelve type-1N2071 diodes mounted on a Vectorbord (64AA32 cut to size). The diodes are soldered to Vector terminals (T9.6) mounted in the prepunched holes in the phenolic board. A pair of long-nose pliers should be used as a heat sink when soldering the diode leads. Grasp the diode lead between the diode body and the joint, permitting the pliers to absorb the soldering heat.

provide 30 μfd at a working voltage of 1800. The negative of the supply is above ground by virtue of the 10-ohm, 10-watt resistor which permits plate-current metering in the negative power lead while the supply and amplifier remain at the same ground potential.

This supply is designed for use with two 811A's in grounded-grid service. The tubes are biased to plate-current cutoff in standby mode by a cathode resistor which is shorted

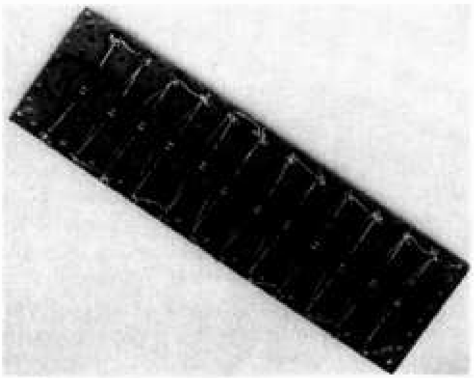


Figure 22

REAR VIEW OF HIGH-VOLTAGE DIODE STACK

The shunt capacitors and resistors are mounted on the rear of the phenolic board. Each diode-resistor-capacitor package has an individual pair of mounting terminals, which are jumpered together to connect the diodes in series. This arrangement provides greatest available heat sink for the components. The assembly is mounted an inch or so away from the chassis by means of 4-40 machine screws and ceramic insulators placed in corners of the board.

out by contacts on the push-to-talk or VOX circuitry. The power supply is built in an inclosed amplifier cabinet, similar to the one shown in figure 16. The B-plus lead is made of a length of RG-8/U coaxial cable, used in conjunction with a high-voltage coaxial connector.

A Heavy Duty Primary Supply This husky power supply provides a nominal 12 volts d-c at a maximum continuous current of 10 amperes. It is useful as a shop supply to test mobile gear, as a battery charger, and as a general-purpose low-voltage power pack. The supply is unregulated and depends solely on the single-section filter for ripple reduction. Regulation is quite good at a current drain over one ampere, as seen in figure 23. The output voltage is controlled by the primary *powerstat*. To alert the user to the unloaded supply voltage (which may rise as high as 30 volts when the primary voltage is high) a meter protection and "alert" circuit is added. The red lamp is lit when more than 20 volts is present at the output terminals of the supply. Below 20 volts, the zener diode is nonconducting. Above 20 volts, the 10-volt zener conducts and the current through it turns the NPN transistor on and lights the warning indicator.

23-9 Regulated Power Supplies

Zener diodes or *voltage-regulator tubes* are commonly used to regulate power supplies to discrete voltages. Electronic voltage regulators have been developed that will handle higher voltage and current variations than the tube and diode devices are capable of handling. The electronic circuits, moreover, may be varied over a wide range of output voltage.

Electronic voltage regulators, in the main, are based on feedback circuits, such as discussed in Chapter 8, Section 7 whereby an error signal is passed through the feedback loop in such a manner as to cause an adjustment to reduce the value of the error signal.

Special integrated circuits have been developed for voltage-regulator service such as the LM300 and the μA-723. The IC regulator provides the gain required for the feedback loop and an auxiliary power transistor passes the major portion of the regular current. The μA-723 and the improved LM305 are shown as series positive regulators with built-in current limiting in figure 25A-B. A negative regulator using an LM304 is shown in figure 25C.

A positive regulator circuit capable of handling several hundred milliamperes (if properly heat-sinked) is shown in figure 26. No external pass transistor is required. This IC regulator is designed for floating regulation and can be powered by a small secondary 25-volt supply that "floats," such as shown

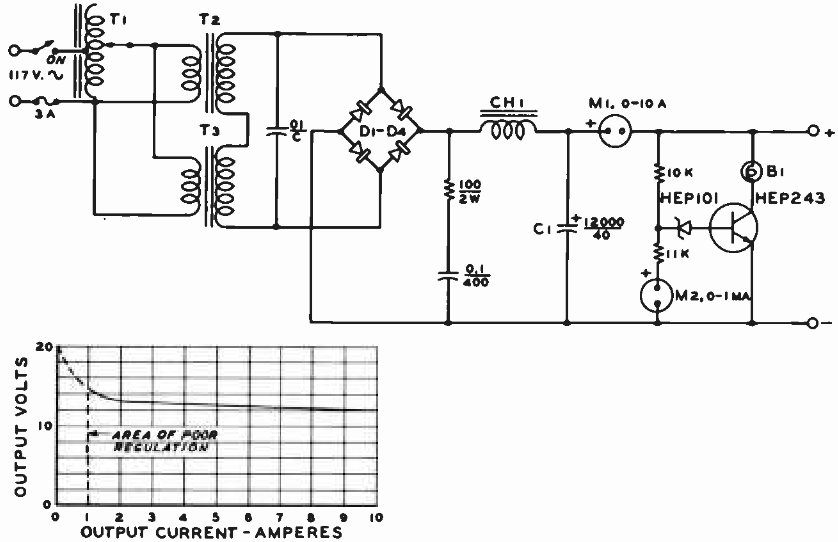


Figure 23

12-VOLT, 10-AMPERE GENERAL PURPOSE PRIMARY SUPPLY

B_1—1 amp, 28 volts. Chicago #327
C_1—12,000 µfd, 40 volts. Sprague 123G040BC
CH_1—.03 Henry, 10-ampere. Triad C-49U
D_1-D_4—Two 1N3209 and two 1N3209R. Use two Thermalloy heatsinks, 6500B-2
T_1—Powerstat, 200 watts, Superior 10B
T_2, T_3—11 volts, 10 amperes. Stancor P-3020
Meters: Weston model 301

in figure 27. In this configuration, the IC never has the main supply voltage across it and the only semiconductor that must stand-off the main supply voltage is the series pass transistor (usually a Darlington Pair). In this manner, the MC1466 may be used to regulate any voltage, high or low, and it also allows the output voltage to be varied from zero to maximum.

A number of small three-terminal IC regulators are available having fixed output voltages for the more commonly used circuit supply voltages. The LM 309, LM 335 and µA7805 are all 5-volt, 1-ampere regulators with built-in current limiting. They are available in a TO-3 can, which is grounded in normal operation, providing a negative return and also a heat sink to the chassis of the equipment. The µA 7800 series, in addition to the 5-volt type, also offers units that regulate at 6, 10, 12, 15, 18, 20, and 24 volts. One ampere is typical

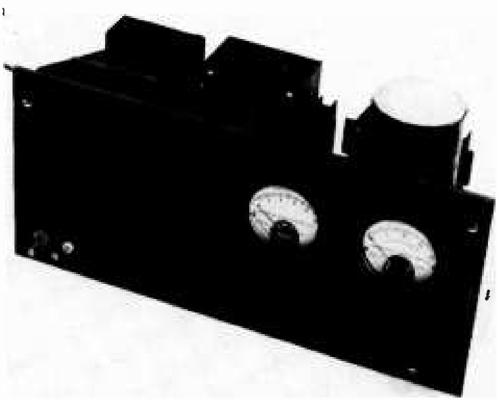


Figure 24

PRIMARY POWER SUPPLY

Handy to test mobile equipment, charge batteries or run surplus equipment, this supply provides 12 volts at 10 amperes with good regulation. Over-voltage lamp for meter protection is included.

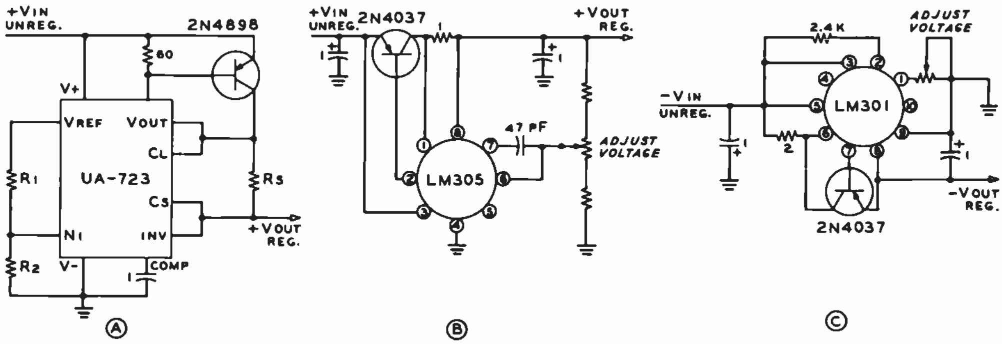


Figure 25

IC REGULATED POWER SUPPLIES

A—μA723 integrated circuit provides gain for feedback loop to 2N4898 pass transistor for series positive regulator. B—LM305 and 2N4037 provides simple adjustable positive voltage regulator. C—LM301 and 2N4037 serve as adjustable negative voltage regulator.

pass current of the 5-volt versions, with somewhat less for the higher-voltage types.

Voltage-Regulator Tubes A voltage-regulator tube (VR tube) is a gaseous device which maintains a constant voltage across its electrodes under conditions of varying supply current. A number of tube types are available which stabilize

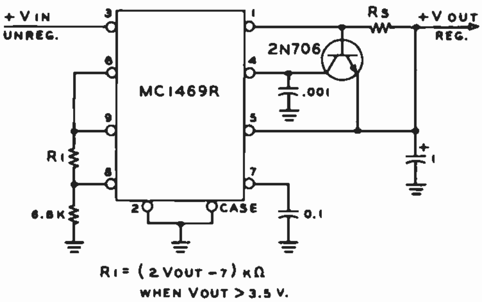


Figure 26

HEAVY-DUTY IC REGULATOR FOR POSITIVE VOLTAGE

Heat sinked MC1469R provides regulated high current for voltages above 3.5 volts. No external pass transistor is required.

the voltage across their terminals at 75, 90, 105, or 150 volts. The regulator tube is connected in series with a current-limiting resistor of such value that will permit the regulator tube to draw from 8 to 40 ma under normal operating conditions. The tube must be supplied from a potential source that is higher than the starting, or ignition voltage of the tube (figure 28). Regulator-tube currents greater than 40 ma will shorten the life of the tube and currents lower than 5 ma or so will result in unstable regulation. A voltage excess of about 15 percent is required to ignite the tube and this is usually taken care of by the no-load voltage rise of the source supply.

The value of the limiting resistor must permit minimum tube current to flow, and at the same time allow maximum regulator-tube current to flow under conditions of no load current, as shown in the illustration.

When a VR tube is to be used to regulate the voltage applied to a circuit drawing *less than 15 ma* normal or average current, the simplest method of adjusting the series resistance is to remove the load and vary the series resistor until the VR tube draws about 40 ma. Then connect the load, and that is all there is to it. This method is particularly recommended when the load is a heater-type vacuum tube, which may not draw current for several seconds after the power supply is turned on. Under these conditions, the current through the VR tube will never exceed 40 ma even when it is running unloaded (while the heater tube is warming up and the power-supply rectifier has already reached operating temperature).

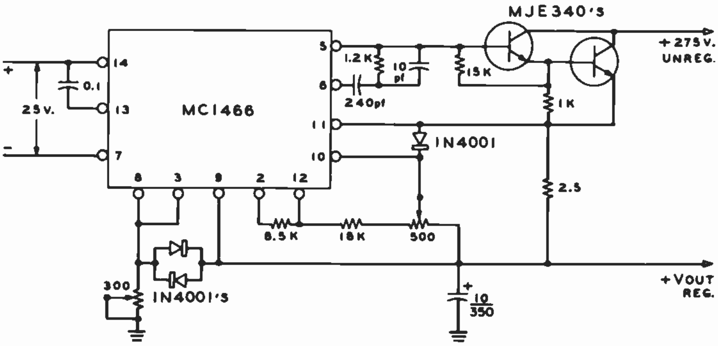
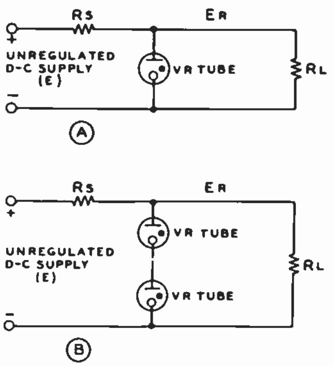


Figure 27

"FLOATING" IC REGULATOR

High-voltage IC regulator uses "floating" 25-volt supply. Series-pass transistors stand-off the main supply voltage. This circuit also allows the output voltage to be varied from zero to maximum value.



$$R_S = \frac{(E - E_R)}{I}$$

Figure 28

VOLTAGE-REGULATOR TUBE CIRCUITS

A—Single regulator tube stabilizes voltage at discrete intervals between 90 and 300 volts. B—Series-connected tubes offer stabilization up to 300 volts. Series resistor (R_S) is a function of supply voltage (E) and regulated voltage (E_R).

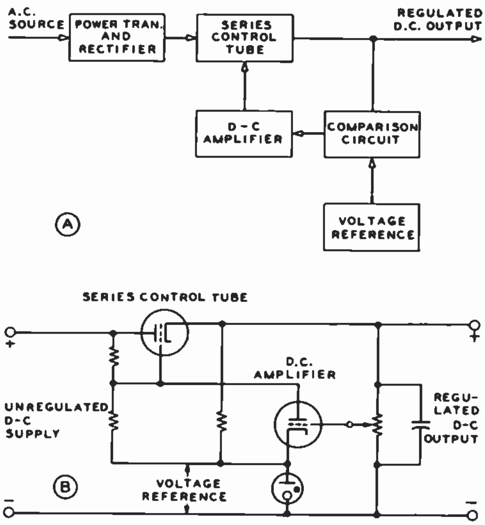


Figure 29

SERIES-REGULATED D-C POWER SUPPLY

D-c amplifier compares the output voltage of power supply to a voltage reference source. Voltage drop through series control tube is adjusted to balance circuit, providing voltage regulation of 1% or better

The Vacuum-Tube Regulator Voltage regulation may be accomplished by the use of *series control tube* and a voltage sensing and comparison circuit, as shown in figure 29. The series tube must be capable of dissipating power represented by the difference between the input voltage from the supply and the output voltage from the regulator at the maximum current flow to the load. In many cases, tubes are operated in parallel to obtain the required plate dissipation. The ouptut voltage of the electronically regulated supply may be changed over a wide range by varying the grid voltage of the d-c amplifier tube. The reference voltage may be supplied from a battery or voltage-regulator tube.

The d-c amplifier compares the output voltage to that of the reference source. When the output voltage drops, the d-c amplifier is unbalanced and the tube draws less plate current, thus raising the grid voltage on the series-connected control tube. The voltage drop through the control tube becomes less and the output voltage from the supply is raised, compensating for the original voltage reduction.

Practical electronic regulated supplies usually employ tetrode tubes in the d-c amplifier for higher amplifier gain and low-μ series control tubes for better control of regulation, providing regulation of the order of plus or minus 1 percent or so.

Three Regulated Supplies Shown in figure 30 are three small, inexpensive regulated power supplies designed by W6GXN that are useful for work with solid-state equipment. The first low-voltage supply (figure 31A) provides regulated 9 volts at 300 ma and may be used to power the whole gamut of little transistorized consumer electronic devices normally powered by batteries as well as some specialized f-m and vhf receivers operating in this power range. The supply provides a nominal 9 volts, regulated to 0.2 volt up to approximately 300 ma current drain.

A compact 5-volt, 1-ampere regulated supply suitable for operating digital IC circuits is shown in figure 31B and figure 32. Since DTL (diode-transistor-logic) and TTL (transistor-transistor-logic) both operate from +5 volts and represents the most popular two of the various IC logic families, this supply should take care of powering most digital systems. The supply includes current limiting at 1 ampere. The NE550L regulator is the heart of the supply and yields more "regulation per dollar" than almost any discrete circuit that can be built. The value of the series resistor (shown as 0.5 ohm) determines the current-limit point. This is about 1 ampere and decreases as the resistor is increased in value. A 5-ohm resistor will current-limit the supply at 100 ma. The ratio of the series-connected resistors across the output of the supply and the voltage impressed on pin #2 of the IC determines the value of the output voltage.

For powering a wide variety of linear ICs, especially operational amplifiers, the supply of figure 31C and figure 33 provides plus and minus 15 volts at 300 ma. A dual regulator IC, the SG 3501D, is used. As with the IC supply previously described, current limiting is provided for each of the two outputs. The two 2-ohm series resistors in the circuit are the controlling elements for current limiting, which is set at 300 ma because of the current capability of the particular transformer used. Note the use of the IC silicon bridge rectifier as a plus-and-minus full-wave rectifier. The center tap of the transformer is used, unlike the ordinary bridge connection.

In both the 5-volt and the plus-and-minus 15-volt regulated supplies the voltage output is constant until the current-limit point is reached, then the voltage value decreases abruptly.

A Variable-Voltage Supply With Current Limiting Although the simpler supplies described in the previous section are very useful for the specific voltage requirements most often encountered, it is helpful to have a continuously variable power supply for experimental purposes. Shown in figure 34 is a "bench supply" which provides 0 to 20 volts with current limiting up to 200 ma. The small

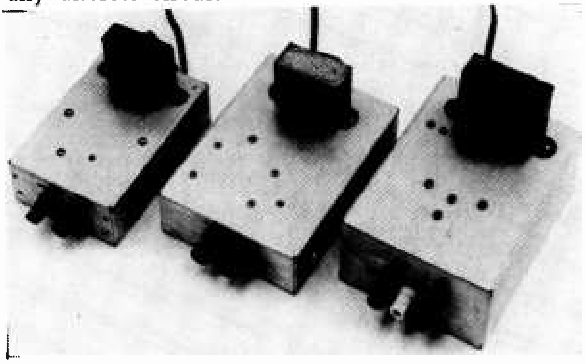


Figure 30

THREE HANDY REGULATED POWER SUPPLIES

Compact supply at left provides 9 volts at 300 ma for small transistorized equipment. Supply at center provides 5 volts at 1 ampere for digital IC circuitry. Supply at right provides +15 and −15 volts for linear ICs and operational amplifiers.

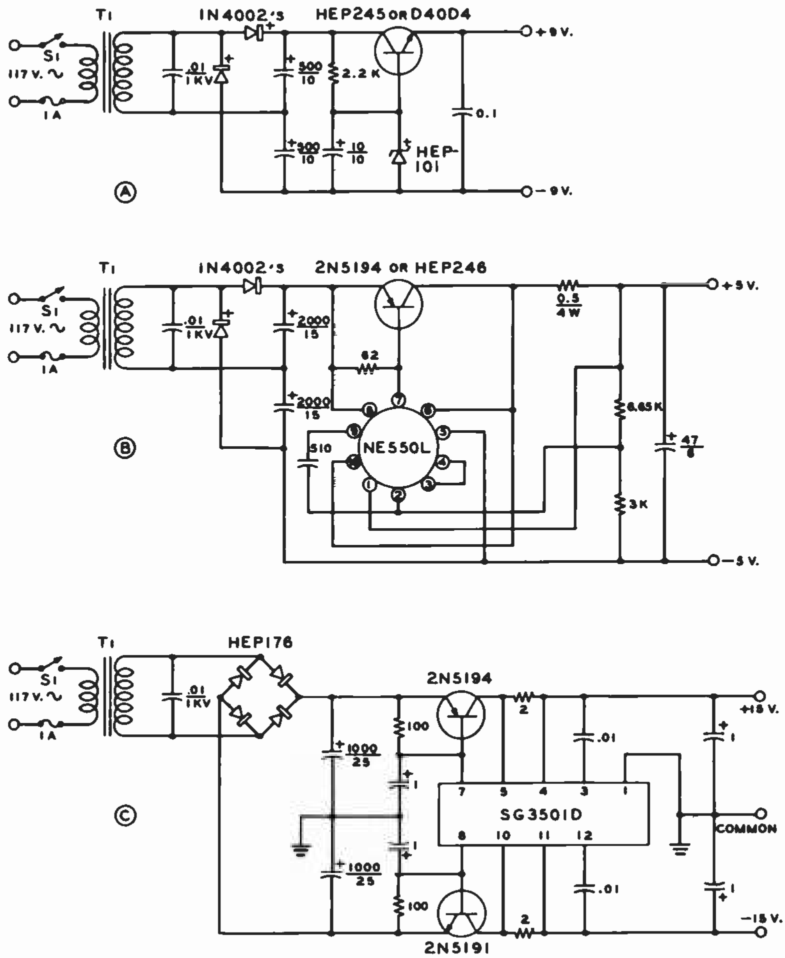


Figure 31

THREE REGULATED POWER SUPPLIES

A—9 volts, 300 milliamperes. T_1 is 6.3 volts, 0.6 ampere. Triad F-13X. B—5 volts, 1 ampere. T_1 is 6.3 volts, 3 amperes. Stancor P-6466. C—15-volt, dual supply. T_1 is 40/20 volts, ct., 300 ma. Triad F-91X.

size of the supply makes it convenient to use even if the builder has only a tiny corner of his operating desk on which to make experimental gear.

The supply is designed around the MC 1466L regulator IC which operates from a "floating' 25-volt source to control another supply of arbitrary voltage. This concept is especially useful where the supply covers the range down to *zero* volts. A small dual-winding transformer that mounts on a printed-circuit board is used (figure 35).

The supply is sufficiently complex so that use of a printed-circuit board is suggested and an etched and drilled board is available from *Southwest Technical Products, Inc.*, 219 West Rhapsody, San Antonio, Texas, as well as a complete kit of parts.

Switch S_{2A} places a 39-ohm resistor in series with the pass transistor, Q_1, which limits the collector dissipation of the device when operating at low voltage and high current. The other section of the switch selects the correct multiplier for the voltmeter to provide either 10 or 20 volts full scale. The switch should be set to the lower

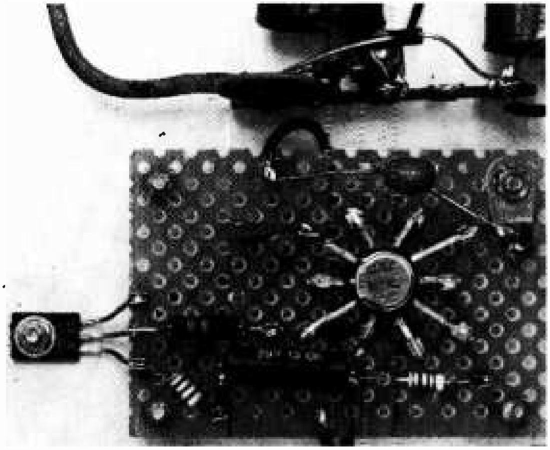


Figure 32

UNDER-CHASSIS VIEW OF 5-VOLT SUPPLY

The HEP 246 pass transistor is insulated from the chassis with mica washers. NE 550L IC and associated components are soldered to pins mounted in small perforated circuit board.

voltage when the supply is used below a 10-volt output level.

The supply is placed within a 4″ × 4″ × 4″ aluminum utility box chassis. The Darlington Pair pass transistor (Q_1) is heat-sinked to the front panel of the box with a mica washer and a nylon 4-40 screw, while the fuse holder and a-c power switch are on the rear of the box to keep their field away from the high-gain circuitry at the front of the assembly (figure 36).

A "Mobile" Power Supply This compact, regulated power supply provides 12.6 volts at 2 amperes and is designed to be used with 10-watt, 2- and 6-meter f-m transceivers, auto radios, and other d-c powered devices in the 20-watt primary power range (figure 37).

At the maximum current limit of 2 amperes a warning light (B_1) is turned on, showing the user that he is getting close to the maximum power capability of the supply. At current levels below the maximum, regulation is in effect and the output voltage remains within 0.5 volt of the nominal value of 12.6 volts. A germanium transistor (Q_1) is used in the indicator circuit, allowing a smaller value of overcurrent-sensing resistor (R_1) to be used. The main pass-current transistor (Q_2), is an inexpensive germanium unit and can be used in this positive regulator circuit because it is used in a complementary pair with an NPN transistor (Q_3). Unlike the Darlington Pair, this configuration has only one emitter-base drop between the output and the controlling base. The output is adjustable around a

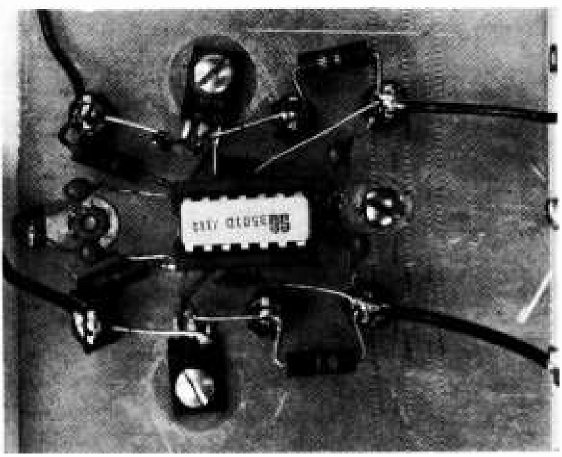


Figure 33

UNDER-CHASSIS VIEW OF 15-VOLT SUPPLY

The two 2N5194 (HEP 246) transistors are insulated from the chassis with mica washers. The IC is supported by its leads from various nearby components.

Figure 34

COMPACT 20-VOLT REGULATED SUPPLY FOR LABORATORY WORK

The supply provides 0 to 20 volts at 200 milliamperes with current limiting. Meter range may be switched between 10 and 20 volts full scale. A "floating" regulator circuit is used to allow the range to be extended down to zero volts.

nominal 12 volts by plus or minus 1 volt. At full load, there is less than 0.2-volt ripple in the output, which is usually tolerable in primary sources. The complete schematic is shown in figure 38 and the various under-chassis components are shown in figure 39.

A Medium-Voltage Regulated Supply (150 to 250 volts) A stable, voltage-regulated power supply is a useful adjunct to the ex-experimenters workshop for use with receivers, test equipment, and other devices requiring controlled voltage. Shown in figure 40 is a small power supply that is well suited to this task. The unit delivers 250 volts at 60 ma and may be controlled down to 150 volts, at which point the maximum current is limited to 40 ma. A single 6JZ8 *Compactron* tube serves as a series regulator and d-c amplifier. A small NE-2 neon lamp connected in the cathode circuit of the triode section of the 6JZ8 provides reference voltage and may be used as a pilot light.

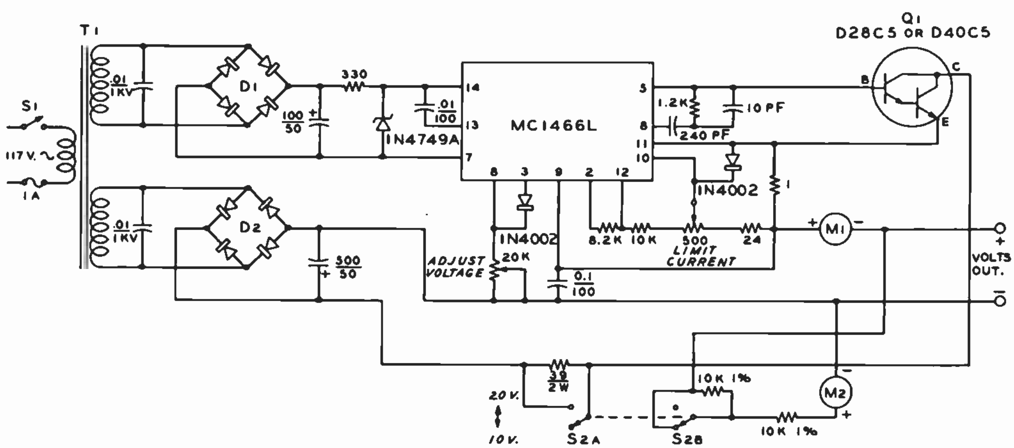


Figure 35

VARIABLE-VOLTAGE SUPPLY WITH CURRENT LIMITING

D_1, D_2—HEP 176. T_1—29, 20 volts, 250 milliamperes. Signal PC-40-250. Box chassis—LMB 444N.

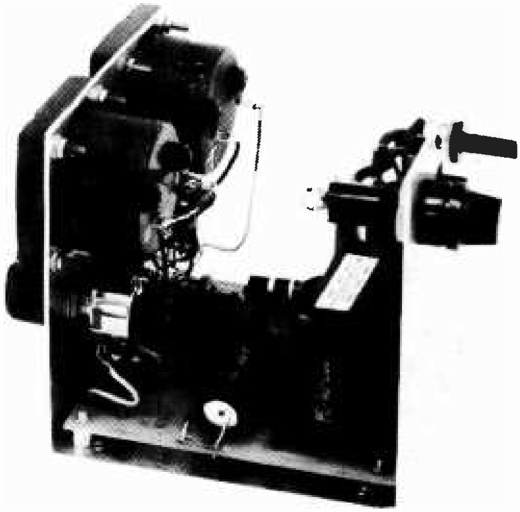


Figure 36

INTERIOR OF VARIABLE-VOLTAGE SUPPLY

Small components are mounted on printed-circuit board. Darlington Pair transistor Q_1 is heat-sinked to front panel with a mica washer and nylon screw.

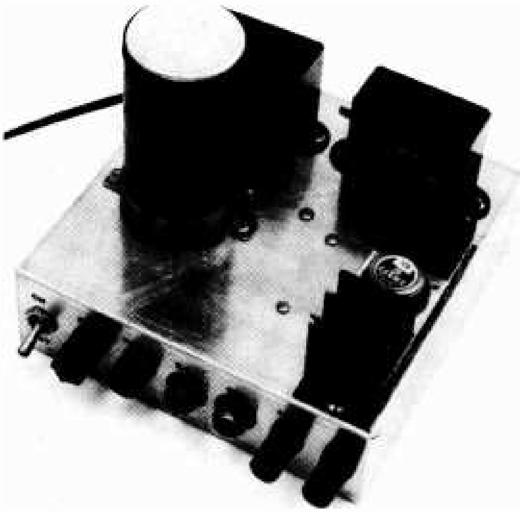


Figure 37

MOBILE (12-VOLT) REGULATED SUPPLY

This supply provides 12.6 volts at maximum current of 2 amperes and is designed to be used with 2- and 6-meter f-m equipment. A red warning light turns on when user draws too much current from the supply. 2N277 (HEP 231) pass transistor is mounted on a heat sink with mica washers.

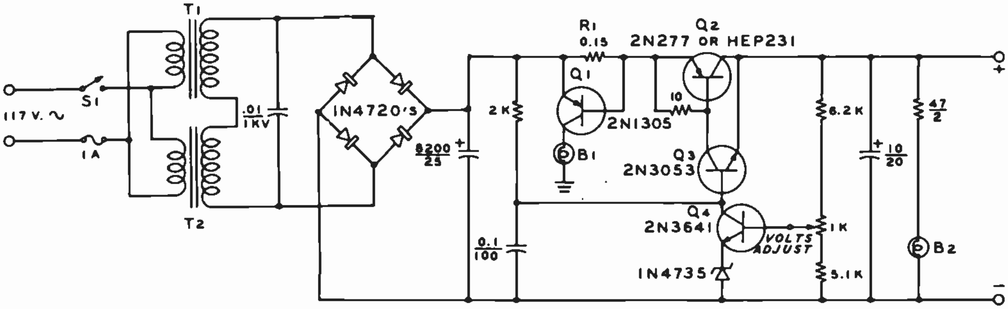


Figure 38

SCHEMATIC, 12-VOLT "MOBILE" POWER SUPPLY

T_1, T_2—6.3-volt, 3-ampere. Stancor P-6466. B_1—12-volt pilot lamp.

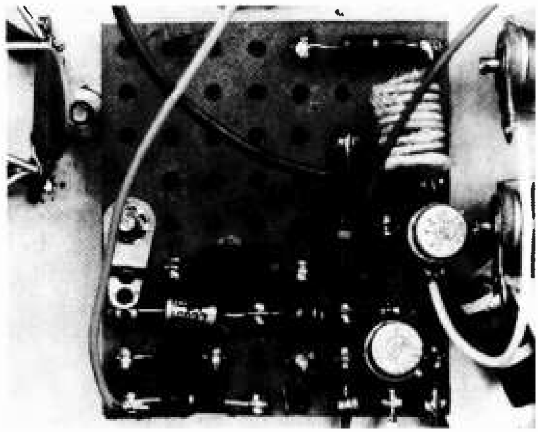


Figure 39

COMPONENT BOARD OF 12-VOLT SUPPLY

Resistor R_1 (0.15 ohm), made of a small coil of resistance wire, is seen in the upper right corner of the board.

23-10 Transceiver Power Supplies

Single-sideband transceivers require power supplies that provide several values of high voltage, bias voltage, filament voltage, and d-c control-circuit voltage. The supply may provide up to 600 watts of d-c power in intermittent voice service. The use of high-storage "computer"-type electrolytic capacitors permits maximum power to be maintained during voice peaks, while still permitting the power transformer to be operated within an average power rating of about 50-percent peak power capabality, even for extended periods of time.

Two transceiver power supplies are shown in this section. The first is designed around a power transformer specially built for SSB service. The second supply is designed around a heavy-duty "TV replacement" type power transformer. The former supply is capable of a PEP power level of better than 600 watts, while the latter design is limited to about 300 watts PEP.

A schematic of the 600-watt PEP power supply is shown in figure 42. A multiple-winding transformer is used which has sufficient capacity to run the largest transceivers on a continuous voice-operated basis. The transformer weighs 16 pounds and has great reserve capacity. The power supply provides 800 volts at an intermittent current of 800 milliamperes, 250 volts at an intermittent current of 200 milliamperes, an adjustable bias voltage at a continuous current of 100 milliamperes, and either 6.3 volts or 12.6 volts filament supply at 12 or 6 amperes, respectively. An additional circuit provides 12 volts d.c. for operation of auxiliary VOX or switching relays. Controlled-avalanche diodes are used in the bridge-rectifier circuit, in conjunction with RC shunt networks and transient suppression across the power-supply secondary winding.

Additional transient protection is afforded by large bypass capacitors placed on the primary winding of the power transformer.

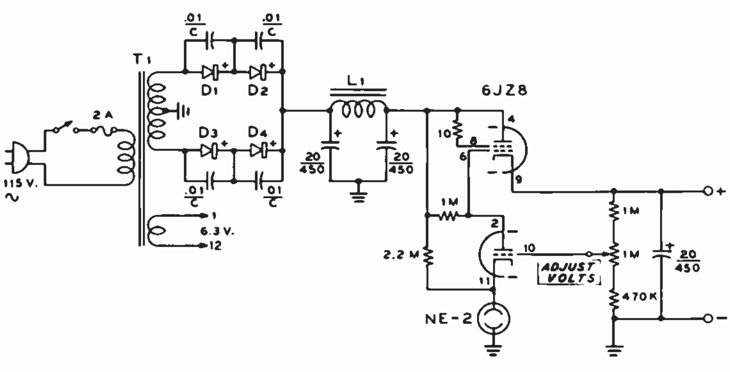


Figure 40

MEDIUM-VOLTAGE REGULATED SUPPLY

D_1 thru D_4—IN4005 or equivalent
T_1—480 volts, c.t. at 70 ma, 6.3 volts at 3 amps
L_1—8 henrys, 75 ma

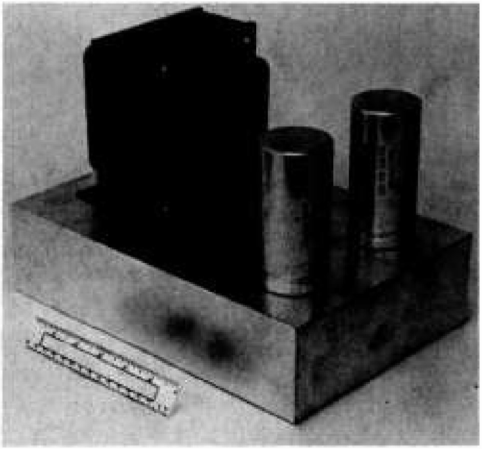


Figure 41

600-WATT IVS POWER SUPPLY FOR SSB TRANSCEIVERS

Special transceiver power supply provides heavy-duty capacity to run largest of SSB transceivers. Power transformer and filter choke are to the left, with bias-adjustment potentiometer in foreground. Multiwire cable connects supply to transceiver.

The supply is actuated by a remote-power-line switch, usually located in the transceiver.

The construction of the supply is shown in figure 41. The aluminum chassis is small enough to fit within the speaker cabinet of the transceiver, and parts layout it not critical. The rectifier bridge is assembled on a phenolic board, and mounted below the chassis in a clear area. The filter capacitors are mounted to a phenolic board, their terminals protruding into the under-chassis area.

All voltage connections are terminated on a connector strip, and a single power cable may be run from the power supply to the transceiver. The leads carrying the filament voltage should be doubled up, using two wires for each lead to reduce voltage drop within the cable to a minimum. The 6.3-volt filament windings of the transformer may be arranged in either series or parallel configuration, according to the requirements of the transceiver.

Complete filter-capacitor discharge takes about 10 seconds once the supply is turned

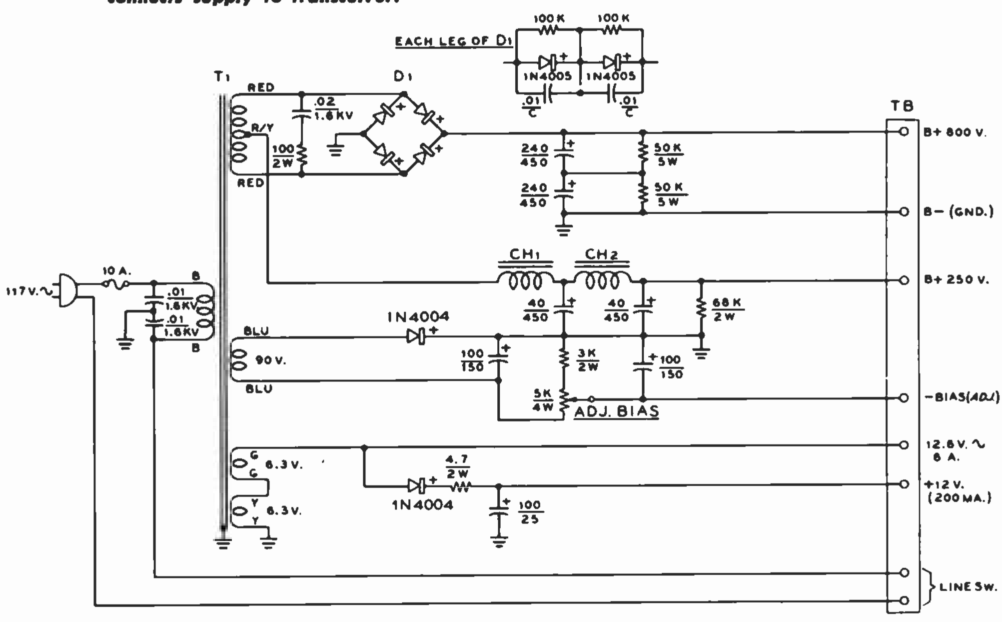


Figure 42

SCHEMATIC, 600-WATT TRANSCEIVER SUPPLY

T_1—600 volts, 400 ma; 250 volts, 100 ma; 6.3 volts, 6 amps; 6.3 volts, 6 amps, 117-volt primary. Triad P-31A
CH_1—1 henry, 300 ma
CH_2—3 henrys, 300 ma

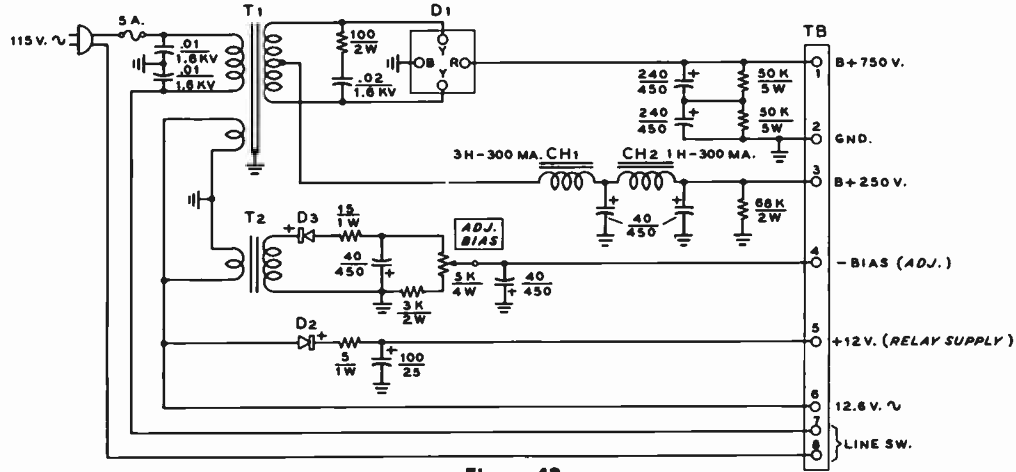


Figure 43

SCHEMATIC, 300-WATT IVS TRANSCEIVER POWER SUPPLY

Various replacement power transformers may be used with this power supply. Suggested units are: (1) 650-volt c.t. at 225 ma.; 12.6-volt at 5.25 amp. (Stancor P-8339), for 650-volt d-c output. (2) 750-volt c.t. at 325 ma.; 12.6 volt at 6.0 amp. (Stancor P-8365), for 750-volt d-c output. (3) 540-volt c.t. at 260 ma.; 6.3-volt at 8.8 amp. (Stancor P-8356), for 600-volt d-c output and 6.3 volt filament supply.

Transformer T_2 6.3 volts at 1 amp. (Stancor P-8389). CH_1: 3 henrys at 300 ma (Stancor C-2334). CH_2: 1 henry at 300 ma (Stancor C-2343). D_1: Diode bridge, 1400-volt rms, 1.5 amp (2000-volt PIV). Diodes Inc. #BR-820A. D_2, D_3: 1N2070.

off, and it is recommended that the capacitor stack be shorted with a 1000-ohm 100-watt resistor before any work is done on the supply.

An inexpensive utility power supply may be constructed about a "TV replacement" transformer, using auxiliary transformers, as needed, for filament and bias supplies, as shown in figure 43. The filament voltage is stepped up to 117 volts by a reverse-connected filament transformer (T_2) and is rectified to provide adjustable bias voltage. The power supply delivers 600 to 750 volts at 400 milliamperes peak current, and about 250 volts at 200 milliamperes peak current. Depending on choice of power transformer,

This compact IVS-rated power supply provides all operating voltages necessary to operate most popular SSB transceivers. The supply uses a "TV-replacement" power transformer in conjunction with a bridge-rectifier circuit. The unit is designed to be placed in the speaker cabinet of the transceiver, and the chassis should be shaped to custom-fit the particular speaker cabinet in use. If desired, the supply may be built on a chassis with a dust cover and placed beneath the station console.

The power transformer is to the left, with the 240-µfd, 450-volt filter capacitors in the foreground. The capacitors are mounted to a phenolic plate which is bolted to the chassis. The two filter chokes are to the rear, along with the low-voltage filter capacitors and the "adjust-bias" potentiometer. The reverse-connected filament transformer is at the rear of the chassis. Semiconductor rectifiers are placed beneath the chassis.

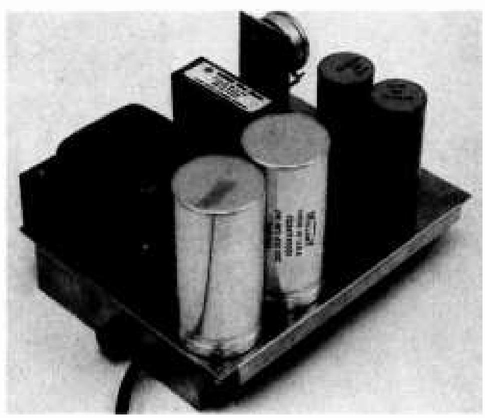


Figure 44

300-WATT IVS POWER SUPPLY FOR SSB TRANSCEIVERS

either 6.3- or 12.6-volt filament supply may be provided, in addition to low-voltage d.c. for operation of VOX or control relays. Layout of the supply is shown in figure 44. The unit is constructed on a home-made aluminum chassis contoured to fit within a speaker cabinet.

CHAPTER TWENTY-FOUR

Radiation, Propagation, and Transmission Lines

Radio waves are electromagnetic waves similar in nature to, but much lower in frequency than, light waves or heat waves. Such waves represent electric energy traveling through space. Radio waves travel in free space with the velocity of light and can be reflected and refracted much the same as light waves.

24-1 Radiation from an Antenna

Alternating current passing through a conductor creates an alternating electromagnetic field around that conductor. Energy is alternately stored in the field, and then returned to the conductor. As the frequency is raised, more and more of the energy does not return to the conductor, but instead is radiated off into space in the form of electromagnetic waves, called radio waves. Radiation from a wire, or wires, is materially increased whenever there is a sudden *change* in the *electrical constants* of the line. These sudden changes produce reflection, which places *standing waves* on the line.

When a wire in space is fed radio-frequency energy having a wavelength of approximately 2.1 times the length of the wire in meters, the wire *resonates* as a *half-wave dipole* antenna at that wavelength or frequency. The greatest possible change in the electrical constants of a line is that which occurs at the open end of a wire. Therefore, a dipole has a great mismatch at each end, producing a high degree of reflection. We say that the ends of a dipole are terminated in an *infinite impedance.*

A returning wave which has been reflected meets the next incident wave, and the voltage and current at any point along the antenna are the vector sum of the two waves. At the ends of the dipole, the voltages add, while the currents of the two waves cancel, thus producing *high voltage* and *low current* at the *ends* of the dipole or half-wave section of wire. In the same manner, it is found that the currents add while the voltages cancel at the center of the dipole. Thus, at the *center* there is *high current* but *low voltage.*

Inspection of figure 1 will show that the current in a dipole decreases sinusoidally toward either end, while the voltage similarly increases. The voltages at the two ends of the antenna are 180° out of phase, which means that the polarities are opposite, one being plus while the other is minus at any instant. A curve representing either the voltage or current on a dipole represents a *standing wave* on the wire.

Radiation From Sources Other Than Antennas Radiation can and does take place from sources other than antennas. Undesired radiation can take place from open-wire transmission lines, both from sin-

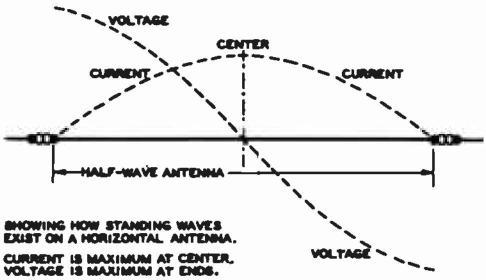


Figure 1

STANDING WAVES ON A RESONANT ANTENNA

gle-wire lines and from lines comprised of more than one wire. In addition, radiation can be made to take place in a very efficient manner from electromagnetic horns, from plastic lenses or from electromagnetic lenses made up of spaced conducting planes, from slots cut in a piece of metal, from dielectric wires, or from the open end of a waveguide.

Directivity of Radiation The radiation from any physically practical radiating system is directive to a certain degree. The degree of directivity can be enhanced or altered when desirable through the combination of radiating elements in a prescribed manner, through the use of reflecting planes or curved surfaces, or through the use of such systems as mentioned in the preceding paragraph. The construction of directive antenna arrays is covered in detail in the chapters which follow.

Polarization Like light waves, radio waves can have a definite polarization. In fact, while light waves ordinarily have to be reflected or passed through a polarizing medium before they have a definite polarization, a radio wave leaving a simple radiator will have a definite polarization, the polarization being indicated by the orientation of the electric-field component of the wave. This, in turn, is determined by the orientation of the radiator itself, as the magnetic-field component is always at right angles to a linear radiator, and the electric-field component is always in the same plane as the radiator. Thus we see that an antenna that is vertical with respect to the earth will transmit a vertically polarized wave, as the electrostatic lines of force will be vertical. Likewise, a simple horizontal antenna will radiate horizontally polarized waves.

Because the orientation of a simple linear radiator is the same as the polarization of the waves emitted by it, the radiator itself is referred to as being either vertically or horizontally polarized. Thus, we say that a horizontal antenna is horizontally polarized.

Figure 2A illustrates the fact that the polarization of the electric field of the radiation from a vertical dipole is vertical. Figure 2B, on the other hand, shows that the polarization of electric-field radiation from a vertical *slot* radiator is horizontal. This fact has been utilized in certain commercial f-m antennas where it is desired to have horizontally polarized radiation but where it is more convenient to use an array of vertically stacked slot arrays. If the metallic sheet is bent into a cylinder with the slot on one side, substantially omnidirectional horizontal coverage is obtained with horizontally polarized radiation when the cylinder with the slot in one side is oriented vertically. An arrangement of this type is shown in figure 2C. Several such cylinders may be stacked vertically to reduce high-angle radiation and to concentrate the radiated energy at the useful low radiation angles.

In any event the polarization of radiation from a radiating system is parallel to the electric field as it is set up inside or in the vicinity of the radiating system.

24-2 General Characteristics of Antennas

All antennas have certain general characteristics to be enumerated. It is the result of differences in these general characteristics which makes one type of antenna system most suitable for one type of application and another type best for a different application. Six of the more important characteristics are: (1) polarization, (2) radiation resistance, (3) horizontal directivity, (4) vertical directivity, (5) bandwidth, and (6) effective power gain.

The *polarization* of an antenna or radiating system is the direction of the electric field and has been defined in Section 20-1.

The *radiation resistance* of an antenna system is normally referred to the feed point

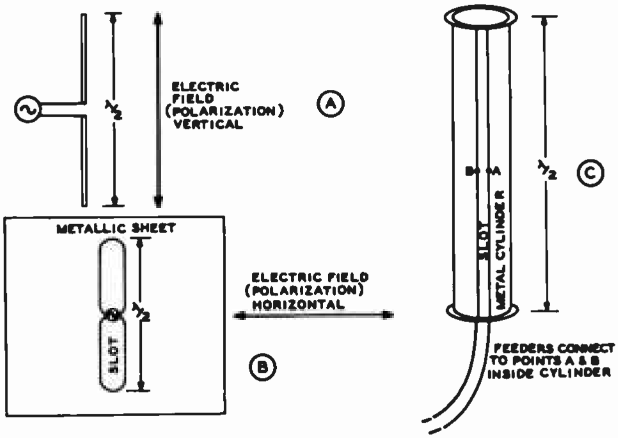


Figure 2

ANTENNA POLARIZATION

The polarization (electric field) of the radiation from a resonant dipole such as shown at A is parallel to the length of the radiator. In the case of a resonant slot cut in a sheet of metal and used as a radiator, the polarization (of the electric field) is perpendicular to the length of the slot. In both cases, however, the polarization of the radiated field is parallel to the potential gradient of the radiator; in the case of the dipole the electric lines of force are from end to end, while in the case of the slot the field is across the sides of the slot. The metallic sheet containing the slot may be formed into a cylinder to make up the radiator shown at C. With this type of radiator the radiated field will be horizontally polarized even though the radiator is mounted vertically.

in an antenna fed at a current loop, or it is referred to a current loop in an antenna system fed at another point. The radiation resistance is that value of resistance which, when substituted for the antenna at a current loop, would dissipate the same energy as is actually radiated by the antenna if the antenna current at the feed point were to remain the same.

The *horizontal* and *vertical directivity* can best be expressed as a *directive pattern* which is a graph showing the relative radiated field intensity against *azimuth* angle for horizontal directivity and field intensity against *elevation* angle for vertical directivity.

The *bandwidth* of an antenna is a measure of its ability to operate within specified limits over a range of frequencies. Bandwidth can be expessed as either *operating frequency plus or minus a specified percent of operating frequency*, or *operating frequency plus or minus a specified number of MHz* for a certain standing-wave-ratio limit on the transmission line feeding the antenna system.

The *effective power gain* or *directive gain* of an antenna is the ratio between the power required in the specified antenna and the power required in a reference antenna (usually a half-wave dipole) to attain the same field strength in the favored direction of the antenna under measurement. Directive gain may be expressed either as an actual power ratio, or as is more common, the power ratio may be expressed in decibels.

Physical Length of a Half-Wave Antenna If the cross section of the conductor which makes up the antenna is kept very small with respect to the antenna length, an electrical half wave is a fixed percentage shorter than a physical half wavelength. This percentage is approximately 5 percent. Therefore, most linear half-wave antennas are close to 95 percent of a half wavelength long physically. Thus, a half-wave antenna resonant at exactly 80 meters would be one-half of 0.95 times 80 meters in length. Another way of saying the same thing is that a wire resonates at a wavelength of about 2.1 times its length in meters. If the diameter of the conductor begins to be an appreciable fraction of a wavelength, as when tubing is used as a vhf radiator, the factor becomes slightly less than 0.95. For the use of wire and not tubing on frequencies below 30 MHz, however, the figure of 0.95 may be taken as accurate. This assumes a radiator removed from surrounding objects, and with *no bends*.

Simple conversion into feet can be obtained by using the factor 1.56. To find the physical length of a half-wave 80-meter antenna, we multiply 80 times 1.56, and get 124.8 feet for the length of the radiator.

It is more common to use frequency than wavelength when indicating a specific spot in the radio spectrum. For this reason, the relationship between wavelength and frequency must be kept in mind. As the velocity of radio waves through space is constant at the speed of light, it will be seen that the more waves that pass a point per second (higher frequency), the closer together the peaks of those waves must be (shorter wavelength). Therefore, the higher the frequency, the lower will be the wavelength.

A radio wave in space can be compared to a wave in water. The wave, in either case, has peaks and troughs. One peak and one trough constitute a *full wave*, or *one wavelength.*

Frequency describes the number of wave cycles or peaks passing a point per second. *Wavelength* describes the distance the wave travels through space during one cycle or oscillation of the antenna current; it is the distance in meters between adjacent peaks or adjacent troughs of a wave train.

As a radio wave travels 300,000,000 meters a second (speed of light), a frequency of 1 cycle per second (1 Hz) corresponds to a wavelength of 300,000,000 meters. So, if the frequency is multiplied by a million, the wavelength must be divided by a million, in order to maintain their correct ratio.

A frequency of 1,000,000 cycles per second (1000 kHz) equals a wavelength of 300 meters. Multiplying frequency by 10 and dividing wavelength by 10, we find: a frequency of 10,000 kHz equals a wavelength of 30 meters. Multiplying and dividing by 10 again, we get: a frequency of 100,000 kHz equals 3 meters wavelength. Therefore, to change wavelength to frequency (in kilohertz), simply divide 300,000 by the wavelength in meters (λ).

$$F_{kHz} = \frac{300,000}{\lambda}$$

$$\lambda = \frac{300,000}{F_{kHz}}$$

Now that we have a simple conversion formula for converting wavelength to frequency and vice versa, we can combine it with our wavelength-versus-antenna length formula, and we have the following:

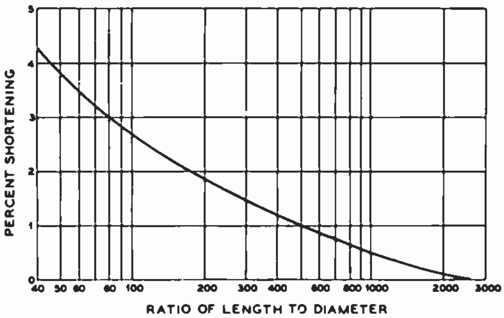


Figure 3

CHART SHOWING SHORTENING OF A RESONANT ELEMENT IN TERMS OF RATIO OF LENGTH TO DIAMETER

The use of this chart is based on the basic formula where radiator length in feet is equal to 475/frequency in MHz. This formula applies to frequencies below perhaps 30 MHz when the radiator is made from wire. On higher frequencies, or on 14 and 28 MHz when the radiator is made of large-diameter tubing, the radiator is shortened from the value obtained with the above formula by an amount determined by the ratio of length to diameter of the radiator. The amount of this shortening is obtainable from the chart shown above.

Length of a half-wave radiator made from wire (No. 14 to No. 10):

3.5-MHz to 30-MHz bands

$$\textit{Length in feet} = \frac{475}{\textit{Freq. in MHz}}$$

50-MHz band

$$\textit{Length in feet} = \frac{466}{\textit{Freq. in MHz}}$$

$$\textit{Length in inches} = \frac{5600}{\textit{Freq. in MHz}}$$

144-MHz band

$$\textit{Length in inches} = \frac{5600}{\textit{Freq. in MHz}}$$

Length-to-Diameter Ratio

When a half-wave radiator is constructed from tubing or rod whose diameter is an appreciable fraction of the length of the radiator, the resonant length of a half-wave antenna will be shortened. The amount of shortening can be determined with the aid of the chart of figure 3. In this chart the amount of additional shortening over the values given in the previous paragraph is plotted against the ratio of the length to the diameter of the half-wave radiator.

The length of a wave in free space is somewhat longer than the length of an antenna for the same frequency. The actual free-space half wavelength is given by the following expressions:

$$\textit{Half wavelength} = \frac{492}{\textit{Freq. in MHz}}\ \textit{in feet}$$

$$\textit{Half wavelength} = \frac{5905}{\textit{Freq. in MHz}}\ \textit{in inches}$$

Harmonic Resonance A wire in space can resonate at more than one frequency. The *lowest* frequency at which it resonates is called its *fundamental* frequency, and at that frequency it is approximately a half wavelength long. A wire can have two, three, four, five, or more standing waves on it, and thus it resonates at approximately the integral harmonics of its fundamental frequency. However, the higher harmonics are not exactly integral multiples of the lowest resonant frequency as a result of *end effects*.

A harmonic-operated antenna is somewhat longer than the corresponding integral number of dipoles, and for this reason, the dipole length formula cannot be used simply by multiplying by the corresponding harmonic. The intermediate half-wave sections do not have *end effects*. Also, the current distribution is disturbed by the fact that power can reach some of the half-wave sections only by flowing through other sections, the latter then acting not only as radiators, but also as transmission lines. For the latter reason, the resonant length will be dependent to an extent on the method of feed, as there will be less attenuation of the current along the antenna if it is fed at or near the center than if fed toward or at one end. Thus, the antenna would have to be somewhat longer if fed near one end than if fed near the center. The difference would be small, however, unless the antenna were many wavelengths long.

The length of a center-fed harmonically operated doublet may be found from the formula:

$$L = \frac{(K - .05) \times 492}{\textit{Freq. in MHz}}$$

where,

K equals number of ½ waves on antenna,
L equals length in feet.

Under conditions of severe current attenuation, it is possible for some of the nodes, or loops, actually to be slightly greater than a physical half wavelength apart. Practice has shown that the most practical method of resonating a harmonically operated antenna accurately is by cut and try, or by using a feed system in which both the feedline and antenna are resonated at the station end as an integral system.

A dipole or half-wave antenna is said to operate on its fundamental or first harmonic. A full-wave antenna, 1 wavelength long, operates on its second harmonic. An antenna with five half wavelengths on it would be operating on its fifth harmonic. Observe that the fifth harmonic antenna is 2½ *wavelengths* long, *not* 5 wavelengths.

Antenna Resonance Most types of antennas operate most efficiently when tuned, or *resonated*, to the frequency of operation. This consideration of course does not apply to the rhombic antenna and to the parasitic elements of Yagi arrays. However,

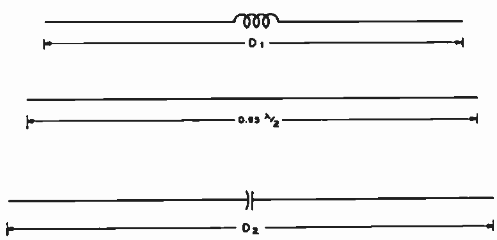


Figure 4

EFFECT OF SERIES INDUCTANCE AND CAPACITANCE ON THE LENGTH OF A HALF-WAVE RADIATOR

The top antenna has been electrically lengthened by placing a coil in series with the center. In other words, an antenna with a lumped inductance in its center can be made shorter for a given frequency than a plain wire radiator. The bottom antenna has been capacitively shortened electrically. In other words, an antenna with a capacitor in series with it must be made longer for a given frequency since its effective electrical length as compared to plain wire is shorter.

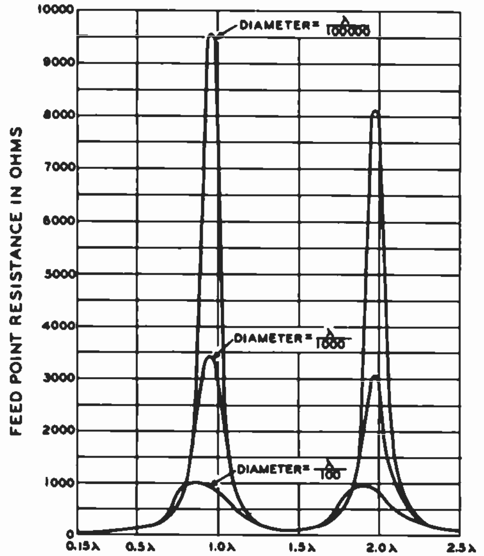


Figure 5

FEED-POINT RESISTANCE OF A CENTER-DRIVEN RADIATOR AS A FUNCTION OF PHYSICAL LENGTH IN TERMS OF FREE SPACE WAVELENGTH

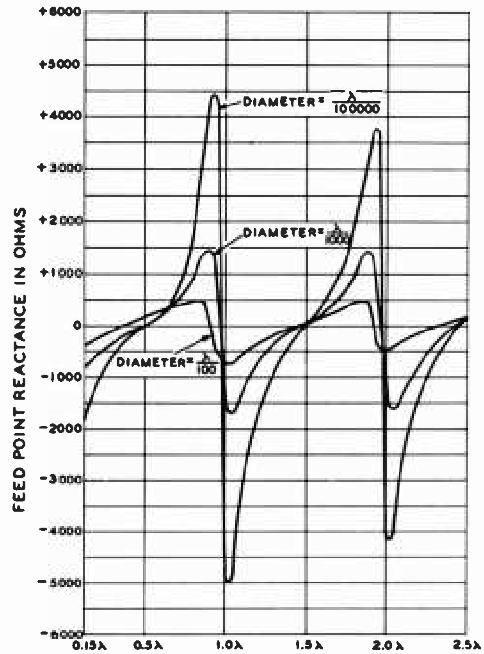


Figure 6

REACTIVE COMPONENT OF THE FEED-POINT IMPEDANCE OF A CENTER-DRIVEN RADIATOR AS A FUNCTION OF PHYSICAL LENGTH IN TERMS OF FREE-SPACE WAVELENGTH

in practically every other case it will be found that increased efficiency results when the entire antenna system is resonant, whether it be a simple dipole or an elaborate array. The radiation efficiency of a resonant wire is many times that of a wire which is not resonant.

If an antenna is slightly too long, it can be resonated by series insertion of a variable capacitor at a high-current point. If it is slightly too short, it can be resonated by means of a variable inductance. These two methods, illustrated schematically in figure 4, are generally employed when part of the antenna is brought into the operating room.

With an antenna array, or an antenna fed by means of a transmission line, it is more common to cut the elements to exact resonant length by "cut-and-try" procedure. Exact antenna resonance is more important when the antenna system has low radiation resistance; an antenna with low radiation resistance has higher Q (tunes sharper) than an antenna with high radiation resistance. The higher Q does not indicate greater efficiency; it simply indicates a sharper resonance curve.

24-3 Radiation Resistance and Feed-point Impedance

In many ways, a half-wave antenna is like a tuned tank circuit. The main difference lies in the fact that the elements of inductance, capacitance, and resistance are *lumped* in the tank circuit, and are *distributed* throughout the length of an antenna. The center of a half-wave radiator is effective at ground potential as far as r-f voltage is concerned, although the current is highest at that point.

When the antenna is resonant, and it always should be for best results, the impedance at the center is substantially resistive, and is termed the *radiation resistance*. Radiation resistance is a fictitious term; it is that value of resistance (referred to the current loop) which would dissipate the same amount of power as being radiated by the antenna, when fed with the current flowing at the current loop.

The radiation resistance depends on the antenna length and its proximity to nearby objects which either absorb or re-radiate power, such as the ground, other wires, etc.

The Marconi Antenna Before going too far with the discussion of radiation resistance, an explanation of the *Marconi* (grounded quarter-wave) antenna is in order. The Marconi antenna is a special type of Hertz antenna in which the earth acts as the "other half" of the dipole. In other words, the current flows into the earth instead of into a similar quarter-wave section. Thus, the current loop of a Marconi antenna is at the *base* rather than in the *center*.

A half-wave dipole far from ground and other reflecting objects has a radiation resistance at the center of about 73 ohms. A Marconi antenna is simply one-half of a dipole. For that reason, the radiation resistance is roughly half the 73-ohm impedance of the dipole, or 36.5 ohms. The radiation resistance of a Marconi antenna, such as a mobile whip, will be lowered by the proximity of the automobile body.

Antenna Impedance Because the power throughout the antenna is the same, the *impedance* of a resonant antenna at any point along its length merely expresses the ratio between voltage and current at that point. Thus, the lowest impedance occurs where the current is highest, namely, at the center of a dipole, or a quarter wave from the end of a Marconi. The impedance rises uniformly toward each end, where it is about 2000 ohms for a dipole remote from ground, and about twice as high for a vertical Marconi.

If a vertical half-wave antenna is set up so that its lower end is at the ground level, the effect of the ground reflection is to increase the radiation resistance to approximately 100 ohms. When a horizontal half-wave antenna is used, the radiation resistance (and, of course, the amount of energy radiated for a given antenna current) depends on the height of the antenna above

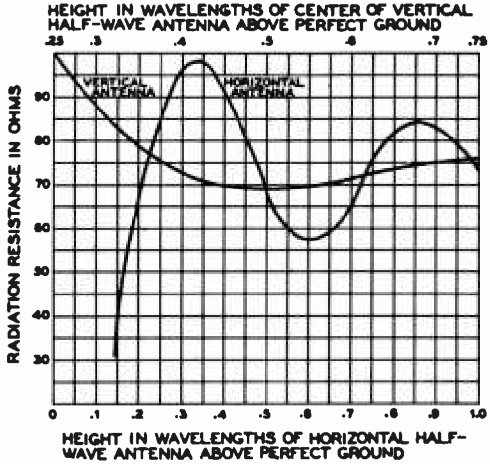


Figure 7

EFFECT OF HEIGHT ON THE RADIATION RESISTANCE OF A DIPOLE SUSPENDED ABOVE PERFECT GROUND

ground, since the height determines the phase and amplitude of the wave reflected from the ground back to the antenna. Thus the resultant current in the antenna for a given power is a function of antenna height.

Center-Fed Feed-Point Impedance When a linear radiator is series fed at the center, the resistive and reactive components of the driving-point impedance are dependent on both the length and diameter of the radiator expressed in wavelengths. The manner in which the resistive component varies with the physical dimensions of the radiator is illustrated in figure 5. The manner in which the reactive component varies is illustrated in figure 6.

Several interesting things will be noted with respect to these curves. The reactive component disappears when the over-all physical length is slightly less than any number of half waves long, the differential increasing with conductor diameter. For over-all lengths in the vicinity of an odd number of half wavelengths, the center feed point looks like a series-resonant lumped circuit to the generator or transmission line, while for over-all lengths in the vicinity of an *even* number of half wavelengths, it looks like a parallel-resonant or antiresonant lumped circuit. Both the feed-point resistance and the feed-point reactance *change more slowly* with over-all radiator length (or with frequency with a fixed length) *as the conductor diameter is increased*, indicating that the effective Q is lowered as the diameter is increased. However, in view of the fact that the damping resistance is nearly all radiation resistance rather than loss resistance, the lower Q does not represent lower efficiency. Therefore, the lower Q is desirable, because it permits use of the radiator over a wider frequency range without resorting to means for eliminating the reactive component. Thus, the use of a large diameter conductor makes the over-all system less frequency sensitive. If the diameter is made sufficiently large in terms of wavelengths, the Q will be low enough to qualify the radiator as a *broadband* antenna.

The curves of figure 7 indicate the theoretical center-point radiation resistance of a half-wave antenna for various heights above perfect ground. These values are of importance in matching untuned radio-frequency feeders to the antenna, in order to obtain a good impedance match and an absence of standing waves on the feeders.

Ground Losses Above *average* ground, the actual radiation resistance of a dipole will vary from the exact value of figure 7 since the latter assumes a hypothetical, perfect ground having no loss and perfect reflection. Fortunately, the curves for the radiation resistance over most types of earth will correspond rather closely with those of the chart, except that the radiation resistance for a horizontal dipole does not fall off as rapidly as is indicated for heights below an eighth wavelength. However, with the antenna so close to the ground and the soil in a strong field, much of the radiation resistance is actually represented by ground loss; this means that a good portion of the antenna power is being dissipated in the earth, which, unlike the hypothetical perfect ground, has resistance. In this case, an appreciable portion of the *radiation resistance* actually is loss resistance. The type of soil also has an effect upon the radiation *pattern*, especially in the vertical plane, as will be seen later.

The radiation resistance of an antenna generally increases with length, although this increase varies up and down about a constantly increasing average. The peaks and dips are caused by the reactance of the antenna, when its length does not allow it to resonate at the operating frequency.

Antenna Efficiency Antennas have a certain loss resistance as well as a radiation resistance. The loss resistance defines the power lost in the antenna due to ohmic resistance of the wire, ground resistance (in the case of a Marconi), corona discharge, and insulator losses.

The approximate effective radiation efficiency (expressed as a decimal) is equal to:

$$N_r = \frac{R_a}{R_a + R_L}$$

where,
R_a equals the radiation resistance,
R_L equals loss resistance of antenna.

The loss resistance will be of the order of 0.25 ohm for large-diameter tubing conductors such as are most commonly used in multi-element parasitic arrays, and will be of the order of 0.5 to 2.0 ohms for arrays of normal construction using copper wire.

When the radiation resistance of an antenna or array is very low, the current at a voltage node will be quite high for a given power. Likewise, the voltage at a current node will be very high. Even with a heavy conductor and excellent insulation, the losses due to the high voltage and current will be appreciable if the radiation resistance is sufficiently low.

Usually, it is not considered desirable to use an antenna or array with a radiation resistance of less than approximately 5 ohms unless there is sufficient directivity, compactness, or other advantage to offset the losses resulting from the low radiation resistance.

Ground Resistance The radiation resistance of a Marconi antenna, especially, should be kept as high as possible. This will reduce the antenna current for a given power, thus minimizing loss resulting from the series resistance offered by the earth connection. The radiation resistance can be kept high by making the Marconi radiator somewhat longer than a quarter wave, and shortening it by series capacitance to an electrical quarter wave. This reduces the current flowing in the earth connection. It also should be removed from ground as much as possible (vertical being ideal). Methods of minimizing the resistance of the earth connection will be found in the discussion of the Marconi antenna.

24-4 Antenna Directivity

All practical antennas radiate better in some directions than others. This characteristic is called *directivity.* The more *directive* an antenna is, the more it concentrates the radiation in a certain direction, or directions. The more the radiation is concentrated in a certain direction, the greater will be the field strength produced in that direction for a given amount of total radiated power. Thus the use of a directional antenna or *array* produces the same result in the favored direction as an increase in the power of the transmitter.

The increase in radiated power in a certain direction with respect to an antenna in free space as a result of inherent directivity is called the *free-space directivity power gain* or just *space directivity gain* of the antenna (referred to a hypothetical *isotropic radiator* which is assumed to radiate equally well in all directions). Because the fictitious isotropic radiator is a purely academic antenna, not physically realizable, it is common practice to use as a reference antenna the simplest ungrounded resonant radiator, the half-wave dipole, or resonant doublet. As a half-wave doublet has a space directivity gain of 2.15 db over an isotropic radiator, the use of a resonant dipole as the comparison antenna reduces the *gain figure* of an array by 2.15 db. However, it should be understood that power gain can be expressed with regard to *any* antenna, just so long as it is specified.

As a matter of interest, the directivity of an *infinitesimal dipole* provides a free-space directivity power gain of 1.5 (or 1.76 db) over an isotropic radiator. This means that *in the direction of maximum radiation* the infinitesimal dipole will produce the same field of strength as an isotropic radiator which is radiating 1.5 times as much total power.

A half-wave resonant doublet, because of its different current distribution and significant length, exhibits slightly more free-space power gain as a result of directivity than does the infinitesimal dipole, for reasons which will be explained in a later section. The space-directivity power gain of a half-wave resonant doublet is 1.63 (or 2.15 db) referred to an isotropic radiator.

Horizontal Directivity When choosing and orienting an antenna system, the radiation patterns of the various common types of antennas should be given careful consideration. The directional characteristics are of still greater importance when a directive antenna array is used.

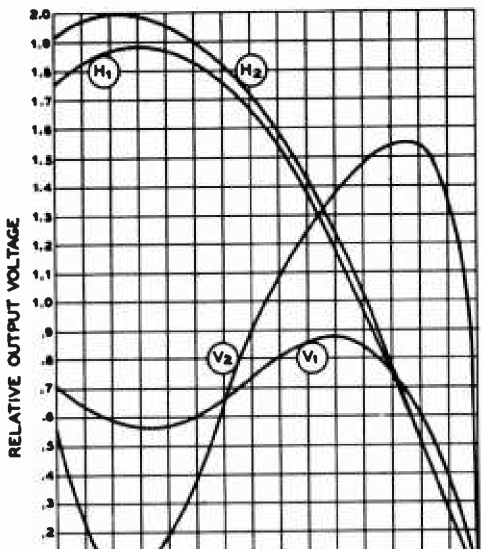


Figure 8

VERTICAL-PLANE DIRECTIONAL CHARACTERISTICS OF HORIZONTAL AND VERTICAL DOUBLETS ELEVATED 0.6 WAVELENGTH AND ABOVE TWO TYPES OF GROUND

H_1 represents a horizontal doublet over typical farmland. H_2 over salt water. V_1 is a vertical pattern of radiation from a vertical doublet over typical farmland, V_2 over salt water. A salt water ground is the closest approach to an extensive ideally perfect ground that will be met in actual practice.

Horizontal directivity is always desirable on any frequency for point-to-point work. However, it is not always attainable with reasonable antenna dimensions on the lower frequencies. Further, when it is attainable, as on the frequencies above perhaps 7 MHz, with reasonable antenna dimensions, operating convenience is greatly furthered if the maximum lobe of the horizontal directivity is controllable. It is for this reason that rotatable antenna arrays have come into such common usage.

Considerable horizontal directivity can be used to advantage when: (1) only point-to-point work is necessary, (2) several arrays are available so that directivity may be changed by selecting or reversing antennas, (3) a single rotatable array is in use. Signals follow the great-circle path, or within 2 or 3 degrees of that path under all normal propagation conditions. However, under turbulent ionospheric conditions, or when unusual propagation conditions exist, the deviation from the great-circle path for greatest signal intensity may be as great as 90°. Making the array rotatable overcomes these difficulties, but arrays having extremely high horizontal directivity become too cumbersome to be rotated, except perhaps when designed for operation on frequencies above 50 MHz.

Vertical Directivity Vertical directivity is of the greatest importance in obtaining satisfactory communication above 14 MHz whether or not horizontal directivity is used. This is true simply because *only* the energy radiated between certain definite *elevation* angles is useful for communication. Energy radiated at other elevation angles is lost and performs no useful function.

Optimum Angle of Radiation The optimum angle of radiation for propagation of signals between two points is dependent on a number of variables. Among these significant variables are: (1) height of the ionosphere layer which is providing the reflection, (2) distance between the two stations, (3) number of hops for propagation between the two stations. For communication on the 14-MHz band it is often possible for different modes of propagation to provide signals between two points. This means, of course, that more than one angle of radiation can be used. If *no* elevation directivity is being used under this condition of propagation, selective fading will take place because of interference between the waves arriving over the different paths.

On the 28-MHz band it is by far the most common condition that only one mode of propagation will be possible between two points at any one time. This explains, of course, the reason why rapid fading in general and selective fading in particular are almost absent from signals heard on the 28-MHz band (except for fading caused by local effects).

Measurements have shown that the angles useful for communication on the 14-MHz band are from 3° to about 30°, angles above about 15° being useful only for local work. On the 28-MHz band, measurements have shown that the useful angles range from about 3° to 18°; angles above about 12° being useful only for local (less than 3000 miles) work. These figures assume normal propagation by virtue of the F_2 layer.

Angle of Radiation of Typical Antennas and Arrays It now becomes of interest to determine the amount of radiation available at these useful lower angles of radiation from commonly used antennas and antenna arrays. Figure 8 shows relative output voltage plotted against elevation angle (wave angle) in degrees above the horizontal, for horizontal and vertical doublets elevated 0.6 wavelength above two types of ground. It is obvious by inspection of the curves that a horizontal dipole mounted at this height above ground (20 feet on the 28-MHz band) is radiating only a small amount of energy at angles useful for communication on the 28-MHz band. Most of the energy is being radiated uselessly upward. The vertical antenna above a good reflecting surface appears much better in this respect—and this fact has been proven many times by actual installations.

It might immediately be thought that the amount of radiation from a horizontal or vertical dipole could be increased by raising the antenna higher above the ground. This is true to an extent in the case of the hori-

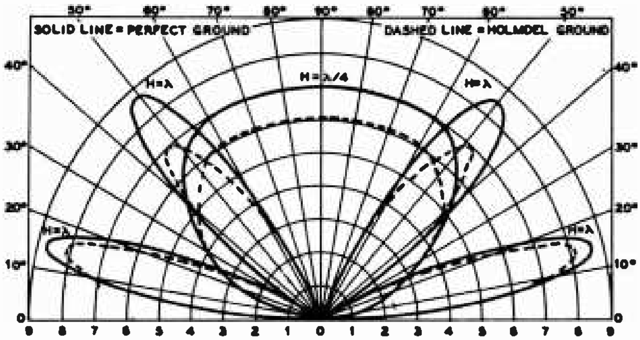


Figure 9

VERTICAL RADIATION PATTERNS

Showing the vertical radiation patterns for half-wave antennas (or collinear half-wave or extended half-wave antennas) at different heights above average ground and perfect ground. Note that such antennas one-quarter wave above ground concentrate most of the radiation at the very high angles which are useful for communication only on the lower-frequency bands. Antennas one-half wave above ground are not shown, but the elevation pattern shows one lobe on each side at an angle of 30° above horizontal.

zontal dipole; the low-angle radiation does increase *slowly* after a height of 0.6 wavelength is reached but at the expense of greatly increased high-angle radiation and the formation of a number of nulls in the elevation pattern. No signal can be transmitted or received at the elevation angles where these nulls have been formed. Tests have shown that a center height of 0.6 wavelength for a vertical dipole (0.35 wavelength to the bottom end) is about optimum for this type of array.

Figure 9 shows the effect of placing a horizontal dipole at various heights above ground. It is easily seen by reference to figure 9 (and figure 10 which shows the radiation from a dipole at ¾ wave height) that a large percentage of the total radiation from the dipole is being radiated at relatively high angles which are useless for communication on the 14-MHz and 28-MHz bands. Thus we see that in order to obtain a worthwhile increase in the ratio of low-angle radiation to high-angle radiation it is necessary to place the antenna high above ground, *and in addition* it is necessary to use additional means for suppressing high-angle radiation.

Suppression of High-angle Radiation High-angle radiation can be suppressed, and this radiation can be added to that going out at low angles, only through the use of some sort of *directive* antenna system. There are three general types of antenna arrays composed of dipole elements commonly used which concentrate radiation at the lower more effective angles for high-frequency communication. These types are: (1) The close-spaced out-of-phase system as exemplified by the "flat-top" beam, or W8JK array. Such configurations are classified as *end-fire arrays* .(2) The wide-spaced in-phase arrays, as exemplified by the "Lazy H" antenna. These configurations are classified as *broadside arrays*. (3) The close-

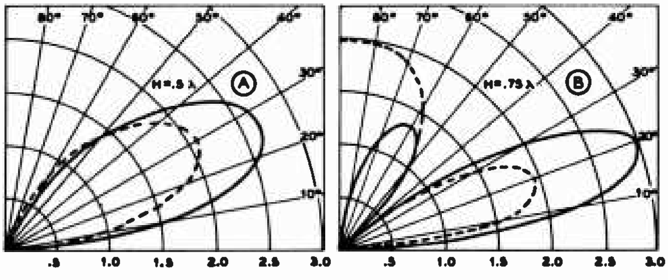


Figure 10

VERTICAL RADIATION PATTERNS

Showing vertical-plane radiation patterns of a horizontal single-section flat-top beam with one-eighth wave spacing (solid curves) and a horizontal half-wave antenna (dashed curves) when both are 0.5 wavelength (A) and 0.75 wavelength (B) above ground.

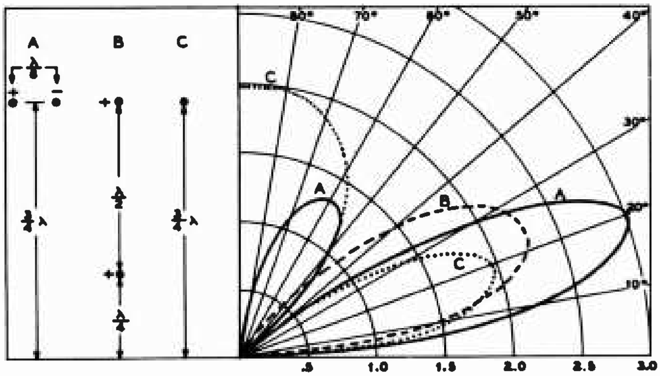


Figure 11

COMPARATIVE VERTICAL RADIATION PATTERNS

Showing the vertical radiation patterns of a horizontal single-section flat-top beam (A), an array of two stacked horizontal in-phase half-wave elements — half of a "Lazy H"—(B), and a horizontal dipole (C). In each case the top of the antenna system is 0.75 wavelength above ground, as shown to the left of the curves.

spaced parasitic systems, as exemplified by the three-element rotary beam.

A comparison between the radiation from a dipole, a "flat-top beam" and a pair of dipoles stacked one above the other (half of a "lazy H"), in each case with the top of the antenna at a height of ¾ wavelength is shown in figure 11. The improvement in the amplitude of low-angle radiation at the expense of the useless high-angle radiation with these simple arrays as contrasted to the dipole is quite marked.

Figure 12 compares the patterns of a 3-element beam and a dipole radiator at a height of 0.75 wavelength. It will be noticed that although there is more energy in the lobe of the beam as compared to the dipole, the axis of the beam is at the same angle above the horizontal. Thus, although more radiated energy is provided by the beam at low angles, the average angle of radiation of the beam is no lower than the average angle of radiation of the dipole.

24-5 Bandwidth

The *bandwidth* of an antenna or an antenna array is a function primarily of the radiation resistance and of the shape of the conductors which make up the antenna system. For arrays of essentially similar construction the bandwidth (or the deviation in frequency which the system can handle without mismatch) is increased with increasing radiation resistance, and the bandwidth is increased with the use of conductors of large diameter (smaller ratio of length to diameter). This is to say that if an array of any type is constructed of large diameter tubing or spaced wires, its bandwidth will be greater than that of a similar array constructed of single wires.

The radiation resistance of antenna arrays of the types mentioned in the previous paragraphs may be increased through the use of wider spacing between elements. With increased radiation resistance in such arrays the *radiation efficiency* increases since the ohmic losses within the conductors become a smaller percentage of the radiation resistance, and the bandwidth is increased proportionately.

24-6 Propagation of Radio Waves

The preceding sections have discussed the manner in which an electromagnetic-wave

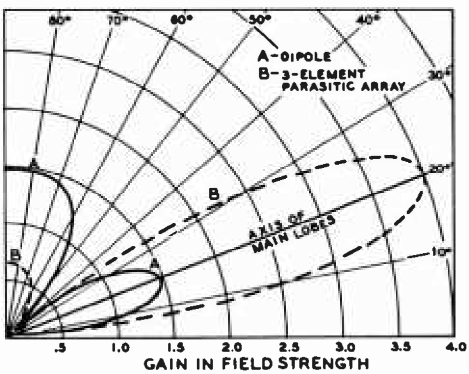


Figure 12

VERTICAL RADIATION PATTERNS

Showing vertical radiation patterns of a horizontal dipole (A) and a horizontal 3-element parasitic array (B) at a height above ground of 0.75 wavelength. Note that the axes of the main radiation lobes are at the same angle above the horizontal. Note also the suppression of high angle radiation by the parasitic array.

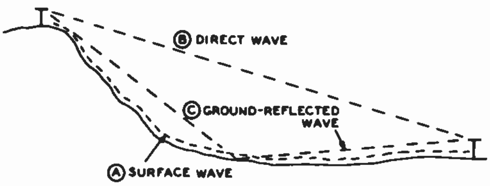


Figure 13

GROUND-WAVE SIGNAL PROPAGATION

The illustration above shows the three components of the ground wave: A, the surface wave; B, the direct wave; and C, the ground-reflected wave. The direct wave and the ground-reflected wave combine at the receiving antenna to make up the space wave.

or radio-wave field may be set up by a radiating system. However, for this field to be useful for communication it must be propagated to some distant point where it can be received, or where it may be reflected so that it can be received at some other point. Radio waves may be propagated to a remote point by either or both of two general methods. Propagation may take place as a result of the *ground wave*, or as a result of the *sky wave* or *ionospheric wave*.

The Ground Wave The term *ground wave* actually includes several different types of waves which usually are called: (1) the *surface wave*, (2) the *direct wave*, and (3) the *ground-reflected wave*. The latter two waves combine at the receiving antenna to form the *resultant wave* or the *space wave*. The distinguishing characteristic of the components of the ground wave is that all travel along or over the surface of the earth, so that they are affected by the conductivity and terrain of the earth's surface.

The Ionospheric Wave or Sky Wave Intense bombardment of the upper regions of the atmosphere by radiations from the sun results in the formation of ionized layers. These ionized layers, which form the *ionosphere*, have the capability of reflecting or refracting radio waves which impinge on them. A radio wave which has been propagated as a result of one or more reflections from the ionosphere is known as an *ionospheric wave* or a *sky wave*. Such waves make possible long distance radio communication. Propagation of radio signals by ionospheric waves is discussed in detail in Section 20-8.

24-7 Ground-Wave Communication

As stated in the preceding paragraph, the term *ground wave* applies both to the *surface wave* and to the *space wave* (the resultant wave from the combination of the direct wave and the ground-reflected wave) or to a combination of the two. The three waves which may combine to make up the ground wave are illustrated in figure 13.

The Surface Wave The surface wave is that wave which we normally receive from a standard broadcast station. It travels directly along the ground and terminates on the earth's surface. Since the earth is a relatively poor conductor, the surface wave is attenuated quite rapidly. The surface wave is attenuated less rapidly as it passes over sea water, and the attenuation decreases for a specific distance as the frequency is decreased. The rate of attenuation with distance becomes so large as the frequency is increased above about 3 MHz that the surface wave becomes of little value for communication.

The Space Wave The resultant wave or space wave is illustrated in figure 13 by the combination of B and C. It is this wave path, which consists of the combination of the direct wave and the ground-reflected wave at the receiving antenna, which is the *normal* path of signal propagation for line-of-sight or near line-of-sight communication or f-m and TV reception on frequencies above about 40 MHz.

Below line-of-sight over plane earth or water, when the signal source is effectively at the horizon, the ground-reflected wave does not exist, so that the *direct* wave is the only component which goes to make up the space wave. But when both the signal source and the receiving antenna are elevated with respect to the intervening terrain, the ground-reflected wave is present and adds vectorially to the direct wave at the receiving antenna. The vectorial addition of the two waves, which travel over different path lengths (since one of the waves has been reflected from the ground) results in an interference pattern. The interference between the two waves brings about a cyclic variation in signal strength as the receiving antenna is raised above the ground. This effect is illustrated in figure 14. From this figure it can be seen that best spacewave reception of a vhf signal often will be obtained with the receiving antenna quite close to the ground.

The distance from an elevated point to the geometrical horizon is given by the approximate equation: $d = 1.22\sqrt{H}$ where distance d is in miles and antenna height H is in feet. This equation must be applied separately to the transmitting and receiving antennas and the results added. However,

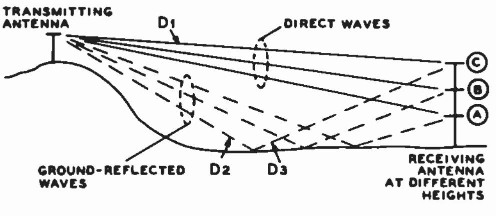


Figure 14

WAVE INTERFERENCE WITH HEIGHT

When the source of a horizontally-polarized space-wave signal is above the horizon, the received signal at a distant location will go through a cyclic variation as the antenna height is progressively raised. This is due to the difference in total path length between the direct wave and the ground-reflected wave, and to the fact that this path length difference changes with antenna height. When the path length difference is such that the two waves arrive at the receiving antenna with a phase difference of 360° or some multiple of 360°, the two waves will appear to be in phase as far as the antenna is concerned and maximum signal will be obtained. On the other hand, when the antenna height is such that the path length difference for the two waves causes the waves to arrive with a phase difference of an odd multiple of 180° the two waves will substantially cancel, and a null will be obtained at that antenna height. The difference between D_1 and D_2 plus D_3 is the path-length difference. Note also that there is an additional 180° phase shift in the ground-reflected wave at the point where it is reflected from the ground. It is this latter phase shift which causes the space-wave field intensity of a horizontally polarized wave to be zero with the receiving antenna at ground level.

refraction and diffraction of the signal around the spherical earth cause a smaller reduction in field strength than would occur in the absence of such bending, so that the average radio horizon is somewhat beyond the geometrical horizon. The equation $d = 1.4\sqrt{H}$ is sometimes used for determining the radio horizon.

Tropospheric Propagation Propagation by signal bending in the lower atmosphere, called *tropospheric propagation,* can result in the reception of signals over a much greater distance than would be the case if the lower atmosphere were homogeneous. In a homogeneous or well-mixed lower atmosphere, called a *normal,* or *standard,* atmosphere, there is a gradual and uniform decrease in index of refraction with height. This effect is due to the combined effects of a decrease in temperature, pressure, and water-vapor content with height.

This gradual decrease in refractive index with height causes waves radiated at very low angles with respect to the horizontal to be bent downward slightly in a curved path. The result of this effect is that such waves will be propagated beyond the *true,* or *geometrical,* horizon. In a so-called standard atmosphere the effect of the curved path is the same as though the radius of the earth were increased by approximately one-third. This condition extends the horizon by approximately 30 percent for normal propagation, and the extended horizon is known as the *radio-path horizon,* mentioned before.

Conditions Leading to Tropospheric Stratification When the temperature, pressure, or water-vapor content of the atmosphere does not change smoothly with rising altitude, the discontinuity or stratification will result in the reflection or refraction of incident vhf signals. Ordinarily this condition is more prevalent at night and in the summer. In certain areas, such as along the west coast of North America, it is frequent enough to be considered normal. Signal strength decreases slowly with distance and, if the favorable condition in the lower atmosphere covers sufficient area, the range is limited only by the transmitter power, antenna gain, receiver sensitivity, and signal-to-noise ratio. There is no skip distance. Usually, transmission due to this condition is accompanied by slow fading, although fading can be violent at a point where direct waves of about the same strength are also received.

Bending in the troposphere, which refers to the region from the earth's surface up to about 10 kilometers, is more likely to occur on days when there are stratus clouds than on clear, cool days with a deep blue sky. The temperature or humidity discontinuities may be broken up by vertical convection currents over land in the daytime but are more likely to continue during the day over water. This condition is in some degree predictable from weather information several days in advance. It does not depend on the sunspot cycle. Like direct communication, best results require similar antenna polarization or orientation at both the transmitting and receiving ends, whereas in transmission via reflection in the ionosphere (that part of the atmosphere between about 50 and 500 kilometers high) it makes little difference whether antennas are similarly polarized.

Duct Formation When bending conditions are particularly favorable they may give rise to the formation of a *duct* which can propagate waves with very little attenuation over great distances in a manner similar to the propagation of waves through a waveguide. *Guided propagation* through a duct in the atmosphere can give quite remarkable transmission conditions (figure 15). However, such ducts usually are formed only on an over-water path. The depth of the duct over the water's surface may be only 20 to 50 feet, or it may be 1000 feet deep or more. Ducts exhibit a low-frequency cutoff characteristic similar to a waveguide. The cutoff frequency is determined by depth of the duct and by the strength of the discontinuity in refractive index at the upper surface of the duct. The lowest frequency that can be propagated by such a duct seldom goes below 50 MHz, and usually will not be greater than 450 MHz even along the Pacific Coast.

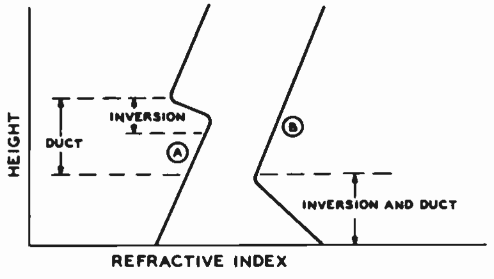


Figure 15

ILLUSTRATING DUCT TYPES

Showing two types of variation in refractive index with height which will give rise to the formation of a duct. An elevated duct is shown at A, and a ground-based duct is shown at B. Such ducts can propagate ground-wave signals far beyond their normal range.

Stratospheric Reflection Communication by virtue of stratospheric reflection can be brought about during magnetic storms, aurora borealis displays, and during meteor showers. DX communication during extensive meteor showers is characterized by frequent bursts of great signal strength followed by a rapid decline in strength of the received signal. The motion of the meteor forms an ionized trail of considerable extent which can bring about effective reflection of signals. However, the ionized region persists only for a matter of seconds so that a shower of meteors is necessary before communication becomes possible.

The type of communication which is possible during visible displays of the aurora borealis and during magnetic storms has been called *aurora-type DX*. These conditions reach a maximum somewhat after the sunspot cycle peak, possibly because the spots on the sun are nearer to its equator (and more directly in line with the earth) in the latter part of the cycle. Ionospheric storms generally accompany magnetic storms. The normal layers of the ionosphere may be churned or broken up, making radio transmission over long distances difficult or impossible on high frequencies. Unusual conditions in the ionosphere sometimes modulate vhf waves so that a definite tone or noise modulation is noticed even on transmitters located only a few miles away.

A peculiarity of this type of auroral propagation of vhf signals in the northern hemisphere is that directional antennas usually must be pointed in a northerly direction for best results for transmission or reception. regardless of the direction of the other station being contacted. Distances out to 700 or 800 miles have been covered during magnetic storms, using 30- and 50-MHz transmitters, with little evidence of any silent zone between the stations communicating with each other. Generally, voice-modulated transmissions are difficult or impossible due to the tone or noise modulation on the signal. Most of the communication of this type has taken place by c.w. or by tone-modulated waves with a keyed carrier.

24-8 Ionospheric Propagation

Propagation of radio waves for communication on frequencies between perhaps 3 and 30 MHz is normally carried out by virtue of *ionospheric reflection* or *refraction.* Under conditions of abnormally high ionization in the ionosphere, communication has been known to have taken place by ionospheric reflection on frequencies higher than 50 MHz.

The ionosphere consists of layers of ionized gas located above the stratosphere, and extending up to possibly 300 miles above

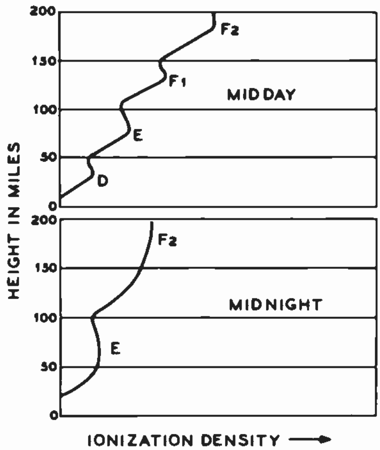


Figure 16

IONIZATION DENSITY IN THE IONOSPHERE

Showing typical ionization density of the ionosphere in midsummer. Note that the F_1 and D layers disappear at night, and that the density of the E layer falls to such a low value that it is ineffective.

the earth. Thus we see that high-frequency radio waves may travel over short distances in a direct line from the transmitter to the receiver, or they can be radiated upward into the ionosphere to be bent downward in an indirect ray, returning to earth at considerable distance from the transmitter. The wave reaching a receiver via the ionosphere route is termed a *sky wave.* The wave reaching a receiver by traveling in a direct line from the transmitting antenna to the receiving antenna is commonly called a *ground wave.*

The amount of bending at the ionosphere which the sky wave can undergo depends on its frequency, and the amount of *ionization* in the ionosphere, which is in turn dependent on radiation from the sun. The sun increases the density of the ionosphere layers (figure 16) and lowers their effective height. For this reason, the ionosphere acts very differently at different times of day, and at different times of the year.

The higher the frequency of a radio wave, the farther it penetrates the ionosphere, and the less it tends to be bent back toward the earth. The lower the frequency, the more easily the waves are bent, and the less they penetrate the ionosphere. 160-meter and 80-meter signals will usually be bent back to earth even when sent straight up, and may be considered as being *reflected* rather than *refracted.* As the frequency is raised beyond about 5000 kHz (dependent on the critical frequency of the ionosphere at the moment), it is found that waves transmitted at angles higher than a certain critical angle *never return to earth.* Thus, on the higher frequencies, it is necessary to confine radiation to low angles, since the high-angle waves simply penetrate the ionosphere and are lost.

The F_2 Layer The higher of the two major reflection regions of the ionosphere is called the F_2 layer. This layer has a virtual height of approximately 175 miles at night, and in the daytime it splits up into two layers, the upper one being called the F_2 layer and the lowerbeing called the F_1 layer. The height of the F_2 layer during daylight hours is normally about 250 miles on the average and the F_1 layer often has a height of as low as 140 miles. It is the F_2 layer which supports all nighttime DX communication and nearly all daytime DX propagation.

The E Layer Below the F_2 layer is another layer, called the *E* layer, which is of importance in daytime communication over moderate distances in the frequency range between 3 and 8 MHz. This layer has an almost constant height of about 70 miles. Since the recombination time of the ions at this height is rather short, the *E* layer disappears almost completely a short time after local sunset.

The D Layer Below the *E* layer at a height of about 35 miles is an *absorbing* layer, called the *D* layer, which exists in the middle of the day in the summertime. The layer also exists during midday in winter during periods of high solar activity, but the layer disappears completely at night. It is this layer which causes high absorption of signals in the medium- and high-frequency range during the middle of the day.

Critical Frequency The *critical frequency* of an ionospheric layer is the highest frequency which will be reflected when the wave strikes the layer at vertical incidence. The critical frequency of the most highly ionized layer of the ionosphere may be as low as 2 MHz at night and as high as 12 to 13 MHz in the middle of the day. The critical frequency is directly of interest in that a *skip-distance zone* will exist on all frequencies *greater* than the highest critical frequency at that time. The critical frequency is a measure of the density of ionization of the reflecting layers. The higher the critical frequency the greater the density of ionization.

Maximum Usable Frequency The *maximum usable frequency* or *m.u.f.* is of great importance in long-distance communication since this frequency is the highest that can be used for communication *between any two specified areas*. The m.u.f. is the highest frequency at which a wave projected into space in a certain direction will be returned to earth in a specified region by ionospheric reflection. The m.u.f. is highest at noon or in the early afternoon and is highest in periods of greatest sunspot activity, often going to frequencies higher than 30 MHz. (figure 17).

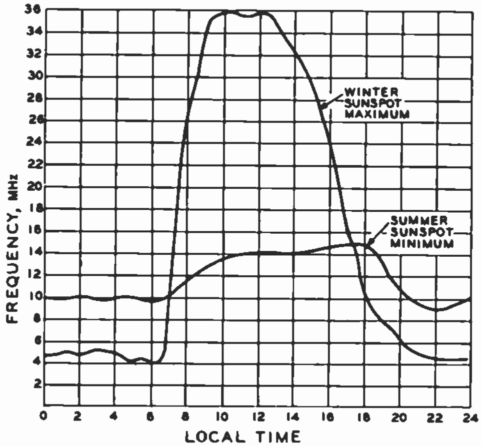


Figure 17

TYPICAL CURVES SHOWING CHANGE IN M.U.F. AT MAXIMUM AND MINIMUM POINTS IN SUNSPOT CYCLE

The m.u.f. often drops to frequencies below 10 MHz in the early morning hours. The high m.u.f. in the middle of the day is brought about by reflection from the F_2 layer. M.u.f. data is published periodically in the magazines devoted to amateur work, and the m.u.f. can be calculated with the aid of *Basic Radio Propagation Predictions*, CRPL-D, published monthly by the Government Printing Office, Washington, D.C.

Absorption and Optimum Working Frequency The *optimum working frequency* for any particular direction and distance is usually about 15 percent less than the m.u.f. for contact with that particular location. The absorption by the ionosphere becomes greater and greater as the operating frequency is progressively lowered below the m.u.f. It is this condition which causes signals to increase tremendously in strength on the 14- and 28-MHz bands just before the signals drop completely out. At the time when the signals are greatest in amplitude the operating frequency is equal to the m.u.f. Then as the signals drop out the m.u.f. has become lower than the operating frequency.

Skip Distance The shortest distance from a transmitting location at which signals reflected from the ionosphere can be returned to the earth is called the *skip distance.* As was mentioned above under *critical frequency,* there is no skip distance for a frequency below the critical frequency of the most highly ionized layer of the ionosphere at the time of transmission. However, the skip distance is always present on the 14-MHz band and is almost always present on the 3.5 and 7 MHz bands at night. The actual measure of the skip distance is the distance between the point where the ground wave falls to zero and the point where the sky wave begins to return to earth. This distance may vary from 40 to 50 miles on the 3.5-MHz band to thousands of miles on the 28-MHz band.

The Sporadic-E Layer Occasional patches of extremely high ionization density appear at intervals throughout the year at a height approximately equal to that of the *E* layer. These patches, called the *sporadic-E* layer may be very small or may be up to several hundred miles in extent. The critical frequency of the *sporadic-E* layer may be greater than twice that of the normal ionosphere layers which exist at the same time.

It is this *sporadic-E* condition which provides "short-skip" contacts from 400 to perhaps 1200 miles on the 28-MHz band in the evening. It is also the *sporadic-E* condition which provides the more common type of "band opening" experienced on the 50-MHz band when very loud signals are received from stations from 400 to 1200 miles distant.

Cycles in Ionosphere Activity The ionization density of the ionosphere is determined by the amount of radiation (probably ultraviolet) which is being received from the sun. Consequently, ionosphere activity is a function of the amount of radiation of the proper character being emitted by the sun and is also a function of the relative aspect of the regions in the vicinity of the location under discussion to the sun. There are four main cycles in ionosphere activity. These cycles are: the daily cycle which is brought about by the rotation of the earth, the 27-day cycle which is caused by the rotation of the sun, the seasonal cycle which is caused by the movement of the earth in its orbit, and the 11-year cycle which is a cycle in sunspot activity. The effects of these cycles are superimposed insofar as ionosphere activity is concerned. Also, the cycles are subject to short term variations as a result of magnetic storms and similar terrestrial disturbances.

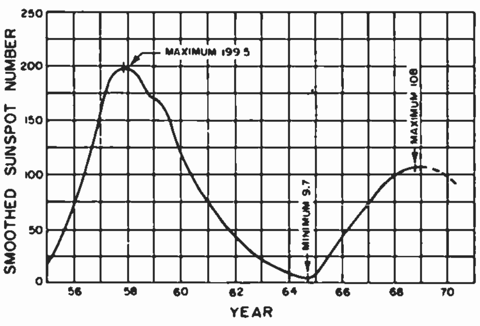


Figure 18

THE YEARLY TREND OF THE SUNSPOT CYCLE. RADIO CONDITIONS IN GENERAL IMPROVED DURING 1966-1969 AS THE CYCLE INCREASED

The most recent minimum of the 11-year sunspot cycle occurred during the winter of 1964-1965, and we are currently moving along the slope of a new cycle, the maximum of which occurred during 1969. The current cycle is pictured in figure 18.

Fading The lower the angle of radiation of the wave, with respect to the horizon, the farther away will the wave return to earth, and the greater the skip distance. The wave can be reflected back up into the ionosphere by the earth, and then be reflected back down again, causing a second skip distance area. The drawing of figure 19 shows the multiple reflections possible. When the receiver receives signals which have traveled over more than one path between transmitter and receiver, the signal impulses will not all arrive at the same instant, since they do not all travel the same distance. When two or more signals arrive in the same phase at the receiving antenna, the resulting signal in the receiver will be quite strong. On the other hand, if the signals arrive 180° out of phase, so they tend to cancel each other, the received signal will drop—perhaps

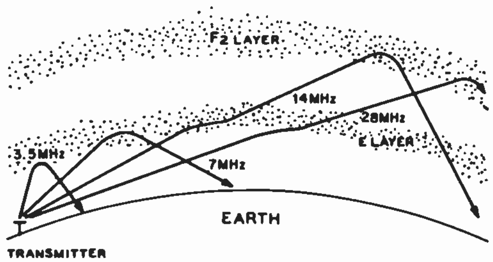


Figure 19

IONOSPHERE-REFLECTION WAVE PATHS

Showing typical ionosphere-reflection wave paths during daylight hours when ionization density is such that frequencies as high as 28 MHz will be returned to earth. The distance between ground-wave range and that range where the ionosphere-reflected wave of a specific frequency first will be returned to earth is called the skip distance.

to zero if perfect cancellation occurs. This explains why high-frequency signals are subject to fading.

Fading can be greatly reduced on the high frequencies by using a transmitting antenna with sharp vertical directivity, thus cutting down the number of possible paths of signal arrival. A receiving antenna with similar characteristics (sharp vertical directivity) will further reduce fading. It is desirable, when using antennas with sharp vertical directivity, to use the lowest vertical angle consistent with good signal strength for the frequency used.

Scattered Reflections Scattered reflections are random, diffused, substantially isotropic reflections which are partly responsible for reception within the skip zone, and for reception of signals from directions off the great circle path.

In a heavy fog or mist, it is difficult to see the road at night because of the bright glare caused by scattered reflection of the headlight beam by the minute droplets. In fact, the road directly to the side of the car will be weakly illuminated under these conditions, whereas it would not on a clear night (assuming flat, open country). This is a good example of propagation of waves by scattered reflections into a zone which otherwise would not be illuminated.

Scattering occurs in the ionosphere at all times, because of irregularities in the medium (which result in "patches" corresponding to the water droplets) and because of random-phase radiation due to the collision or recombination of free electrons. However, the nature of the scattering varies widely with time, in a random fashion. Scattering is particularly prevalent in the *E* region, but scattered reflections may occur at any height, even well out beyond the virtual height of the F_2 layer.

There is no "critical frequency" or "lowest perforating frequency" involved in the scattering mechanism, though the intensity of the scattered reflections due to typical scattering in the *E* region of the ionosphere decreases with frequency.

When the received signal is due primarily to scattered reflections, as is the case in the skip zone or where the great circle path does not provide a direct sky wave (due to low critical or perforation frequency, or to an ionosphere storm) very bad distortion will be evident, particularly a "flutter fade" and a characteristic "hollow" or echo effect.

Deviations from a great circle path are especially noticeable in the case of great circle paths which cross or pass near the auroral zones, because in such cases there often is complete or nearly complete absorption of the direct sky wave, leaving off-path scattered reflections the only mechanism of propagation. Under such conditions the predominant wave will appear to arrive from a direction closer to the equator, and the signal will be noticeably, if not considerably, weaker than a direct sky wave which is received under favorable conditions.

Irregular reflection of radio waves from "scattering patches" is divided into two categories: *short scatter* and *long scatter*.

Short scatter is the scattering that occurs when a radio wave first reaches the scattering patches or media. Ordinarily it is of no particular benefit, as in most cases it only serves to fill in the inner portion of the skip zone with a weak, distorted signal.

Long scatter occurs when a wave has been refracted from the F_2 layer and strikes scattering patches or media on the way down. When the skip distance exceeds several hundred miles, long scatter is primarily responsible for reception within the skip zone, particularly the outer portion of the skip zone. Distortion is much less severe than in the case of short scatter, and while the signal is likewise weak, it sometimes can be utilized for satisfactory communication.

During a severe ionosphere disturbance in the north auroral zone, it sometimes is possible to maintain communication between the Eastern United States and Northern Europe by the following mechanism: That portion of the energy which is radiated in the direction of the great circle path is completely absorbed on reaching the auroral zone. However, the portion of the wave leaving the United States in a *southeasterly* direction is refracted downward from the F_2 layer and encounters scattering patches or media on its downward trip at a distance of approximately 2000 miles from the transmitter. There it is reflected by "long scatter" in all directions, this scattering region acting like an isotropic radiator fed with a very small fraction of the original transmitter power. The great circle path from this southerly point to northern Europe does not encounter unfavorable ionosphere conditions, and the wave is propagated the rest of the trip as though it had been radiated from the scattering region.

Another type of scatter is produced when a sky wave strikes certain areas of the earth. On striking a comparatively smooth surface such as the sea, there is little scattering, the wave being shot up again by what could be considered specular, or mirror, reflection. But on striking a mountain range, for instance, the reradiation or reflected energy is scattered, some of it being directed back toward the transmitter, thus providing another mechanism for producing a signal within the skip zone.

Meteors and "Bursts" When a meteor strikes the earth's atmosphere, a cylindrical region of free electrons is formed at approximately the height of the *E* layer. This slender ionized column is quite long, and when first formed is sufficiently dense to reflect radio waves back to earth most readily, *including vhf waves which are not ordinarily returned by the F_2 layer.*

The effect of a single meteor, or normal size, shows up as a sudden "burst" of signal of short duration at points not ordinarily reached by the transmitter. After a period of from 10 to 40 seconds, recombination and diffusion have progressed to the point where the effect of a *single* fairly large meteor is not perceptible. However, there are many *small* meteors impinging on earth's atmosphere every minute, and the *aggregate* effect of their transient ionized trails, including the small amount of residual ionization that exists for several minutes after the original flash but is too weak and dispersed to prolong a "burst," is believed to contribute to the existence of the *nighttime-E* layer, and perhaps also to *sporadic-E* patches.

While there are many of these very small meteors striking the earth's atmosphere every minute, meteors of normal size (sufficiently large to produce individual "bursts") do not strike nearly so frequently except during some of the comparatively rare meteor "showers." During one of these displays a "quivering" ionized layer is produced which is intense enough to return signals in the lower vhf range with good strength, but with a type of "flutter" distortion which is characteristic of this type of propagation.

24-9 Transmission Lines

For many reasons it is desirable to place an antenna or radiating system as high and in the clear as is physically possible, utilizing some form of nonradiating transmission line to carry energy with as little loss as possible from the transmitter to the radiating antenna, and conversely from the antenna to the receiver.

There are many different types of transmission lines and, generally speaking, practically any type of transmission line or feeder system may be used with any type of antenna. However, mechanical or electrical considerations often make one type of transmission line better adapted for use to feed a particular type of antenna than any other type.

Transmission lines for carrying r-f energy are of two general types: *nonresonant* and *resonant.* A nonresonant transmission line is one on which a successful effort has been made to eliminate reflections from the termination (the antenna in the transmitting case and the receiver for a receiving antenna) and hence one on which standing waves do not exist or are relatively small in magnitude. A resonant line, on the other hand, is a transmission line on which standing waves of appreciable magnitude do appear, either through inability to match the characteristic impedance of the line to the termination or through intentional design.

The principal types of transmission line in use or available at this time include the open-wire line (two-wire and four-wire types), two-wire solid-dielectric line (*twin-lead* and similar ribbon or tubular types), two-wire polyethylene-filled shielded line, coaxial line of the solid-dielectric, beaded, stub-supported, or pressurized type, rectangular and cylindrical waveguide, and the single-wire feeder operated against ground. The significant characteristics of the more popular types of transmission line available at this time are given in the chart of figure 21.

24-10 Nonresonant Transmission Lines

A nonresonant or untuned transmission line is a line with negligible standing waves. Hence, a nonresonant line is a line carrying r-f power only in one direction—from the source of energy to the load.

Physically, the line itself should be *identical throughout its length.* There will be a smooth distribution of voltage and current throughout its length, both tapering off very slightly towards the load end of the line as a result of line losses. The attenuation (loss) in certain types of untuned lines can be kept very low for line lengths up to several thousand feet. In other types, particularly where the dielectric is not air (such as in the twisted-pair line), the losses may become excessive at the higher frequencies, unless the line is relatively short.

Transmission-Line Impedance All transmission lines have distributed inductance, capacitance, and resistance. Neglecting the resistance, as it is of minor importance in short lines, it is found that the *inductance and capacitance per unit length* determine the characteristic or surge impedance of the line. Thus, the surge impedance depends upon the nature and spacing of the conductors, and the dielectric separating them.

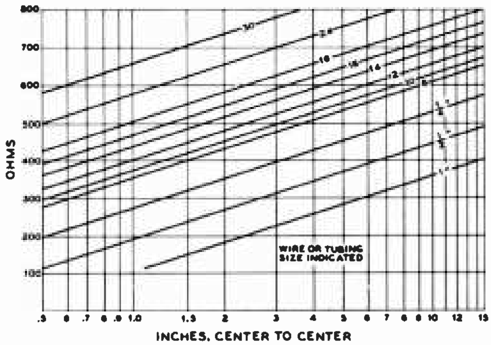


Figure 20

CHARACTERISTIC IMPEDANCE OF TYPICAL TWO-WIRE OPEN LINES

Speaking in electrical terms, the characteristic impedance of a transmission line is simply the ratio of the voltage across the line to the current which is flowing, the same as is the case with a simple resistor: $Z_0 = E/I$. Also, in a substantially lossless line (one whose attenuation per wavelength is small) the energy stored in the line will be equally divided between the electric field and the magnetic field which serve to propagate the energy along the line. Hence the characteristic impedance of a line may be expressed as:

$$Z_0 = \sqrt{L/C}.$$

Two-Wire Open Line A two-wire transmission system is easy to construct. Its surge impedance can be calculated quite easily, and when properly adjusted and balanced to ground, with a conductor spacing which is negligible in terms of the wavelength of the signal carried, undesirable feeder radiation is minimized; the current flow in the adjacent wires is in opposite directions, and the magnetic fields of the two wires are in opposition to each other. When a two-wire line is terminated with the equivalent of a pure resistance equal to the characteristic impedance of the line, the line becomes a nonresonant line.

Expressed in physical terms, the characteristic impedance of a two-wire open line is equal to:

$$Z_0 = 276 \log_{10} \frac{2S}{d}$$

CHARACTERISTICS OF COMMON TRANSMISSION LINES

	Attenuation db/100 feet VSWR = 1.0			Velo-city factor V	pf per ft.	REMARKS
	30 MHz	100 MHz	300 MHz			
Open wire line, No. 12 copper.	0.15	0.3	0.8	0.96-0.99	—	Based on 4" spacing below 50 MHz; 2" spacing above 50 MHz. Radiation losses included. Clean, low-loss ceramic insulation assumed. Radiation high above 150 MHz.
Ribbon line, rec. type, 300 ohms. (7/28 conductors)	0.86	2.2	5.3	0.82*	6*	For clean dry line, wet weather performance rather poor, best line is slightly convex. Avoid line that has concave dielectric. Suitable for low-power transmitting applications. Losses increase as line weathers. Handles 400 watts at 30 MHz if SWR is low.
Tubular "twin-lead" rec. type, 300 ohms, 5/16" O.D., (Amphenol type 14-271)	—	—	—	—	—	Characteristics similar to receiving-type ribbon line except for much better wet-weather performance.
Ribbon line, trans. type, 300 ohms.	—	—	—	—	—	Characteristics vary somewhat with manufacturer, but approximate those of receiving-type ribbon except for greater power-handling capability and slightly better wet weather performance.
Tubular "twin-lead" trans. type, 7/16 O.D. (Amphenol 14-076)	0.85	2.3	5.4	0.79	6.1	For use where receiving-type tubular "twin-lead" does not have sufficient power-handling capability. Will handle 1 kw at 30 MHz if SWR is low.
Ribbon line, trans. type, 75 ohms.	1.5	3.9	8.0	0.71*	18*	Very satisfactory for transmitting applications below 30 MHz at powers up to 1 kw. Not significantly affected by wet weather.
RG-8/U coax (52 ohms)	1.0	2.1	4.2	0.66	29.5	Will handle 2 kw at 30 MHz if SWR is low. 0.4" O.D. 7/21 conductor.
RG-11/U coax (75 ohms)	0.94	1.9	3.8	0.66	20.5	Will handle 1.4 kw at 30 MHz if SWR is low. 0.4" O.D. 7/26 conductor.
RG-17/U coax (52 ohms)	0.38	0.85	1.8	0.66	29.5	Will handle 7.8 kw at 30 MHz if SWR is low. 0.87" O.D. 0.19" dia. conductor.
RG-58/U coax (53 ohms)	1.95	4.1	8.0	0.66	28.5	Will handle 430 watts at 30 MHz if SWR is low. 0.2" O.D. No. 20 conductor.
RG-59/U coax (73 ohms)	1.9	3.8	7.0	0.66	21	Will handle 680 watts at 30 MHz if SWR is low. 0.24" O.D. No. 22 conductor.
TV-59 coax (72 ohms)	2.0	4.0	7.0	0.66	22	"Commercial" version of RG-59/U for less exacting applications. Less expensive.
RG-22/U shielded pair (95 ohms)	1.7	3.0	5.5	0.66	16	For shielded, balanced-to-ground applications. Very low noise pickup. 0.4" O.D.
K-111 shielded pair (300 ohms)	2.0	3.5	6.1	—	4	Designed for TV lead-in in noisy locations. Losses higher than regular 300-ohm ribbon, but do not increase as much from weathering.

* Approximate. Exact figure varies slightly with manufacturer.

FIGURE 21

Older type coaxial lines have a useful life of three to six years after which the cable attenuation gradually rises, especially under conditions of heat. Newer cables (designated by the suffix A: RG-8A/U for example) have useful life up to twelve years or so. The 52-ohm series cables have been recently replaced with 50-ohm cables, RG-8A/U now being designated RG-213/U. Long-life versions of the RG-58 family are: RG-58B/U (53.5-ohm) and RG-58C/U (50-ohm).

where,

S is the exact distance between wire centers in some convenient unit of measurement,

d is the diameter of the wire measured in the same units as the wire spacing, S.

Since $\frac{2S}{d}$ expresses a ratio only, the units of measurement may be centimeters, millimeters, or inches. This makes no difference in the answer, so long as the substituted values for S and d are in the *same units*.

The equation is accurate *so long as the wire spacing is relatively large as compared to the wire diameter.*

Surge impedance values of less than 200 ohms are seldom used in the open-type two-wire line, and, even at this rather high value of Z_o the wire spacing S is uncomfortably close, being only 2.7 times the wire diameter.

Figure 20 gives in graphical form the surge impedance of practical two-wire lines. The chart is self-explanatory, and is sufficiently accurate for practical purposes.

Ribbon and Tubular Transmission Line Instead of using spacer insulators placed periodically along the transmission line it is possible to mold the line conductors into a ribbon or tube of flexible low-loss dielectric material. Such line, with polyethylene dielectric, is used in enormous quantities as the lead-in transmission line for f-m and TV receivers. The line is available from several manufacturers in the ribbon and tubular configuration, with characteristic impedance values from 75 to 300 ohms. Receiving types, and transmitting types of power levels up to one kilowatt in the hf range, are listed with their pertinent characteristics, in the table of figure 21.

Coaxial Line Several types of coaxial cable have come into wide use for feeding power to an antenna system. A cross-sectional view of a coaxial cable (sometimes called concentric cable or line) is shown in figure 22.

As in the parallel-wire line, the power lost in a properly terminated coaxial line is the sum of the effective resistance losses along the length of the cable and the dielectric losses between the two conductors.

Of the two losses, the effective resistance loss is the greater; since it is largely due to the skin effect, the line loss (all other conditions the same) will increase directly as the square root of the frequency.

Figure 22 shows that, instead of having two conductors running side by side, one of the conductors is placed *inside* the other. Since the outside conductor completely shields the inner one, no radiation takes place. The conductors may both be tubes, one within the other; the line may consist of a solid wire within a tube, or it may consist of a stranded or solid inner conductor with the outer conductor made up of one or two wraps of copper shielding braid.

In the type of cable most popular for military and noncommercial use the inner conductor consists of a heavy stranded wire, the outer conductor consists of a braid of copper wire, and the inner conductor is supported within the outer by means of a semisolid dielectric of exceedingly low-loss characteristics called polyethylene. The Army-Navy designation on one size of this cable suitable for power levels up to one kilowatt at frequencies as high as 30 MHz is RG-8/U. The outside diameter of this type of cable is approximately one-half inch. The characteristic impedance of this cable type is 52 ohms, but other similar types of greater and smaller power-handling capacity are available in impedances of 52, 75, and 95 ohms.

When using solid dielectric coaxial cable it is necessary that precautions be taken to ensure that moisture cannot enter the line. If the better grade of connectors manufactured for the line are employed as terminations, this condition is automatically satisfied. If connectors are not used, it is necessary that some type of moisture-proof sealing com-

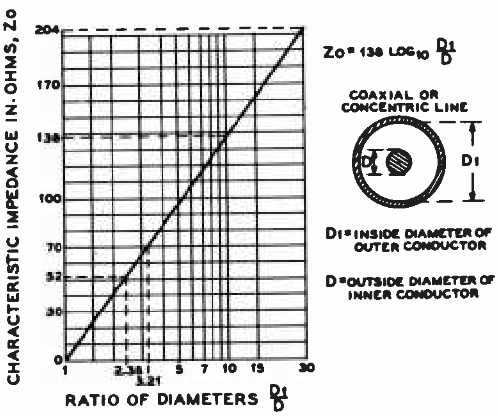


Figure 22

CHARACTERISTIC IMPEDANCE OF AIR-FILLED COAXIAL LINES

If the filling of the line is a dielectric material other than air, the characteristic impedance of the line will be reduced by a factor proportional to the square-root of the dielectric constant of the material used as a dielectric within the line.

pound be applied to the end of the cable where it will be exposed to the weather.

Nearby metallic objects cause no loss, and coaxial cable may be run up air ducts or elevator shafts, inside walls, or through metal conduit. Insulation troubles can be forgotten. The coaxial cable may be buried in the ground or suspended above ground.

Standing Waves Standing waves on a transmission line *always* are the result of the reflection of energy. The only significant reflection which takes place in a normal installation is that at the load end of the line. But reflection can take place from discontinuities in the line, such as caused by insulators, bends, or metallic objects adjacent to an unshielded line.

When a uniform transmission line is terminated in an impedance equal to its surge impedance, reflection of energy does not occur, and no standing waves are present. Thus, for proper operation of an untuned line (with standing waves eliminated), some form of impedance-matching arrangement must be used between the transmission line and the antenna, so that the radiation resistance of the antenna is reflected back into the line as a nonreactive impedance equal to the line impedance.

The termination at the antenna end is the only critical characteristic about the untuned line fed by a transmitter. It is the reflection from the antenna end which starts waves moving back toward the transmitter end. When waves moving in both directions along a conductor meet, standing waves are set up.

Semiresonant Parallel-Wire Lines A well-constructed open-wire line has acceptably low losses when its length is less than about two wavelengths even when the voltage standing-wave ratio is as high as 10 to 1. A transmission line constructed of ribbon or tubular line, however, should have the standing-wave ratio kept down to not more than about 3 to 1 both to reduce power loss and because the energy dissipation on the line will be localized, causing overheating of the line at the points of maximum current.

Because moderate standing waves can be tolerated on open-wire lines without much loss, a standing-wave ratio of 2/1 or 3/1 is considered acceptable with this type of line, *even when used in an untuned system.* Strictly speaking, a line is untuned, or nonresonant, only when it is perfectly *flat*, with a standing-wave ratio of 1 (no standing waves). However, some mismatch can be tolerated with open-wire untuned lines, so long as the reactance is not objectionable, or is eliminated by cutting the line to approximately resonant length.

24-11 Tuned or Resonant Lines

If a transmission line is terminated in its *characteristic surge impedance*, there will be no reflection at the end of the line, and the current and voltage distribution will be uniform along the line. If the end of the line is either open-circuited or short-circuited, the reflection at the end of the line will be 100 percent, and *standing waves* of very great amplitude will appear on the line. There will still be practically no radiation from the line if it is closely spaced, but voltage nodes will be found every half wavelength, the voltage loops corresponding to current nodes (figure 23).

If the line is terminated in some value of resistance other than the characteristic surge impedance, there will be some reflection, the amount being determined by the amount of mismatch. With reflection, there will be standing waves (excursions of current and voltage) along the line, though not to the same extent as with an open-circuited or short-circuited line. The current and voltage loops will occur at the same *points* along the line as with the open- or short-circuited line, and as the terminating impedance is made to approach the characteristic impedance of the line, the current and voltage along the line will become more uniform. The foregoing assumes, of course, a purely resistive (nonreactive) load. If the load is reactive, standing waves also will be formed. But with a reactive load the nodes will occur at different locations from the node locations encountered with improper resistive termination.

A well built 500- to 600-ohm transmission line may be used as a resonant feeder for lengths up to several hundred feet with very low loss, so long as the amplitude of the standing waves (ratio of maximum to minimum voltage along the line) is *not too great.*

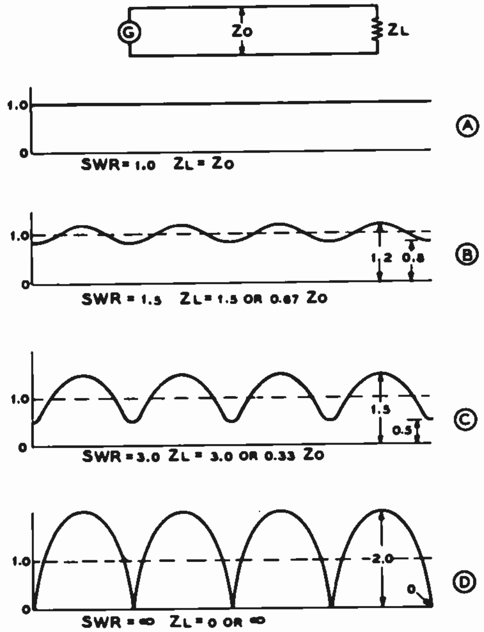


Figure 23

STANDING WAVES ON A TRANSMISSION LINE

As shown at A, the voltage and current are constant on a transmisson line which is terminated in its characteristic impedance, assuming that losses are small enough so that they may be neglected. B shows the variation in current or in voltage on a line terminated in a load with a reflection coefficient of 0.2 so that a standing-wave ratio of 1.5 to 1 is set up. At C the reflection coefficient has been increased to 0.5, with the formation of a 3-to-1 standing-wave ratio on the line. At D the line has been terminated in a load which has a reflection coefficient of 1.0 (short, open circuit, or a pure reactance) so that all the energy is reflected with the formation of an infinite standing-wave ratio.

The amplitude, in turn, depends on the mismatch at the line termination. A line of No. 12 wire, spaced 6 inches with good ceramic or plastic spreaders, has a surge impedance of approximately 600 ohms, and makes an excellent tuned feeder for feeding anything between 60 and 6000 ohms (at frequencies below 30 MHz). If used to feed a load of higher or lower impedance than this, the standing waves become great enough in amplitude that some loss will occur unless the feeder is kept short. At frequencies above 30 MHz, the spacing becomes an appreciable fraction of a wavelength, and radiation from the line no longer is negligible. Hence, coaxial line or close-spaced parallel-wire line is recommended for vhf work.

If a transmission line is not perfectly matched, it should be made *resonant*, even though the amplitude of the standing waves (voltage variation) is not particularly great. This prevents reactance from being coupled into the final amplifier. A feed system having moderate standing waves may be made to present a nonreactive load to the amplifier either by tuning or by pruning the feeders to approximate resonance.

Usually it is preferable with tuned feeders to have a current loop (voltage minimum) at the transmitter end of the line. This means that when voltage-feeding an antenna, the tuned feeders should be made an odd number of quarter wavelengths long, and when current-feeding an antenna, the feeders should be made an even number of quarter wavelengths long. Actually, the feeders are made about 10 percent of a quarter wave longer than the calculated value (the value given in the tables) when they are to be series tuned to resonance by means of a capacitor, instead of being trimmed and pruned to resonance.

When tuned feeders are used to feed an antenna on more than one band, it is necessary to compromise and make provision for both series and parallel tuning, inasmuch as it is impossible to cut a feeder to a length that will be optimum for several bands. If a voltage loop appears at the transmitter end of the line on certain bands, parallel tuning of the feeders will be required in order to get a transfer of energy. It is impossible to transfer energy by inductive coupling unless current is flowing. This is effected as a voltage loop by the presence of the resonant tank circuit formed by parallel tuning of the antenna coil.

24-12 Line Discontinuities

In the previous discussion we have assumed a transmission line which was uniform throughout its length. In actual practice, this is usually not the case.

Whenever there is any sudden change in the characteristic impedance of the line, partial reflection will occur at the point of discontinuity. Some of the energy will be transmitted and some reflected, which is essentially the same as having some of the energy absorbed and some reflected in so far as the effect on the line from the generator to that point is concerned. The discontinuity can be ascribed a reflection coefficient just as in the case of an unmatched load.

In a simple case, such as a finite length of uniform line having a characteristic impedance of 500 ohms feeding into an infinite length of uniform line having a characteristic impedance of 100 ohms, the behavior is easily predicted. The infinite 100-ohm line will have no standing waves and will accept the same power from the 500-ohm line as would a 100-ohm resistor, and the rest of the energy will be reflected at the discontinuity to produce standing waves from there back to the generator. However, in the case of a complex discontinuity placed at an odd distance down a line terminated in a complex impedance, the picture becomes complicated, especially when the discontinuity is neither sudden nor gradual, but intermediate between the two. This is the usual case with amateur lines that must be erected around buildings and trees.

In any case, when a discontinuity exists somewhere on a line and is not a smooth, gradual change embracing several wavelengths, it is not possible to avoid standing waves throughout the entire length of the line. If the discontinuity is sharp enough and is great enough to be significant, standing waves must exist on one side of the discontinuity, and may exist on both sides in many cases.

24-13 A Broadband 50-Ohm Balun

Many triband high-frequency beam antennas feature a balanced input system having a 50-ohm feed point. In order to reduce line discontinuities and to provide a better match between the antenna and an unbalanced transmission line, a *balun* (balance to unbalance) r-f transformer should be used. Shown in figure 24 is a broadband balun that is effective over the range of 6 to 30 MHz. The balun is an inexpensive coil made of a length of coaxial cable and is designed to be installed directly at the terminals of the antenna.

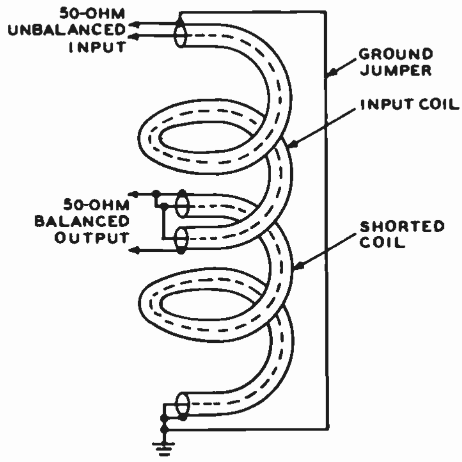


Figure 24

AN EFFECTIVE BROADBAND BALUN FOR MULTIBAND BEAMS

This lumped-constant balun is self-resonant near the center design frequency which, in this case, is about 15 MHz. The balun coil is made of a 16'8"-length of 50-ohm coaxial line (RG-213/U or RG-8A/U) closewound into a coil of 9 turns having an inside diameter of 6¾ inches. At one end of the coil the inner and outer conductors of the line are shorted together and grounded to the common ground point of the antenna assembly. The unbalanced coaxial transmission line is attached to the other end of the coil and a ground jumper is run between the outer ends of the braided conductor. At the center of the winding, the outer braid of the coaxial line is severed for a distance of about one inch, and a connection is made to the inner conductor at this point. In addition, the inner conductor is jumpered to the outer braid of the *shorted* coil section. A second connection is made to the outer braid of the input coil section, as shown in the illustration. These connections are wrapped with vinyl tape and coated with an aerosol plastic spray to protect the joint against the weather. A coaxial plug may be attached to the input terminals of the balun. Connection to the balanced antenna element is made at the center connections of the balun coil, using low-impedance copper straps about ¼ inch wide.

CHAPTER TWENTY-FIVE

Antennas and Antenna Matching

Antennas for the lower-frequency portion of the high-frequency spectrum (from 1.8 to 7.0 MHz), and temporary or limited-use antennas for the upper portion of the high-frequency range, usually are of a relatively simple type in which directivity is not a prime consideration. Also, it often is desirable, in amateur work, that a single antenna system be capable of operation at least on the 3.5- and 7.0-MHz ranges, and preferably on other frequency ranges. Consequently, the first portion of this chapter will be devoted to a discussion of such antenna systems. The latter portion of the chapter is devoted to the general problem of matching the antenna transmission line to antenna systems of the fixed type. Matching the antenna transmission line to the rotatable directive array is discussed in Chapter 26.

25-1 End-Fed Half-Wave Horizontal Antennas

The half-wave horizontal dipole is the most common and the most practical antenna for the 3.5- and 7-MHz amateur bands. The form of the dipole, and the manner in which it is fed are capable of a large number of variations. Figure 2 shows a number of practical forms of the simple dipole antenna along with methods of feed.

Usually a high-frequency doublet is mounted as high and as much in the clear as possible, for obvious reasons. However, it is sometimes justifiable to bring part of the radiation system directly to the transmitter, feeding the antenna without benefit of a transmission line. This is permissible when (1) there is insufficient room to erect a 75- or 80-meter horizontal dipole and feed line, (2) when a long wire is also to be operated on one of the higher-frequency bands on a harmonic. In either case, it is usually possible to get the main portion of the antenna in the clear because of its length. This means that the power lost by bringing the antenna directly to the transmitter is relatively small.

End-Fed Antennas The end-fed antenna has no form of transmission line to couple it to the transmitter, but brings the radiating portion of the antenna right down to the transmitter, where some form of coupling system is used to transfer energy to the antenna.

Figure 1, shows two common methods of feeding the *end-fed Hertz*. Some harmonic-attenuating provision (in addition to the usual low-pass TVI filter) must be included in the coupling system, since an end-fed antenna itself offers no discrimination against harmonics, either odd or even.

As there is voltage at the point where the antenna enters the operating room, the insulation at that point should be several times as effective as the insulation commonly used with low-voltage feeder systems. This antenna can be operated on all of its higher harmonics with good efficiency, and can be operated at half-frequency against ground as a quarter-wave Marconi.

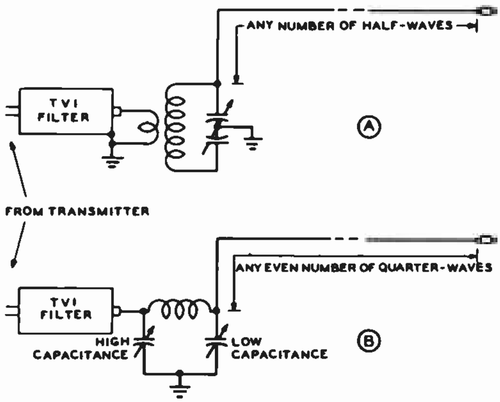


Figure 1

THE END-FED HERTZ ANTENNA

Showing the manner in which an end-fed Hertz antenna may be fed through a low-impedance line and low-pass filter by using a resonant tank circuit as at A, or through the use of a reverse-connected pi-network as at B.

The Zepp Antenna System The *zeppelin*, or *zepp antenna system*, illustrated in figure 2A is very convenient when it is desired to operate a single radiating wire on a number of harmonically related frequencies.

The zepp antenna system is easy to tune, and can be used on several bands by merely retuning the feeders. As the radiating portion of the zepp antenna system must always be some multiple of a half wave long, there is always high voltage present at the point where the live zepp feeder attaches to the end of the radiating portion of the antenna. Thus, this type of zepp antenna system is *voltage-fed*.

Stub-Fed Zepp-Type Radiator Figure 2C shows a modification of the zepp-type antenna system to allow the use of a nonresonant transmission line between the radiating portion of the antenna and the transmitter. The *zepp* portion of the antenna is resonated as a quarter-wave stub and the nonresonant feeders are connected to the stub at a point where standing waves on the feeder are minimized. The procedure for making these adjustments is described in detail in Section 25-8.

25-2 Center-Fed Half-Wave Horizontal Antennas

A center-fed half-wave antenna system is usually to be desired over an end-fed system since the center-fed system is inherently balanced to ground and is therefore less likely to be troubled by feeder radiation. A number of center-fed systems are illustrated in figure 2.

The Tuned Doublet The current-fed doublet with spaced feeders, sometimes called a *center-fed zepp*, is an inherently balanced system if the two legs of the radiator are electrically equal. This fact holds true regardless of the frequency, or of the harmonic, on which the system is operated. The system can successfully be operated over a wide range of frequencies if the system as a whole (both tuned feeders and the center-fed flat top) can be resonated to the operating frequency. It is usually possible to tune such an antenna system to resonance with the aid of a tapped coil and a tuning capacitor that can optionally be placed either in series with the antenna coil or in parallel with it (figure 2D).

The antenna has a different radiation pattern when operated on its harmonics, as would be expected. The arrangement used on the second harmonic is better known as the *Franklin collinear array* and is described in Chapter 26.

Figures 2E and 2F show alternative arrangements for using an untuned transmission line between the transmitter and the tuned-doublet radiator. In figure 2E a half-wave shorted line is used to resonate the radiating system, while in figure 2F a quarter-wave open line is utilized. The adjustment of quarter-wave and half-wave stubs is discussed in Section 25-8.

Doublets with Quarter-Wave Transformers The average value of feed impedance for a center-fed half-wave doublet is 75 ohms. The actual value varies with height and is shown in Chapter 24. Other methods of matching this rather low value

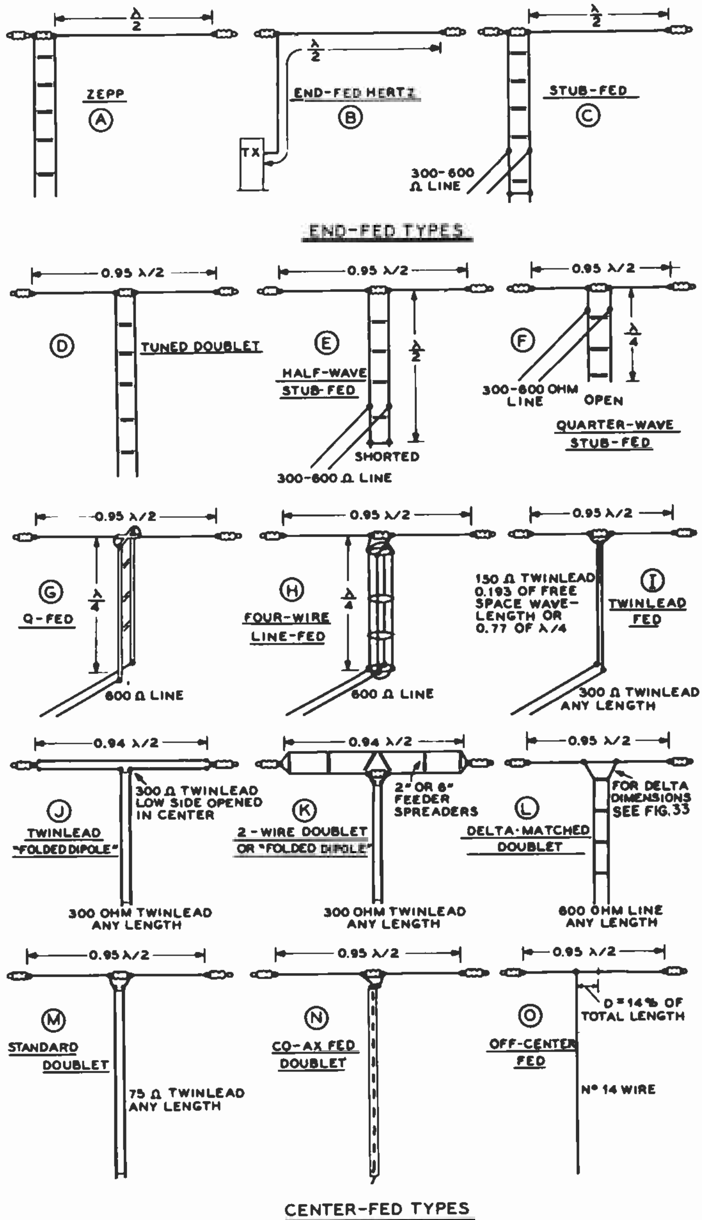


Figure 2

FEED SYSTEMS FOR A HALF-WAVE DIPOLE ANTENNA

The half-wave dipole antenna may be either center- or end-fed, as discussed in the text. For the hf region (below 30 MHz), the length of a simple dipole is computed by: length (feet) = 468/f, with f in MHz. For the folded dipole, length is computed by: length (feet) = 462/f, with f in MHz. Above 30 MHz, the length of the dipole is affected to an important degree by the diameter of the element and the method of supporting the dipole (see VHF and UHF Antennas and Radiation, Propagation, and Lines chapters).

of impedance to a medium-impedance transmission line are shown in G, H, and I of figure 2. Each of these three systems uses a quarter-wave transformer to accomplish the impedance transformation. The only difference between the three systems lies in the type of transmission line used in the quarter-wave transformer. G shows the *Q-match system* whereby a line made up of ½-inch dural tubing is used for the low-impedance linear transformer. A line made up in this manner is frequently called a set of *Q bars*. Illustration H shows the use of a four-wire line as the linear transformer, and I shows the use of a piece of 150-ohm twin-lead electrically ¼-wave in length as the transformer between the center of the dipole and a piece of 300-ohm twin-lead. In any case the impedance of the quarter-wave transformer will be of the order of 150 to 200 ohms. The use of sections of transmission line as linear transformers is discussed in detail in Section 25-8.

Multiwire Doublets An alternative method for increasing the feed-point impedance of a dipole so that a medium-impedance transmission line may be used is shown in figures 2J and 2K. This system utilizes more than one wire in parallel for the radiating element, but only one of the wires is broken for attachment of the feeder. The most common arrangement uses two wires in the flat top of the antenna so that an impedance multiplication of four is obtained.

The antenna shown in figure 2J is the so-called *twin-lead folded dipole* which is a commonly used antenna system on the medium-frequency amateur bands. In this arrangement both the antenna and the transmission line to the transmitter are constructed of 300-ohm twin-lead. The flat top of the antenna is made slightly less than the conventional length ($462/F_{MHz}$ instead of $468/F_{MHz}$ for a single-wire flat top) and the two ends of the twin-lead are joined together at each end. The center of one of the conductors of the twin-lead flat top is broken and the two ends of the twin-lead feeder are spliced into the flat-top leads.

Better bandwidth characteristics can be obtained with a folded dipole made of ribbon line if the two conductors of the ribbon line are shorted a distance of 0.82 (the velocity factor of ribbon line) of a free-space quarter-wave-length from the center or feed point. This procedure is illustrated in figure 3A. An alternative arrangement for a twin-lead folded dipole is illustrated in figure 3B. This type of half-wave antenna system is convenient for use on the 3.5-MHz band when the 116- to 132-foot distance required for a full half-wave is not quite available in a straight line, since the single-wire end pieces may be bent away or downward from the direction of the main section of the antenna.

Figure 2K shows the basic type of two-wire doublet or *folded dipole* wherein the radiating section of the system is made up of standard antenna wire spaced by means of feeder spreaders. The feeder again is made of 300-ohm twin-lead since the feed-point impedance is approximately 300 ohms, the same as that of the twin-lead folded dipole.

The folded-dipole type of antenna has the broadest response characteristics (greatest

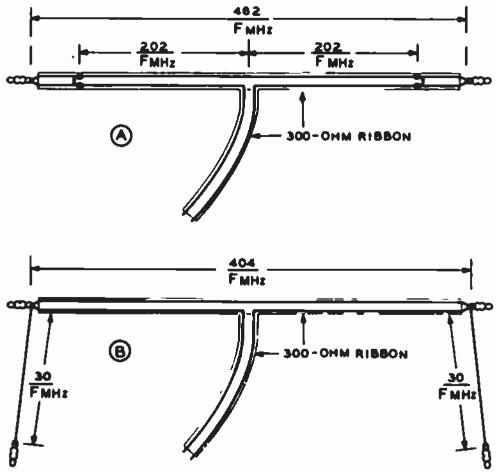


Figure 3

FOLDED DIPOLE WITH SHORTING STRAPS

The impedance match and bandwidth characteristics of a folded dipole may be improved by shorting the two wires of the ribbon a distance out from the center equal to the velocity factor of the ribbon times the half-length of the dipole as shown at A. An alternative arrangement with bent down ends for space conservation is illustrated at B.

bandwidth) of any of the conventional half-wave antenna systems constructed of small wires or conductors. Hence such an antenna may be operated over the greatest frequency range, without serious standing waves, of any common half-wave antenna types.

Delta-Matched Doublet and Standard Doublet These two types of radiating elements are shown in figure 2L and figure 2M. The delta-matched doublet is described in detail in Section eight of this chapter. The standard doublet, shown in figure 2M, is fed in the center by means of 75-ohm transmitting type twin-lead.

The coaxial-fed doublet shown in figure 2N is a variation on the system shown in figure 2M. Either 52-ohm or 75-ohm coaxial cable may be used to feed the center of the dipole, although the 52-ohm type will give a somewhat better impedance match at lower antenna heights.

Off-Center–Fed Doublet The system shown in figure 20 is sometimes used to feed a half-wave dipole, especially when it is desired to use the same antenna on a number of harmonically related frequencies. The feeder wire (No. 14 enameled wire should be used) is tapped a distance of 14 percent of the total length of the antenna either side of center. The feeder wire, operating against ground for the return current, has an impedance of approximately 300 ohms.

25-3 The Half-Wave Vertical Antenna

The half-wave vertical antenna with its bottom end from 0.01 to 0.2 wavelength above ground is an effective transmitting antenna for low-angle radiation, where ground conditions in the vicinity of the antenna are good. Such an antenna is not good for short-range *sky-wave* communication, such as is the normal usage of the 3.5-MHz amateur band, but is excellent for short-range *ground-wave* communication such as on the standard broadcast band and on the amateur 1.8-MHz band. The vertical antenna may cause greater BCI than an equivalent horizontal antenna, due to the much greater ground-wave field intensity. Also, the vertical antenna is poor for receiving under conditions where man-made interference is severe, since such interference is predominantly of vertical polarization.

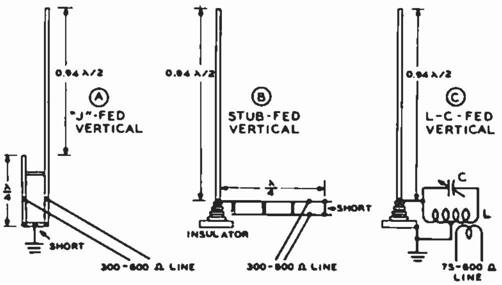


Figure 4

HALF-WAVE VERTICAL ANTENNA SHOWING ALTERNATIVE METHODS OF FEED

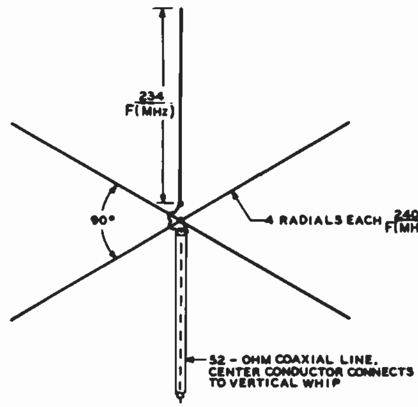


Figure 5

THE LOW-FREQUENCY GROUND PLANE ANTENNA

The radials of the ground plane antenna should lie in a horizontal plane, although slight departures from this caused by nearby objects is allowable. The whip may be mounted on a short post, or on the roof of a building. The wire radials may slope downwards toward their tips, acting as guy wires for the installation.

Three ways of feeding a half-wave vertical antenna with an untuned transmission line are illustrated in figure 4. The J-fed system shown in figure 4A is obviously not practical except on the higher frequencies where the extra length for the stub may easily be obtained.

25-4 The Ground-Plane Antenna

An effective low-angle radiator for any amateur band is the *ground-plane antenna*, shown in figure 5. So named because of the radial ground wires, the ground-plane antenna is not affected by soil conditions in its vicinity due to the creation of an artificial ground system by the radial wires. The base impedance of the ground plane is of the order of 30 to 35 ohms, and it may be fed with 52-ohm coaxial line with only a slight impedance mismatch. For a more exact match, the ground-plane antenna may be fed with a 72-ohm coaxial line and a quarter-wave matching section made of 52-ohm coaxial line.

The angle of radiation of the ground-plane antenna is quite low, and the antenna will be found more effective for communication over 400 miles or so on the 80 and 40 meter bands than a high-angle radiator, such as a dipole.

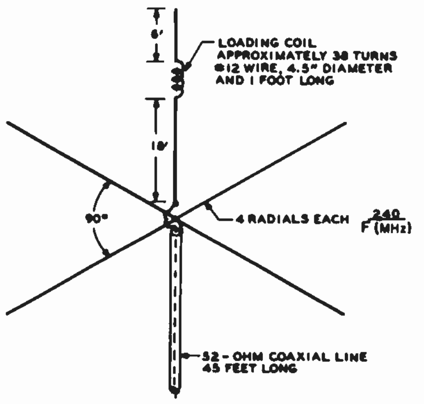


Figure 6

80-METER LOADED GROUND PLANE ANTENNA

Number of turns in loading coil to be adjusted until antenna system resonates at desired frequency in 80-meter band.

The 80-Meter Loaded Ground Plane A vertical antenna of 66 feet in height presents quite a problem on a small lot, as the supporting guy wires will tend to take up quite a large portion of the lot. Under such conditions, it is possible to shorten the length of the vertical radiator of the ground plane by the inclusion of a loading coil in the vertical whip section. The ground-plane antenna can be artificially loaded in this manner so that a 25-foot vertical whip may be used for the radiator. Such an antenna is shown in figure 6. The loaded ground plane tends to have a rather high Q and operates only over a narrow band of frequencies. An operating range of about 100 kHz with a low SWR is possible on 80 meters. Operation over a larger frequency range is possible if a higher standing wave ratio is tolerated on the transmission line. The radiation resistance of a loaded 80-meter ground plane is about 15 ohms.

25-5 The Marconi Antenna

A grounded quarter-wave *Marconi antenna*, widely used on frequencies below 3 MHz, is sometimes used on the 3.5-MHz band, and is also used in vhf mobile services where a compact antenna is required. The Marconi type antenna allows the use of half the length of wire that would be required for a half-wave Hertz radiator. The ground acts as a mirror, in effect, and takes the place of the additional quarter wave of wire that would be required to reach resonance if the end of the wire were not returned to ground.

The fundamental practical form of the Marconi antenna system is shown in figure 7. Other Marconi antennas differ from this type primarily in regard to the method of feeding the energy to the radiator. The feed method shown in figure 7B can often be used to advantage, particularly in mobile work.

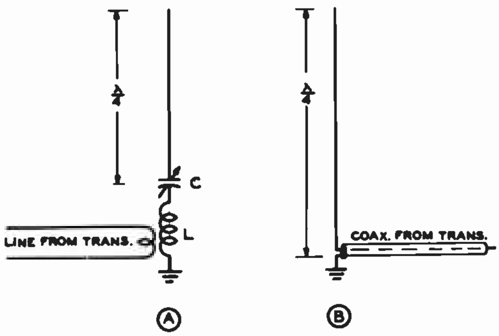


Figure 7

FEEDING A QUARTER-WAVE MARCONI ANTENNA

When an open-wire line is to be used, it may be link-coupled to a series-resonant circuit between the bottom end of the Marconi and ground, as at A. Alternatively, a reasonably good impedance match may be obtained between 52-ohm coaxial line and the bottom of a resonant quarter-wave antenna, as illustrated at B above.

Variations on the basic Marconi antenna are shown in the illustrations of figure 8. Figures 8B and 8C show the L-type and T-type Marconi antennas. These arrangements have been more or less superseded by the top-loaded forms of the Marconi antenna shown in figures 8D, 8E, and 8F. In each of these latter three configurations an antenna somewhat less than one quarter wave in length has been *loaded* to increase its effective length by the insertion of a *loading coil* at or near the top of the radiator. The arrangement shown at figure 8D gives the least loading but is the most practical mechanically. The system shown at figure 8E gives an intermediate amount of loading, while that shown at figure 8F, utilizing a "hat" just above the loading coil, gives the greatest amount of loading. The object of all the top-loading methods shown is to produce an increase in the effective length of the radiator, and thus to raise the point of maximum current in the radiator as far as possible above ground. Raising the maximum-current point in the radiator above ground has two desirable results: The percentage of low-angle radiation is increased and the amount of ground current at the base of the radiator is reduced, thus reducing the ground losses.

Amateurs primarily interested in the higher-frequency bands, but liking to work 80 meters occasionally, can usually manage to resonate one of their antennas as a Marconi by working the whole system (feeders and all) against a water pipe ground, and resorting to a loading coil if necessary.

Importance of Ground Connection With a quarter-wave antenna and a ground, the antenna current generally is measured with a meter placed in the antenna circuit close to the ground connection. If this current flows through a resistor, or if the ground itself presents some resistance, there will be a power loss in the form of heat. Improving the ground connection, therefore, provides a definite means of reducing this power loss, and thus increasing the radiated power.

The best possible ground consists of as many wires as possible, each at least a quarter wave long, buried just below the surface of the earth, and extending out from a common point in the form of *radials*. Copper wire of any size larger than No. 16 is satisfactory, and the larger sizes will take longer to disintegrate. In fact, the radials need not even be buried; they may be supported just above the earth, and insulated from it. This arrangement is called a *counterpoise*, and operates by virtue of its high capacitance to ground.

If the antenna is physically shorter than a quarter wavelength, the antenna current is higher, due to lower radiation resistance; consequently, the power lost in resistive soil is greater. The importance of a good ground with short, inductive-loaded Marconi radiators is, therefore, quite obvious. With a good ground system, even very short (one-eighth wavelength) antennas can be expected to give a high percentage of the efficiency of a quarter-wave antenna used with the same ground system. This is especially true when the short radiator is *top loaded* with a high-Q (low-loss) coil.

Water-Pipe Grounds Copper water pipe, because of its comparatively large surface and cross section, has a relatively low r-f resistance. If it is possible to attach to a junction of several water pipes a satisfactory ground connection will be obtained. If one of the pipes attaches to a lawn or garden sprinkler system in the immediate vicinity of the antenna, the effectiveness of the system will approach that of buried copper radials.

The main objection to iron water-pipe grounds is the possibility of high-resistance joints in the pipe, due to the "dope' put on the coupling threads. By attaching the ground wire to a junction with three or more legs, the possibility of requiring the main portion of the r-f current to flow through a high resistance connection is greatly reduced.

Marconi Dimensions A Marconi antenna is an odd number of electrical quarter waves long (usually only one quarter wave in length), and is always resonated to the operating frequency. The correct loading of the final amplifier is accomplished by varying the coupling, rather than by detuning the antenna from resonance.

Physically, a quarter-wave Marconi may be made anywhere from one-eighth to three-eighths wavelength overall, including the total length of the antenna wire and ground lead from the end of the antenna to the point where the ground lead attaches to the junction of the radials or counterpoise wires, or where the water pipe enters the ground. The longer the antenna is made physically, the lower will be the current flowing in the ground connection, and the greater will be the over-all radiation efficiency. However, when the antenna length

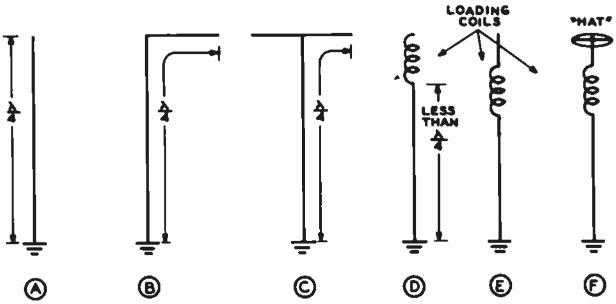


Figure 8

LOADING THE MARCONI ANTENNA

The various loading systems are discussed in the accompanying text.

exceeds three-eights wavelength, the antenna becomes difficult to resonate by means of a series capacitor, and it begins to take shape as an end-fed Hertz, requiring a method of feed such as a pi-network.

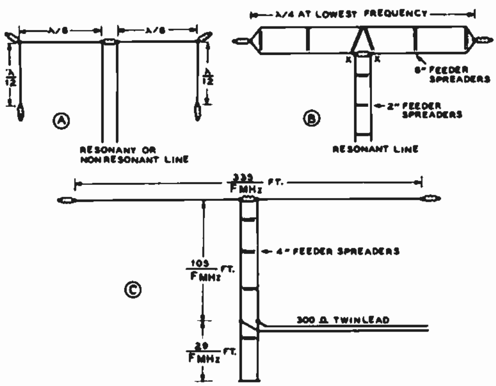


Figure 9

THREE EFFECTIVE SPACE-CONSERVING ANTENNAS

The arrangements shown at A and B are satisfactory where resonant feed line can be used. However, nonresonant 75-ohm feed line may be used in the arrangement at A when the dimensions in wavelengths are as shown. In the arrangement shown at B, low standing waves will be obtained on the feed line when the over-all length of the antenna is a half wave. The arrangement shown at C may be tuned for any reasonable length of flat top to give a minimum of standing waves.

The Radial Ground Wire The ground termination for a Marconi or other unbalanced antenna system can be improved by the addition of a radial ground wire which is connected in parallel with the regular ground connection. The radial wire consists simply of a quarter wavelength of insulated wire connected to the ground terminal of the transmitter. The opposite end of the radial wire is left disconnected, or "floating." The radial wire may be run about the baseboard of the operating room or out the window and a foot or two above the ground. A high-impedance point is esablished at the end of the wire and a corresponding low-impedance (ground) point at the transmitter end which simulates a ground connection. While it may be used by itself as a ground termination, the radial ground wire works best when used in combination with a regular ground connection. Its use is highly recommended with all the antennas shown in this Handbook which require an external ground connection. Since the radial wire is a tuned device, separate radial wires cut to length are required for each amateur band. Several such radials can be connected in parallel at the transmitter ground point for multiband operation.

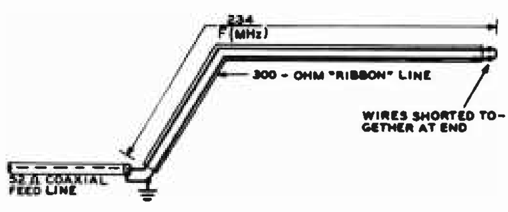


Figure 10

TWIN-LEAD MARCONI ANTENNA FOR THE 80- AND 160-METER BANDS

25-6 Space-Conserving Antennas

In many cases it is desired to undertake a considerable amount of operation on the 80- or 40-meter band, but sufficient space is simply not available for the installation of a half-wave radiator for the desired frequency of operation. This is a common experience of apartment dwellers.

One technique of producing an antenna for lower-frequency operation in restricted space is to erect a short radiator which is balanced with respect to ground and which is therefore independent of ground for its operation. Several antenna types meeting this set of conditions are shown in figure 9. Figure 9A shows a conventional center-fed doublet with bent-down ends. This type of antenna can be fed with 75-ohm twin-lead in the center, or it may be fed with a resonant line for operation on several bands. The over-all length of the radiating wire will be a few percent greater than the normal length for such an antenna since the wire is bent at a position intermediate between a current loop and a voltage loop.

Figure 9B shows a method for using a two-wire doublet on one-half of its normal operating frequency. It is recommended that spaced open conductor be used both for the radiating portion of the folder dipole and for the feed line. The reason for this lies in the fact that the two wires of the flat top are *not* at the same potential throughout their length when the antenna is operated on one-half frequency. Twin-lead may be used for the feed line if operation on the frequency where the flat top is one half-wave in length is most common, and operation on half frequency is infrequent. However, if the antenna is to be used primarily on the half frequency as shown, it should be fed by means of an open-wire line. If it is desired to feed the antenna with a nonresonant line, a quarter-wave stub may be connected to the antenna at the points X—X in figure 9B. The stub should be tuned and the transmission line connected to it in the normal manner.

The antenna system shown in figure 9C may be used when not quite enough length is available for a full half-wave radiator. The dimensions in terms of frequency are given on the drawing. An antenna of this type is 93 feet long for operation on 3600 kHz and 86 feet long for operation on 3900 kHz. This type of antenna has the additional advantage that it may be operated on the 7- and 14-MHz bands, when the flat top has been cut for the 3.5-MHz band, simply by changing the position of the shorting bar and the feeder line on the stub.

A sacrifice which must be made when using a shortened radiating system (as for example the types shown in figure 9), is in the bandwidth of the radiating system. The frequency range which may be covered by a shortened antenna system is approximately in proportion to the amount of shortening which has been employed.

The Twin-Lead Marconi Antenna Much of the power loss in the Marconi antenna is a result of low radiation resistance and high ground resistance. If the radiation resistance of the Marconi antenna is raised, the amount of power lost in the ground resistance is proportionately less. If a Marconi antenna is made out of 300-ohm TV-type ribbon line, as shown in figure 10, the radiation resistance of the antenna is raised from a low value of 10 or 15 ohms to a more reasonable value of 40 to 60 ohms. The ground losses are now reduced by a factor of 4. In addition, the antenna may be directly fed from a 50-ohm coaxial line, or directly from the unbalanced output of a pi-network transmitter.

A Broadband Dipole System Shown in figures 11 and 12 are broadband dipoles for the 40- and 80-meter amateur bands. These fan-type dipoles have excellent broadband response, and are designed to be fed with a 52-ohm unbalanced coaxial line. The antenna system consists of a fan-type dipole, a balun matching section, and a suitable coaxial feedline. The Q of the half-wave 80-meter doublet is lowered by decreasing the effective length-to-diameter ratio. The frequency range of operation of the doublet is increased considerably by this change. A typical SWR curve for the 80-meter doublet is shown in figure 13.

The balanced doublet is matched to the unbalanced coaxial line by the quarter-wave balun. If desired, a shortened balun may be used (figure 14). The short balun is capacitance loaded at the junction between the balun and the broadband dipole.

The Inverted-V Antenna The *Inverted-V* antenna is a center-fed dipole with the ends lower than the middle. The radiation pattern is similar to a dipole,

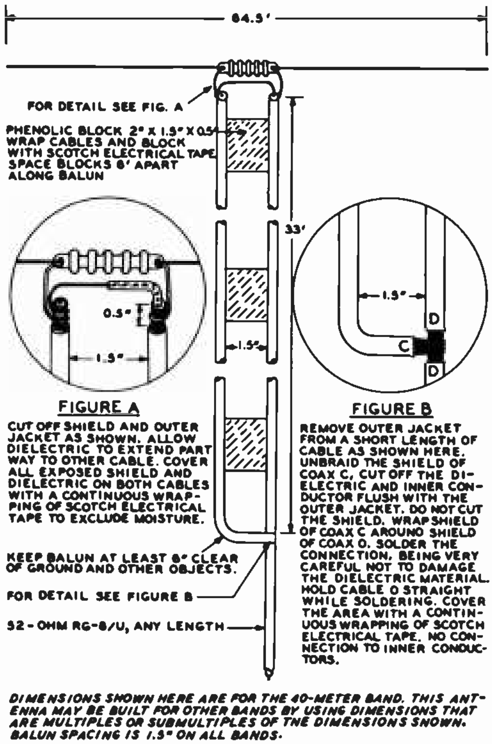


Figure 11

HALF-WAVE ANTENNA WITH QUARTER-WAVE UNBALANCED-TO-BALANCED TRANSFORMER (BALUN) FEED SYSTEM FOR 40-METER OPERATION

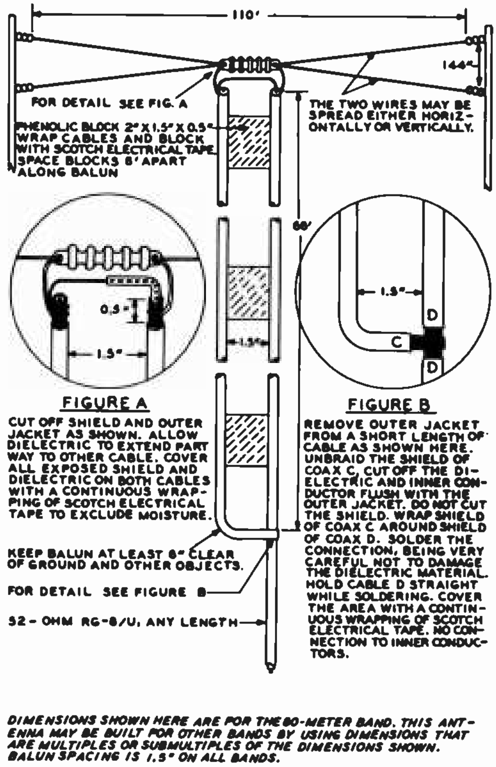


Figure 12

BROADBAND ANTENNA WITH QUARTER-WAVE UNBALANCED-TO-BALANCED TRANSFORMER (BALUN) FEED SYSTEM FOR 80-METER OPERATION

except that more radiation is apparent off the ends of the antenna. The main advantage of this antenna is that it may be hung, or supported, at the center from an existing tower, with the ends tied off near the surface of the ground. For 40- or 80-meter inverted V's, the center support should be from 40 to 60 feet above ground and the ends should clear the ground by at least 10 feet.

The impedance of an inverted V is less than that of an equivalent dipole placed at the apex of the array, however, a good match may be had to 50-ohm coaxial transmission line. Bandwidth is about equal to that of a regular horizontal dipole.

The length of the inverted V is greater than that of a dipole and may be computed from the following formula:

$$\text{Over-all length (feet)} = \frac{485}{f_{\mathrm{MHz}}}$$

Objects near the end of the inverted V will affect the resonant frequency to some extent and the installation may require end trimming to bring it on or near the desired frequency, especially in the presence of nearby buildings, telephone wires or other conductors. For best results, a balun should be used between the inverted V and the coaxial feedline.

25-7 Multiband Antennas

The availability of a *multiband antenna* is a great operating convenience to an amateur station. In most cases it will be found best to install an antenna which is optimum for the band which is used for the majority of the available operating time, and then to have an additional multiband antenna which may be pressed into service for operation on another band when propagation conditions on the most frequently used band are not suitable.

The choice of a multiband antenna depends on a number of factors such as the amount of space available, the band which is to be used for the majority of operation with the antenna, the radiation efficiency which is desired, and the type of antenna tuning network to be used at the transmitter. A number of recommended types are shown on the next pages.

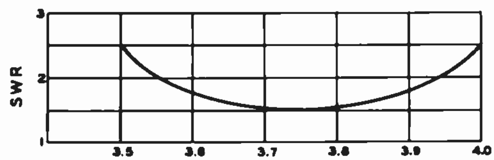


Figure 13

SWR CURVE OF 80-METER BROADBAND DIPOLE

The ¾-Wave Folded Doublet Figure 15 shows an antenna type which will be found to be very effective when a moderate amount of space is available, when most of the operating will be done on one band with occasional operation on the second harmonic. The system is quite satisfactory for use with high-power transmitters since a 600-ohm nonresonant line is used from the antenna to the transmitter and since the antenna system is balanced

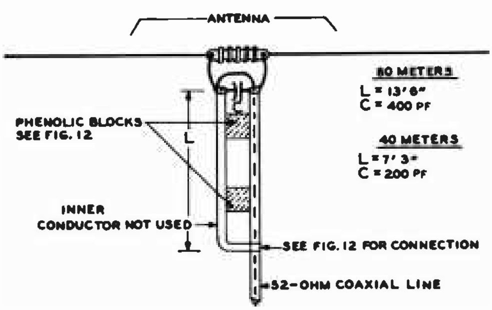


Figure 14

SHORT BALUN FOR 40 AND 80 METERS

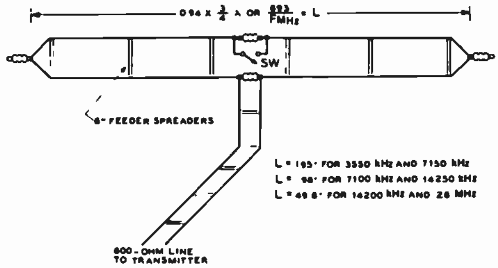


Figure 15

THE THREE-QUARTER WAVE FOLDED DOUBLET

This antenna arrangement will give very satisfactory operation with a 600-ohm feed line for operation with the switch open on the fundamental frequency and with the switch closed on twice frequency. A balun may be used to match the 600-ohm line to the transmitter.

with respect to ground. With operation on the fundamental frequency of the antenna where the flat top is ¾ wave long the switch SW is left open. The system affords a very close match between the 600-ohm line and the feed point of the antenna. A standing-wave ratio of approximately 1.2 to 1 over the 14-MHz band exists when the antenna is located approximately one-half wave above ground.

For operation on the second harmonic the switch SW is *closed*. The antenna is still an effective radiator on the second harmonic but the pattern of radiation will be different from that on the fundamental, and the standing-wave ratio on the feed line will be greater. The flat top of the antenna must be made of open wire rather than ribbon or tubular line.

For greater operating convenience, the shorting switch may be replaced with a section of transmission line. If this transmission line is made one-quarter wavelength long for the fundamental frequency, and the free end of the line is shorted, it will act as an open circuit across the center insulator. At the second harmonic, the transmission line is one-half wavelength long, and reflects the low impedance of the shorted end across the center insulator. Thus the switching action is automatic as the frequency of operation is changed. Such an installation is shown in figure 16.

The End-Fed Hertz The end-fed Hertz antenna shown in figure 17 is not as effective a radiating system as many other antenna types, but it is particularly convenient when it is desired to install an antenna in a hurry for a test, or for field-day work. The flat top of the radiator should be as high and in the clear as possible. In any event at least three quarters of the total wire length should be in the clear.

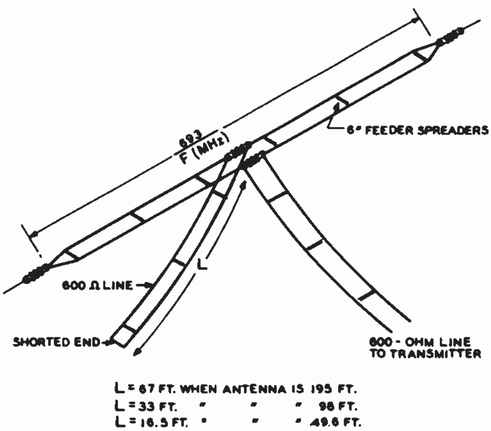


Figure 16

AUTOMATIC BANDSWITCHING STUB FOR THE THREE-QUARTER WAVE FOLDED DOUBLET

The antenna of Figure 15 may be used with a shorted stub line in place of the switch normally used for second-harmonic operation.

The End-Fed Zepp The end-fed zepp is convenient for multiband operation. It is shown in figure 18 along with recommended dimensions for operation on various amateur band groups. Since this antenna type is an unbalanced radiating system, its use is not recommended with high-power transmitters where interference to broadcast listeners is likely to be encountered.

The coupling coil at the transmitter end of the feeder system should be link-coupled to the output of the low-pass TVI filter in order to reduce harmonic radiation.

The Two-Band Marconi Antenna A three-eighths wavelength Marconi antenna may be operated on its harmonic

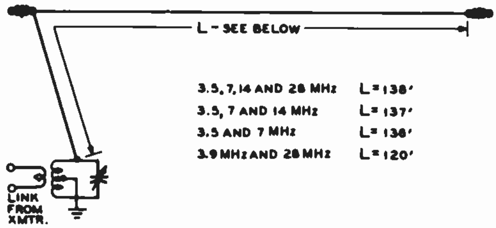
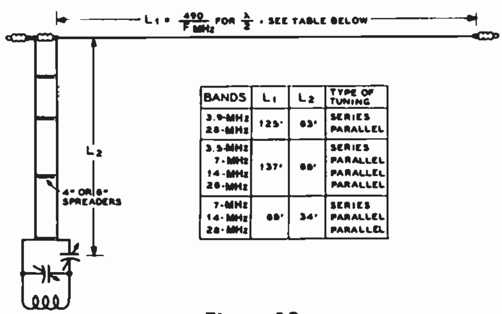


Figure 17

RECOMMENDED LENGTHS FOR THE END-FED HERTZ ANTENNA

frequency, providing good two band performance from a simple wire. Such an arrangement for operation on 160-80 meters, and 80-40 meters is shown in figure 19. On the fundamental (lowest) frequency, the antenna acts as a three-eighths wavelength series-tuned Marconi. On the second harmonic, the antenna is a current-fed three-quarter wavelength antenna operating against ground. For proper operation, the antenna should be resonated on its second harmonic by means of a grid-dip oscillator to the operating frequency most used on this particular band. The Q of the antenna is relatively low, and the antenna will perform well over a frequency range of several hundred kHz.



BANDS	L1	L2	TYPE OF TUNING
3.9-MHz 28-MHz	125'	63'	SERIES PARALLEL
3.5-MHz 7-MHz 14-MHz 28-MHz	137'	68'	SERIES PARALLEL PARALLEL PARALLEL
7-MHz 14-MHz 28-MHz	68'	34'	SERIES PARALLEL PARALLEL

Figure 18

END-FED ZEPP

The over-all length of the antenna may be varied slightly to place its self-resonant frequency in the desired region. Bends or turns in the antenna tend to make it resonate higher in frequency, and it may be necessary to lengthen it a bit to resonate it at the chosen frequency. For fundamental operation, the series capacitor is inserted in the circuit, and the antenna may be resonated to any point in the lower-frequency

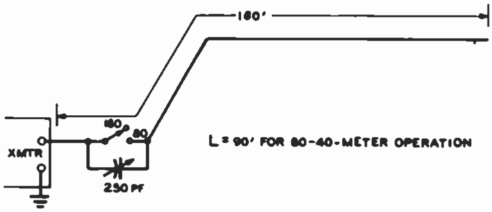


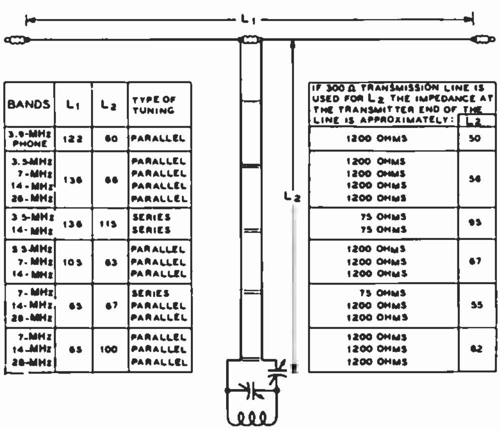
Figure 19

A TWO-BAND MARCONI ANTENNA FOR 160-80 METER OPERATION

band. As with any Marconi-type antenna, the use of a good ground is essential. This antenna works well with transmitters employing coaxial antenna feed, since its transmitting impedance on both bands is in the neighborhood of 40 to 60 ohms. It may be attached directly to the output terminal of a pi-network transmitter coupling circuit. The use of a low-pass TVI filter is of course recommended.

The Center-Fed Multiband Antenna For multiband operation, the center-fed antenna is without doubt the best compromise. It is a balanced system on all bands, it requires no ground return, and when properly tuned has good rejection properties for the higher harmonics generated in the transmitter. It is well suited for use with the various multiband 150-watt transmitters that are currently so popular. For proper operation with these transmitters, an antenna tuning unit *must* be used with the center-fed antenna. In fact, some sort of tuning unit is necessary for any type of efficient, multiband antenna.

Various dimensions for center-fed antenna systems are shown in figure 20. If the feed line is made up in the conventional manner of No. 12 or No. 14 wire spaced 4 to 6 inches, the antenna system is sometimes called a *center-fed zepp*. With this type of feeder the impedance at the transmitter end of the feeder varies from about 70 ohms to approximately 5000 ohms, the same range encountered in an end-fed zepp antenna. This great impedance ratio requires provision for either series or parallel tuning of the feeders at the transmitter, and involves quite high r-f voltages at various points along the feed line.



BANDS	L1	L2	TYPE OF TUNING
3.9-MHz PHONE	122	60	PARALLEL
3.5-MHz 7-MHz 14-MHz 28-MHz	136	66	PARALLEL PARALLEL PARALLEL PARALLEL
3.5-MHz 14-MHz	136	115	SERIES SERIES
3.5-MHz 7-MHz 14-MHz	105	63	PARALLEL PARALLEL PARALLEL
7-MHz 14-MHz 28-MHz	65	67	SERIES PARALLEL PARALLEL
7-MHz 14-MHz 28-MHz	65	100	PARALLEL PARALLEL PARALLEL

IF 300 Ω TRANSMISSION LINE IS USED FOR L2 THE IMPEDANCE AT THE TRANSMITTER END OF THE LINE IS APPROXIMATELY:	L2
1200 OHMS	50
1200 OHMS 1200 OHMS 1200 OHMS 1200 OHMS	56
75 OHMS 75 OHMS	93
1200 OHMS 1200 OHMS 1200 OHMS	67
75 OHMS 1200 OHMS 1200 OHMS	55
1200 OHMS 1200 OHMS 1200 OHMS	62

Figure 20

DIMENSIONS FOR CENTER-FED MULTI-BAND ANTENNA

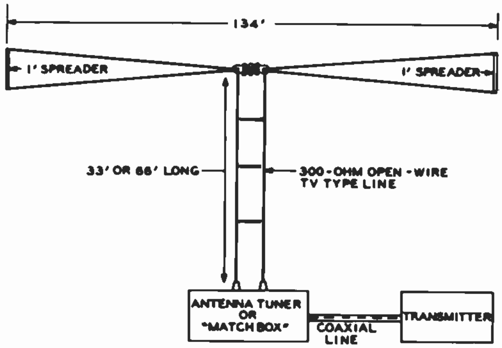


Figure 21

MULTIBAND ANTENNA USING FAN-DIPOLE TO LIMIT IMPEDANCE EXCURSIONS ON HARMONIC FREQUENCIES

To increase operating bandwidth and to limit impedance excursions, a two-wire flat top may be employed for the radiator, as shown in figure 21. The use of such a radiator will limit the impedance excursions on the harmonic frequencies of the antenna and make the operation of the antenna matching unit much less critical. The use of a two-wire radiator is highly recommended for any center-fed multiband antenna.

Folded Flat Top Dual-Band Antenna As has been mentioned earlier, most amateurs use rotary or fixed arrays for the 14-MHz band and those higher in frequency. In order to afford complete coverage of the amateur bands it is then desirable to have an additional system which will operate with equal effectiveness on the 3.5- and 7-MHz bands, but this low frequency antenna system will not be required to operate on any bands higher in frequency than the 7-MHz band. The antenna system shown in figure 22 has been developed to fill this need.

This system consists essentially of an open-wire folded dipole for the 7-MHz band with a special feed system which allows the antenna to be fed with minimum standing waves on the feed line on both the 7-MHz and 3.5-MHz bands. The feed-point impedance of a folded dipole on its fundamental frequency is approximately 300 ohms. Hence the 300-ohm twin lead shown in figure 22 can be connected directly into the center of the system for operation only on

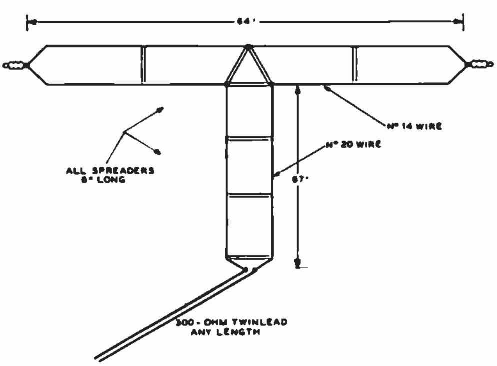


Figure 22

FOLDED-TOP DUAL-BAND ANTENNA

the 7-MHz band and standing waves on the feeder will be very small. However, it is possible to insert an electrical half wave of transmission line of any characteristic impedance into a feeder system such as this and the impedance at the far end of the line will be exactly the same value of impedance which the half-wave line "sees" at its termination. Hence this has been done in the antenna system shown in figure 22; an electrical half wave of line has been inserted between the feed point of the antenna and the 300-ohm transmission line to the transmitter.

The characteristic impedance of this additional half-wave section of transmission line has been made about 715 ohms (No. 20 wire spaced 6 inches), but since it is an electrical half wave long at 7 MHz and operates into a load of 300 ohms at the antenna the 300-ohm twin lead at the bottom of the half-wave section still "sees" an impedance of 300 ohms. The additional half-wave section of transmission line introduces a negligible amount of loss since the current flowing in the section of line is the same which would flow in a 300-ohm line at each end of the half-wave section, and at all other points it is *less* than the current which would flow in a 300-ohm line since the effective impedance is *greater* than 300 ohms in the center of the half-wave section. This means that the loss is less than it would be in an equivalent length of 300-ohm twin lead since this type of manufactured transmission line is made up of conductors which are equivalent to No. 20 wire.

So we see that the added section of 715-ohm line has substantially no effect on the operation of the antenna system on the 7-MHz band. However, when the flat top of the antenna is operated on the 3.5 MHz band the feed-point impedance of the flat top is approximately 3500 ohms. Since the section of 715-ohm transmission line is an electrical *quarter-wave* in length on the 3.5-MHz band, this section of line will have the effect of transforming the approximately 3500 ohms feed-point impedance of the antenna down to an impedance of about 150 ohms which will result in a 2:1 standing-wave ratio on the 300-ohm twin lead transmission line from the transmitter to the antenna system.

An antenna tuner or 4:1 balun may be placed at the station end of the 300-ohm line to transform it to a lower impedance value suitable for pi-network output circuits.

The Multee Antenna An antenna that works well on 160 and 80 meters, or 80 and 40 meters and is sufficiently compact to permit erection on the average city lot is the *Multee* antenna, illustrated in figure 23. The antenna evolves from a vertical two-wire radiator, fed on one leg only. On the low-frequency band the top portion does little radiating, so it is folded down to form a radiator for the higher-frequency band. On the lower-frequency band, the antenna acts as a top-loaded vertical radiator, while on the higher-frequency band, the flat top does the radiating rather than the vertical portion. The vertical portion acts as a quarter-wave linear transformer, matching the 6000-ohm antenna impedance to the 50-ohm impedance of the coaxial transmission line.

The earth below a vertical radiator must be of good conductivity not only to provide a low-resistance ground connection, but also to provide a good reflecting surface for the waves radiated downward toward the ground. For best results, a radial system should be installed beneath the antenna. For 160/80-meter operation, six radials 50 feet in length, made of No. 16 copper wire should be buried just below the surface of the ground. While an ordinary water-pipe ground system with no radials may be used, a system of radials will provide a worthwhile increase in signal strength. For 80/40-meter operation, the length of the radials may be reduced to 25 feet. As with all multiband antennas that employ no lumped tuned circuits, this antenna offers no attenuation to harmonics of the transmitter. When operating on the lower-frequency band, it would be wise to check the transmitter for second-harmonic emission, since this antenna will effectively radiate this harmonic.

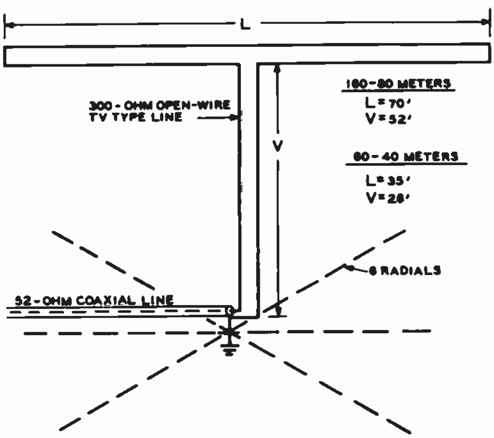
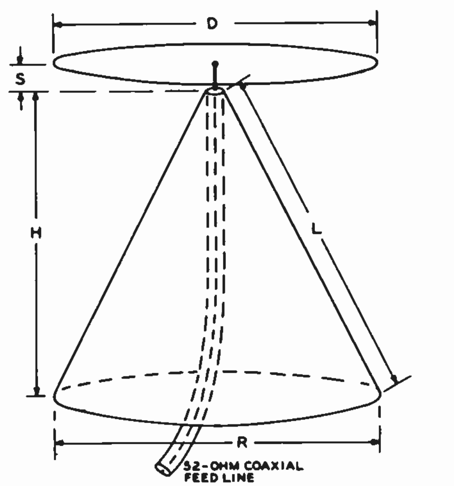


Figure 23

THE MULTEE TWO-BAND ANTENNA

This compact antenna can be used with excellent results on 160/80 and 80/40 meters. The feedline should be held as vertical as possible, since it radiates when the antenna is operated on its fundamental frequency.

The Low-Frequency Discone The *discone* antenna is widely used on the vhf bands, but until recently it has not been put to any great use on the lower-frequency bands. Since the discone is a broadband device, it may be used on several harmonically related amateur bands. Size is the limiting factor in the use of a discone, and the 20-meter band is about the lowest practical frequency for a discone of reasonable dimensions. A discone designed for 20-meter operation may be used on 20, 15, 11, 10, and 6 meters with excellent results. It affords a good match to a 50-ohm coaxial feed system on all of these bands. A practical discone antenna is shown in figure 24, with a SWR curve for its operation over the frequency range of 13 to 55 MHz shown in figure 25. The discone antenna radiates a vertically polarized wave and has a very



DIMENSIONS

20, 15, 11, 10, 6 METERS	15, 11, 10, 6 METERS	11, 10, 6, 2 METERS
D=12' L=18'	D=6' L=12'	D=6' L=9'6"
S=10" R=18'	S=6" R=12'	S=4" R=9'6"
H=15'7"	H=10'5"	H=6'3"

Figure 24

DIMENSIONS OF DISCONE ANTENNA FOR LOW-FREQUENCY CUTOFF AT 13.2 MHz, 20.1 MHz, AND 26 MHz

The Discone is a vertically polarized radiator, producing an omnidirectional pattern similar to a ground plane. Operation on several amateur bands with low SWR on the coaxial feed line is possible.

low angle of radiation. For vhf work the discone is constructed of sheet metal, but for low-frequency work it may be made of copper wire and aluminum angle stock. A suitable mechanical layout for a low-frequency discone is shown in figure 26. Smaller versions of this antenna may be constructed for 15, 11, 10, and 6 meters, or for 11, 10, 6, and 2 meters as shown in figure 24.

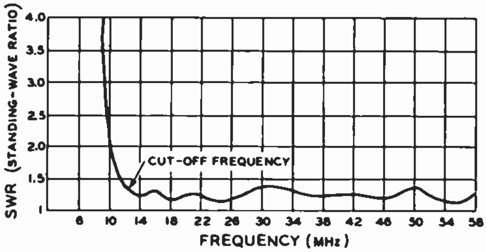


Figure 25

SWR CURVE FOR A 13.2-MHz DISCONE ANTENNA. SWR IS BELOW 1.5 TO 1 FROM 13.0 MHz TO 58 MHz

For minimum wind resistance, the top "hat" of the discone is constructed from three-quarter inch aluminum angle stock, the rods being bolted to an aluminum plate at the center of the structure. The tips of the rods are all connected together by lengths of No. 12 enamelled copper wire. The cone elements are made of No. 12 copper wire and act as guy wires for the discone structure. A very rigid arrangement may be made from this design, one that will give no trouble in high winds. A 4″ × 4″ post can be used to support the discone structure.

The discone antenna may be fed by a length of 50-ohm coaxial cable directly from the transmitter, with a very low SWR on all bands.

The Single-Wire-Fed Antenna The old favorite *single-wire-fed antenna* system is quite satisfactory for an impromptu all-band antenna system. It is widely used for portable installations and "Field Day" contests where a simple, multi-band antenna is required. A single-wire feeder has a characteristic impedance of approximately 300 ohms, depending on the wire size and the point of attachment to the antenna. The earth losses are comparatively low over ground of good conductivity.

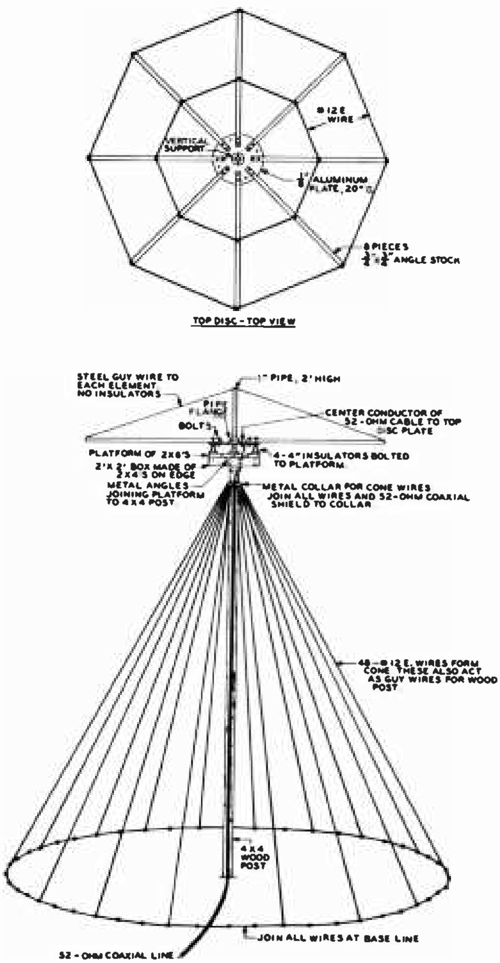


Figure 26

MECHANICAL CONSTRUCTION OF 20-METER DISCONE

Since the single-wire feeder radiates, it is necessary to bring it away from the antenna at right angles to the antenna wire for at least one-half the length of the antenna.

The correct point for best impedance match on the fundamental frequency is not suitable for harmonic operation of the antenna. In addition, the correct length of the antenna for fundamental operation is not correct for harmonic operation. Consequently, a compromise must be made in antenna length and point of feeder connection to enable the single-wire-fed antenna to operate on more than one band. Such a compromise introduces additional reactance into the single-wire feeder, and might cause loading difficulties with pi-network transmitters. To minimize this trouble, the single-wire feeder should be made a multiple of 33 feet long (figure 27).

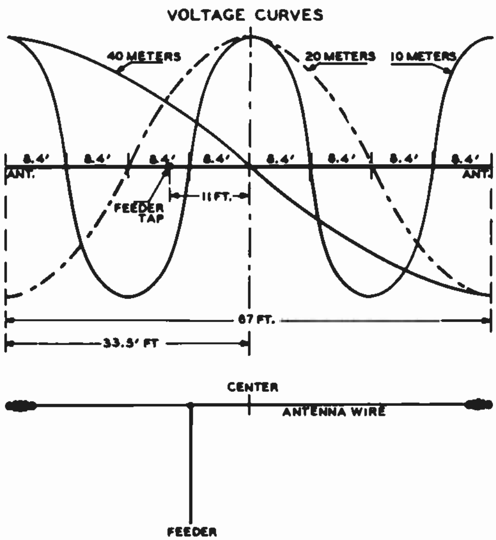


Figure 27

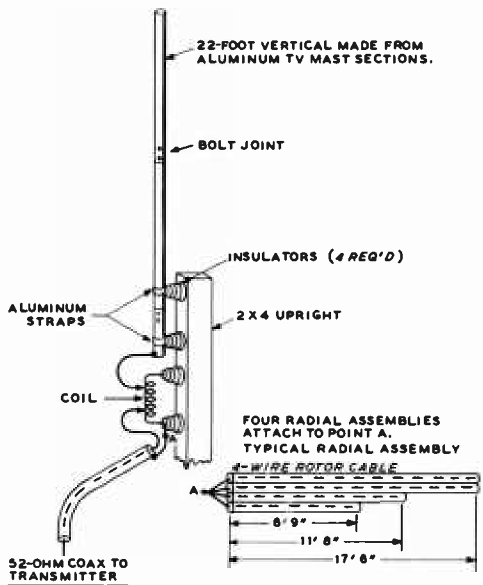
SINGLE-WIRE-FED ANTENNA FOR ALL-BAND OPERATION

An antenna of this type for 40-, 20- and 10-meter operation would have a radiator 67 feet long, with the feeder tapped 11 feet off center. The feeder can be 33, 66 or 99 feet long. The same type of antenna for 80-, 40-, 20- and 10-meter operation would have a radiator 134 feet long, with the feeder tapped 22 feet off center. The feeder can be either 66 or 132 feet long. This system should be used only with those coupling methods which provide good harmonic attenuation.

Multiband Vertical Antennas A vertical radiator can be used on several amateur bands either by employing a variable base-loading inductor or by the inclusion of trap elements in the radiator. In either case, tuned radial wires should be used for lowest ground loss at the higher frequencies. Shown in figure 28 is a 22-foot vertical antenna designed for operation on amateur bands from 80 through 10 meters. The height is chosen to present a ¾-wavelength vertical for low angle radiation at the highest frequency of operation. Radial wires are used for the 10-, 15-, and 20-meter bands and an external ground connection is used on 40 and 80 meters. If the antenna is mounted on the roof of a building, it may be possible to use the metal rain gutter system as a ground.

Four-wire TV rotator cable can be used to construct the radial system, each cable including a radial wire for one of the three higher bands. The fourth wire may be extended for 40 meters, or two of the four wires can be cut for 20 meters, and one each for 15 and 10 meters. At least three and preferably four such radial assemblies should be used. These can be laid out on the roof, or possibly hidden in the attic.

The radiator is made from two ten-foot sections of aluminum TV mast, plus one five-foot section cut to the the proper length. The mast sections are assembled and self-tapping sheet-metal screws are run



BAND	80	40	20	15
COAX TAP	7	5	3	2
ANTENNA TAP	25	12	6	3

TAPS MEASURED FROM GROUND END OF COIL. COIL NOT USED ON 10 METERS.

Figure 28

"ALL-BAND" VERTICAL ANTENNA

Base-loaded whip and multiple radial system may be used on all bands from 80 through 10 meters. Loading-coil taps are adjusted for lowest SWR on each band. The SWR on 10 meters may be improved by placing a 250-pf capacitor in series with the feedline connection to the base of the antenna and adjusting the capacitor for minimum SWR. Coil is 40 turns, 2" in diameter, 4" long (Air-Dux 1610).

through each joint to make a good electrical connection. The radiator and base coil are attached to sturdy ceramic "beehive" insulators, using strips of aluminum bent to form clamps to encircle the tubing. The insulators are mounted to a vertical section of "two-by-four" lumber bolted to the frame of the building. If securely mounted, no guy wires are required for the vertical radiator.

The antenna is resonated to the center of each operative band with the aid of a SWR meter placed in the 52-ohm feedline. The taps are adjusted as indicated in the chart and sufficient power is applied to the antenna to cause a reading on the SWR meter. The number of active turns in the coil and the feedline tap are varied a turn at a time until proper transmitter loading is

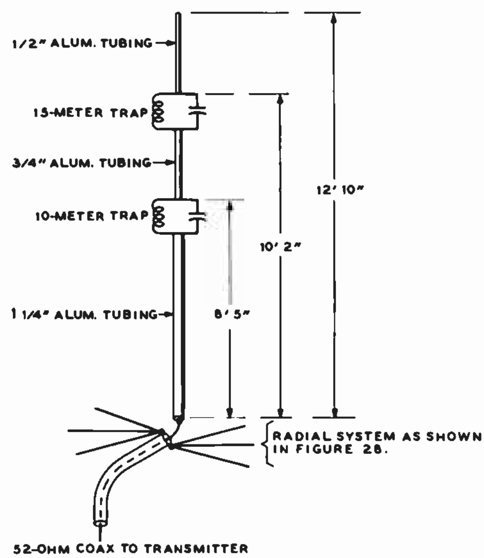


Figure 29

TRIBAND TRAP VERTICAL ANTENNA

Parallel-tuned trap assemblies are used in this vertical antenna designed for 20-, 15-, and 10-meter operation. The radial system of figure 28 is used. Automatic trap action electrically switches antenna for proper operation on each band.

achieved with a reasonably low value of SWR on the transmission line (below 1.5/1 or so at the center frequency in each band).

The trap technique described in the *Directive Antennas* chapter can be used for a three-band vertical antenna as shown in figure 29. This antenna is designed for operation on 10, 15, and 20 meters and uses a separate radial system for each band. No adjustments need be made to the antenna when changing frequency from one band to another. Substitution of a ground connection for the radials is not recommended because of the high ground loss normally encountered at these frequencies. Typical trap construction is discussed in the reference chapter, and the vertical radiator is built of sections of aluminum tubing, as described earlier.

Each trap is built and grid-dipped to the proper frequency before it is placed in the radiator assembly. The 10-meter trap is self-resonant at about 27.9 MHz and the 15-meter trap is self-resonant at about 20.8 MHz. Once resonated, the traps need no further adjustment and do not enter into later adjustments made to the antenna. The complete antenna is resonated to each amateur band by placing a single-turn coil between the base of the vertical radiator and the radial connection and coupling the grid-dip oscillator to the coil. The coaxial line is removed for this test. The lower section of the vertical antenna is adjusted in length for 10-meter resonance at about 28.7 MHz, followed by adjustment of the center section for resonance at 21.2 MHz. The last adjustment is to the top section for resonance at about 14.2 MHz.

It must be remembered that trap, or other multifrequency antennas are capable of radiating harmonics of the transmitter that may be coupled to them via the transmission line. It is well to check for harmonic radiation with a nearby radio amateur. If such harmonics are noted, an antenna tuner similar to the one described later in this chapter should be added to the installation to reduce unwanted harmonics to a minimum.

The Trap Dipole The trap principle discussed in Chapter 26 may be applied to dipoles as well as to vertical antennas. Shown in figure 30 are designs for trap dipoles for various amateur bands. For portable or Field Day use, the antennas may be fed directly with either 50- or 72-ohm coaxial line or 72-ohm TV-type ribbon line. For fixed station use, insertion of a balun between the antenna and the coaxial transmission line is recommended. A 20- and

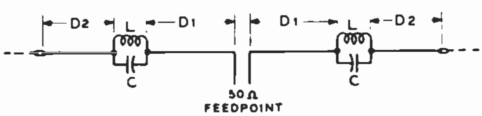
15-meter trap is shown in figure 31. It is designed to be left unprotected and is water-resistant. If desired, it may be covered with a plastic "overcoat" made from a section of a flexible squeeze bottle, such as bleach or laundry soap containers.

Operational bandwidth on the lower-frequency band is somewhat less than that of a comparable dipole, since a portion of the antenna is wound up in the trap element and does not radiate. Typical bandwidth for an 80- and 40-meter dipole, as measured between the 2/1 SWR points on the transmission line is: 80 meters, 180 kHz; 40 meters, 250 kHz.

Operational bandwidth of the 40- and 20-meter antenna is typically: 40 meters, 300 kHz; 20 meters, 350 kHz. In addition, the antenna may be operated over the lower 1 MHz of the 10-meter band with an SWR figure of less than 1.5/1.

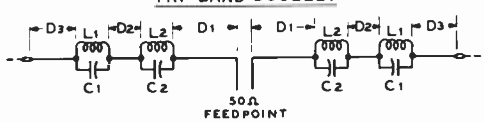
Data is also given in figure 30 for a tri-band doublet covering the 20-, 15-, and 10-meter amateur bands. Operational bandwidth is sufficient to cover all the included bands with a maximum SWR figure at the band edges of less than 2/1 on the transmission line. As with any antenna configuration, bandwidth and minimum SWR indication are a function of the height of the antenna above the ground.

DUO-BAND DOUBLET



BANDS	D1	D2	L(µH)	C(PF)	FR
80-40	32' 0"	22' 0"	8.2	60	6.95
40-20	16' 8"	10' 6"	4.7	25	13.8
20-15	10' 5"	3' 7 1/2"	2.9	20	20.7
15-10	8' 0"	1' 11"	1.65	20	27.8

TRI-BAND DOUBLET



BAND	D1	D2	D3	L1(µH)	C1(PF)	L2(µH)	C2(PF)
20-15-10	8' 0"	1' 10"	2' 9"	2.9	20	1.65	20

Figure 30

MULTIBAND TRAP DOUBLETS

Trap doublet dimensions for duoband and triband antennas. Traps are assembled in the manner shown in figure 31, and antenna dimensions are based on an over-all trap length of two inches. Highest-band resonant frequency many be varied by changing dimension D_1. Lower band is also affected and dimension D_2 must be adjusted to compensate for change in D_1. Sequence of adjustment, then, is D_1, D_2, and D_3. Dimensions shown are for center-of-band resonance. Parallel-tuned traps are adjusted to trap frequency (f_R) outside low-frequency end of band. Trap doublet is preferably fed with balun and 50-ohm transmission line.

Figure 31

TRAP CONSTRUCTION

Fifteen-meter trap is shown here. Trap is designed for power level of 500 watts, PEP. Trap is built around strain insulator which removes pull of antenna from coil and capacitor. Capacitor is Centralab 853A-20Z (20 pf) and coil is 14½ turns #16, 1" diameter and 2" long (8 turns per inch), Air-Dux 808T. Trap is about 2" long with 1½" leads. Before placement in the antenna, it is grid-dipped to 20.7 MHz on the bench and adjusted to frequency by removal or addition of a fraction of a turn. Traps for other bands are constructed in similar manner. For 2 kW PEP level, coil should be #12 wire, about 2" diameter, and capacitor should be Centralab type 850S.

25-8 Matching Nonresonant Lines to the Antenna

While ribbon or tubular molded 300-ohm line is often used in vhf antenna installations, coaxial line is universally used in h-f antenna systems and also into the lower portion of the vhf spectrum. Open-wire lines are occasionally used for portable installations or for fixed antenna arrays, but even these are usually matched to a coaxial transmission line for ultimate connection to the transmitting equipment.

These transmission systems generally require some sort of matching device to make an efficient transition from the characteristic impedance of the line to the radiation resistance value of the antenna, otherwise severe standing waves can occur on the transmission line system.

Standing Waves As was discussed earlier, standing waves on the antenna transmission line, in the transmitting case, are a result of reflection from the point where the feed line joins the antenna system. The magnitude of the standing waves is determined by the degree of mismatch between the characteristic impedance of the transmission line and the input impedance of the antenna system. When the feed-point impedance of the antenna is resistive and of the same value as the characteristic impedance of the feed line, standing waves will not exist on the feeder. It may be well to repeat at this time that there is no adjustment which can be made at the transmitter end of the feed line which will change the magnitude of the standing waves on the antenna transmission line.

Delta-Matched Dipole Antenna The *delta-type matched-impedance dipole* antenna is shown in figure 32. The impedance of the transmission line is transformed gradually into a higher value by the fanned-out Y portion of the feeders, and the Y portion is tapped on the antenna at points where the Y portion is a compromise between the impedance at the antenna and the impedance of the unfanned portion of the line.

The constants of the system are rather critical, and the antenna must resonate at the operating frequency in order to minimize standing waves on the line. Some slight readjustment of the taps on the antenna is desirable, if appreciable standing waves persist in appearing on the line.

The constants for a dipole are determined by the following formulas:

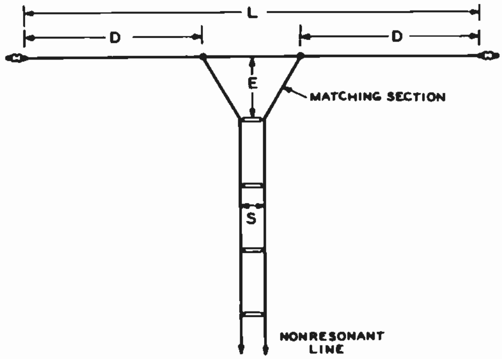


Figure 32

THE DELTA-MATCHED DIPOLE ANTENNA

The dimensions for the portions of the antenna are given in the text.

$$L_{feet} = \frac{467.4}{F_{MHz}}$$

$$D_{feet} = \frac{175}{F_{MHz}}$$

$$E_{feet} = \frac{147.6}{F_{MHz}}$$

where,

L is antenna length,

D is the distance *in from each end* at which the Y taps on,

E is the height of the Y section.

Since these constants are correct only for a 600-ohm transmission line, the spacing S of the line must be approximately 75 times the diameter of the wire used in the transmission line. For No. 14 wire, the spacing will be slightly less than 5 inches.

Multiwire Dipoles When a dipole antenna or the driven element in an array consists of more than one wire or tubing conductor the radiation resistance of the antenna or array is increased slightly as a result of the increase in the effective diameter of the element. Further, if one wire of such a radiator is split, as shown in figure 33, the effective *feed-point* resistance of the antenna or array will be increased by a factor of N^2 where N is equal to the number of conductors, all in parallel, of the same diameter in the array. Thus if there are two conductors of the same diameter in the driven element or the antenna the feed-point resistance will be multiplied by 2^2, or 4. If the antenna has a radiation resistance of 75 ohms its feed-point resistance will be 300 ohms. This is the case of the conventional *folded dipole* as shown in figure 33B.

If three wires are used in the driven radiator the feed-point resistance is increased by a factor of 9; if four wires are used the impedance is increased by a factor of 16, etc. In certain cases when feeding a parasitic array it is desirable to have an impedance step up different from the value of 4:1 obtained with two elements of the same

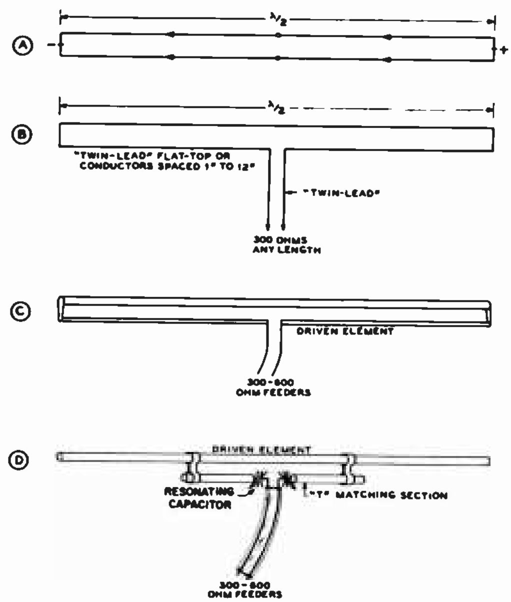


Figure 33

FOLDED-ELEMENT MATCHING SYSTEMS

Drawing A above shows a half-wave made up of two parallel wires. If one of the wires is broken as in B and the feeder connected, the feed-point impedance is multiplied by four; such an antenna is commonly called a "folded doublet." The feed-point impedance for a simple half-wave doublet fed in this manner is approximately 300 ohms, depending on antenna height. Drawing C shows how the feed-point impedance can be multiplied by a factor greater than four by making the half of the element that is broken smaller in diameter than the unbroken half. An extension of the principles of B and C is the arrangement shown at D where the section into which the feeders are connected is considerably shorter than the driven element.

diameter and 9:1 with three elements of the same diameter. Intermediate values of impedance step up may be obtained by using two elements of different diameter for the complete driven element as shown in figure 33C. If the conductor that is broken for the feeder is of *smaller* diameter than the

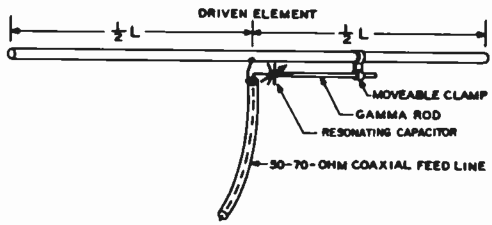


Figure 34

THE GAMMA MATCH FOR CONNECTING AN UNBALANCED COAXIAL LINE TO A BALANCED DRIVEN ELEMENT

other conductor of the radiator, the impedance step up will be *greater* than 4:1. On the other hand if the *larger* of the two elements is broken for the feeder the impedance step up will be *less* than 4:1.

The "T" Match A method of matching a balanced low-impedance transmission line to the driven element of a parasitic array is the *T match* illustrated in figure 33D. This method is an adaptation of the multiwire doublet principle which is more practical for lower-frequency parasitic arrays such as those for use on the 14- and 28-MHz bands. In the system a section of tubing of approximately one-quarter the diameter of the driven element is spaced about four inches below the driven element by means of clamps which hold the T-section mechanically and which make electrical connection to the driven element. The length of the T-section is normally between 15 and 30 inches each side of the center of the dipole for transmission lines of 300 to 600 ohms impedance, assuming 28-MHz operation. In series with each leg of the T-section and the transmission line is a series resonating capacitor. These two capacitors tune out the reactance of the T-section. If they are not used, the T-section will detune the dipole when the T-section is attached to it. The two capacitors may be ganged together, and once adjusted for minimum detuning action, they may be locked. A suitable housing should be devised to protect these capacitors from the weather. Additional information on the adjustment of the T-match is given in the chapter covering rotary beam antennas.

The Gamma Match An unbalanced version of the T-match may be used to feed a dipole from an unbalanced coaxial line. Such a device is called a *Gamma match*, and is illustrated in figure 34.

The length of the Gamma rod and the spacing of it from the dipole determine the impedance level at the transmission line end of the rod. The series capacitor is used to tune out the reactance introduced into the system by the Gamma rod. The adjustment of the Gamma match is discussed in the chapter covering rotary beam antennas.

Matching Stubs By connecting a resonant section of transmission line (called a *matching stub*) to either a voltage or current loop and attaching parallel-wire nonresonant feeders to the resonant stub at a suitable voltage (impedance) point, standing waves on the line may be virtually eliminated. The stub is made to serve as an autotransformer. Stubs are particularly adapted to matching an open line to certain directional arrays, as will be described later.

Voltage Feed When the stub attaches to the antenna at a voltage loop, the stub should be a quarter wavelength long electrically, and be shorted at the bottom end. The stub can be resonated by sliding the shorting bar up and down before the nonresonant feeders are attached to the stub, the antenna being shock-excited from a separate radiator during the process. Slight errors in the length of the radiator can be compensated for by adjustment of the stub if both sides of the stub are connected to the radiator in a symmetrical manner. Where only one side of the stub connects to the radiating system, as in the zepp and in certain antenna arrays, the radiator length must be exactly right in order to prevent excessive unbalance in the untuned line.

Current Feed When a stub is used to current-feed a radiator, the stub should either be left *open* at the bottom end instead of shorted, or else made a *half wave* long. The open stub should be resonated in the same manner as the shorted stub before attaching the transmission line; however, in this case, it is necessary to prune the stub to resonance, as there is no shorting bar.

Sometimes it is handy to have a stub hang from the radiator to a point that can be reached from the ground, in order to facilitate adjustment of the position of the transmission-line attachment. For this reason, a quarter-wave stub is sometimes made three-quarters wavelength long at the higher frequencies, in order to bring the bottom nearer the ground. Operation with any *odd* number of quarter waves is the same as for a quarter-wave stub.

Stub Length Electrical)	Current-Fed Radiator	Voltage-Fed Radiator
¼-¾-1¼-etc. wavelengths	Open Stub	Shorted Stub
½-1-1½-2-etc. wavelengths	Shorted Stub	Open Stub

Any number of half waves can be added to either a quarter-wave stub or a half-wave stub without disturbing the operation, though losses and frequency sensitivity will be lowest if the shortest usable stub is employed (see chart).

Linear R-F Transformers A resonant quarter-wave line has the unusual property of acting much as a transformer. Let us take, for example, a section consisting of No. 12 wire spaced 6 inches, which happens to have a surge impedance of 600 ohms. Let the far end be terminated with a pure resistance, and let the near end be fed with radio-frequency energy at the frequency for which the line is a *quarter wavelength* long. If an impedance measuring set is used to measure the impedance at the near end while the impedance at the far end is varied, an interesting relationship between the 600-ohm characteristic surge impedance of this particular quarter-wave matching line, and the impedance at the ends will be discovered.

When the impedance at the far end of the line is the same as the characteristic surge impedance of the line itself (600 ohms), the impedance measured at the near end of the quarter-wave line will also be found to be 600 ohms.

Under these conditions, the line would not have any standing waves on it, since

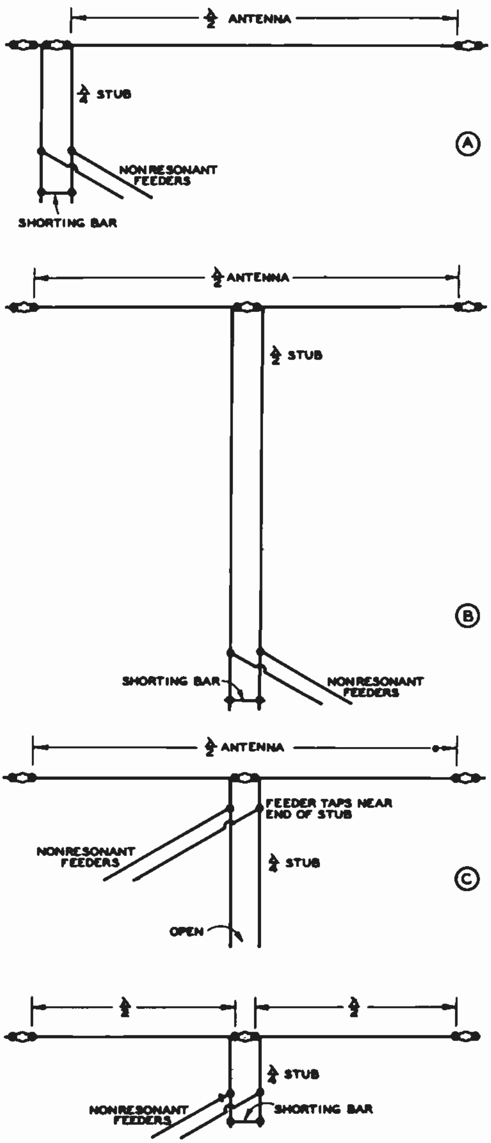


Figure 35

MATCHING STUB APPLICATIONS

An end-fed half-wave antenna with a quarter-wave shorted stub is shown at A. B shows the use of a half-wave shorted stub to feed a relatively low impedance point such as the center of the driven element of a parasitic array, or the center of a half-wave dipole. The use of an open-ended quarter-wave stub to feed a low impedance is illustrated at C. D shows the conventional use of a shorted quarter-wave stub to voltage-feed two half-wave antennas with a 180° phase difference.

it is terminated in its characteristic impedance. Now, let the resistance at the far end of the line be doubled, or changed to 1200 ohms. The impedance measured at the near end of the line will be found to have been cut in half (to 300 ohms). If the resistance at the far end is made half the original value of 600 ohms or 300 ohms, the impedance at the near end doubles the original value of 600 ohms, and becomes 1200 ohms. As one resistance goes up, the other goes down proportionately.

It will always be found that the characteristic surge impedance of the quarter-wave matching line is the geometric mean between the impedance at both ends. This relationship is shown by the following formula:

$$Z_{MS} = \sqrt{Z_A Z_L}$$

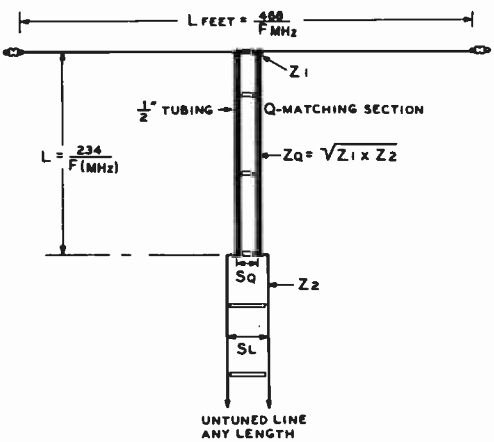
where,

Z_{MS} equals impedance of matching section,
Z_A equals antenna resistance,
Z_L equals line impedance.

Quarter-Wave Matching Transformers The impedance inverting characteristic of a quarter-wave section of transmission line is widely used by making such a section of line act as a *quarter-wave transformer*. The quarter-wave transformer may be used in a wide number of applications wherever a transformer is required to match two impedances whose geometric mean is somewhere between perhaps 25 and 750 ohms when transmission-line sections can be used. Paralleled coaxial lines may be used to obtain the lowest impedance mentioned, and open-wire lines composed of small conductors spaced a moderate distance may be used to obtain the higher impedance. A short list of impedances, which may be matched by quarter-wave sections of transmission line having specified impedances, follows.

Load or Ant. Impedance ↓	300	480	600	← Feed-Line Impedance
20	77	98	110	Quarter-
30	95	120	134	Wave
50	110	139	155	Transformer
75	150	190	212	Impedance
100	173	220	245	

Q-Section Feed System The standard form of *Q-section* feed to a doublet is shown in figure 36. An impedance match is obtained by utilizing a matching section, the surge impedance of which is the geometric mean between the transmission-line surge impedance and the radiation resistance of the radiator. A sufficiently good match usually can be obtained by either designing or adjusting the matching section for a dipole to have a surge impedance that is the geometric mean between the line impedance and 72 ohms, the latter being the theoretical radiation resistance of a half-wave doublet either infinitely high or a half wave above a perfect ground.



Center to Center Spacing in Inches	Impedance in Ohms for ½" Diameters	Impedance in Ohms for ¼" Diameters
1.0	170	250
1.25	188	277
1.5	207	298
1.75	225	318
2.0	248	335

Figure 36
HALF-WAVE RADIATOR FED BY "Q BARS"

The Q matching section is simply a quarter-wave transformer whose impedance is equal to the geometric mean between the impedance at the center of the antenna and the impedance of the transmission line to be used to feed the bottom of the transformer.

The Inducto-Match The driven element of a beam antenna or a simple dipole antenna may form a portion of a network whose input impedance is close to 50 ohms over a small frequency range (figure 37). It is necessary that the radiation resistance of the element be less than the impedance of the tranmission line, and this condition is met under most circumstances.

The radiation resistance of the antenna element is made to appear as a capacitive reactance at the driving point by shortening the element past the normal resonant length. The inductive portion of the network takes the form of a hairpin or coil placed across the terminals of the driven element. The L/C ratio of the combination determines the transformation ratio of the network when the LC product is resonant at the center frequency of antenna operation. Inductance of the hairpin or coil is best determined by experiment. Measurements made at 14 MHz, point to a shortening effect of about six inches in the over-all length of the driven element, and an inductance of about 0.5 μH in the hairpin. Complete information on this compact and efficient matching system is given in the *Beam Antenna Handbook*, published by Radio Publications, Inc., Wilton, Conn.

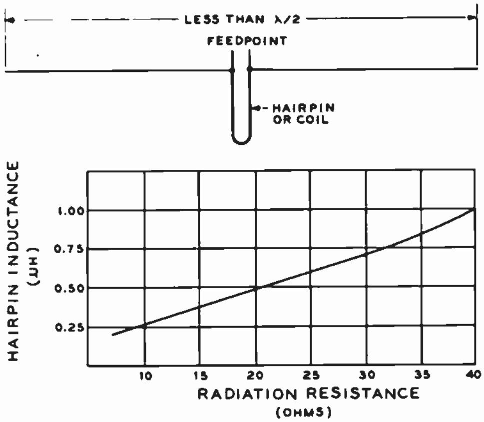


Figure 37
THE INDUCTO-MATCH

Dipole element acts as matching transformer by placing inductor at the center and shortening element to provide capacitive reactance across feedpoint. Typical three-element Yagi antenna has feedpoint impedance of about 20 ohms and calls for 0.5 μH inductor. Impedance match is made by varying inductor and length of dipole.

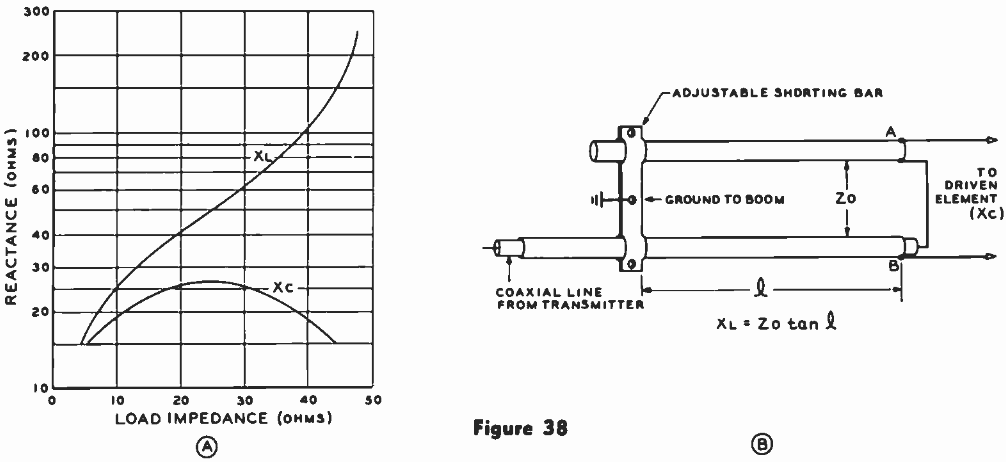


Figure 38

ADJUSTABLE BALUN TRANSFORMER

A practical balun transformer to match a 50-ohm coaxial line to the low-impedance balanced load presented by a beam antenna is shown here. Coaxial line passes through one leg of balun. Outer conductor of the line is trimmed short to the point where the line enters balun tube, and is soldered to tube at this point. Inner conductor of the line passes along the balun tube and emerges at the antenna end, where it is cross-connected to the opposite tube as shown in the illustration. If the load impedance is known, the balun transformer may be set to length by the use of chart (A) and formula (B).

The Balun Transformer The *Inducto-Match* described in the previous section may be modified into a balun by passing a coaxial line down one leg, as shown in figure 38. Points A and B are balanced to ground and the inner conductor of the coaxial line is cross-connected to the opposite balun leg to provide the proper phase reversal. The impedance transformation is adjusted by varying the length of the balun and the length of the driven element.

A practical balun transformer can be made of 3/8-inch diameter hard-drawn copper tubing. Two tubes, about 4 feet long, spaced about 3 inches will serve for 20, 15 and 10 meter work.

Balun length as a function of terminal impedance may be determined by connecting 1-watt composition resistors of various values between 10 and 50 ohms across the end of the device and adjusting the length for minimum SWR on a 50-ohm line, feeding the balun with a few hundred milliwatts of r-f power. The calibrated balun may then be used in an antenna system to determine the feedpoint impedance. Balun length and driven-element length are adjusted to provide a 1:1 SWR at the measuring frequency.

A permanent balun of this style may be mounted on a beam antenna to provide a good match between the driven element and a coaxial feedline. The balun is run parallel to the boom for mounting convenience at a distance of about six inches. Positioning the balun closer to the boom may necessitate a change in setting. The driven element, for a starter, should be shortened about 3 inches on each tip (for 20 meters). Balun and driven element are then adjusted for a good impedance match at the center of the band.

25-9 Antenna Supports

The foregoing portion of this chapter has been concerned primarily with the *electrical* characteristics and considerations of antennas. Some of the physical aspects and mechanical problems incident to the actual erection of antennas and arrays will be discussed in the following section.

Up to 30 feet, there is little point in using mast-type antenna supports unless guy wires either must be eliminated or kept to a minimum. While a little more difficult to erect, because of their floppy nature, fabricated wood poles of the type to be described will be just as satisfactory as more rigid types, *provided* many guy wires are used.

Rather expensive when purchased through

the regular channels, 40- and 50-foot telephone poles sometimes can be obtained quite reasonably. In the latter case, they are hard to beat, inasmuch as they require no guying if set in the ground six feet (standard depth), and the resultant pull in any lateral direction is not in excess of a hundred pounds or so.

For heights of 80 to 100 feet, either three- or four-sided lattice-type masts are most practical. They can be made self-supporting, but a few guys will enable one to use a smaller cross section without danger from high winds. The torque exerted on the base of a high self-supporting mast is terrific during a strong wind.

The "A-Frame" Mast Figures 39A and 39B show the standard method of construction of the *A-frame* type of mast. This type of mast is quite frequently used since there is only a moderate amount of work involved in the construction of the assembly and since the material cost is relatively small. The three pieces of selected 2 by 2 are first set up on three sawhorses or boxes and the holes drilled for the three ¼-inch bolts through the center of the assembly. Then the base legs are spread out to about 6 feet and the bottom braces installed. Finally the upper braces and the cross pieces are installed and the assembly given several coats of good-quality paint as a protection against weathering.

Figure 39C shows another common type of mast which is made up of sections of 2 × 4 placed end-to-end with stiffening sections of 1 by 6 bolted to the edge of the 2 by 4 section. Both types of mast will require a set of top guys and another set of guys about one-third of the way down from the top. Two guys spaced about 90 to 100 degrees and pulling against the load of the antenna will normally be adequate for the top guys. Three guys are usually used at the lower level, with one directly behind the load of the antenna and two more spaced 120 degrees from the rear guy.

Raising the mast is made much easier if a gin pole about 20 feet high is installed about 30 or 40 feet to the rear of the direction in which the antenna is to be raised. A line from a pulley on the top of the gin pole is then run to the top of the pole to be raised. The gin pole comes into play when the center of the mast has been raised 10 to 20 feet above the ground and an additional elevated pull is required to keep the top of the mast coming up as the center is raised further above ground.

Using TV Masts Steel tubing masts of the telescoping variety are widely available at a moderate price for use in supporting television antenna arrays. These masts usually consist of several 10-foot lengths of electrical metal tubing (EMT) of sizes such that the sections will telescope. The 30- and 40-foot lengths are well suited as masts for supporting antennas and arrays of the type used on the amateur bands. The masts are constructed in such a manner that the bottom 10-foot length may be guyed permanently before the other sections are raised. Then the upper sections may be extended, beginning with the top-mast section, until the mast is at full length (provided a strong wind is not blowing) following which all the guys may be anchored. It is important that there be no load on the top of the mast when the "vertical" raising method is to be employed.

Guy Wires Guy wires should never be pulled taut; a *small* amount of slack is desirable. Galvanized wire, somewhat heavier than seems sufficient for the job, should be used. The heavier wire is a little harder to handle, but costs only a little more and takes longer to rust through. Care should be taken to make sure that no kinks exist when the pole or tower is ready for erection, as the wire will be greatly weakened at such points if a kink is pulled tight, even if it is later straightened.

Stranded aluminum wire, which is corrosion resistant, may be used in place of galvanized wire guys for light weight towers.

If "dead men" are used for the guy wire terminations, the wire or rod reaching from the dead men to the surface should be of nonrusting material, such as brass, or given a heavy coating of asphalt or other protective substance to prevent destructive action by the damp soil. Galvanized iron wire will last only a short time when buried in moist soil.

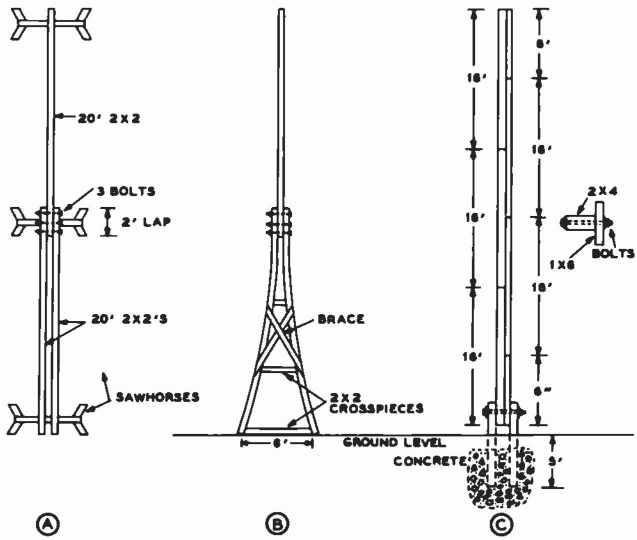


Figure 39

TWO SIMPLE WOOD MASTS

Shown at A is the method of assembly, and at B is the completed structure, of the conventional "A-frame" antenna mast. At C is shown a structure which is heavier but more stable than the A-frame for heights above about 40 feet.

Only strain-type (compression) insulators should be used for guy wires. Regular ones might be sufficiently strong for the job, but it is not worth taking chances, and egg-type strain halyard insulators are no more expensive.

Only a brass or bronze pulley should be used for the halyard, as a high pole with a rusted pulley is truly a sad affair. The bearing of the pulley should be given a few drops of heavy machine oil before the pole or tower is raised. The halyard itself should be of good material, preferably waterproofed. Hemp rope of good quality is better than window sash cord from several standpoints, and is less expensive. Soaking it thoroughly in engine oil of medium viscosity, and then wiping it off with a rag, will not only extend its life but minimize shrinkage in wet weather. Because of the difficulty of replacing a broken halyard it is a good idea to replace it periodically, without waiting for it to show excessive deterioration.

It is an excellent idea to tie both ends of the halyard line together in the manner of a flag-pole line. Then the antenna is tied onto the place where the two ends of the halyard are joined. This procedure of making the halyard into a loop prevents losing the top end of the halyard should the antenna break near the end, and it also prevents losing the halyard completely should the end of the halyard carelessly be allowed to go free and be pulled through the pulley at the top of the mast by the antenna load. A somewhat longer piece of line is required but the insurance is well worth the cost of the additional length of rope.

Trees as Supports Often a tall tree can be used to support one end of an antenna, but one should not attempt to attach anything to the top, as the swaying of the top of the tree during a heavy wind will complicate matters.

If a tree is utilized for support, provision should be made for keeping the antenna taut without submitting it to the possibility of being severed during a heavy wind. This can be done by the simple expedient of using a pulley and halyard, with weights attached to the lower end of the halyard to keep the antenna taut. Only enough weight to avoid excessive sag in the antenna should be tied to the halyard, as the continual swaying of the tree submits the pulley and halyard to considerable wear.

Painting The life of a wood mast or pole can be increased several hundred percent by protecting it from the elements with a coat or two of paint. And, of course, the appearance is greatly enhanced. The wood should first be given a primer coat of flat white outside house paint, which can be thinned down a bit to advantage with second-grade linseed oil. For the second coat, which should not be applied until the first is thoroughly dry, *aluminum paint* is not only the best from a preservative standpoint, but looks very well. This type of paint, when purchased in quantities, is considerably cheaper than might be gathered from the price asked for quarter-pint cans.

Portions of posts or poles below the surface of the soil can be protected from termites and moisture by painting with *cresote*. While not so strong initially, redwood will deteriorate much more slowly when buried than will the white woods, such as pine.

After the base of the pole or post has been treated, it should be given a wrapping of heavy aluminum foil paper to insulate it against ground water.

Antenna Wire The antenna or array itself presents no special problem. A few considerations should be borne in mind, however, For instance, soft-drawn copper should not be used, as even a short span will stretch several percent after whipping around in the wind a few weeks, thus affecting the resonant frequency. Enameled copper wire, as ordinarily available at radio stores, is usually soft-drawn, but by tying one end to some object such as a telephone pole and the other to the frame of an auto, a few husky tugs can be given and the wire, after stretching a bit, is equivalent to hard-drawn.

Where a long span of wire is required, or where heavy insulators in the center of the span result in considerable tension, copper-clad steel wire is somewhat better than hard-drawn copper. It is a bit more expensive, though the cost is far from prohibitive. The use of such wire, in conjunction with strain insulators is advisable where the antenna would endanger persons or property should it break.

For transmission lines and tuning stubs steel-core or hard-drawn wire will prove awkward to handle, and soft-drawn copper should, therefore, be used. If the line is long, the strain can be eased by supporting it at several points.

More important from an electrical standpoint than the actual size of wire used is the soldering of joints, especially at current loops in an antenna of low radiation resistance. In fact, it is good practice to solder *all* joints, thus ensuring quiet operation when the antenna is used for receiving.

Insulation A question that often arises is that of insulation. It depends, of course, on the r-f voltage at the point at which the insulator is placed. The r-f voltage, in turn, depends on the distance from a current node, and the radiation resistance of the antenna. Radiators having low radiation resistance have very high voltage at the voltage loops; consequently, better than usual insulation is advisable at those points.

Open-wire lines operated as nonresonant lines have little voltage across them; hence the most inexpensive ceramic types are sufficiently good electrically. With tuned lines, the voltage depends on the amplitude of the standing waves. If they are very great, the voltage will reach high values at the voltage loops, and the best spacers available are none too good. At the current loops the voltage is quite low, and almost anything will suffice.

When insulators are subject to very high r-f voltages, they should be cleaned occasionally if in the vicinity of sea water or smoke. Salt scum and soot are not readily dislodged by rain, and when the coating becomes heavy enough, the efficiency of the insulators is greatly impaired.

If a very pretentious installation is to be made, it is wise to check up on both Underwriter's rules and local ordinances which might be applicable. If you live anywhere near an airport, and are contemplating a tall pole, it is best to investigate possible regulations and ordinances pertaining to towers in the district, before starting construction.

25-10 Coupling to the Antenna System

When coupling an antenna feed system to a transmitter the most important considerations are as follows: (1) means should be provided for varying the load on the amplifier; (2) the load presented to the final amplifier should be resistive (nonreactive) in character; and (3) means should be provided to reduce harmonic coupling between the final amplifier plate tank circuit and the antenna or antenna transmission line to an *extremely low* value.

Transmitter Loading and TVI The problem of coupling the power output of a high-frequency or vhf transmitter to the radiating portion of the antenna system has been complicated by the virtual necessity for eliminating interference to TV reception. However, the TVI-elimination portion of the problem may *always* be accomplished by adequate shielding of the transmitter, by filtering of the control and power leads which enter the transmitter enclosure, and by the inclusion of a harmonic-attenuating filter between the output of the transmitter and the antenna system.

Although TVI may be eliminated through inclusion of a filter between the output of a shielded transmitter and the antenna system, the fact that such a filter should be included in the link between transmitter and antenna makes it necessary that the transmitter-loading problem be re-evaluated in terms of the necessity for inclusion of such a filter.

Harmonic-attenuating filters must be operated at an impedance level which is close to their design value; therefore they must operate *into* a resistive termination substantially equal to the characteristic impedance of the filter. If such filters are operated into an impedance which is not resistive and approximately equal to their characteristic impedance: (1) the capacitors used in the filter sections will be subjected to high peak voltages and may be damaged, (2) the harmonic-attenuating properties of the filter will be decreased, and (3) the impedance at the input end of the filter will be different from that seen by the filter at the load end (except in the case of the half-wave type of filter). It is therefore important that the filter be included in the transmitter-to-antenna circuit at a point where the impedance is close to the nominal value of the filter, and at a point where this impedance is likely to remain fairly constant with variations in frequency.

Block Diagrams of Transmitter-to-Antenna Coupling Systems There are two basic arrangements which include all the provisions required in the transmitter-to-antenna coupling system, and which permit the harmonic-attenuating filter to be placed at a position in the coupling system where it can be operated at an impedance level close to its nominal value. These arrangements are illustrated in block diagram form in figures 40 and 41.

The arrangement of figure 40 is recommended for use with a single-band antenna system, such as a dipole or a rotatable array, wherein an impedance matching system is included within or adjacent to the antenna. The feed line coming down from the antenna system should have a characteristic impedance equal to the nominal impedance of the harmonic filter, and the impedance matching at the antenna should be such that the standing-wave ratio on the antenna feed line is less than 2 to 1 over the range of frequency to be fed to the antenna.

The arrangement of figure 40 is more or less standard for commercially manufactured equipment for amateur and commercial use in the high-frequency and vhf range.

The arrangement of figure 41 merely adds an antenna coupler between the output of the harmonic attenuating filter and the antenna transmission line. The antenna coupler will have some harmonic-attenuating action, but its main function is to transform the impedance at the station end of the antenna transmission line to the nominal value of the harmonic filter. Hence the arrangement of figure 41 is more general than the figure 40 system, since the inclusion of the antenna coupler allows the system to feed an antenna transmission line of any reasonable impedance value, and also without regard to the standing-wave ratio which might exist on the antenna transmission line. Antenna couplers are discussed in a following section.

Output Coupling Adjustment It will be noticed by reference to both figure 40 and figure 41 that a box labeled *Coupling Adjustment* is included in the block diagram. Such an element is necessary in the complete system to afford an adjustment in the value of load impedance presented to the tubes in the final amplifier stage of the transmitter. The impedance at the input terminal of the harmonic filter is established by the antenna, through its matching system and the antenna coupler,

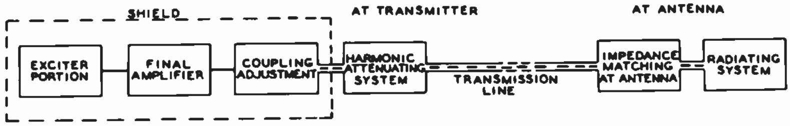


Figure 40

ANTENNA COUPLING SYSTEM

The harmonic suppressing antenna coupling system illustrated above is for use when the antenna transmission line has a low standing-wave ratio, and when the characteristic impedance of the antenna transmission line is the same as the nominal impedance of the low-pass harmonic-attenuating filter.

if used. In any event the impedance at the input terminal of the harmonic filter should be very close to the nominal impedance of the filter. Then the *Coupling Adjustment* provides means for transforming this impedance value to the correct operating value of load impedance which should be presented to the final amplifier stage.

There are two common ways for accomplishing the antenna coupling adjustment, as illustrated in figures 42 and 43. Figure 42 shows the variable-link arrangement often used in home-constructed equipment, while the pi-network coupling arrangement is illustrated in figure 43. Either method may be used, and each has its advantages.

Variable-Link Coupling The variable-link method illustrated in figure 42 provides good rejection to subharmonics. For greatest bandwidth of operation of the coupling circuit, the reactance of link coil L and the reactance of link tuning capacitor C should both be between 3 and 4 times the nominal load impedance of the harmonic filter. This is to say that the inductive reactance of coupling link L should be tuned out or resonated by capacitor C, and the operating Q of the LC link circuit should be between 3 and 4. If L and C are made resonant at the center of a band, with a link circuit Q of 3 to 4, and coupling adjustment is made by physical adjustment of L with respect to the final amplifier tank coil, it usually will be possible to operate over an entire amateur band without change in the coupling system. Capacitor C normally may have a low voltage rating, even with a high-power transmitter, due to the low Q and low impedance of the coupling circuit.

Pi-Network Coupling The pi-network coupling system offers two advantages: (1) a *mechanical* coupling variation is not required to vary the loading of the final amplifier, and (2) the pi-network (if used with an operating Q of about 10) offers within itself a harmonic attenuation of 30 db or more, in addition to the harmonic attenuation provided by the additional harmonic attenuating filter. Some commercial equipment incorporates an L-network in addition to the pi-network, for accomplishing the impedance transformation in two steps to provide additional harmonic attenuation.

25-11 Antenna Couplers

As stated in the previous section, an antenna coupler is not required when the impedance of the antenna transmission line is

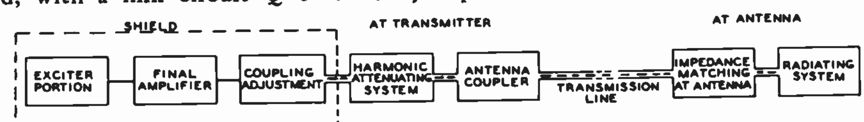


Figure 41

ANTENNA COUPLING SYSTEM

The antenna coupling system illustrated above is for use when the antenna transmission line does not have the same characteristic impedance as the TVI filter, and when the standing-wave ratio on the antenna transmission line may or may not be low.

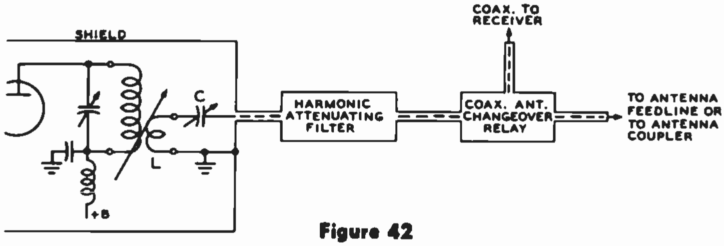


Figure 42

TUNED-LINK OUTPUT CIRCUIT

Capacitor C should be adjusted so as to tune out the inductive reactance of the coupling link, L. Amplifier loading is controlled by varying the coupling between the plate tank of the final amplifier and the antenna link.

the same as the nominal impedance of the harmonic filter, and the antenna feed line is being operated with a low standing-wave ratio. However, there are many cases where it is desirable to feed a multiband antenna from the output of the harmonic filter, where a tuned line is being used to feed the antenna, or where a long wire without a separate feed line is to be fed from the output of the harmonic filter. In such cases an *antenna coupler* is required.

In certain cases when a pi-network is being used at the output of the transmitter, the addition of an antenna coupler will provide sufficient harmonic attenuation. But in all normal cases it is prudent to include a harmonic filter between the output of the transmitter and the antenna coupler.

Function of an Antenna Coupler The function of the antenna coupler is, basically, to transform the impedance of the antenna system being used to the correct value of *resistive* impedance for the harmonic filter, and hence for the transmitter. Thus the antenna coupler may be used to resonate the feeders or the radiating portion of the antenna system, in addition to its function of impedance transformation.

It is important to remember that there is *nothing* that can be done at the antenna coupler which will eliminate standing waves on the antenna transmission line. Standing waves are the result of reflection from the antenna, and the coupler can do nothing about this condition. However, the antenna coupler can resonate the feed line (by introducing a conjugate impedance) in addition to providing an impedance transformation. Thus, a resistive impedance of the correct value can be presented to the harmonic filter, as in figure 41, regardless of any *reasonable* value of standing-wave ratio on the antenna transmission line.

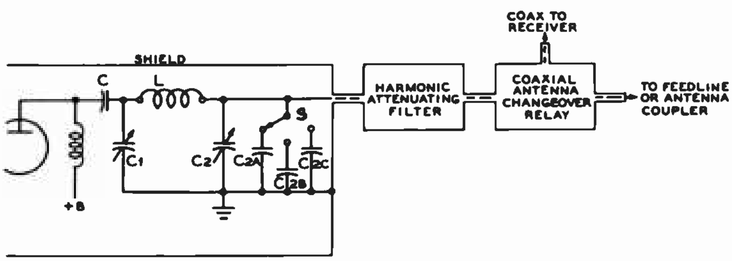
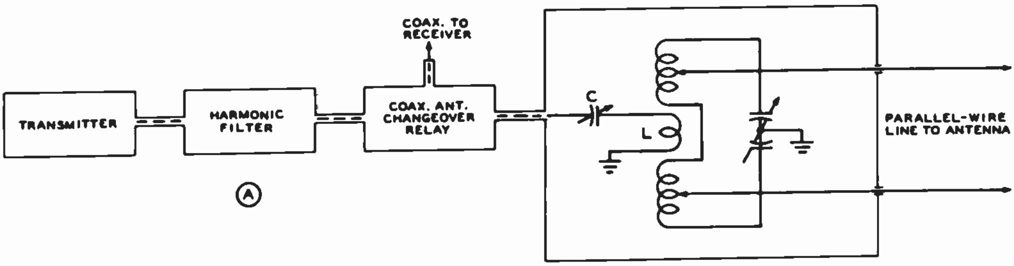


Figure 43

PI-NETWORK ANTENNA COUPLER

The design of pi-network circuits is discussed in Chapter Fifteen. The additional output-end shunting capacitors selected by switch S are for use on the lower frequency ranges. Inductor L may be selected by a tap switch; it may be continuously variable; or plug-in inductors may be used.



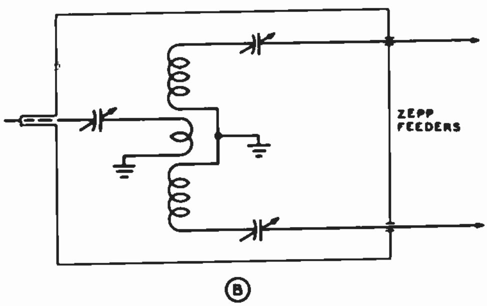


Figure 44

ALTERNATIVE ANTENNA COUPLER CIRCUITS

Plug-in coils, one or two variable capacitors of the split-stator variety, and a system of switches or plugs and jacks may be used in the antenna coupler to accomplish the feeding of different types of antennas and antenna transmission lines from the coaxial input line from the transmitter or from the antenna changeover relay. Link L should be resonated with capacitor C at the operating frequency of the transmitter so that the harmonic filter will operate into a resistive load impedance of the correct nominal value.

Types of Antenna Couplers All usual types of antenna couplers fall into two classifications: (1) inductively coupled resonant system as exemplified by those shown in figure 44, and (2) conductively coupled pi-network systems such as shown in figure 45. The inductively coupled system is convenient for feeding a balanced line from the coaxial output of the usual harmonic filter. The pi-network system is most useful for feeding a length of wire from the output of a transmitter.

Several general methods for using the inductively coupled resonant types of antenna coupler are illustrated in figure 44. The coupling between link coil L and the main tuned circuit need not be variable; in fact it is preferable that the correct link size and placement be determined for the tank coil which will be used for each band, and then that the link be made a portion of the plug-in coil. Capacitor C then can be adjusted to a predetermined value for each band so that it will resonate with the link coil for that band. The reactance of the link coil (and hence the reactance of the capacitor setting which will resonate the coil) should be about 3 or 4 times the impedance of the transmission line between the antenna coupler and the harmonic filter, so that the link coupling circuit will have an operating Q of 3 or 4.

The pi-network type of antenna coupler, as shown in figure 45, is useful for certain applications, but is primarily useful in feeding a single-wire antenna from a low-impedance transmission line. In such an application the operating Q of the pi-network may be somewhat lower than that of a pi-network in the plate circuit of the final amplifier of a transmitter, as shown in figure 40. An operating Q of 3 or 4 in such an application

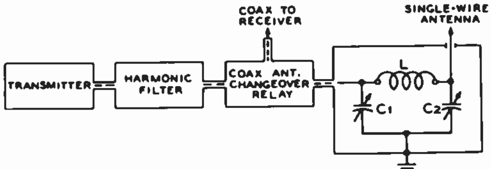


Figure 45

PI-NETWORK ANTENNA COUPLER

An arrangement such as illustrated above is convenient for feeding an end-fed Hertz antenna, or a random length of wire for portable or emergency operation, from the nominal value of impedance of the harmonic filter.

will be found to be adequate, since harmonic attenuation has been accomplished ahead of the antenna coupler.

An alternative arrangement shown in figure 46 utilizes the antenna-coupling tank circuit only when feeding the coaxial output of the transmitter to the open-wire feed line (or similar multiband antenna) of the 40- and 80-meter antenna. The coaxial lines to the 10-meter beam and to the 20-meter beam would be fed directly from the output of the coaxial antenna-changeover relay through switch S.

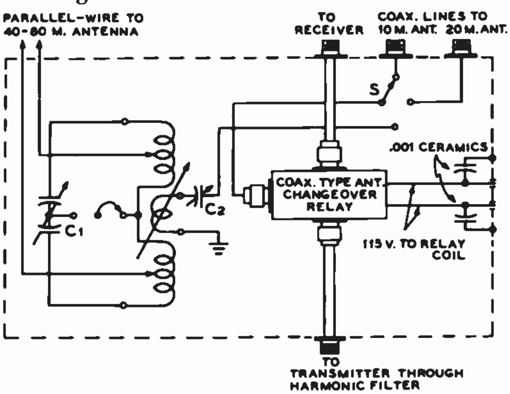


Figure 46

ALTERNATIVE COAXIAL ANTENNA COUPLER

This circuit is recommended for coaxial lines with low SWR used to feed antenna systems such as rotatable beams, and when it is desired to feed open-wire line to some sort of multiband antenna for the lower-frequency ranges. The tuned circuit of the antenna coupler is operative only when using the open-wire feed, and then it is in operation both for transmit and receive.

25-12 A Single-Wire Antenna Tuner

One of the simplest and least expensive antennas for transmision and reception is the single-wire, end-fed Hertz antenna. When used over a wide range of frequencies, this type of antenna exhibits a very great range of input impedance. At the low-frequency end of the spectrum such an antenna may present a resistive load of less than one ohm to the transmitter, combined with a large positive or negative value of reactance. As the frequency of operation is raised, the resistive load may rise to several thousand ohms (near half-wave resonance) and the reactive component of the load can rapidly change from positive to negative values, or vice-versa.

To provide indication for tuning the network, a radio-frequency bridge (SWR meter) is included to indicate the degree of mismatch (standing-wave ratio) existing at the input to the tuner. All adjustments to the tuner are made with the purpose of reaching unity standing-wave ratio on the coaxial feed system between the tuner and the transmitter.

A Practical Antenna Tuner A simple antenna tuner for use with transmitters of 250 watts power or less is shown in figures 47 through 49. An SWR-bridge circuit is used to indicate tuner resonance. The resistive arm of the bridge consists of ten 10-ohm, 1-watt carbon resistors connected in parallel to form a 1-ohm resistor (R_1). The other pair of bridge arms are capacitive rather than resistive. The bridge detector is a simple r-f voltmeter employing a 1N56 crystal diode and a 0-1 d-c milliammeter. A sensitivity control is incorporated to prevent overloading the meter when power is first applied to the tuner. Final adjustments are made with the sensitivity control at its maximum (clockwise) position. The bridge is balanced when the input impedance of the tuner is 52 ohms

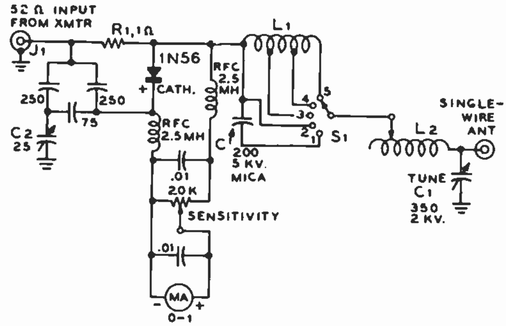
Figure 47

ANTENNA TUNER IS HOUSED IN METAL CABINET 7 INCHES $\times$ 8 INCHES IN SIZE

Inductance switch S_1 and sensitivity control are at left with counter dial for L_1 at center. Output tuning capacitor C_1 is at right. SWR meter is mounted above S_1.

resistive. This is the condition for maximum energy transfer between transmission line and antenna. The meter is graduated in arbitrary units, since actual SWR value is not required.

Tuner Construction Major parts placement in the tuner is shown in figures 47 and 49. Tapped coil L_1 is mounted on ½-inch ceramic insulators, and all major components are mounted above deck with the exception of the SWR bridge (figure 50). The components of the bridge are placed below deck, adjacent to the coaxial input plug mounted on the rear apron of the chassis. The ten 10-ohm resistors are soldered to two 1-inch rings made of copper wire as shown in the photograph. The bridge capacitors are attached to this assembly with extremely short leads. The 1N56 crystal mounts at right angles to the resistors to ensure minimum amount of capacitive coupling between the resistors and the detector.



L1-35 TURNS #18, 2" DIA., 3.5" LONG (*AIR-DUX*) TAP AT 15 T., 27 T., FROM POINT A

L2-*JOHNSON 229-201* VARIABLE INDUCTOR (*10 µH*)

C1-*JOHNSON 350E20*

C2-*CENTRALAB TYPE 822*

J1-TYPE SO-239 RECEPTACLE

R1-TEN 10-OHM 1-WATT CARBON RESISTORS IN PARALLEL. *IRC TYPE BTA*

Figure 48

SCHEMATIC OF A SINGLE-WIRE ANTENNA TUNER

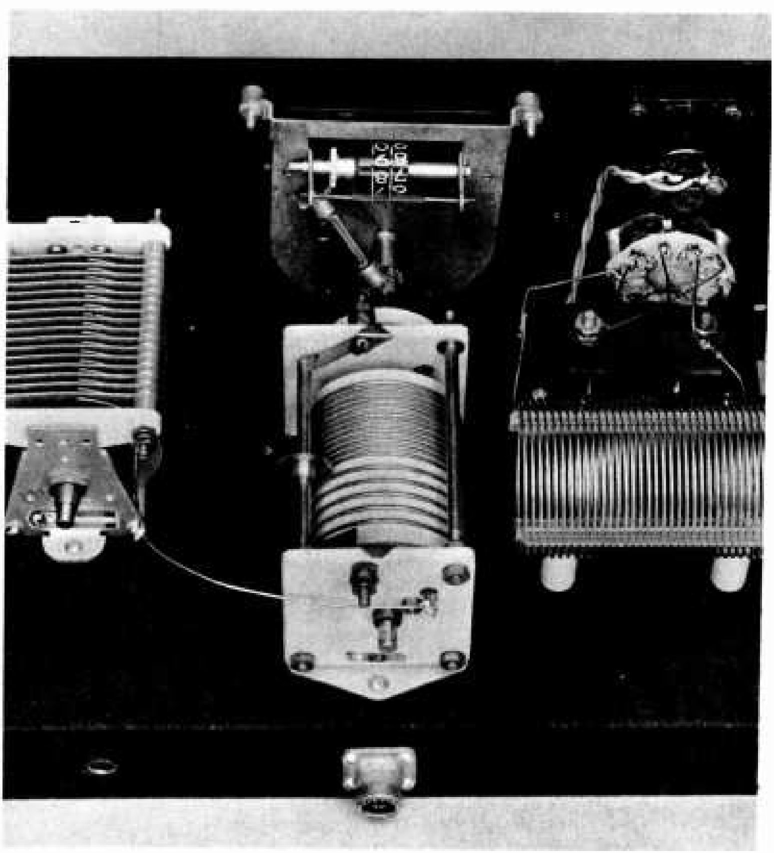


Figure 49

REAR VIEW OF TUNER SHOWING PLACEMENT OF MAJOR COMPONENTS

Rotary inductor is driven by Johnson 116-208-4 counter dial. Coaxial input receptacle J_1 is mounted directly below rotary inductor.

The output lead from the bridge passes through a ceramic feedthrough insulator to the top side of the chassis.

Connection to the antenna is made by means of a large feedthrough insulator mounted on the back of the tuner cabinet. This insulator is not visible in the photographs.

Bridge Calibration The SWR bridge must be calibrated for 52-ohm service. This can be done by temporarily disconnecting the lead between the bridge and the antenna tuner and connecting a 2-watt, 52 ohm carbon resistor to the junction of R_1 and the negative terminal of the 1N56 diode. The opposite lead of the carbon resistor is grounded to the chassis of the bridge. A small amount of r-f energy is fed to the input of the bridge until a reading is obtained on the r-f voltmeter. The 25-pf bridge-balancing capacitor C_2 (see figure 50) is then adjusted with a fiber-blade screwdriver until a zero reading is obtained on the meter. The sensitivity control is advanced as the meter null grows, in order to obtain the exact point of bridge balance. When this point is found, the carbon resistor should be removed and the bridge attached to the antenna tuner. The bridge capacitor is sealed with a drop of nail polish to prevent misadjustment.

Tuner Adjustments All tuning adjustments are made to obtain proper transmitter loading with a balanced (zero-meter-reading) bridge condition. The tuner is connected to the transmitter through a random length of 52-ohm coaxial line, and the single-wire antenna is attached to the output terminal of the tuner. Transmitter loading controls are set to approximate a 52-ohm termination. The transmitter is turned on (preferably at reduced input) and resonance is established in the amplifier tank circuit. The sensitivity control of the tuner is adjusted to provide near full-scale deflection on the bridge meter. Various settings of S_1, L_2, and C_1 should be tried to obtain a reduction of bridge reading. As tuner resonance is approached, the meter reading will decrease and the sensitivity control should be advanced. When the system is in resonance, the meter will read zero. All loading adjustments may then be made with the transmitter controls. The tuner should be readjusted whenever the frequency of the transmitter is varied by an appreciable amount.

CHAPTER TWENTY-SIX

High-Frequency Directive Antennas

It is becoming of increasing importance in most types of radio communication to be capable of concentrating the radiated signal from the transmitter in a certain desired direction and to be able to discriminate at the receiver against reception from directions other than the desired one. Such capabilities involve the use of directive antenna arrays.

Few simple antennas, except the single vertical element, radiate energy equally well in all azimuth (horizontal or compass) directions. All horizontal antennas, except those specifically designed to give an omnidirectional azimuth radiation pattern such as the turnstile, have some directive properties. These properties depend on the length of the antenna in wavelengths, the height above ground, and the slope of the radiator.

The various forms of the half-wave horizontal antenna produce maximum radiation at right angles to the wire, but the directional effect is not great. Nearby objects also minimize the directivity of a dipole radiator, so that it hardly seems worth while to go to the trouble to rotate a simple half-wave dipole in an attempt to improve transmission and reception in any direction.

The half-wave doublet, folded-dipole, zepp, single-wire-fed, matched-impedance, and Q-section antennas all have practically the same radiation pattern *when properly built and adjusted*. They all are dipoles, and the feeder system, if it does not radiate in itself, will have no effect on the radiation pattern.

26-1 Directive Antennas

When a multiplicity of radiating elements is located and phased so as to reinforce the radiation in certain desired directions and to neutralize radiation in other directions, a *directive antenna array* is formed.

The function of a directive antenna when used for transmitting is to give an increase in signal strength in some direction at the expense of radiation in other directions. For reception, one might find useful an antenna giving little or no gain in the direction from which it is desired to receive signals if the antenna is able to discriminate against interfering signals and static arriving from other directions. A good directive transmitting antenna, however, can also be used to good advantage for reception.

If radiation can be confined to a narrow beam, the signal intensity can be increased a great many times in the desired direction of transmission. This is equivalent to increasing the power output of the transmitter. On the higher frequencies, it is more economical to use a directive antenna than to increase transmitter power, if more than a few watts of power is being used.

Directive antennas for the high-frequency range have been designed and used commercially with gains as high as 23 db over a simple dipole radiator. Gains as high as 35

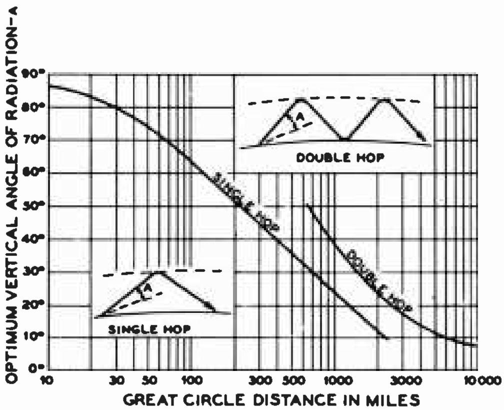


Figure 1

OPTIMUM ANGLE OF RADIATION WITH RESPECT TO DISTANCES

Shown above is a plot of the optimum angle of radiation for one-hop and two hop communication. An operating frequency close to of radiation for one-hop and two-hop communication distance is assumed.

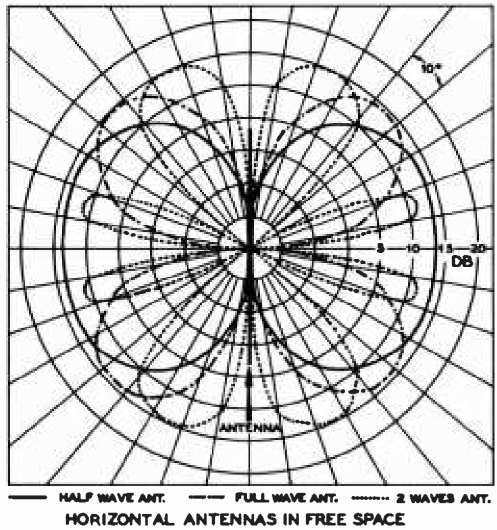


Figure 2

FREE-SPACE FIELD PATTERNS OF LONG-WIRE ANTENNAS

The presence of the earth distorts the field pattern in such a manner that the azimuth pattern becomes a function of the elevation angle.

db are common in direct-ray microwave communication and radar systems. A gain of 23 db represents a power gain of 200 times and a gain of 35 db represents a power gain of almost 3500 times. However, an antenna with a gain of only 15 to 20 db is so sharp in its radiation pattern that it is usable to full advantage only for point-to-point work.

The increase in radiated power in the desired direction is obtained at the expense of radiation in the undesired directions. Power gains of 3 to 12 db seem to be most practical for amateur communication, since the width of a beam with this order of power gain is wide enough to sweep a fairly large area. Gains of 3 to 12 db represent effective transmitter power increases from 2 to 16 times.

Horizontal Pattern versus Vertical Angle There is a certain optimum vertical angle of radiation for sky-wave communication, this angle being dependent on distance, frequency, time of day, etc. Energy radiated at an angle much lower than this optimum angle is largely lost, while radiation at angles much higher than this optimum angle is often not nearly so effective.

For this reason, the horizontal directivity pattern as measured on the ground is of no import when dealing with frequencies and distances dependent on sky-wave propagation. It is the horizontal directivity (or gain, or discrimination) *measured at the most useful vertical angles of radiation* that is of consequence. The horizontal radiation pattern, as measured on the ground, is considerably different from the pattern obtained at a vertical angle of 15 degrees, and still more different from a pattern obtained at a vertical angle of 30 degrees. In general, the energy which is radiated at angles higher than approximately 30 degrees above the earth, is effective only for local work at any frequency.

For operation at frequencies in the vicinity of 14 MHz, the most effective angle of radiation is usually about 15 degrees above the horizon, from *any* kind of antenna. The most effective angles for 10-meter operation are those in the vicinity of 10 degrees. Figure 1 is a chart giving the optimum vertical angle of radiation for sky-wave propagation in terms of the great-circle distance between the transmitting and receiving antennas.

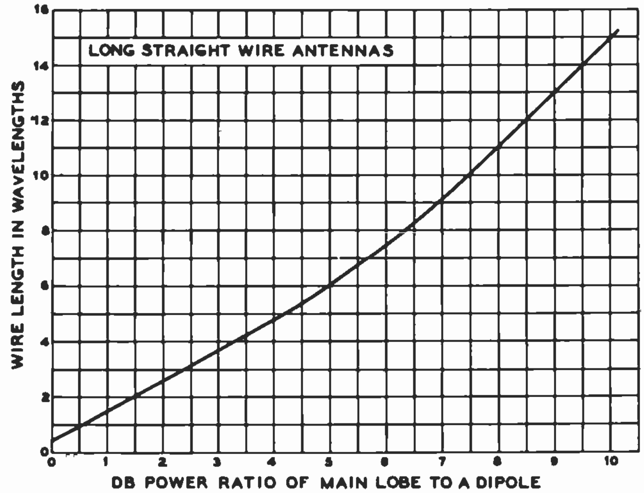


Figure 3
DIRECTIVE GAIN OF LONG-WIRE ANTENNAS

Types of Directive Arrays There is an enormous variety of directive antenna arrays that can give a substantial power gain in the desired direction of transmission or reception. However, some are more effective than others which require the same space. In general it may be stated that long-wire antennas of various types, such as the single long wire, the V beam, and the rhombic, are less effective for a given space than arrays composed of resonant elements, but the long-wire arrays have the significant advantage that they may be used over a relatively large frequency range while resonant arrays are usable only over a quite narrow frequency band.

26-2 Long-Wire Radiators

Harmonically operated long wires radiate better in certain directions than others, but cannot be considered as having appreciable directivity unless several wavelengths long. The current in adjoining half-wave elements flows in opposite directions at any instant, and, thus, the radiation from the various elements adds in certain directions and cancels in others.

A half-wave doublet in free space has a "doughnut" of radiation surrounding it. A full wave has 2 lobes, 3 half waves 3, etc. When the radiator is made more than 4 half wavelengths long, the *end* lobes (cones of radiation) begin to show noticeable power gain over a half-wave doublet, while the broadside lobes get smaller and smaller in amplitude, even though numerous (figure 2).

The horizontal radiation pattern of such antennas depends on the vertical angle of radiation being considered. If the wire is more than 4 wavelengths long, the maximum radiation at vertical angles of 15° to 20° (useful for DX) is in line with the wire, being slightly greater a few degrees either side of the wire than directly off the ends. The directivity of the main lobes of radiation is not particularly sharp, and the minor lobes fill in between the main lobes to permit working stations in nearly all directions, though the power radiated broadside to the radiator will not be great if the radiator is more than a few wavelengths long. The directive gain of long-wire antennas, in terms of the wire length in wavelengths is given in figure 3.

To maintain the out-of-phase condition in adjoining half-wave elements throughout the length of the radiator, it is necessary that a harmonic antenna be fed either at *one end* or at a *current* loop. If fed at a voltage loop, the adjacent sections will be fed *in phase*, and a different radiation pattern will result.

The directivity of a long wire does not increase very much as the length is increased beyond about 15 wavelengths. This is due to the fact that all long-wire antennas are adversely affected by the r-f resistance of the wire, and because the current

LONG-ANTENNA DESIGN TABLE Approximate Length in Feet—End-Fed Antennas								
Frequency In MHz	1λ	1½λ	2λ	2½λ	3λ	3½λ	4λ	4½λ
29 28	33 34	50 52	67 69	84 87	101 104	118 122	135 140	152 157
21.4 21.2 21.0	45 45 ¼ 45 ½	68 68 ¼ 68 ½	91 ½ 91 ¾ 92	114 ½ 114 ¾ 115	136 ½ 136 ¾ 137	160 ½ 160 ¾ 161	185 ½ 185 ¾ 186	209 ½ 209 ¾ 210
14.2 14.0	67 ½ 68 ½	102 103 ½	137 139	171 174	206 209	240 244	275 279	310 314
7.3 7.15 7.0	136 136 ½ 137	206 207 207 ½	276 277 277 ½	346 347 348	416 417 418	486 487 488	555 557 558	625 627 628
4.0 3.8 3.6 3.5	240 252 266 274	362 381 403 414	485 511 540 555	618 640 676 696	730 770 812 835	853 900 950 977	977 1030 1090 1120	1100 1160 1220
2.0 1.9 1.8	480 504 532	725 763 805	972 1020 1080	1230 1280	1475			

amplitude begins to become unequal at different current loops as a result of attenuation along the wire caused by radiation and losses. As the length is increased, the tuning of the antenna becomes quite broad. In fact, a long wire about 15 waves long is practically *aperiodic*, and works almost equally well over a wide range of frequencies.

One of the most practical methods of feeding a long-wire antenna is to bring one end of it into the radio room for direct connection to a tuned antenna circuit which is link-coupled through a harmonic-attenuating filter to the transmitter. The antenna can be tuned effectively to resonance for operating on any harmonic by means of the tuned circuit which is connected to the end of the antenna. A ground is sometimes connected to the center of the tuned coil.

If desired, the antenna can be opened and current-fed at a point of *maximum current* by means of low-impedance ribbon line, or by a quarter-wave matching section and open line.

26-3 The V Antenna

If two long-wire antennas are built in the form of a V, it is possible to make two of the maximum lobes of one leg shoot in the same direction as two of the maximum lobes of the other leg of the V. The resulting antenna is bidirectional (two opposite directions) for the main lobes of radiation. Each side of the V can be made any odd or

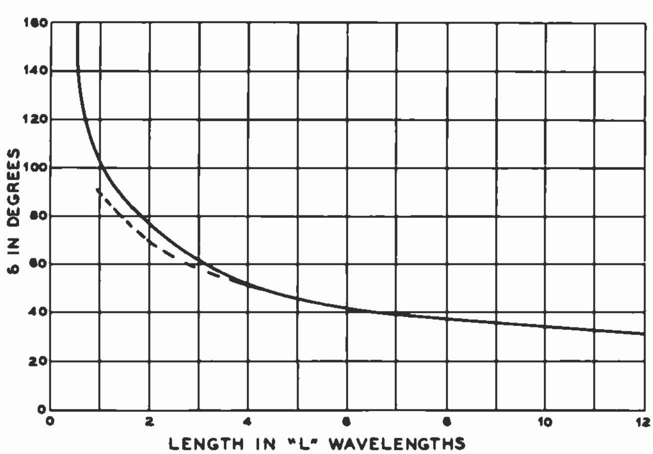


Figure 4

INCLUDED ANGLE FOR A V BEAM

Showing the included angle between the legs of a V beam for various leg lengths. For optimum alignment of the radiation lobe at the correct vertical angle with leg lengths less than three wavelengths, the optimum included angle is shown by the dashed curve.

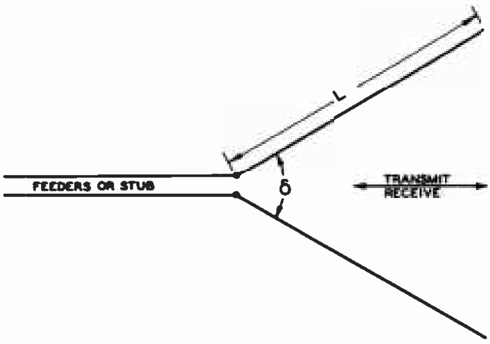


Figure 5

TYPICAL V BEAM ANTENNA

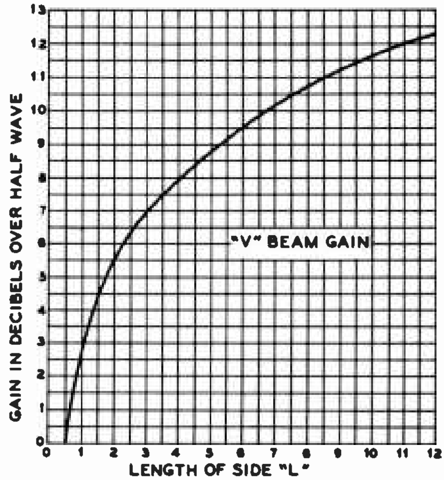


Figure 6

DIRECTIVE GAIN OF A V BEAM

This curve shows the approximate directive gain of a V beam with respect to a half-wave antenna located the same distance above ground, in terms of the side length L.

even number of quarter wavelengths, depending on the method of feeding the apex of the V. The complete system must be a multiple of half waves. If each leg is an even number of quarter waves long, the antenna must be voltage-fed at the apex; if an odd number of quarter waves long, current feed must be used.

By choosing the proper apex angle (figure 4 and figure 5) the lobes of radiation from the two long-wire antennas aid each other to form a bidirectional beam. Each wire by itself would have a radiation pattern similar to that for a long wire. The reaction of one on the other removes two of the four main lobes, and increases the other two in such a way as to form two lobes of still greater magnitude.

The correct wire lengths and the degree of the angle δ are listed in the *V-Antenna Design Table* for various frequencies in the 10-, 15-, 20- and 40-meter amateur bands. Apex angles for all side lengths are given in figure 4. The gain of a V beam in terms of the side length when optimum apex angle is used is given in figure 6.

The legs of a very long V antenna are usually so arranged that the included angle is twice the angle of the major lobe from a single wire if used alone. This arrangement concentrates the radiation of each wire along the bisector of the angle, and permits part of the other lobes to cancel each other.

With legs shorter than 3 wavelengths, the best directivity and gain are obtained with a somewhat smaller angle than that determined by the lobes. Optimum directivity for a one-wave V is obtained when the angle is 90° rather than 180°, as determined by the ground pattern alone.

V-ANTENNA DESIGN TABLE

Frequency in kHz	L = λ, δ = 90°	L = 2λ, δ = 70°	L = 4λ, δ = 52°	L = 8λ, δ = 39°
28000	34'8"	69'8"	140'	280'
29000	33'6"	67'3"	135'	271'
21100	45'9"	91'9"	183'	366'
21300	45'4"	91'4"	182'6"	365'
14050	69'	139'	279'	558'
14150	68'6"	138'	277'	555'
14250	68'2"	137'	275'	552'
7020	138'2"	278'	558'	1120'
7100	136'8"	275'	552'	1106'
7200	134'10"	271'	545'	1090'

If very long wires are used in the V, the angle between the wires is almost unchanged when the length of the wires in wavelengths is altered. However, an error of a few degrees causes a much larger loss in directivity and gain in the case of the longer V than in the shorter one.

The vertical angle at which the wave is best transmitted or received from a horizontal V antenna depends largely on the included angle. The sides of the V antenna should be at least a half wavelength above ground; commercial practice dictates a height of approximately a full wavelength above ground.

26-4 The Rhombic Antenna

The terminated *rhombic* or *diamond* is probably the most effective directional antenna that is practical for amateur communication. This antenna is nonresonant, with the result that it can be used on four amateur bands, such as 10, 15, 20, and 40 meters. When the antenna is nonresonant, i.e., properly terminated, the system is unidirectional, and the wire dimensions are not critical.

Rhombic Termination When the free end is terminated with a resistance of a value between 700 and 800 ohms the rear lobes are eliminated, the forward gain is increased, and the antenna can be used on several bands without changes. The terminating resistance should be capable of dissipating one-third the power output of the transmitter, and should have very little reactance. For medium- or low-power transmitters, the noninductive *plaque* resistors will serve as a satisfactory termination. Several manufacturers offer special resistors suitable for terminating a rhombic antenna. The terminating device should, for technical reasons, present a small amount of inductive reactance at the point of termination.

A compromise terminating device commonly used consists of a terminated 250-foot or longer length of line made of resistance wire which *does not have too much resistance per unit length.* If the latter qualification is not met, the reactance of the line will be excessive. A 250-foot line

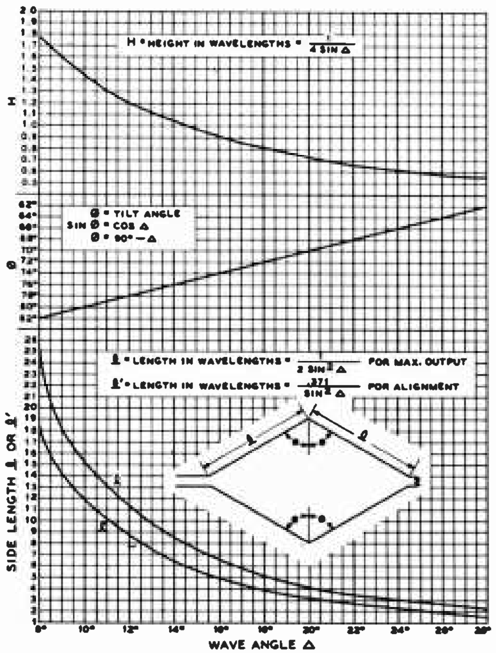


Figure 7

RHOMBIC ANTENNA DESIGN TABLE

Design data is given in terms of the wave angle (vertical angle of transmission and reception) of the antenna. The lengths l are for the "maximum output" design; the shorter lengths (l') are for the "alignment" method which gives approximately 1.5 db less gain with a considerable reduction in the space required for the antenna. The values of side length, tilt angle, and height for a given wave angle are obtained by drawing a vertical line upward from the desired wave angle.

consisting of No. 25 *nichrome* wire, spaced 6 inches and terminated with 800 ohms, will serve satisfactorily. Because of the attenuation of the line, the lumped resistance at the end of the line need dissipate but a few watts even when high power is used. A half dozen 5000-ohm 2-watt carbon resistors in parallel will serve for all except very high power. The attenuating line may be folded back on itself to take up less room.

The determination of the best value of terminating resistor may be made while receiving, if the input impedance of the receiver is approximately 800 ohms. The value of resistor which gives the best di-

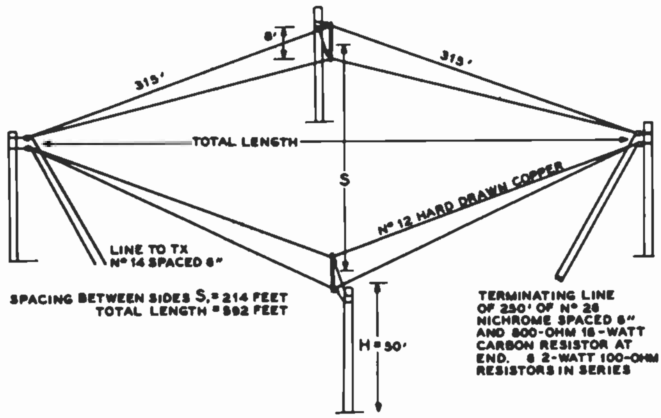


Figure 8

TYPICAL RHOMBIC ANTENNA DESIGN

The antenna system illustrated above may be used over the frequency range from 7 to 29 MHz without change. The directivity of the system may be reversed by the system discussed in the text.

rectivity on reception will not give the most gain when transmitting, but there will be little difference between the two.

The input resistance of the rhombic which is reflected into the transmission line that feeds it is always somewhat less than the terminating resistance, and is around 700 to 750 ohms when the terminating resistor is 800 ohms.

The antenna should be fed with a nonresonant line having a characteristic impedance of 650 to 700 ohms. The four corners of the rhombic should be at least one-half wavelength above ground for the lowest frequency of operation. For three-band operation the proper tilt angle (ϕ) for the center band should be observed.

The rhombic antenna transmits a horizontally polarized wave at a relatively low angle above the horizon. The angle of radiation (wave angle) decreases as the height above ground is increased in the same manner as with a dipole antenna. The rhombic should not be tilted in any plane. In other words, the poles should all be of the same height and the plane of the antenna should be parallel to the ground.

A considerable amount of directivity is lost when the terminating resistor is left off the end and the system is operated as a resonant antenna. If it is desired to reverse the direction of the antenna it is much better practice to run transmission lines to both ends of the antenna and then run the terminating line to the operating position. Then with the aid of two dpdt switches it will be possible to connect either feeder to the antenna changeover switch and the other feeder to the terminating line, thus reversing the direction of the array and maintaining the same termination for either direction of operation.

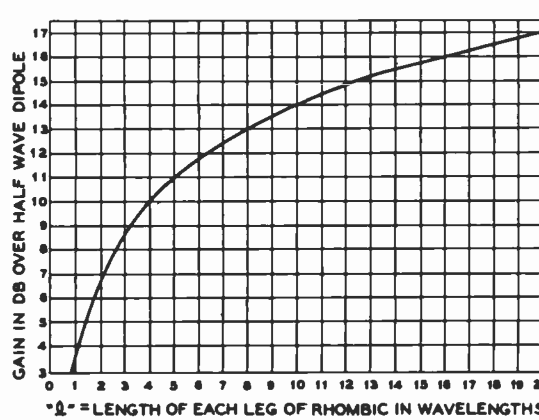


Figure 9

RHOMBIC ANTENNA GAIN

Showing the theoretical gain of a rhombic antenna, in terms of the side length, over a half-wave antenna mounted at the same height above the same type of soil.

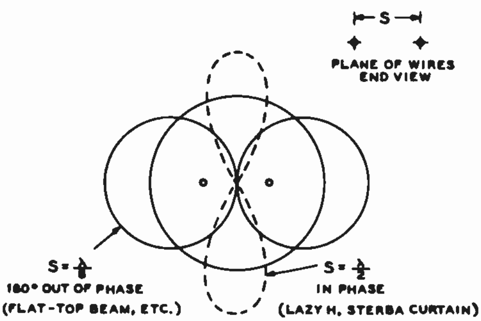


Figure 10

RADIATION PATTERNS OF A PAIR OF DIPOLES OPERATING WITH IN-PHASE EXCITATION, AND WITH EXCITATION 180° OUT OF PHASE

If the dipoles are oriented horizontally most of the directivity will be in the vertical plane; if they are oriented vertically most of the directivity will be in the horizontal plane.

Figure 7 gives curves for optimum-design rhombic attennas for both the maximum-output method and the alignment method. The alignment method is about 1.5 db down from the maximum output method but requires only about 0.74 as much leg length. The height and tilt angle are the same in either case. Figure 8 gives construction data for a recommended rhombic antenna for the 7.0- through 29.7-MHz bands. This antenna will give about 11 db gain in the 14.0-MHz band. The approximate gain of a rhombic antenna over a dipole (both above normal soil) is given in figure 9.

26-5 Stacked-Dipole Arrays

The characteristics of a half-wave dipole already have been described. When another dipole is placed in the vicinity and excited either directly or parasitically, the resultant radiation pattern will depend on the spacing and phase differential, as well as the relative magnitude of the currents. With spacings less than 0.65 wavelength, the radiation is mainly broadside to the two wires (bidirectional) when there is no phase difference, and *through* the wires (end fire) when the wires are 180° out of phase. With phase difference between 0° and 180° (45°,

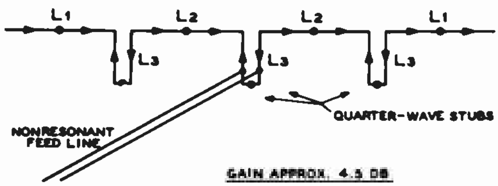


Figure 11

THE FRANKLIN OR COLLINEAR ANTENNA ARRAY

An antenna of this type, regardless of the number of elements, attains all of its directivity through sharpening of the horizontal or azimuth radiation pattern; no vertical directivity is provided. Hence a long antenna of this type has an extremely sharp azimuth pattern, but no vertical directivity.

90°, and 135° for instance), the pattern is unsymmetrical, the radiation being *greater in one direction* than in the opposite direction.

With spacings of more than 0.8 wavelength, more than two main lobes appear for all phasing combinations; hence, such spacings are seldom used.

In-Phase Spacing With the dipoles driven so as to be in phase, the most effective spacing is between 0.5 and 0.7 wavelength. The latter provides greater gain, but minor lobes are present which do not appear at 0.5-wavelength spacing. The radiation is broadside to the plane of the wires, and the gain is slightly greater than can be obtained from two dipoles out of phase. The gain falls off rapidly for spacings less than 0.375 wavelength, and there is little point in using spacing of 0.25 wavelength or less with in-phase dipoles, except where it is desirable to increase the radiation resistance. (See *Multiwire Doublet.*)

Out of Phase Spacing When the dipoles are fed 180° *out of phase*, the directivity is through the plane of the wires, and is greatest with *close spacing*, though there is but little difference in the pattern after the spacing is made less than 0.125 wavelength. The radiation resistance becomes so low for spacings of less than 0.1 wavelength that such spacings are not practical.

In the three foregoing examples, most of the directivity provided is in a plane at a right angle to the wires, though when out

COLLINEAR ANTENNA DESIGN CHART			
Frequency in MHz	L_1	L_2	L_3
28.5	16'8"	17'	8'6"
21.2	22'8"	23'3"	11'6"
14.2	33'8"	34'7"	17'3"
7.15	67'	68'8"	34'4"
4.0	120'	123'	61'6"
3.6	133'	136'5"	68'2"

of phase, the directivity is in a line *through* the wires, and when in phase, the directivity is *broadside* to them. Thus, if the wires are oriented vertically, mostly horizontal directivity will be provided. If the wires are oriented horizontally, most of the directivity obtained will be *vertical* directivity.

To increase the sharpness of the directivity in all planes that include one of the wires, additional identical elements are added *in the line of the wires,* and fed so as to be *in phase.* The familiar lazy-H array is one array utilizing both types of directivity in the manner prescribed. The two-section 8JK flat-top beam is another.

These two antennas in their various forms are directional in a horizontal plane, in addition to being low-angle radiators, and are perhaps the most practical of the *bidirectional* stacked-dipole arrays for amateur use. More phased elements can be used to provide greater directivity in planes including one of the radiating elements. The H then becomes a *Sterba-curtain* array.

For unidirectional work the most practical stacked-dipole arrays for amateur-band use are parasitically excited systems using relatively close spacing between the reflectors and the directors. Antennas of this type are described in detail in a later chapter. The next most practical unidirectional array is an H or a Sterba curtain with a similar system placed approximately one-quarter wave behind. The use of a reflector system in conjunction with any type of stacked-dipole broadside array will increase the gain by 3 db.

Collinear Arrays The simple *collinear antenna* array is a very effective radiating system for the 3.5- and 7.0-MHz bands, but its use is not recommended on higher frequencies since such arrays do not possess any vertical directivity. The elevation radiation pattern for such an array is essentially the same as for a half-wave

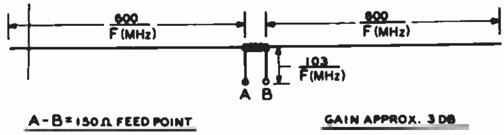


Figure 12

DOUBLE EXTENDED ZEPP ANTENNA

For best results, antenna should be tuned to operating frequency by means of grid-dip oscillator.

dipole. This consideration applies whether the elements are of normal length or are extended.

The collinear antenna consists of two or more radiating sections from 0.5 to 0.65 wavelengths long, with the current in phase in each section. The necessary phase reversal between sections is obtained through the use of resonant tuning stubs as illustrated in figure 11. The gain of a collinear array using half-wave elements (in decibels) is approximately equal to the number of elements in the array. The exact figures are as follows:

Number of Elements	2	3	4	5	6
Gain in Decibels	1.8	3.3	4.5	5.3	6.2

As additional in-phase collinear elements are added to a doublet, the radiation resistance goes up much faster than when additional half waves are added out of phase (harmonic operated antenna).

For a collinear array of from 2 to 6 elements, the terminal radiation resistance in ohms at any current loop is approximately 100 times the number of elements.

It should be borne in mind that the *gain* from a collinear antenna depends on the

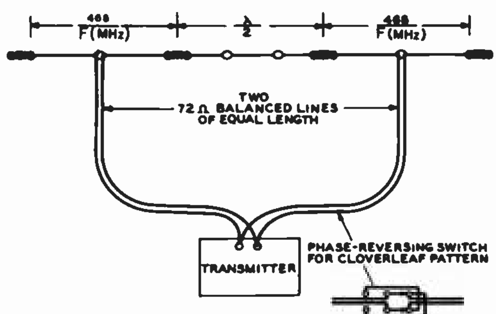


Figure 13

TWO COLLINEAR HALF-WAVE ANTENNAS IN PHASE PRODUCE A 3 DB GAIN WHEN SEPARATED ONE-HALF WAVELENGTH

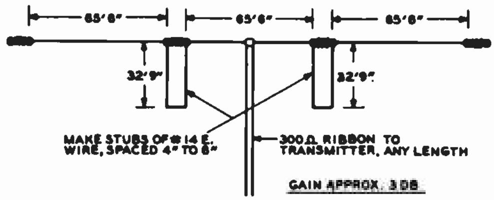


Figure 14

PRECUT LINEAR ARRAY FOR 40-METER OPERATION

sharpness of the horizontal directivity since no vertical directivity is provided. An array with several collinear elements will give considerable gain, but will have a sharp horizontal radiation pattern.

Double Extended Zepp The gain of a conventional two-element •Franklin collinear antenna can be increased to a value approaching that obtained from a three-element Franklin, simply by making the two radiating elements 230° long instead of 180° long. The phasing stub is shortened correspondingly to maintain the whole array in resonance. Thus, instead of having 0.5-wavelength elements and 0.25-wavelength stub, the elements are made 0.64 wavelength long and the stub is approximately 0.11 wavelength long.

Dimensions for the double extended zepp are given in figure 12.

The *vertical* directivity of a collinear antenna having 230° elements is the same as for one having 180° elements. There is little advantage in using extended sections when the total length of the array is to be greater than about 1.5 wavelength over all since the gain of a collinear antenna is proportional to the overall length, whether the individual radiating elements are ¼-wave, ½-wave or ¾-wave in length.

Spaced Half Wave Antennas The gain of two collinear half waves may be increased by increasing the physical spacing between the elements, up to a maximum of about one-half wavelength. If the half-wave elements are fed with equal lengths of transmission line, correctly phased, a gain of about 3.3 db is produced. Such an antenna is shown in figure 13. By means of a phase reversing switch, the two elements may be operated out of phase, producing a cloverleaf pattern with slightly less maximum gain.

A three-element precut array for 40-meter operation is shown in figure 14. It is fed directly with 300-ohm ribbon line, and may be matched to a 52-ohm coaxial output transmitter by means of a balun.

26-6 Broadside Arrays

Collinear elements may be stacked above or below another string of collinear elements to produce what is commonly called a *broadside* array. Such an array, when horizontal elements are used, possesses vertical directivity in proportion to the number of broadsided (vertically stacked) sections which have been used.

Since broadside arrays do have good vertical directivity their use is recommended on the 14-MHz band and on those higher in frequency. One of the most popular of simple broadside arrays is the *Lazy H* array of figure 15. Horizontal collinear elements stacked two above two make up this antenna system which is highly recommended when moderate gain without too much directivity is desired. It has high radiation resistance and a gain of approximately 5.5 db. The high radiation resistance results in low voltages and a broad resonance curve, which permits use of inexpensive insulators and enables the array to be used over a fairly wide range in frequency. For dimensions, see the stacked dipole design table.

Stacked Dipoles Vertical stacking may be applied to strings of collinear elements longer than two half waves. In such arrays, the end quarter wave of each string of radiators usually is bent in to meet a similar bent quarter wave from the opposite-end radiator. This provides better balance and better coupling between the upper and lower elements when the array is current-fed. Arrays of this type are shown in figure 16, and are commonly known as *curtain arrays*.

Correct length for the elements and stubs can be determined for any stacked-dipole array from the *Stacked-Dipole Design Table*.

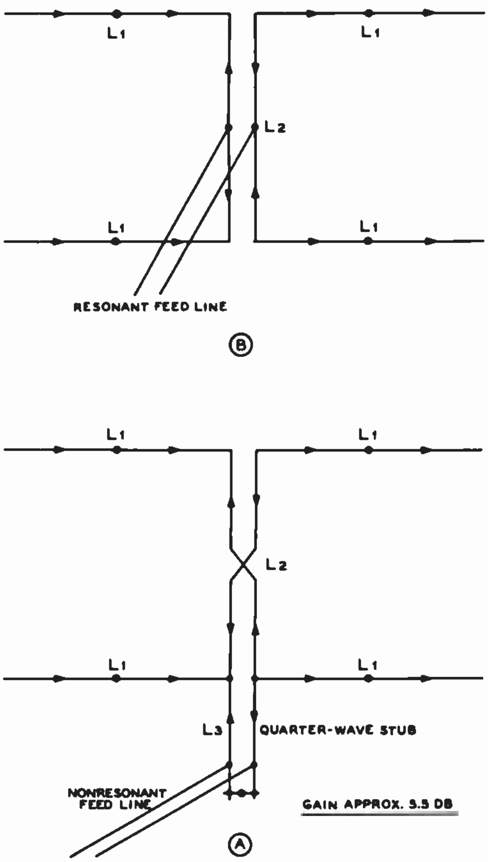


Figure 15

THE LAZY H ANTENNA SYSTEM

Stacking the collinear pairs gives both horizontal and vertical directivity. As shown, the array will give about 5.5 db gain. Note that the array may be fed either at the center of the phasing section or at the bottom; if fed at the bottom the phasing section must be twisted through 180°.

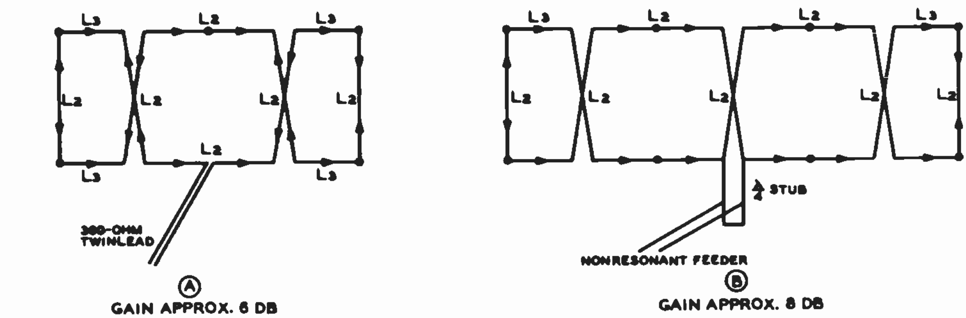
In the arrays of figure 16 the arrowheads represent the direction of current flow at any given instant. The dots on the radiators represent points of maximum current. All arrows should point in the same direction in each portion of the radiating sections of an antenna in order to provide a field in phase for broadside radiation. This condition is satisfied for the arrays illustrated in figure 16. Figures 16A and 16C show simple methods of feeding a short Sterba curtain, while an alternative method of feed is shown in the higher-gain antenna of figure 16B.

In the case of each of the arrays of figure 16, and also the Lazy H of figure 15, the array may be unidirectional and the gain increased by 3 db if an exactly similar array is constructed and placed approximately ¼ wave behind the driven array. A screen or mesh of wires, slightly greater in area than the antenna array, may be used instead of an additional array as a reflector to obtain a unidirectional system. The spacing between the reflecting wires may vary from 0.05 to 0.1 wavelength with the spacing between the reflecting wires the smallest directly behind the driven elements. The wires in the untuned reflecting system should be parallel to the radiating elements of the array, and the spacing of the complete reflector system should be approximately 0.2 to 0.25 wavelength behind the driven elements.

On frequencies below perhaps 100 MHz, it normally will be impractical to use a wire-screen reflector behind an antenna array such as a Sterba curtain or a Lazy H. Parasitic elements may be used as reflectors or directors, but parasitic elements have the disadvantage that their operation is selective with respect to relatively small changes in frequency. Nevertheless, parasitic reflectors for such arrays are quite widely used.

The X-Array In section 22-5 it was shown how two dipoles may be arranged in phase to provide a power gain of about 3 db. If two such pairs of dipoles are stacked in a vertical plane and properly phased, a simplified form of in-phase curtain is formed, providing an over-all gain of about 6 db. Such an array is shown in figure 17. In this *X-array*, the four dipoles are all in phase, and are fed by four sections of 300-ohm line, each one-half wavelength long, the free ends of all four lines being connected in parallel. The feed impedance at the junction of these four lines is about 75 ohms, and a length of 75-ohm twin lead may be used for the feedline to the array.

An array of this type is quite small for the 28-MHz band, and is not out of the question for the 21-MHz band. For best results, the bottom section of the array should be one-half wavelength above ground.



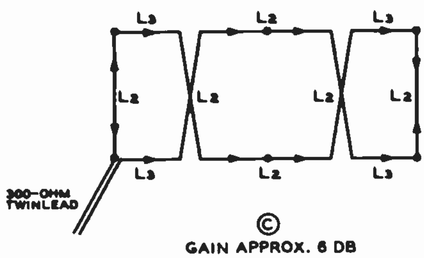


Figure 16

THE STERBA-CURTAIN ARRAY

Approximate directive gains along with alternative feed methods are shown.

LAZY H AND STERBA (STACKED-DIPOLE) DESIGN TABLE

Frequency in MHz	L_1	L_2	L_3
7.0 7.3	68'2" 65'10"	70' 67'6"	35' 33'9"
14.0 14.2 14.4	34'1" 33'8" 33'4"	35' 34'7" 34'2"	17'6" 17'3" 17'
21.0 21.5	22'9" 22'3"	23'3" 22'9"	11'8" 11'5"
27.3 28.0 29.0	17'7" 17' 16'6"	17'10" 17'7" 17'	8'11" 8'9" 8'6"
50.0 52.0 54.0	9'7" 9'3" 8'10"	9'10" 9'5" 9'1"	4'11" 4'8" 4'6"
144.0 146.0 148.0	39.8" 39" 38.4"	40.5" 40" 39.5"	20.3" 20" 19.8"

The Double-Bruce Array The *Bruce Beam* consists of a long wire folded so that vertical elements carry in-phase currents while the horizontal elements carry out-of-phase currents. Radiation from the horizontal sections is low since only a small current flows in this part of the wire, and it is largely phased-out. Since the height of the Bruce Beam is only one-quarter wavelength, the gain per linear foot of array is quite low. Two Bruce Beams may be combined as shown in figure 18 to produce the *Double Bruce* array. A four section *Double Bruce* will give a vertically polarized emission, with a power gain of 5 db over a simple dipole, and is a very simple beam to construct. This antenna, like other so-called broadside arrays, radiates

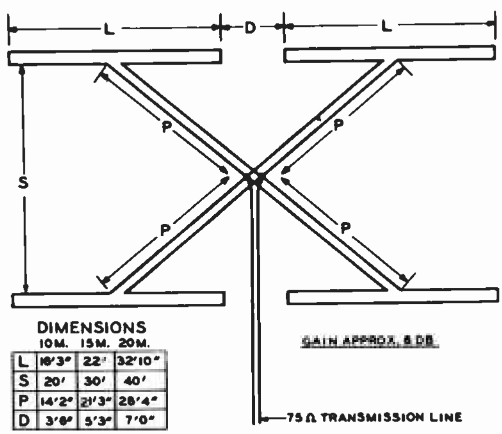


Figure 17

THE X-ARRAY FOR 28, 21, OR 14 MHz

The entire array (with the exception of the 75-ohm feedline) is constructed of 300-ohm ribbon line. Be sure phasing lines (P) are polarized correctly, as shown.

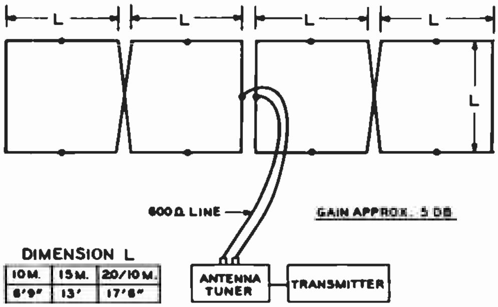


Figure 18

THE DOUBLE-BRUCE ARRAY FOR 10, 15, AND 20 METERS

If a 600-ohm feed line is used, the 20-meter array will also perform on 10 meters as a Sterba curtain, with an approximate gain of 9 db.

maximum power at right angles to the plane of the array.

The feed impedance of the Double Bruce is about 750 ohms. The array may be fed with a quarter-wave stub made of 300-ohm ribbon line and a feedline made of 150-ohm ribbon line. Alternatively, the array may be fed directly with a wide-spaced 600-ohm transmission line (figure 18). The feedline should be brought away from the Double Bruce for a short distance before it drops downward, to prevent interaction between the feedline and the lower part of the center phasing section of the array. For best results, the bottom sections of the array should be one-half wavelength above ground.

Arrays such as the X-array and the Double Bruce are essentially high-impedance devices, and exhibit relatively broadband characteristics. They are less critical of adjustment than a parasitic array, and they work well over a wide frequency range such as is encountered on the 28- to 29.7-MHz band.

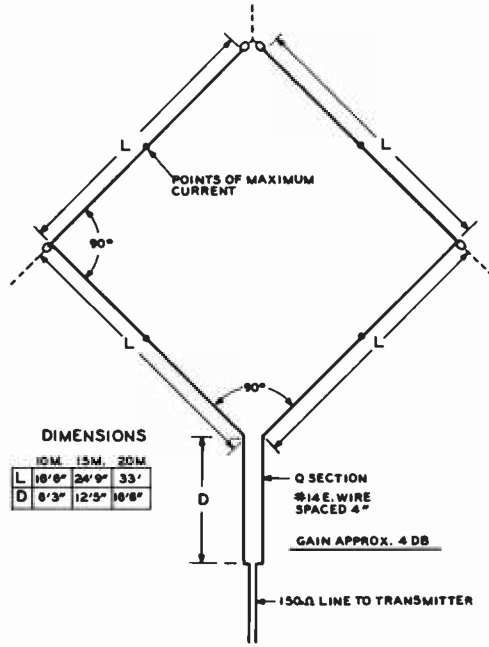


Figure 19

THE BI-SQUARE BROADSIDE ARRAY

This bidirectional array is related to the Lazy H, and in spite of the oblique elements, is horizontally polarized. It has slightly less gain and directivity than the Lazy H, the free-space directivity gain being approximately 4 db. Its chief advantage is the fact that only a single pole is required for support, and two such arrays may be supported from a single pole without interaction if the planes of the elements are at right angles. A 600-ohm line may be substituted for the twin lead, and either operated as a resonant line, or made nonresonant by the incorporation of a matching stub.

The Bi-Square Broadside Array Illustrated in figure 19 is a simple method of feeding a small broadside array. As two arrays of this type can be supported at right angles from a single pole without interaction, it offers a solution to the problem of suspending two arrays in a restricted space with a minimum of erection work. The free space directivity gain is slightly less than that of a Lazy H, but is still worthwhile, being approximately 4 db over a half-wave horizontal dipole at the same average elevation.

When two *Bi-Square* arrays are suspended at right angles to each other (for general coverage) from a single pole, the Q-sections should be well separated or else symmetrically arranged in the form of a square (the diagonal conductors forming one Q-section) in order to minimize coupling between them. The same applies to the line if open construction is used instead of twin lead, but

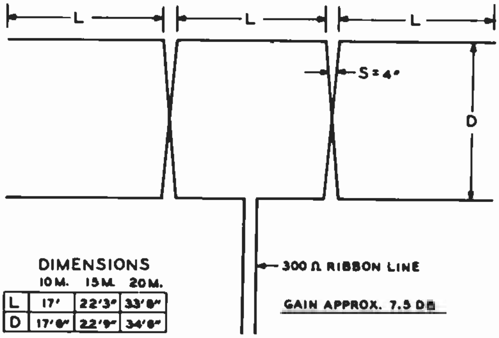


Figure 20

THE SIX-SHOOTER BROADSIDE ARRAY

if twin lead is used the coupling can be made negligible simply by separating the two twin-lead lines by at least two inches and twisting one twin lead so as to effect a transposition every foot or so.

When tuned feeders are employed, the Bi-Square array can be used on half frequency as an end-fire vertically polarized array, giving a slight practical DX signal gain over a vertical half-wave dipole at the same height.

A second Bi-Square serving as a reflector may be placed 0.15 wavelength behind this antenna to provide an over-all gain of 8.5 db. The reflector may be tuned by means of a quarter-wave stub which has a movable shorting bar at the bottom end.

The Six-Shooter Broadside Array The array of figure 20 is recommended for the 10- to 30-MHz range as a good compromise between gain, directivity, compactness, mechanical simplicity, ease of adjustment, and bandwidth, when the additional array width and greater directivity are not obtainable. The free-space directivity gain is approximately 7.5 db over one element, and the practical DX signal gain over one element at the same average elevation is of about the same magnitude when the array is sufficiently elevated. To show up to best advantage the array should be elevated sufficiently to put the lower elements well in the clear, and preferably at least 0.5 wavelength above ground.

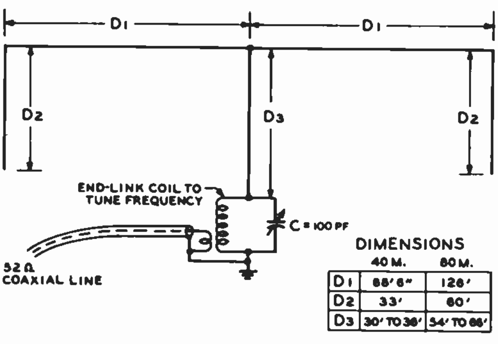


Figure 21

BOBTAIL BIDIRECTIONAL BROADSIDE CURTAIN FOR THE 7-MHz OR THE 4.0-MHz AMATEUR BANDS

This simple vertically polarized array provides low angle radiation and response with comparatively low pole heights, and is very effective for DX work on the 7-MHz band or the 4.0-MHz phone band. Because of the phase relationships, radiation from the horizontal portion of the antenna is effectively suppressed. Very little current flows in the ground lead to the coupling tank; so an elaborate ground system is not required, and the length of the ground lead is not critical so long as it uses heavy wire and is reasonably short.

The Bobtail Bidirectional Broadside Curtain Another application of vertical orientation of the radiating elements of an array in order to obtain low-angle radiation at the lower end of the high-frequency range with low pole heights is illustrated in figure 21. When precut to the specified dimensions this single-pattern array will perform well over the 7-MHz amateur band or the 4-MHz amateur phone band. For the 4-MHz band, the required two poles need be only 70 feet high, and the array will provide a practical signal gain averaging from 7 to 10 db over a horizontal half-wave dipole utilizing the same pole height when the path length exceeds 2500 miles.

The horizontal directivity is only moderate, the beam width at the half-power points being slightly greater than that obtained from three cophased vertical radiators fed with equal currents. This is explained by the fact that the current in each of the two outer radiators of this array carries only about half as much current as the center-driven element. While this *binomial* current distribution suppresses the end-fire lobe that

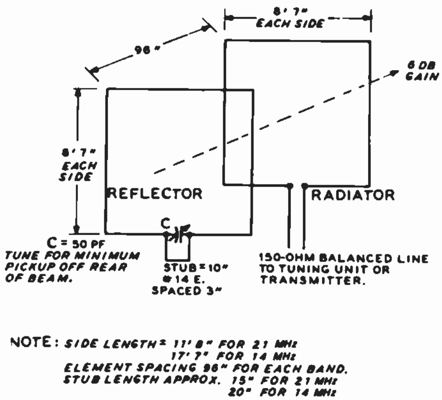


Figure 22

THE CUBICAL QUAD ANTENNA FOR THE 10-, 15-, OR 20-METER BANDS

occurs when an odd number of parallel radiators with half-wave spacing are fed equal currents, the array still exhibits some high-angle radiation and response off the end as a result of imperfect cancellation in the flat-top portion. This is not sufficient to affect the power gain appreciably, but does degrade the discrimination somewhat.

A moderate amount of sag can be tolerated at the center of the flat top, where it connects to the driven vertical element. The poles and antenna tank should be so located with respect to each other that the driven vertical element drops approximately straight down from the flat top.

Normally the antenna tank will be located in the same room as the transmitter, to facilitate adjustment when changing frequency. In this case it is recommended that the link-coupled tank be located across the room from the transmitter if much power is used, in order to minimize r-f feedback difficulties which might occur as a result of the asymmetrical high-impedance feed. If tuning of the antenna tank from the transmitter position is desired, flexible shafting can be run from the antenna tank capacitor to a control knob at the transmitter.

The lower end of the driven element is quite "hot" if much power is used, and the lead-in insulator should be chosen with this in mind. The ground connection need not have very low resistance, as the current flowing in the ground connection is comparatively small. A stake or pipe driven a few feet in the ground will suffice. However, the ground lead should be of heavy wire and preferably the length should not exceed about 10 feet at 7 MHz or about 20 feet at 4 MHz in order to minimize reactive effects due to its inductance. If it is impossible to obtain this short a ground lead, a piece of screen or metal sheet about four feet square may be placed parallel to the earth in a convenient location and used as an artificial ground. A fairly high C/L ratio ordinarily will be required in the antenna tank in order to obtain adequate coupling and loading.

26-7 The Cubical Quad Beam

The *Cubical Quad* may be thought of as a smaller version of the Bi-Square antenna. The radiator loop consists of a wavelength of wire folded into a rectangular or diamond shape, one-quarter of a wavelength on a side, as shown in figure 22. The director is a similar element placed in an end-fire position, and additional directors or a reflector may be added to the driven loop. The Quad, when fed at the bottom radiates a horizontally polarized signal.

A two-element Quad may be composed of driven element and director, with the director loop placed about 0.12-wavelength distance in front of the driven element. Power gain is about 7 decibels over a dipole antenna. A reflector loop placed about 0.12 wavelength behind the two-element assembly forms a three-element Quad having a power gain of about 8.5 decibels. Quad antennas for the 20-, 15-, or 10-meter bands have been built with up to five elements, with correspondingly higher gain figures.

Element lengths for the Quad antenna may be expressed in the circumference of the loop, and the following formulas apply for high-frequency Quads made of wire and having a square or diamond configuration:

Circumference of driven element:

$$(\text{feet})\ \frac{1005}{f_{\text{MHz}}}$$

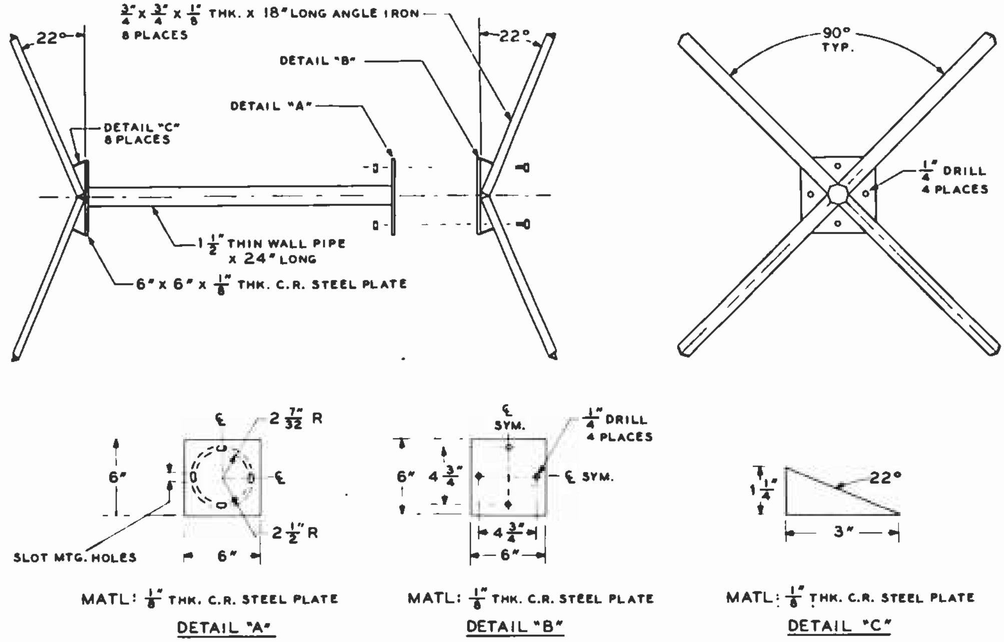


Figure 23

SPIDER CENTER STRUCTURE FOR QUAD ANTENNA

Circumference of director element:

$$(\text{feet})\ \frac{975}{f_{\text{MHz}}}$$

Circumference of reflector element:

$$(\text{feet})\ \frac{1030}{f_{\text{MHz}}}$$

A simple two-element Quad for 20, 15, or 10 meters is illustrated in figure 22. The elements are shortened a bit over those figures derived by formula and adjustable stubs are included in series with the loop wire to permit tuning to frequency. The Quad is fed with a 300-ohm balanced TV-type ribbon line and should employ an antenna tuner at the transmitter end of the line if a pi-network output stage is used in the transmitter. Alternatively, a 72-ohm coaxial line with a balun at the Quad terminals may be used for unbalanced feed. The radiation resistance of the Quad is about 100 ohms and a reasonably low value of SWR is obtainable across any one amateur band.

If a reflector stub is used, the array may be adjusted by aiming the back of the antenna at a nearby field-strength meter and adjusting the stub capacitor for minimum received signal at the operating frequency.

This simple antenna provides high gain for its size. The elements may be made of No. 14 wire with the array built on a light framework of bamboo arms with a wooden center structure. For maximum utility and longer life, the metal support structure of figure 23 is recommended. Built of conduit pipe and angle iron, this "spider" will accommodate bamboo or *Fiberglas* arms of sufficient length for a 20-, 15-, or 10-meter Quad, or an interlaced triband version. The "spider" is made in two parts so the elements may be assembled on the ground and carried to the top of the tower for final assembly. Boom length is only two feet, so the entire antenna can be easily supported by a single man.

Full information on interlaced Quads and complete Quad designs may be found in the handbook *All About Cubical Quad Antennas*, Radio Publications, Wilton, Conn.

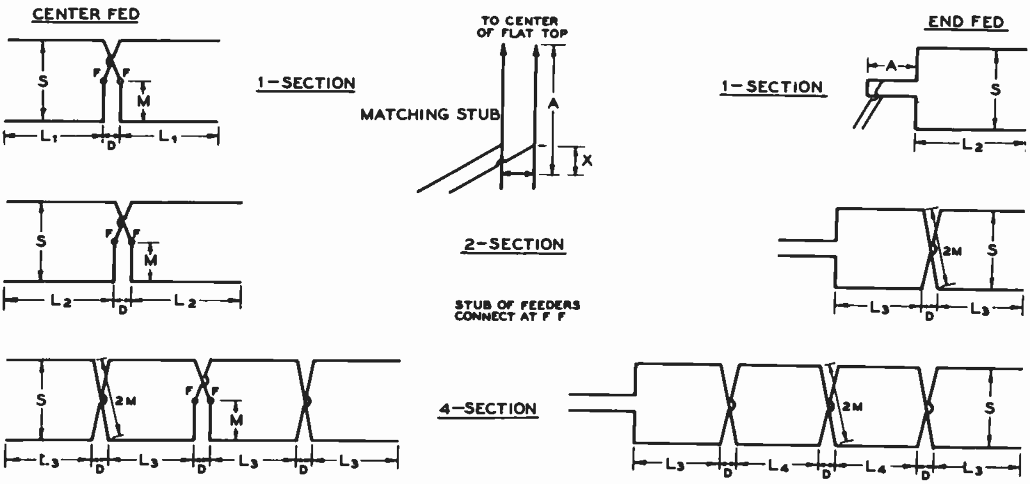


Figure 24

FLAT-TOP BEAM (8JK ARRAY) DESIGN DATA

Band	S	L_1	L_2	L_3	L_4	M	D	A (¼)	A (½)	A (¾)	X
40	17′	33′6″	59′	51′8″	43′1″	8′8″	4′	26′	59′	94′	4′
20	8′8″	17′	30′	26′4″	22′	4′9″	2′	13′	30′	48′	2′
15	7′10″	12′8″	22′6″	18′3″	15′	4′0″	1′8″	10′6″	22′	36′	1′6″
10	5′2″	8′6″	15′	12′7″	10′	2′8″	1′6″	7′	15′	24′	1′

Dimension chart for flat-top beam antennas. The meanings of the symbols are as follows:

L_1, L_2, L_3, L_4, the lengths of the sides of the flap-top sections as shown. L_1 is length of the sides of single-section center-fed, L_2 single-section end-fed and 2-section center fed, L_3 4-section center-fed and end-sections of 4-section end-fed, and L_4 middle sections of 4-section end-fed.

S, the spacing between the flat-top wires.

M, the wire length from the outside to the center of each cross-over.

D, the spacing lengthwise between sections.

A (¼), the approximate length for a quarter-wave stub.

A (½), the approximate length for a half-wave stub.

A (¾), the approximate length for a three-quarter wave stub.

X, the approximate distance above the shorting wire of the stub for the connection of a 600-ohm line. This distance, as given in the table, is approximately correct only for 2-section flat-tops.

For single-section types it will be smaller and for 4-section types it will be larger.

The lengths given for a half-wave stub are applicable only to single-section center-fed flat-tops. To be certain of sufficient stub length, it is advisable to make the stub a foot or so longer than shown in the table, especially with the end-fed types. The lengths, A, are measured from the point where the stub connects to the flat-top.

Both the center and end-fed types may be used horizontally. However, where a vertical antenna is desired, the flat-tops can be turned on end. In this case, the end-fed types may be more convenient, feeding from the lower end.

26-8 End-Fire Directivity

By spacing two half-wave dipoles, or collinear arrays, at a distance of from 0.1 to 0.25 wavelength and driving the two 180° out of phase, directivity is obtained *through the two wires* at right angles to them. Hence, this type of bidirectional array is called *end-fire.* A better idea of end-fire directivity can be obtained by referring to figure 10.

Remember that *end-fire* refers to the radiation with *respect to the two wires* in the array rather than with respect to the array as a whole.

The vertical directivity of an end-fire bidirectional array which is oriented horizontally can be increased by placing a similar end-fire array a half wave below it, and excited in the same phase. Such an array is a combination broadside and end-fire affair.

8JK Flat-Top Beam A very effective bidirectional end-fire array is the *8JK Flat-Top Beam.* Essentially, this antenna consists of two close-spaced dipoles or collinear arrays. Because of the close spacing, it is possible to obtain the proper phase relationships in multisection flat tops by crossing the wires at the voltage loops, rather than by resorting to phasing stubs. This greatly simplifies the array. (See figure 24.) Any number of sections may be used, though the one- and two-section arrangements are the most popular. Little extra gain is obtained by using more than four sections, and trouble from phase shift may appear.

A center-fed single-section flat-top beam cut according to the table, can be used quite successfully on its second harmonic, the pattern being similar except that it is a little sharper. The single-section array can also be used on its fourth harmonic with some success, though there then will be four cloverleaf lobes, much the same as with a full-wave antenna.

If a flat-top beam is to be used on more than one band, tuned feeders are necessary.

The radiation resistance of a flat-top beam is rather low, especially when only one section is used. This means that the voltage will be high at the voltage loops. For this reason, especially good insulators should be used for best results in wet weather.

The exact lengths for the radiating elements are not especially critical, because slight deviations from the correct lengths can be compensated in the stub or tuned feeders. Proper stub adjustment is covered in Chapter Twenty-four. Suitable radiator lengths and approximate stub dimensions are given in the accompanying design table.

Figure 24 shows *top views* of six types of flat-top beam antennas. The dimensions for using these antennas on different bands are given in the design table.

The antennas are tuned to the frequency used, by adjusting the shorting wire on the stub, or tuning the feeders, if no stub is used. The data in the table may be extended to other bands or frequencies by applying the proper factor. Thus, for 50- to 52-MHz operation, the values for 28 to 29 MHz are divided by 1.8.

All of the antennas have a bidirectional horizontal pattern on their fundamental frequency. The maximum signal is broadside to the flat top. The single-section type has this pattern on both its fundamental frequency and second harmonic. The other types have four main lobes of radiation on the second and higher harmonics. The nominal gains of the different types over a half-wave comparison antenna are as follows: single-section, 4 db; two-section, 6 db; four-section, 8 db.

Although the center-fed type of flat-top generally is to be preferred because of its symmetry, the end-fed type often is convenient or desirable. For example, when a flat-top beam is used vertically, feeding from the lower end is in most cases more convenient.

If a multisection flat-top array is end-fed instead of center-fed, and tuned feeders are used, stations off the ends of the array can be worked by tying the feeders together and working the whole affair, feeders and all, as a long-wire harmonic antenna. A single-pole double-throw switch can be used for changing the feeders and directivity.

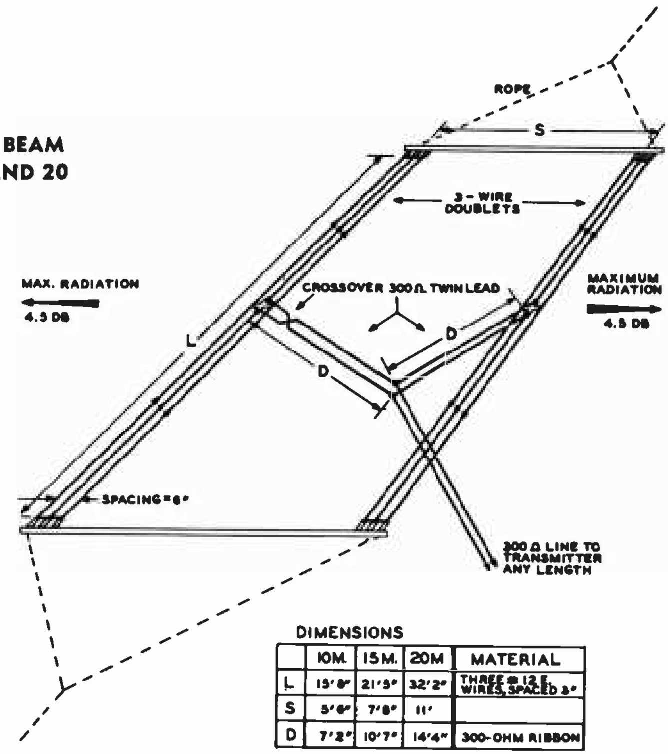
The Triplex Beam The *Triplex beam* is a modified version of the flat-top antenna which uses folded dipoles for the half wave elements of the array. The use of folded dipoles results in higher radiation resistance of the array, and a high over-all system performance. Three wire dipoles are used for the elements, and 300-ohm twin-lead is used for the two phasing sections. A recommended assembly for Triplex beams for 28, 21, and 14 MHz is shown in figure 25. The gain of a Triplex beam is about 4.5 db over a dipole.

26-9 Combination End-Fire and Broadside Arrays

Any of the end-fire arrays previously described may be stacked one above the other or placed end to end (side by side) to give

Figure 24

THE TRIPLEX FLAT-TOP BEAM ANTENNA FOR 10, 15, AND 20 METERS



DIMENSIONS

	10M.	15M.	20M	MATERIAL
L	15'8"	21'3"	32'2"	THREE #12 E. WIRES, SPACED 3"
S	5'6"	7'6"	11'	
D	7'2"	10'7"	14'4"	300-OHM RIBBON

greater directivity gain while maintaining a bidirectional characteristic. However, it must be kept in mind that to realize a worthwhile increase in directivity and gain while maintaining a bidirectional pattern the individual arrays must be spaced sufficiently to reduce the mutual impedances to a negligible value.

When two flat-top beams, for instance, are placed one above the other or end to end, a center spacing on the order of one wavelength is required in order to achieve a worthwhile increase in gain, or approximately 3 db.

Thus it is seen that, while maximum gain occurs with two stacked dipoles at a spacing of about 0.7 wavelength and the space directivity gain is approximately 5 db over one element under these conditions; the case of two flat-top or parasitic arrays stacked one above the other is another story. Maximum gain will occur at a greater spacing, and the gain over one array will not appreciably exceed 3 db.

When two broadside curtains are placed one ahead of the other in end-fire relationship, the aggregate mutual impedance between the two curtains is such that considerable spacing is required in order to realize a gain approaching 3 db (the required spacing being a function of the size of the curtains). While it is true that a space-directivity gain of approximately 4 db can be obtained by placing one half-wave dipole an eighth wavelength ahead of another and feeding them 180 degrees out of phase, a gain of less than 1 db is obtained when the same procedure is applied to two large broadside curtains. To obtain a gain of approximately 3 db and retain a bidirectional pattern, a spacing of many wavelengths is required between two large curtains placed one ahead of the other.

A different situation exists, however, when one driven curtain is placed ahead of an identical one and the two are phased so as to give a unidirectional pattern. When a unidirectional pattern is obtained, the gain over one curtain will be approximately 3 db regardless of the spacing. For instance, two large curtains one placed a quarter wavelength ahead of the other may have a space-directivity gain of only 0.5 db over one curtain when the two are driven 180 degrees out of phase to give a bidirectional pattern (the type of pattern obtained with a single curtain). However, if they are driven in phase quadrature (and with equal currents) the gain is approximately 3 db.

The directivity gain of a composite array also can be explained on the basis of the directivity patterns of the component arrays alone, but it entails a rather complicated picture. It is sufficient for the purpose of this discussion to generalize and simplify by saying that the greater the directivity of an end-fire array, the farther an identical array must be spaced from it in broadside relationship to obtain optimum performance; and the greater the directivity of a broadside array, the farther an identical array must be spaced from it in end-fire relationship to obtain optimum performance and retain the bidirectional characteristic.

It is important to note that while a bidirectional end-fire pattern is obtained with two driven dipoles when spaced anything under a half wavelength, and while the proper phase relationship is 180 degrees regardless of the spacing for all spacings not exceeding one half wavelength, the situation is different in the case of two curtains placed in end-fire relationship to give a bidirectional pattern. For maximum gain at zero wave angle, the curtains should be spaced an odd multiple of one-half wavelength and driven so as to be 180 degrees out of phase, or spaced an even multiple of one half wavelength and driven in the same phase. The optimum spacing and phase relationship will depend on the directivity pattern of the individual curtains used alone, and as previously noted the optimum spacing increases with the size and directivity of the component arrays.

A concrete example of a combination broadside and end-fire array is two Lazy H arrays spaced along the direction of maximum radiation by a distance of four wavelengths and fed in phase. The space-directivity gain of such an arrangement is slightly less than 9 db. However, approximately the same gain can be obtained by juxtaposing the two arrays side by side or one over the other in the same plane, so that the two combine to produce, in effect, one broadside curtain of twice the area. It is obvious that in most cases it will be more expedient to increase the area of a broadside array than to resort to a combination of end-fire and broadside directivity.

One exception, of course, is where two curtains are fed in phase quadrature to obtain a unidirectional pattern and space-directivity gain of approximately 3 db with a spacing between curtains as small as one quarter wavelength. Another exception is where very low angle radiation is desired and the maximum pole height is strictly limited. The two aforementioned Lazy H arrays when placed in endfire relationship will have a considerably lower radiation angle than when placed side by side if the array elevation is low, and therefore may *under some conditions* exhibit appreciably practical signal gain.

CHAPTER TWENTY-SEVEN

VHF and UHF Antennas

The *very-high-frequency* or *vhf* frequency range is defined as that range falling between 30 and 300 MHz. The *ultrahigh-frequency* or *uhf range* is defined as falling between 300 and 3000 MHz. This chapter will be devoted to the design and construction of antenna systems for operation on the amateur 50-, 144-, 235-, and 420-MHz bands. Although the basic principles of antenna operation are the same for all frequencies, the shorter physical length of a wave in this frequency range and the differing modes of signal propagation make it possible and expedient to use antenna systems different in design from those used in the range from 3 to 30 MHz.

27-1 Antenna Requirements

Any type of antenna system usable on the lower frequencies *may* be used in the vhf and uhf bands. In fact, simple nondirective half-wave or quarter-wave vertical antennas are very popular for general transmission and reception from all directions, especially for short-range work. But for serious vhf or uhf work the use of some sort of directional antenna array is a necessity. In the first place, when the transmitter power is concentrated into a narrow beam the apparent transmitter power at the receiving station is increased many times. A "billboard" array or a Sterba curtain having a gain of 16 db will make a 25-watt transmitter sound like a kilowatt at the other station. Even a much simpler and smaller three- or four-element parasitic array having a gain of 7 to 10 db will produce a marked improvement in the received signal at the other station.

However, as all vhf and uhf workers know, the most important contribution of a high-gain antenna array is in reception. If a remote station cannot be heard it obviously is impossible to make contact. The limiting factor in vhf and uhf reception is in almost every case the noise generated within the receiver itself. Atmospheric noise is almost nonexistent and ignition interference can almost invariably be reduced to a satisfactory level through the use of an effective noise limiter. Even with a grounded-grid or neutralized triode first stage in the receiver, the noise contribution of the first tuned circuit in the receiver will be relatively large. Hence it is desirable to use an antenna system which will deliver the greatest signal voltage to the first tuned circuit for a given field strength at the receiving location.

Since the field intensity being produced at the receiving location by a remote transmitting station may be assumed to be constant, the receiving antenna which intercepts the greatest amount of wave front (assuming that the polarization and directivity of the receiving antenna is proper, will be the antenna which gives the best received signal-to-noise ratio. An antenna which has two square wavelengths of effective area will pick up twice as much signal power as one which has one square wavelength area, assuming the same general type of antenna and that both are directed at the station being received. Many instances have been reported where a frequency band sounded completely dead with a simple dipole receiving antenna but when the receiver was switched to a three-element or larger array a considerable amount of activity from 80 to 160 miles distant was heard.

Transmission Lines Transmission lines to vhf and uhf antenna systems may be either of the parallel-conductor or coaxial-conductor type. Coaxial line is recommended for short runs and closely spaced open wire line for longer runs. Waveguides may be used under certain conditions for frequencies greater than perhaps 1500 MHz but their dimensions become excessively great for frequencies much below this value. Nonresonant transmission lines will be found to be considerably more efficient on these frequencies than those of the resonant type. It is wise to use the very minimum length of transmission line possible since transmission-line losses at frequencies above about 100 MHz mount very rapidly.

Open lines should preferably be spaced closer than is common for longer wavelengths, since a few inches are an appreciable fraction of a wavelength at 2 meters. Radiation from the line will be greatly reduced if 1-inch or 1½-inch spacing is used, rather than the wider spacing used in the uhf region.

Ordinary TV-type 300-ohm ribbon or the new coaxial *foamflex* line may be used on the 2-meter band for feeder lengths of about 50 feet or less. For longer runs, either the uhf or vhf TV open-wire lines may be used with good over-all efficiency. The vhf line is satisfactory for use on the amateur 420-MHz band.

Antenna Changeover It is recommended that the same antenna be used for transmitting and receiving in the vhf and uhf range. An ever-present problem in this connection, however, is the antenna changeover relay. Reflections at the antenna changeover relay become of increasing importance as the frequency of transmission is increased. When coaxial cable is used as the antenna transmission line, satisfactory coaxial antenna changeover relays with low reflection can be used.

On the 235- and 420-MHz amateur bands, the size of the antenna array becomes quite small, and it is practical to mount two identical antennas side by side. One of these antennas is used for the transmitter, and the other antenna for the receiver. Separate transmission lines are used, and the antenna relay may be eliminated.

Effect of Feed System on Radiation Angle A vertical radiator for general-coverage uhf use should be made either ¼- or ½-wavelength long. Longer vertical antennas do not have their maximum radiation at right angles to the line of the radiator (unless co-phased), and, therefore, are not practical for use where greatest possible radiation parallel to the earth is desired.

Unfortunately, a feed system which is not perfectly balanced and does some radiating, not only robs the antenna itself of that much power, but *distorts the radiation pattern of the antenna.* As a result, the pattern of a vertical radiator may be so altered that the radiation is bent upwards slightly, and the amount of power leaving the antenna parallel to the earth is greatly reduced. A vertical half-wave radiator fed at the bottom by a quarter-wave stub is a good example of this; the slight radiation from the matching section decreases the power radiated parallel to the earth by nearly 10 db. It is important, therefore, to decouple the transmission line from the antenna with a balun or other matching device to keep current from flowing on the outside of the shield of a coaxial line.

Radiator Cross Section In the vhf region, aluminum tubing is commonly used for dipoles since the radiator length is so short that the expense of large-diameter conductor is relatively small, even though tubing of 1-inch cross section is used. With such conductors, the antenna will tune much more broadly, and often a broad resonance characteristic is desirable. This is particularly true when an antenna or array is to be used over an entire amateur band.

It should be kept in mind that with such large-cross-section radiators, the resonant length of the radiator will be somewhat shorter, being only slightly greater than 0.90 of a half wavelength for a dipole when large-diameter pipe is used above 100 MHz.

Insulation The matter of insulation is of prime importance at very-high frequencies. Many insulators that have very low losses as high as 30 MHz show up rather poorly at frequencies above 100 MHz. Even the low-loss ceramics are none too good where the r-f voltage is high. One of the best and most practical insulators for use at this frequency is polystyrene. It has one disadvantage, however, in that it is subject to fracture and to deformation in the presence of heat.

It is common practice to design vhf and uhf antenna systems so that the various radiators are supported only at points of relatively low voltage; the best insulation, obviously, is air. The voltages on properly operated untuned feed lines are not high, and the question of insulation is not quite so important, though insulation still should be of good grade.

Antenna Polarization Commercial broadcasting in the U.S.A. for both frequency modulation and television in the vhf range has been standardized on horizontal polarization. One of the main reasons for this standardization is the fact that ignition interference is reduced through the use of a horizontally polarized receiving antenna. Amateur practice, however, is divided between horizontal and vertical polarization in the vhf and uhf range. Mobile stations are often vertically polarized due to the physical limitations imposed by the automobile antenna installation. Most of the stations doing intermittent or occasional work on these frequencies use a simple ground-plane vertical antenna for both transmission and reception. However, those stations doing serious work and striving for maximum-range contacts on the 50- and 144-MHz bands almost invariably use horizontal polarization.

TABLE 1. WAVELENGTHS AND ANTENNA DIMENSIONS

Frequency (MHZ)	½ Wave-length	¼ Wave-length	½-Wave Dipole	0.2 Wave-length
50.5	122	61	109.5	47"
51.5	120	60	107.5	
52.5	118	59	105.5	
53.5	116	58	103.5	
144	41.0	20.50	38.8	15½"
145	40.75	20.36	38.6	
146	40.5	20.25	38.4	
147	40.25	20.12	38.2	
148	40.0	20.00	38.0	
221	26.5	13.25	25.3	10¾"
222	26.4	13.20	25.2	
223	26.3	13.15	25.1	
224	26.2	13.10	25.0	
420	13.70	6.90	12.90	5⅜"
430	13.65	6.82	12.75	
440	13.50	6.75	12.60	
450	13.35	6.68	12.45	

1—All dimensions in inches.
2—For parasitic director, multiply dipole length by 0.95.
3—For parasitic reflector, multiply dipole length by 1.05.
4—For additional directors, multiply dipole length by 0.94.
5—Use ½" tubing for 50 MHz, ¼" tubing for 144 MHz, ⅛" tubing for 222 and 432 MHz arrays.

Experience has shown that there is a great attenuation in signal strength when using crossed polarization (transmitting antenna with one polarization and receiving antenna with the other) for all normal ground-wave contacts on these bands. When contacts are being made through sporadic-*E* reflection, however, the use of crossed polarization seems to make no discernible difference in signal strength. So the operator of a station doing vhf work (particularly on the 50-MHz band) is faced with a problem: If contacts are to be made with all stations doing work on the same band, provision must be made for operation on both horizontal and vertical polarization. This problem has been solved in many cases through the construction of an antenna array that may be revolved in the plane of polarization in addition to being capable of rotation in the azimuth plane.

An alternate solution to the problem which involves less mechanical construction is simply to install a good ground-plane vertical antenna for all vertically-polarized work, and then to use a multielement horizontally polarized array for DX work.

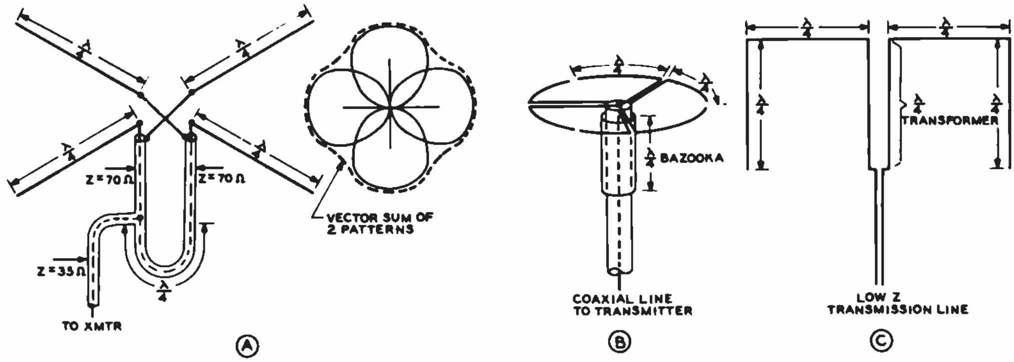


Figure 1

THREE NONDIRECTIONAL HORIZONTALLY POLARIZED ANTENNAS

VHF Antenna Dimensions Listed in Table 1 are representative dimensions for the elements of vhf and uhf antenna arrays of the parasitic or collinear type. Since the length-to-diameter ratio of antennas above 100 MHz or so is somewhat smaller than that of high-frequency arrays and because the arrays are physically smaller, dimensions are generally given in inches, based on the following formula:

$$\text{Dipole length (inches)} = \frac{5600}{f_{\text{MHz}}}$$

The dimensions for small (3, 4, or 5 element) Yagis may be derived from Table 1, based on elements of the listed diameters and using a nominal spacing of 0.2 wavelength. If other element spacings are to be used, the reflector and director elements will have to be readjusted accordingly. Closer reflector driven-element spacing will call for a slightly shorter reflector for optimum gain. Closer director driven-element spacing will call for a slightly longer director for optimum gain. Generally speaking, anything closer to 0.2-wavelength spacing in Yagi arrays tends to reduce the bandwidth, reduce the driven-element impedance, and increase the front-to-back ratio.

Vhf and uhf elements may be made of small-diameter aluminum tubing, or (in the case of the 432-MHz band) ⅛″ diameter aluminum clothesline wire. The parasitic element should not be painted, as this tends to detune the element. A light coat of *Krylon* plastic spray may be used to protect the element against weather.

27-2 Simple Horizontally Polarized Antennas

Antenna systems which do not concentrate radiation at the very low elevation angles are not recommended for vhf and uhf work. It is for this reason that the horizontal dipole and horizontally disposed collinear arrays are generally unsuitable for work on these frequencies. Arrays using broadside or end-fire elements do concentrate radiation at low elevation angles and are recommended for vhf work. Arrays such as the lazy H, Sterba curtain, log-periodic beam, and arrays with parasitically excited elements are recommended for this work. Dimensions for the first two types of arrays may be determined from the data given in the previous chapter, and reference may be made to the *Table of Wavelengths* given in this chapter.

Arrays using vertically stacked horizontal dipoles, such as are used by commercial television and f-m stations, are capable of giving high gain *without* a sharp horizontal radiation pattern. If sets of crossed dipoles, as shown in figure 1A, are fed 90° out of phase the resulting system is called a *turn-*

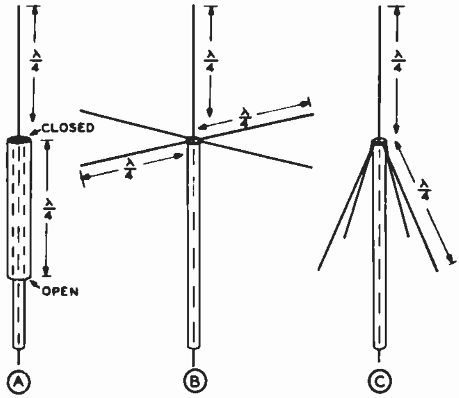


Figure 2

THREE VERTICALLY POLARIZED LOW-ANGLE RADIATORS

Shown at A is the "sleeve" or "hypodermic" type of radiator. At B is shown the ground-plane vertical, and C shows a modification of this antenna system which increases the feed-point impedance to a value such that the system may be fed directly from a coaxial line with no standing waves on the feed line.

stile antenna. The 90° phase difference between sets of dipoles may be obtained by feeding one set of dipoles with a feed line which is one-quarter wave longer than the feed line to the other set of dipoles. The field strength broadside to one of the dipoles is equal to the field from that dipole alone. The field strength at a point at any other angle is equal to the vector sum of the fields from the two dipoles at that angle. A nearly circular horizontal pattern is produced by this antenna.

A second antenna producing a uniform, horizontally polarized pattern is shown in figure 1B. This antenna employs three dipoles bent to form a circle. All dipoles are excited in phase, and are center fed. A bazooka is included in the system to prevent unbalance in the coaxial feed system.

A third nondirectional antenna is shown in figure 1C. This simple antenna is made of two half-wave elements, of which the end quarter wavelength of each is bent back 90 degrees. The pattern from this antenna is very much like that of the turnstile antenna. The field from the two quarter-wave sections that are bent back are additive because they are 180 degrees out of phase and are a half wavelength apart. The advantage of this antenna is the simplicity of its feed system and construction.

27-3 Simple Vertical-Polarized Antennas

For general coverage with a single antenna, a single vertical radiator is commonly employed. A two-wire open transmission line is not suitable for use with this type antenna, and coaxial polyethylene feed line such as RG-8/U is to be recommended. Three practical methods of feeding the radiator with concentric line, with a minimum of current induced in the outside of the line, are shown in figure 2. Antenna A is known as the *sleeve antenna*, the lower half of the radiator being a large piece of pipe up through which the concentric feed line is run. At B is shown the ground-plane vertical, and at C a modification of this latter antenna. In many cases, the antennas of illustrations A and C have a set of quarter-wave radials placed beneath the array to decouple it from the transmission line.

The radiation resistance of the ground-plane vertical is approximately 30 ohms, which is not a standard impedance for coaxial line. To obtain a good match, the first quarter wavelength of feeder may be of 52 ohms surge impedance, and the remainder of the line of approximately 75 ohms impedance. Thus, the first quarter-wave section of line is used as a matching transformer, and a good match is obtained.

In actual practice the antenna would consist of a quarter-wave rod, mounted by means of insulators atop a pole or pipe mast. Elaborate insulation is not required, as the voltage at the lower end of the quarter-wave radiator is very low. Self-supporting rods from 0.25 to 0.28 wavelength are extended out, as shown in the illustration, and connected together. Since the point of connection is effectively at ground potential, no insulation is required; the horizontal rods may be bolted directly to the supporting pole or mast, even if of metal. The coaxial line should be of the low-loss type especially designed for vhf use. The shield connects to the junction of the radials, and the inner

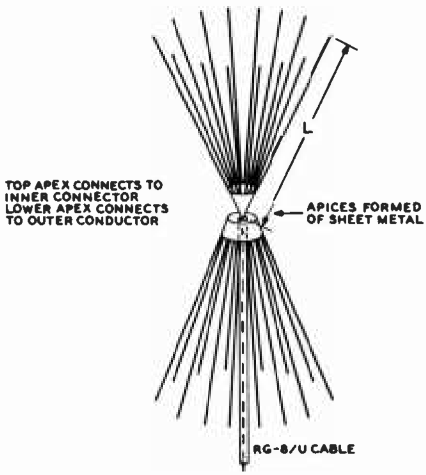


Figure 3

THE DOUBLE SKELETON CONE ANTENNA

A skeleton cone has been substituted for the single element radiator of figure 2C. This greatly increases the bandwidth. If at least 10 elements are used for each skeleton cone and the angle of revolution and element length are optimized, a low SWR can be obtained over a frequency range of at least two octaves. To obtain this order of bandwidth, element length L should be approximately 0.2 wavelength at the lower frequency end of the band, and the angle of revolution optimized for the lowest maximum SWR within the frequency range to be covered. A greater improvement in the impedance-frequency characteristic can be achieved by adding elements than by increasing the diameter of the elements. With only 3 elements per "cone" and a much smaller angle of revolution a low SWR can be obtained over a frequency range of approximately 1.3 to 1.0 when the element lengths are optimized.

conductor to the bottom end of the vertical radiator. An antenna of this type is moderately simple to construct and will give a good account of itself when fed at the lower end of the radiator directly by the 52-ohm RG-8/U coaxial cable. Theoretically the standing-wave ratio will be approximately 1.5-to-1 but in practice this moderate SWR produces no deleterious effects.

The modification shown in figure 2C permits matching to a standard 50- or 70-ohm flexible coaxial cable without a linear transformer. If the lower rods hug the line and supporting mast rather closely, the feed-point impedance is about 70 ohms. If they are bent out to form an angle of about 30° with the support pipe the impedance is about 50 ohms.

The number of radial legs used in a ground-plane antenna of either type has an important effect on the feed-point impedance and on the radiation characteristics of the antenna system. Experiment has shown that three radials is the minimum number that should be used, and that increasing the number of radials above six adds substantially nothing to the effectiveness of the antenna and has no effect on the feed-point impedance. Measurement shows, however, that the radials should be slightly longer than one-quarter wave for best results. A length of 0.26 wavelength has been found to be the optimum value. This means that the radials for a 50-MHz ground-plane vertical antenna should be 62″ in length.

Double Skeleton Cone Antenna The bandwidth of the antenna of figure 2C can be increased considerably by substituting several space-tapered rods for the single radiating element, so that the "radiator" and skirt are similar. If a sufficient number of rods are used in the skeleton cones and the angle of revolution is optimized for the particular type of feed line used, this antenna exhibits a very low SWR over a 2-to-1 frequency range. Such an arangement is illustrated schematically in figure 3.

A Nondirectional Vertical Array Half-wave elements may be stacked in the vertical plane to provide a non-directional pattern with good horizontal gain. An array made up of four half-wave vertical elements is shown in figure 4A. This antenna provides a circular pattern with a gain of about 4.5 db over a vertical dipole. It may be fed with 300-ohm TV-type line. The feed line should be conducted in such a way that the vertical portion of the line is at least one-half wavelength away from the vertical antenna elements. A suitable mechanical assembly is shown in figure 4B for the 144- and 235-MHz amateur bands.

A Stacked Sleeve Antenna for 144 MHz The sleeve antenna makes a good omnidirectional array for 144 MHz in areas

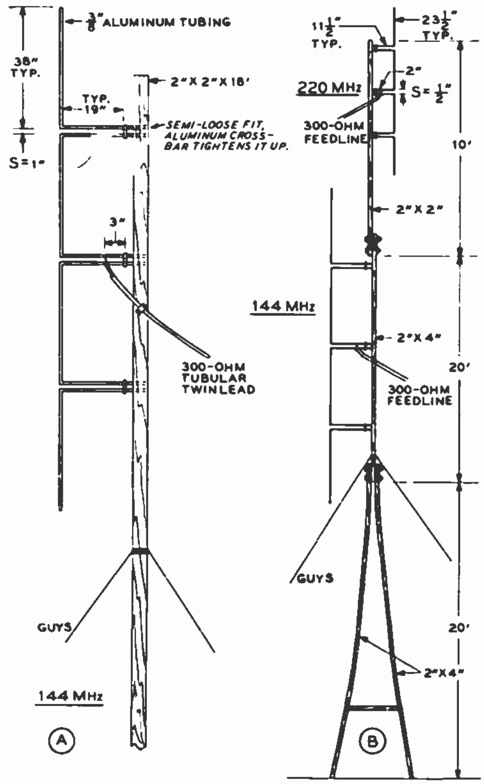


Figure 4

NONDIRECTIONAL ARRAYS FOR 144 AND 235 MHz

On right is shown a two-band installation. For portable use, the whole system may easily be dissembled and carried on a luggage rack atop a car.

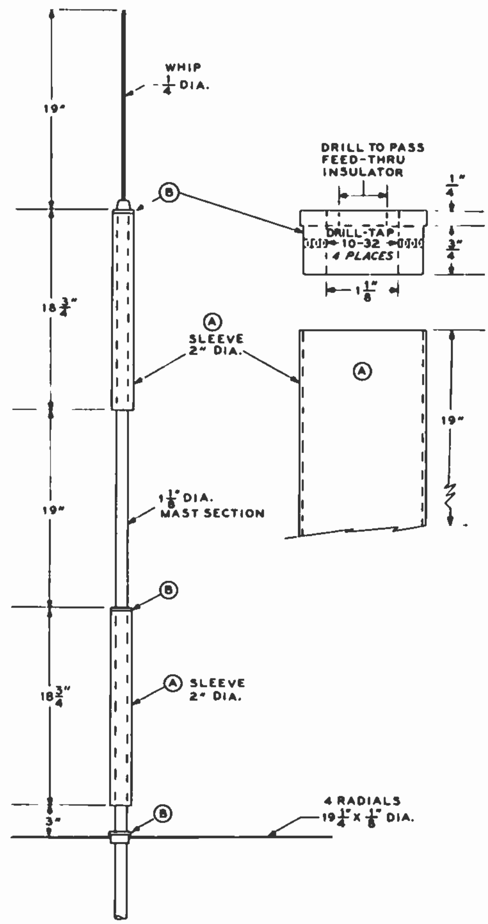


Figure 5

SLEEVE ANTENNA FOR 144 MHz

Stacked dipoles provide nondirectional coverage with low-angle radiation. The top whip is fed by a coaxial line passed up through the mast section and is insulated from remainder of the antenna structure. Lower dipole is composed of mast section and matching skirt which is grounded to the mast at the top. Bottoms of both skirts are free. Radials beneath bottom section impede flow of antenna current on outside of coaxial line.

where vertical polarization is used. A double stack, such as illustrated in figure 5, will provide low-angle radiation and a power gain of about 3 decibels. The array is designed to be fed with a 50-ohm coaxial transmission line.

The antenna is built on an eight-foot length of aluminum TV mast section, 1 1/8" diameter. A quarter-wavelength whip extends from the top of the assembly, and two sleeves are mounted to the mast section below the whip. Both sleeves are electrically connected to the mast at their tops, and the bottom sleeve is shock-excited by the top antenna array, which functions as a simple dipole. Directly below the sleeves are mounted four quarter-wave horizontal radials which decouple the stacked antenna from the outer shield of the coaxial transmission line.

Antenna construction is straightforward and simple. The top of the mast is closed with an aluminum plug (B) having a ceramic feedthrough insulator mounted in it. The vertical whip attaches to the insulator, as does the center conductor of the coaxial feedline. The outer shield of the line is grounded to the mast section at the insulator. The outer sleeve (A) is attached to the mast section by means of machine screws tapped into the aluminum plug.

The lower sleeve is attached to the mast in a similar manner, as shown in the drawing. The radials, made of aluminum clothesline wire are threaded and screwed to an aluminum mounting cylinder (similar to B) which encircles the mast.

Three aluminum fittings (B) are required: one for the top sleeve, one for the lower sleeve, and one for the radials. The top fitting is shown in figure 5. The center one is similar, except that it is drilled to pass the mast section. The fitting for the radials is similar to the center one, except that the ¼-inch lip at the top is omitted.

The length of the fitting is such so that the inner resonant portion of the sleeve is slightly shorter than the outer section. The outer section acts as a portion of the antenna and the inner section acts as a decoupling transformer. The resonant lengths are different for each case, and the length of the fitting makes up the electrical difference.

The sleeves are free at the lower ends, with no connection or support at this point. Care must be taken to make the assembly waterproof, as an accumulation of moisture in the sleeve may detune it. Plugs at the bottom of the sleeves, therefore, are not advised.

The 50-ohm coaxial transmission line runs up the inside of the mast to the top fitting where the outer shield is grounded to the structure by means of a washer placed beneath the feedthrough insulator. The shield is soldered to a lug of the washer, which may be cut from thin brass or copper shim stock.

When fed with a 50-ohm transmission line, the measured SWR across the 144-MHz band is less than 2/1, and better than 1.5/1 at the center frequency of 146 MHz.

27-4 The Discone Antenna

The *Discone* antenna is a vertically polarized omnidirectional radiator which has very broad band characteristics and permits a simple, rugged structure. This antenna presents a substantially uniform feed-point impedance, suitable for direct connection of a coaxial line, over a range of several octaves. Also, the vertical pattern is suitable for ground-wave work over several octaves, the gain varying only slightly over a very wide frequency range.

A Discone antenna suitable for multiband amateur work in the uhf/vhf range is shown schematically in figure 6. The distance (D) should be made approximately equal to a free-space quarter wavelength at the lowest operating frequency. The antenna then will perform well over a frequency range of at least 8 to 1. At certain frequencies within this range the vertical pattern will tend to rise slightly, causing a slight reduction in gain at zero angular elevation, but the reduction is very slight.

Below the frequency at which the slant height of the conical skirt is equal to a free-space quarter wavelength the standing-wave ratio starts to climb, and below a frequency approximately 20 percent lower than this

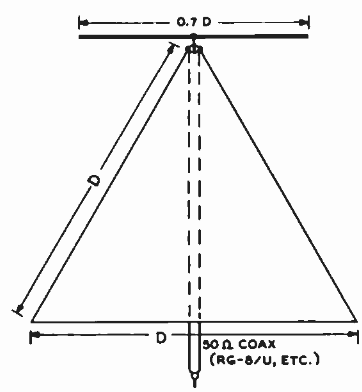


Figure 6

THE DISCONE BROADBAND RADIATOR

This antenna system radiates a vertically polarized wave over a very wide frequency range. The "disc" may be made of solid metal sheet, a group of radials, or wire screen; the "cone" may best be constructed by forming a sheet of thin aluminum. A single antenna may be used for operation on the 50-, 144-, and 220-MHz amateur bands. The dimension D is determined by the lowest frequency to be employed, and is given in figure 7.

the standing-wave ratio climbs very rapidly. This is termed the *cutoff frequency* of the antenna. By making the slant height approximately equal to a free-space quarter wavelength at the lowest frequency employed (refer to figure 7), an SWR of less than 1.5 will be obtained throughout the operating range of the antenna.

The Discone antenna may be considered as a cross between an electromagnetic horn and an inverted ground-plane unipole antenna. It looks to the feed line like a properly terminated high-pass filter.

Construction Details The top disc and the conical skirt may be fabricated either from sheet metal, screen (such as "hardware cloth"), or 12 or more "spine" radials. If screen is used, a supporting framework of rod or tubing will be necessary for mechanical strength except at the higher frequencies. If spines are used, they should be terminated on a stiff ring for mechanical strength, except at the higher frequencies.

The top disc is supported by means of three insulating pillars fastened to the skirt. Either polystyrene or low-loss ceramic is suitable for the purpose. The apex of the conical skirt is grounded to the supporting mast and to the outer conductor of the coaxial line. The line is run down through the supporting mast. An alternative arrangement, one suitable for certain mobile applications, is to fasten the base of the skirt directly to an effective ground plane such as the top of an automobile.

27-5 Helical Beam Antennas

Most vhf and uhf antennas are either vertically polarized or horizontally polarized (plane polarization). However, *circularly* polarized antennas having interesting characteristics which may be useful for certain applications. The installation of such an antenna can effectively solve the problem of horizontal versus vertical polarization.

A circularly polarized wave has its energy divided equally between a vertically polarized component and a horizontally polarized component, the two being 90 degrees out of phase. The circularly polarized wave may be either "left handed" or "right handed," depending on whether the vertically polarized component leads or lags the horizontal component.

A circularly polarized antenna will respond to any plane polarized wave whether horizontally polarized, vertically polarized,

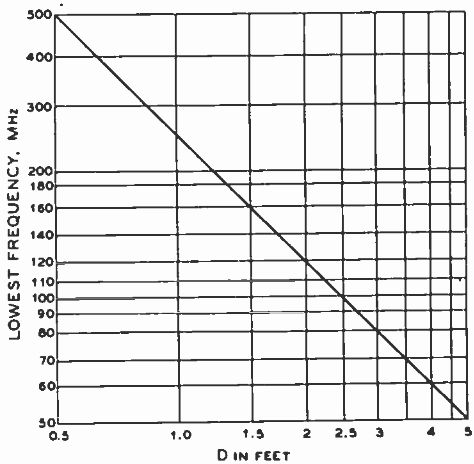


Figure 7

DESIGN CHART FOR THE DISCONE ANTENNA

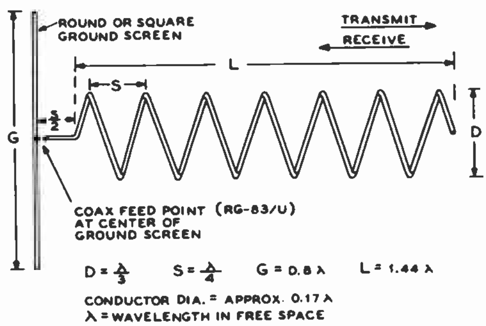


Figure 8

THE HELICAL BEAM ANTENNA

This type of directional antenna system gives excellent performance over a frequency range of 1.7 to 1.8 to 1. Its dimensions are such that it is ordinarily not practical, however, for use as a rotatable array on frequencies below about 100 MHz. The center conductor of the feed line should pass through the ground screen for connection to the feed point. The outer conductor of the coaxial line should be grounded to the ground screen.

or diagonally polarized. Also, a circular polarized wave can be received on a plane polarized antenna, regardless of the polarization of the latter.

When using circularly polarized antennas at *both* ends of the circuit, however, both must be left handed or both must be right handed. This offers some interesting possibilities with regard to reduction of interference. At the time of writing, there has been no standardization of the "twist" for general amateur work.

Perhaps the simplest antenna configuration for a directional beam antenna having circular polarization is the *helical beam* which consists simply of a helix working against a ground plane and fed with coaxial line. In the uhf and the upper vhf range the physical dimensions are sufficiently small to permit construction of a rotatable structure without much difficulty.

When the dimensions are optimized, the characteristics of the helical beam antenna are such as to qualify it as a broadband antenna. An optimized helical beam shows little variation in the pattern of the main lobe and a fairly uniform feed-point impedance averaging approximately 125 ohms over a frequency range of as much as 1.7 to 1. The direction of "electrical twist" (right or left handed) depends on the direction in which the helix is wound.

A six-turn helical beam is shown schematically in figure 8. The dimensions shown will give good performance over a frequency range of plus or minus 20 percent of the design frequency. This means that the dimensions are not especially critical when the array is to be used at a single frequency or over a narrow band of frequencies, such as an amateur band. At the design frequency the beam width is about 50 degrees and the power gain about 12 db, referred to a nondirectional circularly polarized antenna.

The Ground Screen For the frequency range 100 to 500 MHz a suitable ground screen can be made from "chicken wire" poultry netting of 1-inch mesh, fastened to a round or square frame of either metal or wood. The netting should be of the type that is galvanized *after* weaving. A small, sheet-metal ground plate of diameter equal to approximately D/2 should be centered on the screen and soldered to it. Tin, galvanized iron, or sheet copper is suitable. The outer conductor of the RG-63/U (125-ohm) coax is connected to this plate, and the inner conductor contacts the helix through a hole in the center of the plate. The end of the coax should be taped with *Scotch* electrical tape to keep water out.

The Helix It should be noted that the beam proper consists of six full turns. The start of the helix is spaced a distance of S/2 from the ground screen, and the conductor goes directly from the center of the ground screen to the start of the helix.

Aluminum tubing in the 2014 alloy grade is suitable for the helix. Alternatively, lengths of the relatively soft aluminum electrical conduit may be used. In the vhf range it will be necessary to support the helix on either two or four wooden longerons in order to achieve sufficient strength. The longerons should be of the smallest cross section which provides sufficient rigid-

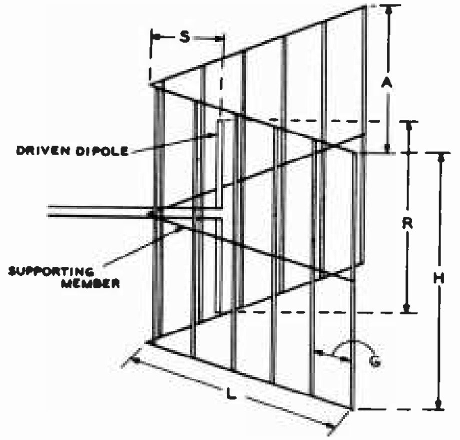


Figure 9

CONSTRUCTION OF THE CORNER REFLECTOR ANTENNA

Such an antenna is capable of giving high gain with a minimum of complexity in the radiating system. It may be used either with horizontal or vertical polarization. Design data for the antenna is given in the Corner-Reflector Design Table.

ity, and should be given several coats of varnish. The ground plane butts against the longerons and the whole assembly is supported from the balance point if it is to be rotated.

Aluminum tubing in the larger diameters ordinarily is not readily available in lengths greater than 12 feet. In this case several lengths can be spliced by means of short telescoping sections and sheet-metal screws.

The tubing is closewound on a drum and then spaced to give the specified pitch. Note that the length of one complete turn when spaced is somewhat greater than the circumference of a circle having the diameter D.

Broad-Band 144- to 225-MHz Helical Beam A highly useful vhf helical beam which will receive signals with good gain over the complete frequency range from 144 through 255 MHz may be constructed by using the following dimensions (180 MHz design center):

D .... 22 in.
S .... 16½ in.
G .... 53 in.
Tubing o.d. .... 1 in.

The D and S dimensions are to the center of the tubing. These dimensions must be held rather closely, since the range from 144 through 255 MHz represents just about the practical limit of coverage of this type of antenna system.

High-Band TV Coverage Note that an array constructed with the above dimensions will give unusually good high-band TV reception in addition to covering the 144- and 220-MHz amateur bands and the taxi and police services.

On the 144-MHz band the beam width is approximately 60 degrees to the half-power points, while the power gain is approximately 11 db over a nondirectional circularly polarized antenna. For high-band TV coverage the gain will be 12 to 14 db, with a beam width of about 50 degrees, and on the 220-MHz amateur band the beam width will be about 40 degrees with a power gain of approximately 15 db.

The antenna system will receive vertically polarized or horizontally polarized signals with equal gain over its entire frequency range. Conversely, it will transmit signals over the same range, which then can be received with equal strength on either horizontally polarized or vertically polarized receiving antennas. The standing-wave ratio will be very low over the complete frequency range if RG-63/U coaxial feed line is used.

27-6 The Corner-Reflector and Horn-Type Antennas

The corner-reflector antenna is a good directional radiator for the vhf and uhf region. The antenna may be used with the radiating element vertical, in which case the directivity is in the horizontal or azimuth plane, or the system may be used with the driven element horizontal, in which case the radiation is horizontally polarized, and most of the directivity is in the vertical plane. With the antenna used as a horizontally polarized radiating system the array is a very good low-angle beam array although the nose of the horizontal pattern is still quite sharp. When the radiator is oriented vertically the corner reflector operates very satisfactorily as a direction-finding antenna.

TABLE 2. CORNER-REFLECTOR DESIGN DATA

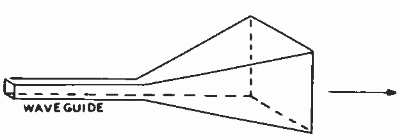
Corner Angle	Freq. Band, MHz	R	S	H	A	L	G	Feed Imped.	Approx. Gain, db
90	50	110″	82″	140″	200″	230″	18″	72	10
60	50	110″	115″	140″	230″	230″	18″	70	12
60	144	38″	40″	48″	100″	100″	5″	70	12
60	220	24.5″	25″	30″	72″	72″	3″	70	12
60	420	13″	14″	18″	36″	36″	screen	74	12

NOTE: Refer to figure 9 for construction of corner-reflector antenna.

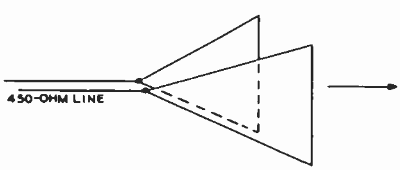
Design data for the corner-reflector antenna is given in figure 9 and in Table 2, *Corner-Reflector Design Data.* The planes which make up the reflecting corner may be made of solid sheets of copper or aluminum for the uhf bands, although spaced wires with the ends soldered together at top and bottom may be used as the reflector on the lower frequencies. Copper screen may also be used for the reflecting planes.

The values of spacing given in the corner-reflector chart have been chosen such that the center impedance of the driven element would be approximately 70 ohms. This means that the element may be fed directly with 70-ohm coaxial line, or a quarter-wave matching transformer such as a Q-section may be used to provide an impedance match between the center impedance of the element and a 460-ohm line constructed of No. 12 wire spaced 2 inches.

In many uhf antenna systems, waveguide transmission lines are terminated by *pyramidal horn* antennas. These horn antennas (figure 10A) will transmit and receive either horizontally or vertically polarized waves. The use of waveguides at 144 and 235 MHz, however, is out of the question because of the relatively large dimensions needed for a waveguide operating at these low frequencies.



(A) UHF HORN ANTENNA

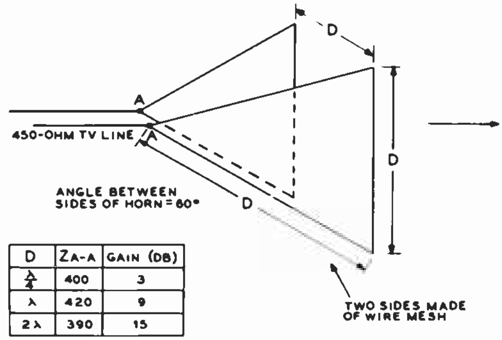


(B) VHF HORIZONTALLY POLARIZED HORN

Figure 10

TWO TYPES OF HORN ANTENNAS

The "two-sided horn" of illustration B may be fed by means of an open-wire transmission line.



D	Z_{A-A}	GAIN (DB)
$\lambda/4$	400	3
λ	420	9
2λ	390	15

Figure 11

THE 60° HORN ANTENNA FOR USE ON FREQUENCIES ABOVE 144 MHz

A modified type of horn antenna may still be used on these frequencies, since only one particular plane of polarization is of interest to the amateur. In this case, the horn antenna can be simplified to two triangular sides of the pyramidal horn. When these two sides are insulated from each other, direct excitation at the apex of the horn by a two-wire transmission line is possible.

In a normal pyramidal horn, all four triangular sides are covered with conducting material, but when horizontal polarization alone is of interest (as in amateur work) only the *vertical* areas of the horn need be used. If vertical polarization is required, only the *horizontal* areas of the horn are employed. In either case, the system is unidirectional, away from the apex of the horn. A typical horn of this type is shown in figure 10B. The two metallic sides of the horn are insulated from each other, and the sides of the horn are made of small mesh "chicken wire" or copper window screening.

A pyramidal horn is essentially a high-pass device whose low-frequency cutoff is reached when a side of the horn is ½ wavelength. It will work up to infinitely high frequencies, the gain of the horn increasing by 6 db every time the operating frequency is doubled. The power gain of such a horn compared to a half-wave dipole at frequencies higher than cutoff is:

$$\text{Power gain (db)} = \frac{8.4\,A^2}{\lambda^2}$$

where A is the frontal area of the mouth of the horn. For the 60-degree horn shown in figure 8B the formula simplifies to:

Power gain (db) $= 8.4\ D^2$, when D is expressed in terms of wavelength.

When D is equal to one wavelength, the power gain of the horn is approximately 9 db. The gain and feed-point impedance of the 60-degree horn are shown in figure 11. A 450-ohm open-wire TV-type line may be used to feed the horn.

27-7 VHF Horizontal Rhombic Antenna

For vhf transmission and reception in a fixed direction, a horizontal *rhombic* permits 10 to 16 db gain with a simpler construction than does a phased dipole array, and has the further advantage of being useful over a wide frequency range.

Except at the upper end of the vhf range a rhombic array having a worthwhile gain is too large to be rotated. However, in locations 75 to 150 miles from a large metropolitan area a rhombic array is ideally suited for working into the city on extended (horizontally polarized) ground wave while at the same time making an ideal antenna for TV reception.

The useful frequency range of a vhf rhombic array is about 2 to 1, or about plus 40% and minus 30% from the *design frequency*. This coverage is somewhat less than that of a high-frequency rhombic used for sky-wave communication. For ground-wave transmission or reception the only effective vertical angle is that of the horizon, and a frequency range greater than 2 to 1 cannot be covered with a rhombic array without an excessive change in the vertical angle of maximum radiation or response.

The dimensions of a vhf rhombic array are determined from the design frequency and figure 12, which shows the proper *tilt angle* (see figure 13) for a given leg length. The gain of a rhombic array increases with leg length. There is not much point in constructing a vhf rhombic array with legs shorter than about 4 wavelengths, and the beam width begins to become excessively sharp for leg lengths greater than about 8 wavelengths. A leg length of 6 wavelengths is a good compromise between beam width and gain.

The tilt angle given in figure 12 is based on a wave angle of zero degrees. For leg

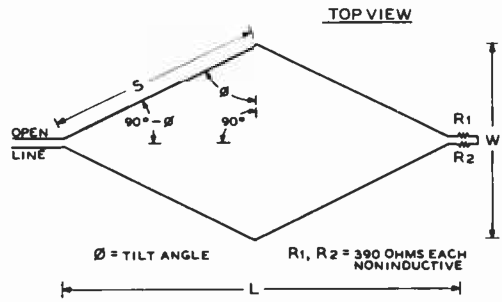


Figure 13

VHF RHOMBIC ANTENNA CONSTRUCTION

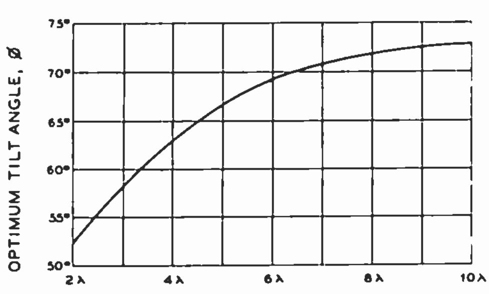


Figure 12

VHF RHOMBIC ANTENNA DESIGN CHART

The optimum tilt angle (see figure 13) for "zero-angle" radiation depends on the length of the sides.

	6 METERS AND LOW-BAND TV	2 METERS, HIGH-BAND TV, AND 1¼ METERS
S (side)	90'	32'
L (length)	166' 10"	59' 4"
W (width)	67' 4"	23' 11"
S = 6 wavelengths at design frequency Tilt angle = 68°		

TABLE 3.

DIMENSIONS FOR TWO DUAL-PURPOSE RHOMBIC ARRAYS

lengths of 4 wavelengths or longer, it will be necessary to elongate the array a few percent (pulling in the sides slightly) if the horizon elevation exceeds about 3 degrees.

Table 3 gives dimensions for two dual purpose rhombic arrays. One covers the 6-meter amateur band and the "low" television band. The other covers the 2-meter amateur band, the "high" television band, and the 1¼-meter amateur band. The gain is approximately 12 db over a matched half wave dipole and the beam width is about 6 degrees.

The Feed Line The recommended feed line is an open-wire line having a surge impedance between 450 and 600 ohms. With such a line the SWR will be less than 2 to 1. A line with two-inch spacing is suitable for frequencies below 100 MHz, but one-inch spacing is recommended for higher frequencies.

The Termination If the array is to be used only for reception, a suitable termination consists of two 390-ohm carbon resistors in series. If 2-watt resistors are employed, this termination also is suitable for transmitter outputs of 10 watts or less. For higher powers, however, resistors having greater dissipation with negligible reactance in the upper vhf range are not readily available.

For powers up to several hundred watts a suitable termination consists of a "lossy" line consisting of stainless-steel wire (corresponding to No. 24 or 26 gauge) spaced 2 inches, which in turn is terminated by two 390-ohm 2-watt carbon resistors. The dissipative line should be at least 6 wavelengths long.

27-8 The Log-Periodic Antenna

Frequency-independent antennas, of which the *Log-periodic* array is an example, are structures that have the same performance at different frequencies by virtue of the fact that the array is self-scaling and has no dimensions that are frequency sensitive. A basic self-scaling structure (shown in figure 14) is described by angles alone, with no characteristic length. Practical structures of this type are finite in size, thus limiting the frequency-independent behavior. Variations of this basic design may take the form of toothed structures, such as illustrated.

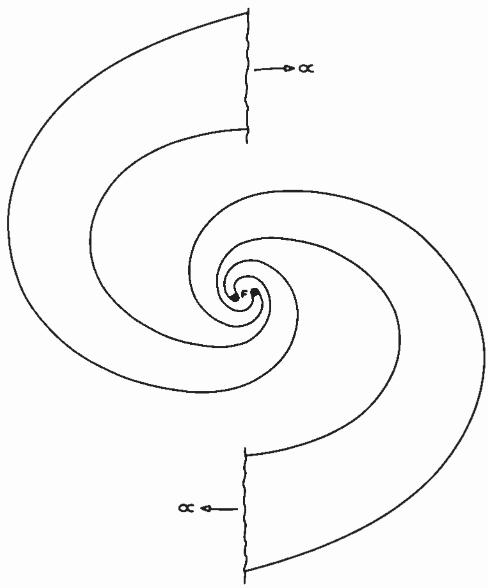
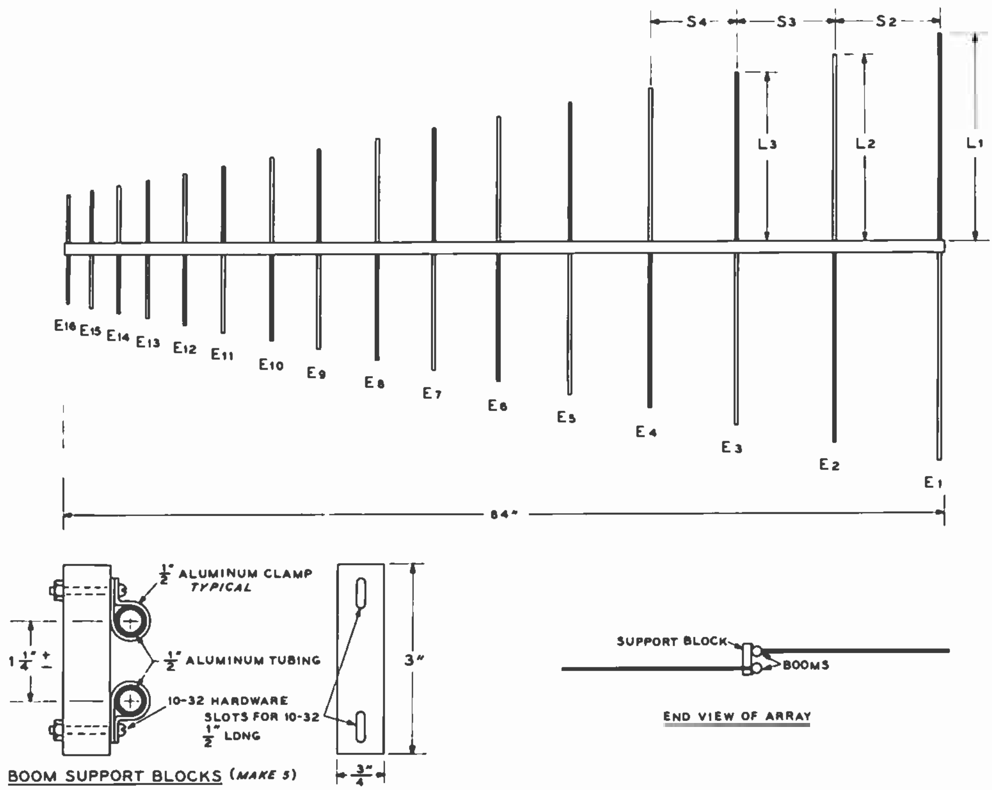


Figure 14

SPIRAL ANTENNA STRUCTURE

This equiangular spiral antenna structure serves as a frequency independent antenna as its shape is entirely specified by angles. The shape of the antenna, when expressed in terms of operating wavelength, is the same for any frequency. The structure is fed at the center (point F) and the arm length is infinite.

An outgrowth of this form of wideband antenna is the *log-periodic dipole array* (figure 15) which is well suited to vhf and uhf work. This interesting antenna is made up of dipole elements whose lengths are determined by the angle they subtend from the apex point, and whose distance from the apex is such as to provide the log-periodic behavior. The dipoles are fed at the center from a parallel-wire line in such fashion that successive dipoles come out from the line in opposite directions, equivalent to a 180° phase shift between elements. A broadband log-periodic structure is thus formed, with most of the radiation coming from



L-P YAGI DIMENSIONS

ELEMENT (L)	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
LENGTH	19½	17½	16½	15	13	11¾	10¾	9¾	8¾	8	7¼	6½	6	5¾	4¾	4½
SPACING (S)	—	10	9½	8¼	7¾	6¾	6¼	5½	5½	4½	4½	3¾	3½	2¾	2½	2

Figure 15

LOG-PERIODIC ANTENNA FOR 140 TO 450 MHz

Vhf log-periodic dipole array is built on double-boom structure made of two lengths of aluminum tubing spaced by insulated support blocks. Elements coded black are attached to the top boom and elements coded white are attached to the lower boom. The coaxial transmission line is inserted in the rear of one boom and passed through the boom, which acts as a balancing device. Center conductor is attached to opposite boom, and shield is attached to balancing boom.

those dipole elements in the vicinity of a half-wavelength long. The bandwidth of the structure is thus limited by the length of the longest and shortest elements, which must be approximately a half-wavelength long at the extreme frequency limits of the antenna array. Gain and bandwidth of the log-periodic antenna thus bear a definite relationship to the included angle of the structure and the length.

An easily constructed log-periodic antenna is the *log-periodic dipole array,* a two-dimensional structure made up of a series of dipoles, fed at the center in such a way that adjacent dipoles are out of phase. The array is fed at the apex and the elements are excited from a parallel-wire transmission line which, if properly designed, may serve as the support structure for the dipoles. The dipole array, in effect, is a balanced transmission line with elements fed from each line, each set of elements reversed in feed polarity. The limiting structure, is a two-element array, and amateur versions of this device are often termed the "ZL-Special" antenna.

The balanced log-periodic dipole structure may be fed with an unbalanced coaxial line by using the support structure as a balun, feeding the coaxial line back from the feed-point through the structure toward the rear.

A L-P Dipole Array for 140–450 MHz A practical L-P dipole array for the vhf spectrum is shown in figure 15. The antenna has a power gain over a dipole of about 7 decibels and may be fed with a 50-ohm coaxial transmission line. The maximum SWR on the transmission line, after adjustment of the boom spacing is better than 2.5/1 over the entire range. The L-P array is built on a twin boom made of ½-inch diameter, heavy-wall aluminum tubing. Two lengths of material are clamped together to form a low-impedance transmission line 86″ long. The clamps may be made of hard wood, or other good insulating material. An impedance match between the array and the transmission line is effected by varying the spacing of the boom, which changes the impedance of the transmission line created by the proximity of the booms to each other.

Alternate halves of successive dipole elements are fastened to a boom section by threading the element, and affixing it to a clamp, as shown in the illustration. Element spacings are measured from the rear of the array and are rounded off to the nearest quarter inch.

When the array is completed, all elements lie in the same plane, with successive elements off center from the supporting structure by virtue of the alternate feed system employed. Boom spacing should be set as shown in the drawing, and later adjusted for minimum SWR on the coaxial transmission line at the various frequencies of interest.

The coaxial line is passed through one boom from the rear and connection to both booms is made at the nose of the array. The outer braid of the line is connected to the boom through which the line passes, and the center conductor connects to the opposite boom. Type-N coaxial connectors are recommended for use in this frequency region.

A L-P Yagi for 50 MHz A yagi antenna consists of a driven element plus parasitic elements to increase the gain and directivity of the radiation pattern over that of a dipole. The number of parasitic elements, their length and spacing with respect to the driven element determine the characteristics of the parasitic yagi antenna. As gain and directivity increase, bandwidth decreases, limiting the ultimate usefulness of this antenna over a complete amateur band, especially at 10 meters and above. To increase the bandwidth of the array, the log-periodic principle used for broadband antennas may be applied to the parasitic beam. The log-periodic yagi array consists of log-periodic elements, interpersed with parasitic reflectors and directors to form individual cells, differing in size by a geometric constant. The driven element in each cell is fed by a common balanced transmission line.

A variation of the log-periodic principle is used in the parasitic antenna described in this section. This L-P yagi antenna is composed of a five element log-periodic section designed to cover the 50- to 52-MHz range and is used in conjunction with three parasitic director elements mounted in front of the log-periodic section. A top view of the antenna is shown in figure 16. The antenna exhibits about 12 decibels forward gain and compares nearly identically with an 8-element yagi mounted on a 30-foot boom. The over-all length of the L-P yagi is only about 18½ feet and it provides improved bandwidth performance and smaller size than the comparable yagi array.

This antenna configuration was designed and developed by the Swan Antenna Co., 646 No. Union St., Stockton, Calif. and is manufactured by that company for amateur and television use. Thanks is given to Mr. Oliver Swan for permission to publish this description of this unusual antenna.

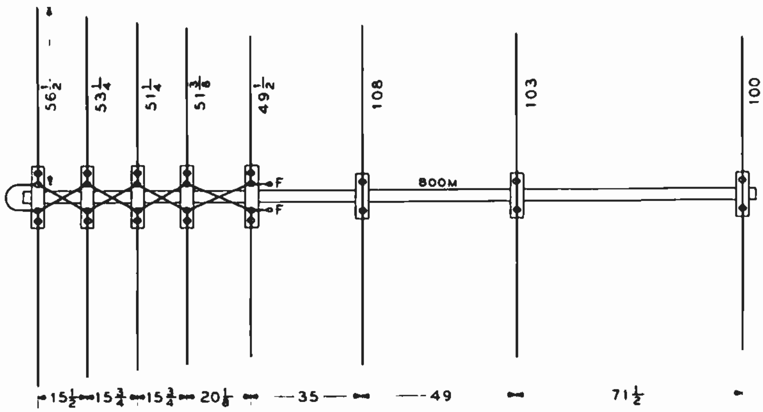


Figure 16

L-P YAGI ANTENNA FOR SIX METERS

This design combines bandwidth of log-periodic structure with gain of yagi antenna. L-P yagi may be built on 1½-inch diameter boom, about 19 feet long. L-P elements are insulated from boom by mounting on insulating blocks. Yagi elements are grounded to boom at their center point. The antenna is fed with a balanced 70-ohm ribbon line at the feedpoint and the L-P transmission line is made up of No. 8 aluminum clothesline wire, criss-cross connected between the elements. Rear element is shorted with six-inch loop of aluminum wire. The spacing between the inner tips of the L-P elements is 3½ inches.

27-9 VHF Yagi Beam Antennas

The multielement rotary beam is undoubtedly the most popular type of vhf antenna in use. In general, the design, assembly and tuning of these antennas follows a pattern similar to that used for the larger rotary arrays used on the lower-frequency amateur bands. The characteristics of the latter antennas are discussed in the next chapter of this Handbook, and the information contained in that chapter applies in general to the vhf beam antennas discussed herewith.

Element Lengths Optimum length for parasitic elements in vhf arrays is a function of element spacing and the diameter of the element. To hold a satisfactory length/diameter ratio, the diameter of the element must decrease as the frequency of operation is raised. At very-high frequencies, element length is so short that the diameter of a self-supporting element becomes a large fraction of the length. Short, large-diameter elements have low Q and are not practical in parasitic arrays. Thus the yagi array becomes critical in adjustment and marginal in operation in the upper reaches of the vhf spectrum. Yagi antennas can be made to work at 432 MHz and higher, but their adjustment is tedious, and preference is given to broadside arrays having relatively large spacings between elements and high impedance. The yagi antenna, however, remains "the antenna to beat" for the 50-, 144-, and 220-MHz amateur bands.

The yagi antennas shown in this section are of all-metal construction with the elements directly grounded to the boom. Either a gamma-match system, T match, or folded-dipole element may be used on the arrays. For short lengths of transmission line, 50-ohm low-loss coaxial cable is recommended for use with a gamma match, or with folded dipole or T match and a coaxial balun. Longer line lengths should be made up of 300-ohm TV-type "ribbon" line or open-wire TV-type transmission line. Care should be taken to keep the ribbon or open-wire lines clear of nearby metallic objects.

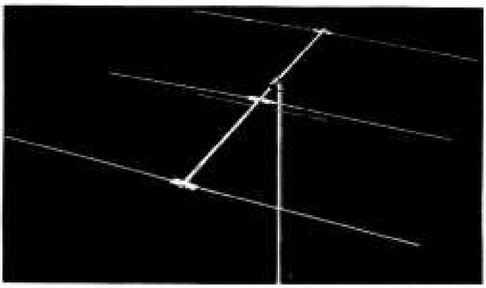


Figure 17

THREE-ELEMENT YAGI BEAM FOR SIX METERS

This all-aluminum array is a popular six-meter antenna. Available in kit form (Hy-Gain), it also may easily be constructed from available aluminum tubing. Elements are clamped to the boom and either a T match, Gamma match, or split-driven-element feed system used. T match with half-wave coaxial balun is recommended system for ease in adjustment. Brass or aluminum hardware should be employed to prevent corrosion of elements due to weather.

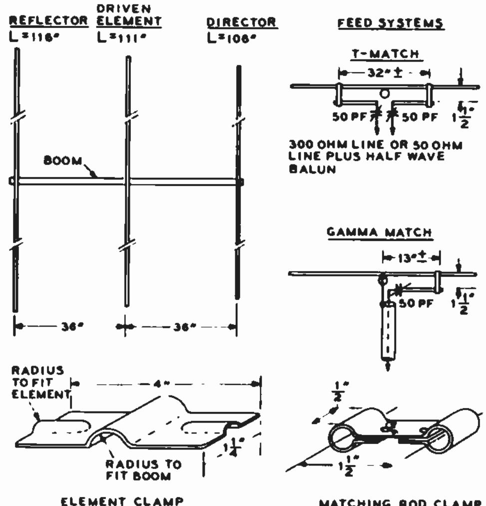


Figure 18

SIX-METER BEAM ASSEMBLY

Element clamps are fabricated from soft aluminum strip. All joints should be cleaned and covered with Penetrox paste to prevent corrosion. Elements may be made of sections of telescoping tubing. Diameters between one inch and one-half inch are recommended.

Yagi Beams for 6 and 2 All-aluminum beam antennas are easy to construct for the 6- and 2-meter amateur bands. The three-element array is very popular for general 6-meter operation, and up to ten elements are often used for DX work on this band. The four-element array is often used on 2 meters, either horizontally or vertically polarized, and arrays having as many as twelve to fifteen elements are used for meteor-scatter and over-horizon work on 144 MHz.

Shown in figures 17 and 18 is a simple three-element array for the 6-meter band. The design frequency is 50.5 MHz, and the beam is capable of operation over the 50- to 51-MHz frequency span. The antenna may be fed from a 50-ohm coaxial line with a half-wave balun and T match as shown in the illustration. The supporting boom is made of a length of 1⅛-inch diameter aluminum TV mast section, and the elements are made of ½-inch diameter aluminum tubing. The elements are mounted in position by drilling the boom to pass the element and then clamping the joint as shown in the drawing.

The T-match system must be properly resonated at the center frequency of antenna operation. To do this, the antenna is temporarily mounted atop a step ladder, in the clear, and fed with a few watts of power from the station transmitter. An SWR meter or reflectometer is placed in the line near the antenna and the length of the T sections and the series capacitors are adjusted to provide the lowest value of SWR on the transmission line. The capacitors are varied in unison to preserve the symmetry of balance. The capacitors should be enclosed in a weatherproof box and mounted at the center of the T section.

A four-element array for the 2-meter band is shown in figures 19 and 20. Dimensions are given for a center frequency of 146 MHz. The antenna provides a power gain of about 9 decibels over a dipole and is capable of good operation over the complete 2-meter band. For optimum operation at the low end of the band, all element lengths should be increased by one-half inch.

Antenna construction is similar to the 6-meter array in that an aluminum section of tubing is used for the boom and the elements are passed through holes drilled in the boom. One-quarter inch aluminum tubing is used for the elements. The T match

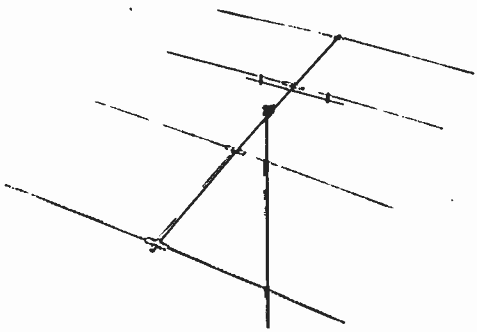


Figure 19

FOUR-ELEMENT YAGI BEAM FOR TWO METERS

Light aluminum is employed for easy-to-build two meter beam. Reynolds "Do It Yourself" aluminum, available at many hardware and building supply stores may be used. Construction is similar to six-meter array. If boom diameter is about one inch, the boom may be drilled for the elements, which are then held in place by a sheet-metal screw through boom and element.

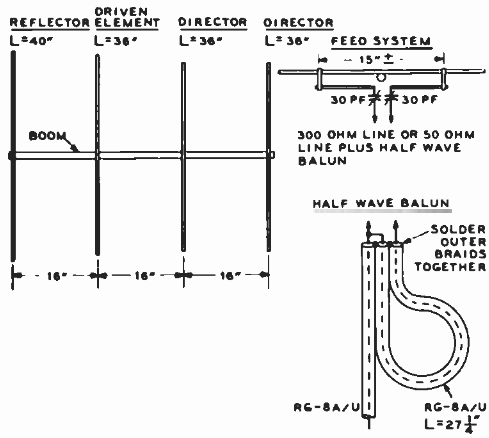


Figure 20

TWO-METER BEAM ASSEMBLY

and coaxial balun are used to match the antenna to a 50-ohm coaxial transmission line.

An 8-Element "Tiltable" Yagi for 144 MHz Two four-element beams may be stacked to double the power gain and to sharpen the pattern of a single beam. Figure 21 illustrates an 8-element rotary beam for 144-MHz use. This array can be tilted to obtain either horizontal or vertical polarization. It is necessary that the transmitting and receiving stations use the same polarization for the ground-wave signal propagation which is characteristic of this frequency range. Although polarization has been loosely standardized in various areas of the country, exceptions are frequent enough so that it is desirable that the polarization of antenna radiation be easily changeable from horizontal to vertical.

The antenna illustrated has shown a signal gain of about 11 db, representing a power gain of about 13. Although the signal gain of the antenna is the same whether it is oriented for vertical or horizontal polarization, the horizontal beam width is smaller when the antenna is oriented for vertical polarization. Conversely, the vertical pattern is sharper when the antenna system is oriented for horizontal polarization.

The changeover from one polarization to the other is accomplished simply by pulling on the appropriate cord. Hence, the operation is based on the offset head sketched in figure 21. Although a wood mast has been used, the same system may be used with a pipe mast.

The 40-inch lengths of RG-59/U cable (electrical ¾-wavelength) running from the center of each folded dipole driven element to the coaxial T-junction allow enough slack to permit free movement of the main boom when changing polarity. Type RG-8/U cable is run from the T-junction to the operating position. Measured standing-wave ratio was less than 2:1 over the 144- to 148-MHz band, with the lengths and spacing given in figure 21.

Construction of the Array Most of the constructional aspects of the antenna array are self-evident from figure 21. However, the pointers given in the following paragraphs will be of assistance to those wishing to reproduce the array.

The drilling of holes for the small elements should be done carefully on accurately marked centers. A small angular error in the drilling of these holes will result in a

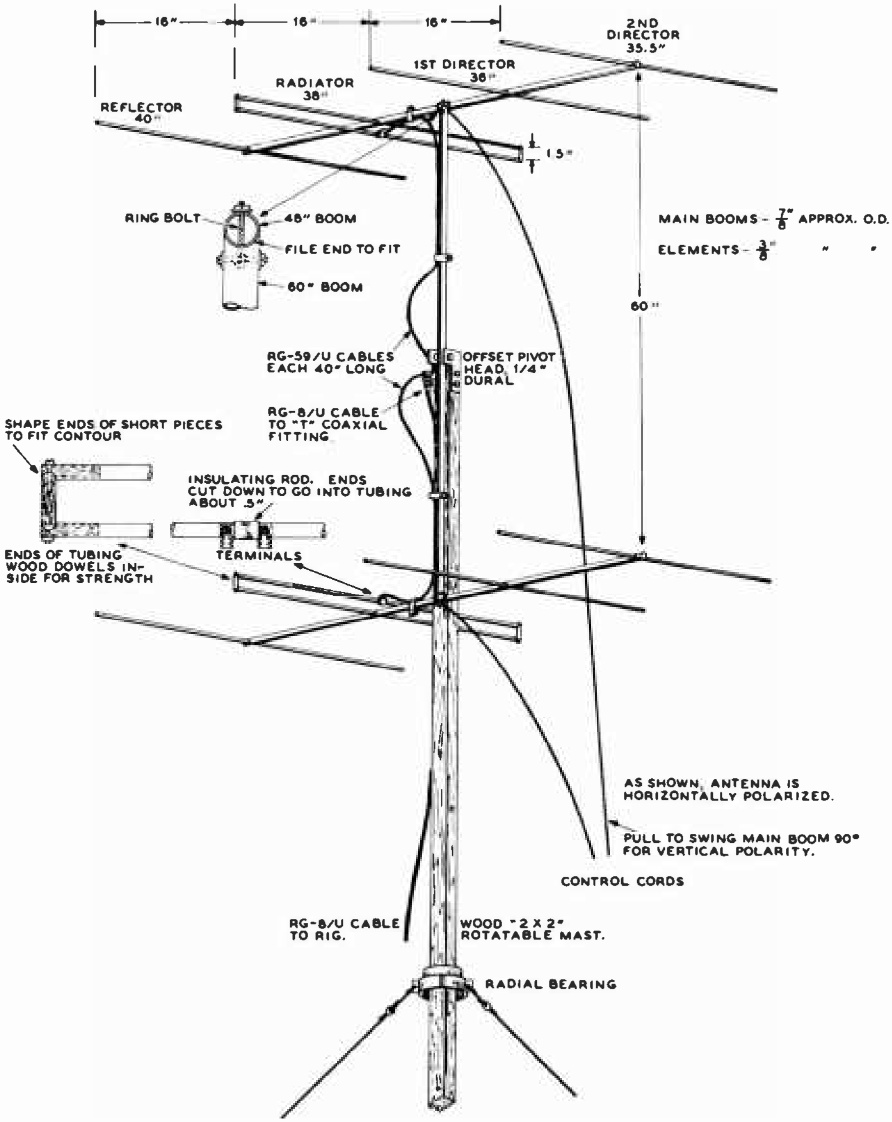
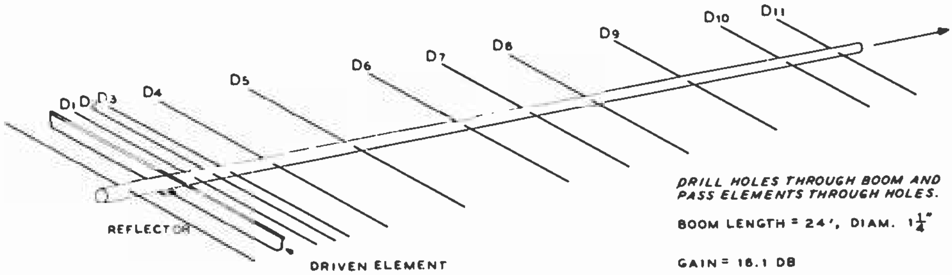


Figure 21
CONSTRUCTION DRAWING OF AN EIGHT-ELEMENT TILTABLE 144-MHz ARRAY

considerable misalignment of the elements after the array is assembled. The same consideration is true of the filing out of the rounded notches in the ends of the main boom for the fitting of the two-antenna booms.

Short lengths of wood dowel are used freely in the construction of the array. The ends of the small elements are plugged with an inch or so of dowel, and the ends of the antenna booms are similarly treated with larger discs pressed into place.

The ends of the folded dipoles are made in the following manner: Drive a length of dowel into the short connecting lengths of aluminum tubing. Then drill down the cen-



ELEMENT DIMENSIONS, 2-METER BAND

ELEMENT (DIAM. 1/8")	LENGTH				SPACING FROM DIPOLE
	144 MHz	145 MHz	146 MHz	147 MHz	
REFLECTOR	41"	40 3/4"	40 7/16"	40 3/16"	19"
DIRECTORS	36 3/4"	36 1/2"	36 3/8"	36 3/16"	D1 = 7"
DRIVEN ELEMENT					D2 = 14.5"
					D3 = 22"
					D4 = 38"
					D5 = 70"
					D6 = 102"
					D7 = 134"
					D8 = 166"
					D9 = 198"
					D10 = 230"
					D11 = 242"

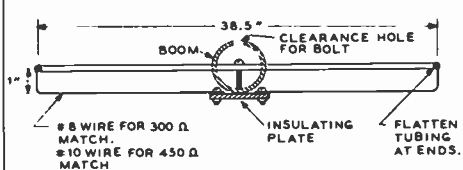


Figure 22

DESIGN DIMENSIONS FOR A 2-METER LONG YAGI ANTENNA

ter of the dowel with a clearance hole for the connecting screw. Then shape the ends of the connecting pieces to fit the sides of the element ends. After assembly the junctions may be dressed with a file and sandpaper until a smooth fit is obtained.

The mast used for supporting the array is a 30-foot spliced 2 by 2. A large discarded ball bearing is used as the radial load bearing and guy-wire termination. Enough of the upper-mast corners were removed with a drawknife to permit sliding the ball bearing down about 9 feet from the top of the mast. The bearing then was encircled by an assembly of three pieces of dural ribbon to form a clamp, with ears for tightening screws and attachment of the guy wires. The bearing then was greased and covered with a piece of auto inner tube to serve as protection from the weather. Another junk-box bearing was used at the bottom of the mast as a thrust bearing.

The main booms were made from ¾-inch aluminum electrical conduit. Any size of small tubing will serve for making the elements. Note that the main boom is mounted at the balance center and not necessarily at the physical center.

In connecting the phasing sections between the T-junction and the centers of the folded dipoles, it is important that the center conductors of the phasing sections be connected to the same side of the driven elements of the antennas. In other words, when the antenna is oriented for horizontal polarization and the center of the coaxial section goes to the left side of the top antenna, the center conductor of the other coaxial phasing section should go to the left side of the bottom antenna.

Long Yagi Antennas For a given power gain, the *Yagi antenna* can be built lighter, more compact, and with less

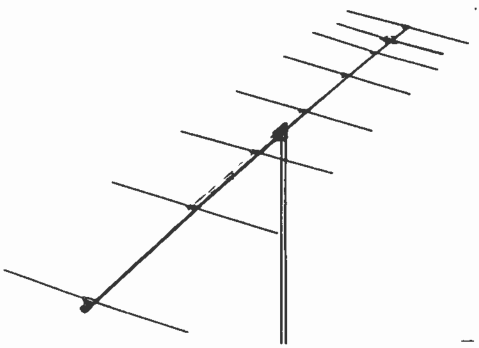


Figure 23

TWO-METER LONG YAGI ARRAY

Elements are mounted atop boom by means of small clamps made of soft aluminum strap. Either folded dipole or T-matching device may be used with antenna. Eight-element beam similar in construction is manufactured by Hy-Gain and sold in kit form.

wind resistance than any other type. On the other hand, if a Yagi array of the same approximate size and weight as another antenna type is built, it will provide a higher order of power gain and directivity than that of the other antenna.

The power gain of a Yagi antenna increases directly with the physical length of the array. The maximum practical length is entirely a mechanical problem of physically supporting the long series of director elements, although when the array exceeds a few wavelengths in length the element lengths, spacings, and Q's becomes more and more critical. The effectiveness of the array depends on a proper combination of the mutual coupling loops between adjacent directors and between the first director and the driven element.

Practically all work on Yagi antennas with more than three or four elements has

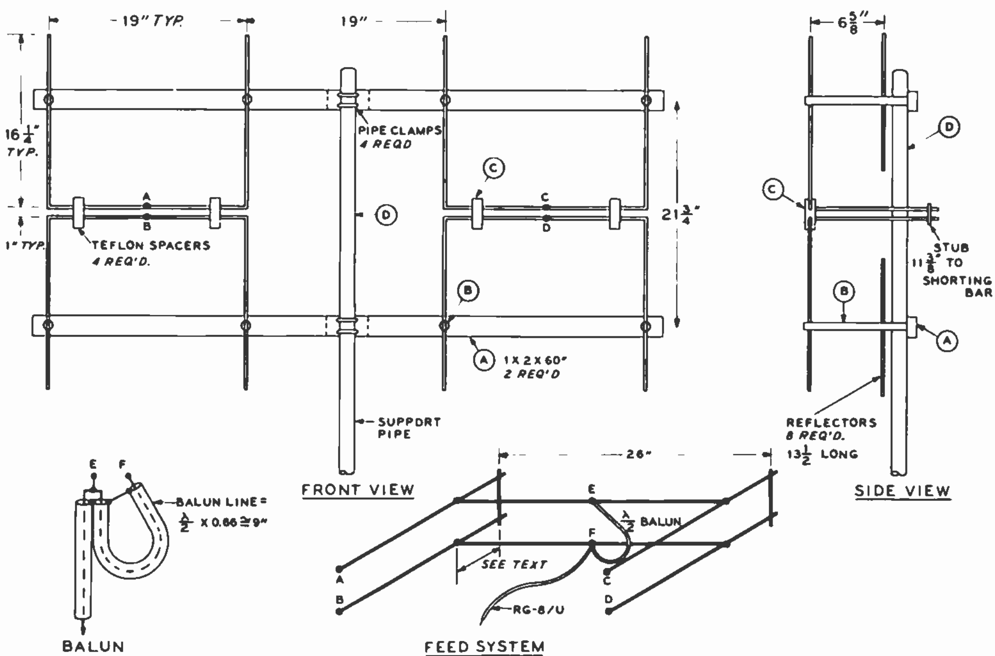


Figure 24

W6GD EXPANDED BROADSIDE ARRAY FOR 432-MHz

The 16-element beam is made of brass rod suspended from a wooden frame at low-voltage points on the antenna. Small ceramic insulators are used to mount the rods. Antenna elements and lines are aligned by means of small teflon or polystyrene spacer blocks passed over the rods before they are bent into shape. Half-wave lines are employed in feed system, together with a full-wavelength transformer and balun to provide a close match to a 50-ohm transmission line. Lines and transformer are made up of brass rod and adjustable shorting bars are used.

been on an experimental, cut-and-try basis. Figure 22 provides dimensions for a typical long Yagi antenna for the 2-meter vhf band. Note that all directors have the same physical length. If the long Yagi is designed so that the directors gradually decrease in length as they progress from the dipole bandwidth will be increased, and both side lobes and forward gain will be reduced. One advantage gained from staggered director length is that the array can be shortened and lengthened by adding or taking away directors without the need for retuning the remaining group of parasitic elements. When all directors are the same length, they must be all shortened *en masse* as the array is lengthened, and vice versa when the array is shortened.

The W6GD Broadside Array for 432 MHz The extended-expanded broadside array was designed by the late W6GD of Stanford University and has consistently out-performed larger and more sophisticated antennas at 432 MHz. The W6GD beam is a 16-element beam and has been measured to have 12 decibels power gain over a dipole. Extended elements are used with ¾-wavelength spacing. The array has a sharp front lobe, with nulls at 19° and 42° each side of center and must be aimed carefully for best results.

All elements are made of 0.175-inch diameter brass rod. The active elements are made of square "U"s bent from four lengths of rod, each 51½ inches long. The half-wavelength reflectors are cut of the same material and are 13⅛-inches long. The W6GD array is built on a wooden framework, so designed as to keep the supporting structure in back of the array. The driven elements are self-supporting except for four insulating blocks placed at low-voltage points. The blocks and spacers are drilled and slipped on the brass rods before the assembly is bent into shape (figure 24).

After assembly, the matching stubs are silver-soldered to the driven elements and the balun and the interconnecting transmission line temporarily connected in place. The line is tapped up each stub to attain a low value of SWR on the coaxial or open-wire transmission line. Placement of the taps is determined by experiment.

A complete discussion of vhf antennas is contained in the *VHF Handbook*, available from Radio Publications, Inc., Wilton, Conn.

High-Frequency Rotary-Beam Antennas

The rotatable antenna array has become almost standard equipment for operation on the 28- and 50-MHz bands and is commonly used on the 14- and 21-MHz bands and on those frequencies above 144 MHz. The rotatable array offers many advantages for both military and amateur use. The directivity of the antenna types commonly employed (particularly the unidirectional arrays) offers a worthwhile reduction in interference from undesired directions. Also, the increase in the ratio of low-angle radiation plus the theoretical gain of such arrays results in a relatively large increase in both the transmitted signal and the signal intensity from a station being received.

There are two normal configurations of radiating elements which, when horizontally polarized, will contribute to obtaining a low angle of radiation. These configurations are the end-fire array and the broadside array. The conventional three- or four-element rotary beam may properly be called a *unidirectional parasitic end-fire array*, and is actually a type of *yagi* array. The flat-top beam is a type of *bidirectional end-fire array*. The *broadside type* of array is also quite effective in obtaining low-angle radiation, and, although widely used in f-m and TV broadcasting, has seen little use by amateurs in rotatable arrays because of its size.

28-1 Unidirectional Parasitic End-Fire Arrays (Yagi Type)

If a single parasitic element is placed on one side of a driven dipole at a distance of from 0.1 to 0.25 wavelength the parasitic element can be tuned to make the array substantially unidirectional.

This simple array is termed a *two-element parasitic beam*.

The Two-Element Beam The two-element parasitic beam provides the greatest amount of gain per unit size of any array commonly used by radio amateurs. Such an antenna is capable of a signal gain of 5 db over a dipole, with a front-to-back ratio of 7 to 15 db, depending on the adjustment of the parasitic element. The parasitic element may be used either as a director or as a reflector.

The optimum spacing for a reflector in a two-element array is approximately 0.13 wavelength and with optimum adjustment of the length of the reflector a gain of approximately 5 db will be obtained, with a feed-point resistance of about 25 ohms.

If the parasitic element is to be used as a director, the optimum spacing between it

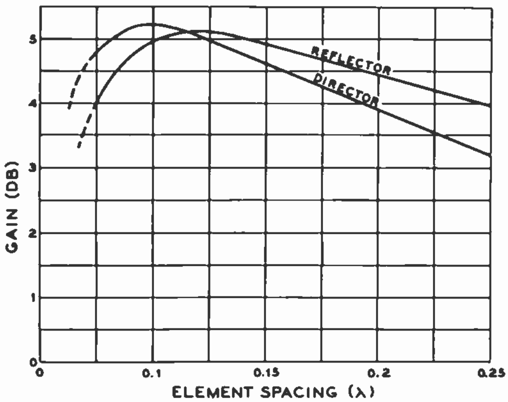


Figure 1

GAIN VERSUS ELEMENT SPACING FOR A TWO-ELEMENT CLOSE-SPACED PARASITIC BEAM ANTENNA WITH PARASITIC ELEMENT OPERATING AS A DIRECTOR OR REFLECTOR

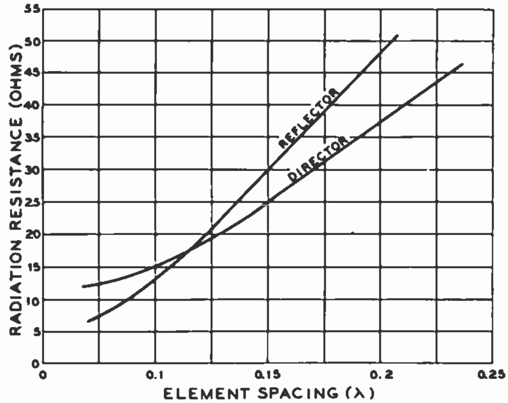


Figure 2

RADIATION RESISTANCE AS A FUNCTION OF THE ELEMENT SPACING FOR A TWO-ELEMENT PARASITIC ARRAY

and the driven element is 0.11 wavelength.

The general characteristics of a two-element parasitic array may be seen in figures 1, 2 and 3. The gain characteristics of a two-element array when the parasitic element is used as a director or as a reflector are shown. It can be seen that the director provides a maximum of 5.3 db gain at a spacing of slightly greater than 0.1 wavelength from the antenna. In the interests of greatest power gain and size conservation, therefore, the choice of a parasitic director would be wiser than the choice of a parasitic reflector, although the gain difference between the two is small.

Figure 2 shows the relationship between the element spacing and the radiation resistance for the two-element parasitic array for both the reflector and the director case. For either type of array, the radiation resistance falls in the 15- to 25-ohm region for typical spacings.

Figure 3 shows the *front-to-back* ratio for the two-element parasitic array for both the reflector and director cases. To produce these curves, the elements were tuned for maximum gain of the array. Better front-to-back ratios may be obtained at the expense of array gain, if desired, but the general shape of the curves remains the same.

It can be readily observed that operation of the parasitic element as a reflector produces relatively poor front-to-back ratios except when the element spacing is greater than 0.15 wavelength. However, at this element spacing, the gain of the array begins to suffer.

Since a radiation resistance of 17 ohms is not unduly hard to match, it can be argued that the best all-around performance may be obtained from a two-element parasitic beam employing 0.11 element spacing, with the parasitic element tuned to operate as a director. This antenna will provide a for-

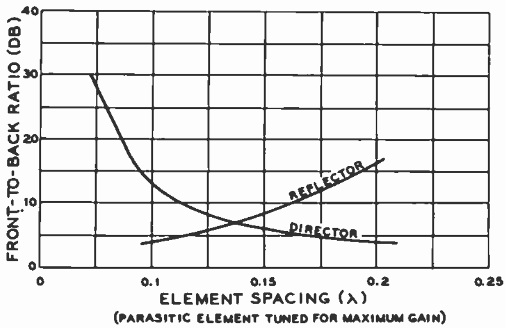


Figure 3

FRONT-TO-BACK RATIO AS A FUNCTION OF ELEMENT SPACING FOR A TWO-ELEMENT PARASITIC ARRAY

ward gain of 5.3 db, with a front-to-back ratio of 10 db, or slightly greater. Closer spacing than 0.11 wavelength may be employed for greater front-to-back ratios, but the radiation resistance of the array becomes quite low, the bandwidth of the array becomes very narrow, and the tuning becomes quite critical. Thus the Q of the antenna system will be *increased* as the spacing between the elements is *decreased*, and smaller optimum frequency coverage will result.

Element Lengths When the parasitic element of a two-element array is used as a director, the following formulas may be used to determine the lengths of the driven element and the parasitic director, assuming an element diameter-to-length ratio of 200 to 400:

$$\text{Driven element length (feet)} = \frac{476}{F_{MHz}}$$

$$\text{Director length (feet)} = \frac{450}{F_{MHz}}$$

$$\text{Element spacing (feet)} = \frac{120}{F_{MHz}}$$

The effective bandwidth taken between the 1.5/1 standing-wave points of an array cut to the above dimensions is about 2.5 percent of the operating frequency. This means that an array precut to a frequency of 14,150 kHz would have a bandwidth of 350 kHz (plus or minus 175 kHz of the center frequency), and therefore would be effective over the whole 20-meter band. In like fashion, a 15-meter array should be precut to 21,200 kHz.

A beam designed for use on the 10-meter band would have an effective bandwidth of some 700 kHz. Since the 10-meter band is 1700 kHz in width, the array should either be cut to 28,500 kHz for operation in the low-frequency portion of the band, or to 29,200 kHz for operation in the high-frequency portion of the band. Operation of the antenna outside the effective bandwidth will increase the SWR on the transmission line, and noticeably degrade both the gain and front-to-back ratio performance. The height above ground also influences the F/B ratio.

28-2 The Three-Element Array

The three-element array using a director, driven element, and reflector will exhibit as much as 30 db front-to-back ratio and 20 db front-to-side ratio for *low-angle radiation*. The theoretical gain is about 9 db over a dipole in free space. In actual practice, the array will often show 7 to 8 db apparent gain over a horizontal dipole placed the same height above ground (at 28 and 14 MHz).

The use of more than three elements is desirable when the length of the supporting structure is such that spacings of approximately 0.15 wavelength between elements becomes possible. Four-element arrays are quite common on the 28- and 50-MHz bands, and five elements are sometimes used for increased gain and discrimination. As the number of elements is increased the gain and front-to-back ratio increases but the bandwidth or frequency range over which the antenna will operate without reduction in effectiveness is decreased.

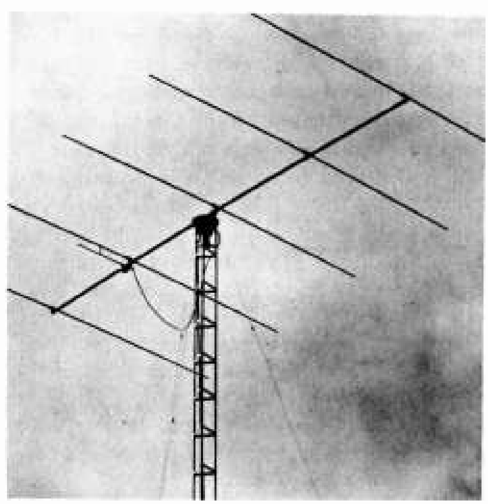


Figure 4

FIVE ELEMENT 28-MHz BEAM ANTENNA AT W6SAI

Antenna boom is made of twenty foot length of three-inch aluminum irrigation pipe. Spacing between elements is five feet. Elements are made of twelve foot lengths of 7/8-inch aluminum tubing, with extension tips made of 3/4-inch tubing. Beam dimensions are taken from figure 5.

Material for Elements The majority of high-frequency beams make use of elements composed of telescoping sections of metal tubing. This configuration is easy to construct and avoids the problem of getting sufficiently good insulation at the ends of the elements. The voltages reach such high values toward the ends of the elements that losses will be excessive, unless the insulation is excellent.

The elements may be fabricated of thin-walled steel conduit, or hard-drawn thin-walled copper tubing, but *dural* tubing is much better. Dural tubing may be obtained in telescoping sizes from large metal-supply houses in many cities. Various manufacturers, moreover, supply beam antenna kits of all types and prices. The majority of these beams employ dural elements because of the good weather-capability of this material.

Element Spacing The optimum spacing for a two-element array is, as has been mentioned before, approximately 0.11 wavelength for a director and 0.13 wavelength for a reflector. However, when both a director and a reflector are combined with the driven element to make up a three-element array the optimum spacing is established by the bandwidth which the antenna will be required to cover. Wide spacing (of the order of 0.25 wavelength between elements) will result in greater bandwidth for a specified maximum standing-wave ratio on the antenna transmission line. Smaller spacings may be used when boom length is an important consideration, but for a specified standing-wave ratio and forward gain the frequency coverage will be smaller. Thus the Q of the antenna system will be *increased* as the spacing between the elements is *decreased,* resulting in smaller frequency coverage, and at the same time the feed-point impedance of the driven element will be decreased.

For broad band coverage, such as the range from 28.0 to 29.7 MHz or from 50 to 54 MHz, 0.2 wavelength spacing from the driven element to each of the parasitic elements is recommended. For narrower bandwidth, such as would be adequate for the 14.0- to 14.4-MHz band or the 21- to 21.45 MHz band, the radiator-to-parasitic element spacing may be reduced to 0.12 wavelength, while still maintaining adequate array bandwidth for the amateur band in question.

Length of the Parasitic Elements Experience has shown that it is practical to cut the parasitic elements of a three-element parasitic array to a predetermined length before the installation of such an antenna. A pretuned antenna such as this will give good signal gain, adequate front-to-back ratio, and good bandwidth factor. By carefully tuning the array after it is in position the gain may be increased by a fraction of a db, and the front-to-back ratio by several db. However the slight improvement in performance is usually not worth the effort expended in tuning time.

The closer the lengths of the parasitic elements are to the resonant length of the driven element, the lower will be the feed-point resistance of the driven element, and the smaller will be the bandwidth of the array. Hence, for wide frequency coverage the director should be considerably shorter, and the reflector considerably longer than the driven element. For example, the directer should still be less than a resonant half-wavelength at the upper frequency limit of the range wherein the antenna is to be operated, and the reflector should still be long enough to act as a reflector at the lower frequency limit. Another way of stating the same thing is to say, in the case of an array to cover a wide frequency range such as the amateur range from 28 to 29.7 MHz that the director should be cut for the upper end of the band and the reflector for the lower end of the band. In the case of the 28- to 29.7-MHz range this means that the director should be about 8 percent shorter than the driven element and the reflector should be about 8 percent longer. Such an antenna will show a relatively constant gain of about 6 db over its range of coverage, and the pattern will not reverse at any point in the range.

Where the frequency range to be covered is somewhat less, such as the 14.0- to 14.4-MHz amateur band, or the lower half of the amateur 28-MHz phone band, the re-

TYPE	DRIVEN ELEMENT LENGTH	REFLECTOR LENGTH	1ST DIRECTOR LENGTH	2ND DIRECTOR LENGTH	3RD DIRECTOR LENGTH	SPACING BETWEEN ELEMENTS	APPROX. GAIN DB	APPROX. RADIATION RESISTANCE (Ω)
3-ELEMENT	$\frac{473}{F\,(MHz)}$	$\frac{501}{F\,(MHz)}$	$\frac{445}{F\,(MHz)}$	—	—	.15-.15	7.5	20
3-ELEMENT	$\frac{473}{F\,(MHz)}$	$\frac{501}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	—	—	.25-.25	8.5	35
4-ELEMENT	$\frac{473}{F\,(MHz)}$	$\frac{501}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	—	.2-.2-.2	9.5	20
5-ELEMENT	$\frac{473}{F\,(MHz)}$	$\frac{501}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	$\frac{450}{F\,(MHz)}$	.2-.2-.2-.2	10.0	15

Figure 5

DESIGN CHART FOR PARASITIC ARRAYS (DIMENSIONS GIVEN IN FEET)

flector should be about 5 percent longer than the driven element, and the director about 5 percent shorter. Such an antenna will perform well over its rated frequency band, will not reverse its pattern over this band, and will show a signal gain of 7 to 8 db. See figure 5 for design figures for 3-element arrays.

More Than Three Elements A small amount of additional gain may be obtained through use of more than two parasitic elements, at the expense of reduced feed-point impedance and lessened bandwidth. One additional director will add about 1 db, and a second additional director (making a total of five elements including the driven element) will add slightly less than 1 db more. In the vhf range, where the additional elements may be added without much difficulty, and where required bandwidths are small, the use of more than two parasitic elements is quite practical.

Stacking of Yagi Arrays Parasitic arrays (yagis) may be stacked to provide additional gain in the same manner that dipoles may be stacked. Thus if an array of six dipoles would give a gain of 10 db, the substitution of yagi arrays for each of the dipoles would add the gain of *one* yagi array to the gain obtained with the dipoles. However, the yagi arrays *must be more widely spaced* than the dipoles to obtain this theoretical improvement. As an example, if six 5-element yagi arrays having a gain of about 10 db were substituted for the dipoles, with appropriate increase in the spacing between the arrays, the gain of the whole system would approach the sum of the two gains, or 20 db. A group of arrays of yagi antennas, with recommended spacing and approximate gains, is illustrated in figure 6.

28-3 Feed Systems for Parasitic (Yagi) Arrays

The table of figure 5 gives, in addition to other information, the approximate radiation resistance referred to the center of the driven element of multielement parasitic arrays. It is obvious, from these low values of radiation resistance, that special care must be taken in materials used and in the construction of the elements of the array to ensure that ohmic losses in the conductors will not be an appreciable percentage of the radiation resistance. It is also obvious that some method of impedance transformation must be used in many cases to match the low radiation resistance of these antenna arrays to the normal range of characteristic impedance used for antenna transmission lines.

Impedance Matching A group of possible methods of impedance matching is shown in figures 7, 8, 9, and 10. All these methods have been used but certain of them offer advantages over some of the other methods. Generally speaking it is not mechanically desirable to break the center of the driven element of an array for feeding the system. Breaking the driven element rules out the practicability of building an all-metal type of array, and imposes mechanical limitations with any type of construction. However, when continuous rotation is desired, an arrangement such as

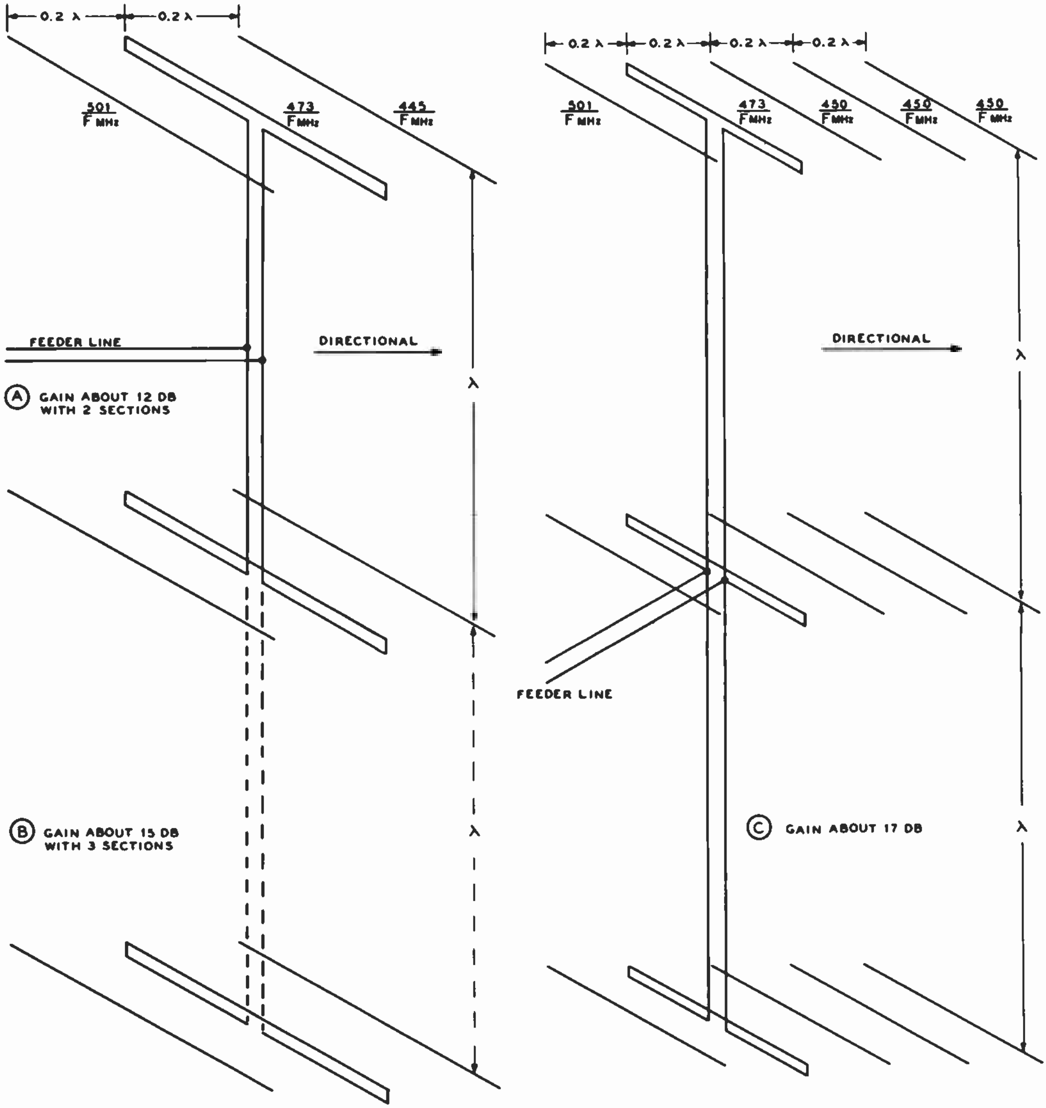


Figure 6

STACKED YAGI ARRAYS

It is possible to attain a relatively large amount of gain over a limited bandwidth with stacked yagi arrays. The two-section array at A will give a gain of about 12 db, while adding a third section will bring the gain up to about 15 db. Adding two additional parasitic directors to each section, as at C will bring the gain up to about 17 db.

shown in figure 9D utilizing a broken driven element with a rotatable transformer for coupling from the antenna transmission line to the driven element has proven to be quite satisfactory.

The feed systems shown in figure 7 will, under normal conditions, show the lowest losses of any type of feed system since the currents flowing in the matching network are the lowest of all the systems commonly

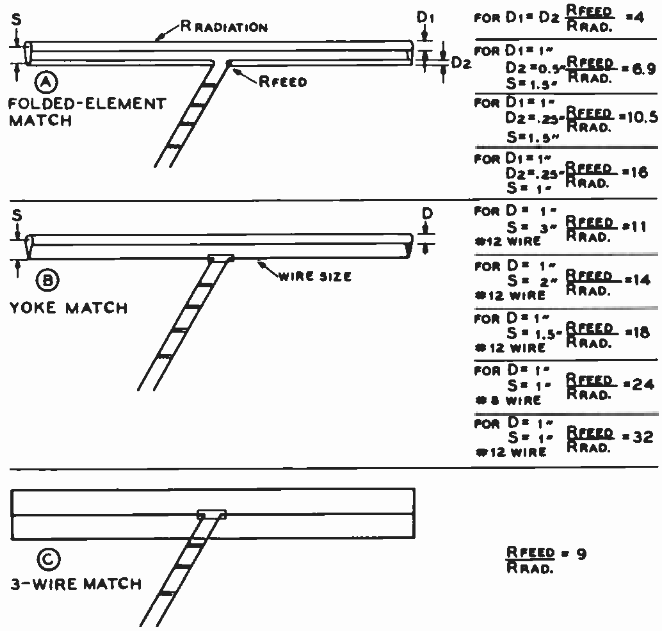


Figure 7

DATA FOR FOLDED-ELEMENT MATCHING SYSTEMS

In all normal applications of the data given the main element as shown is the driven element of a multielement parasitic array. Directors and reflectors have not been shown for the sake of clarity.

used. The *folded-element* match shown in figure 7A and the *Yoke* match shown in figure 7B are the most satisfactory, electrically, of all standard feed methods. However, both methods require the extension of an additional conductor out to the end of the driven element as a portion of the matching system. The folded-element match is best on the 50-MHz band and higher where the additional section of tubing may be supported below the main radiator element without undue difficulty. The yoke-match is more satisfactory mechanically on the 28- and 14-MHz bands since it is only necessary to suspend a wire below the driven element proper. The wire may be spaced below the self-supporting element by means of several small strips of polystyrene which have been drilled for both the main element and the small wire and threaded on the main element.

The Folded-Element Match Calculations The calculation of the operating conditions of the folded-element matching systems and the yoke match, as shown in figures 7A and 7B is relatively simple. A selected group of operating conditions has been shown on the drawing of figure 7. In applying the system it is only necessary to multiply the ratio of feed to radiation resistance (given in the figures to the right of the suggested operating dimensions in figure 7) by the radiation resistance of the antenna system to obtain the impedance of the cable to be used in feeding the array. Approximate values of radiation resistance for a number of commonly used parasitic-element arrays are given in figure 5.

In many cases it will be desired to use the folded-element or yoke matching system with different sizes of conductors or different spacings than those shown in figure 7. Note, then, that the impedance transformation ratio of these types of matching systems is dependent *both on the ratio of conductor diameters and on their spacing*. The following equation may be used for the determination of the impedance transformation when using different diameters in the two sections of a folded element:

$$\text{Transformation ratio} = \left(1 + \frac{Z_1}{Z_2}\right)^2$$

In this equation Z_1 is the characteristic impedance of a line made up of the smaller of the two conductor diameters spaced the center-to-center distance of the two con-

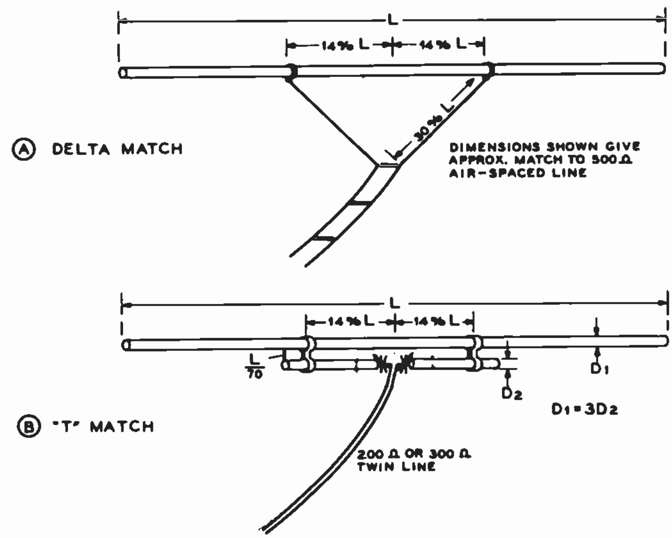


Figure 8

AVERAGE DIMENSIONS FOR THE DELTA AND "T" MATCH

ductors in the antenna, and Z_2 is the characteristic impedance of a line made up of two conductors the size of the larger of the two. This assumes that the feed line will be connected in series with the *smaller* of the two conductors so that an impedance step-up of greater than four will be obtained. If an impedance step-up of less than four is desired, the feed line is connected in series with the *larger* of the two conductors and Z_1 in the above equation becomes the impedance of a hypothetical line made up of the larger of the two conductors and Z_2 is made up of the smaller. The folded vhf unipole is an example where the transmission line is connected in series with the larger of the two conductors.

The Delta Match and T-Match The *delta match* and the *T-match* are shown in figure 8. The delta match has been largely superseded by the newer T-match, however, both these systems can be adjusted to give a low value of SWR on 50- to 600-ohm balanced transmission lines. In the case of the systems shown it will be necessary to make adjustments in the tapping distance along the driven radiator until minimum standing waves on the antenna transmission line are obtained. Since it is sometimes impractical to eliminate completely the standing waves from the antenna transmission line when using these matching systems, it is common practice to cut the feed line, after standing waves have been reduced to a minimum, to a length which will give satisfactory loading of the transmitter over the desired frequency range of operation.

The inherent reactance of the T-match is tuned out by the use of two identical resonating capacitors in series with each leg of the T-rod. These capacitors should each have a maximum capacity of 8 pf per meter of wavelength. Thus for 20 meters, each capacitor should have a maximum capacitance of at least 160 pf. For power up to a kilowatt, 1000-volt spacing of the capacitors is adequate. These capacitors should be tuned for minimum SWR on the transmission line. The adjustment of these capacitors should be made at the same time the correct setting of the T-match rods is made as the two adjustments tend to be interlocking. The use of the standing-wave meter (described in Test Equipment chapter) is recommended for making these adjustments to the T-match.

Feed Systems Using a Driven Element with Center Feed Four methods of exciting the driven element of a parasitic array are shown in figure 9. The system shown at A has proven to be quite satisfa-

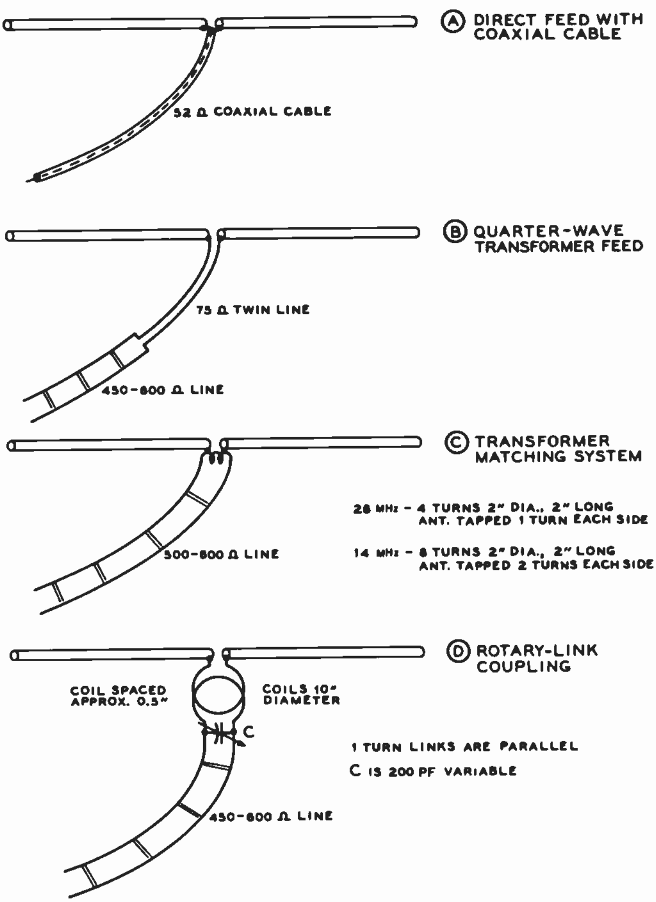


Figure 9

ALTERNATE FEED METHODS WHERE THE DRIVEN ELEMENT MAY BE BROKEN IN THE CENTER

tory in the case of an antenna-reflector two-element array or in the case of a three-element array with 0.2 to 0.25 wavelength spacing between the elements of the antenna system. The feed-point impedance of the center of the driven element is close enough to the characteristic impedance of the 52-ohm coaxial cable that the standing-wave ratio on the 52-ohm coaxial cable is less than 2-to-1. B shows an arrangement for feeding an array with a broken driven element from an open-wire line with the aid of a quarter-wave matching transformer.

Rotary-Link Coupling In many cases it is desirable to be able to allow the antenna array to rotate continuously without regard to snarling of the feed line. If this is to be done some sort of slip rings or rotary joint must be made in the feed line. One relatively simple method of allowing unrestrained rotation of the antenna is to use the method of *rotary-link coupling* shown in figure 9D. The two coupling rings are 10 inches in diameter and are usually constructed of ¼-inch copper tubing supported one from the rotating structure and one from the fixed structure by means of standoff insulators. The capacitor (C in figure 9D) is adjusted, after the antenna has been tuned, for minimum standing-wave ratio on the antenna transmission line. The dimensions shown will allow operation with either 14- or 28-MHz elements, with appropriate adjustment of capacitor C. The rings must of course be parallel and must lie in a plane normal to the axis of rotation of the rotating structure.

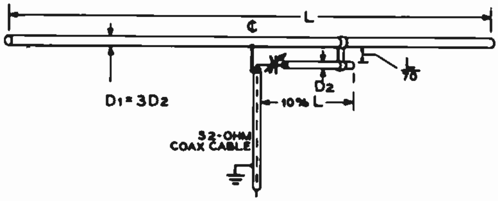


Figure 10

THE GAMMA MATCHING SYSTEM

See text for details of resonating capacitor

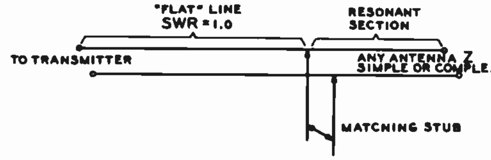


Figure 11

IMPEDANCE MATCHING WITH A CLOSED STUB ON A TWO-WIRE TRANSMISSION LINE

The Gamma Match The use of coaxial cable to feed the driven element of a yagi array is becoming increasingly popular. One reason for this increased popularity lies in the fact that the TVI-reduction problem is simplified when coaxial feed line is used from the transmitter to the antenna system. Radiation from the feed line is minimized when coaxial cable is used, since the outer conductor of the line may be grounded at several points throughout its length and since the intense field is entirely confined within the outer conductor of the coaxial cable. Other advantages of coaxial cable as the antenna feed line lie in the fact that coaxial cable may be run within the structure of a building without danger, or the cable may be run underground without disturbing its operation. Also, transmitting-type low-pass filters for 52-ohm impedance are more widely available and are less expensive than equivalent filters for two-wire line.

The *gamma-match* is illustrated in figure 10, and may be considered as one-half of a T-match. One resonating capacitor is used, placed in series with the gamma rod. The capacitor should have a capacity of 7 pf per meter of wavelength. For 15-meter operation the capacitor should have a maximum capacitance of 105 pf. The *length* of the gamma rod determines the impedance transformation between the transmission line and the driven element of the array, and the gamma capacitor tunes out the inductance of the gamma rod. By adjustment of the length of the gamma rod, and the setting of the gamma capacitor, the SWR on the coaxial line may be brought to a very low value at the chosen operating frequency.

The Matching Stub If an open-wire line is used to feed a low-impedance radiator, a section of the transmission line may be employed as a matching stub as shown in figure 11. The matching stub can transform any complex impedance to the characteristic impedance of the transmission line. While it is possible to obtain a perfect match and good performance with either an open stub or a shorted one by observing appropriate dimensions, a shorted stub is much more readily adjusted. Therefore, the following discussion will be confined to the problem of using a closed stub to match a low-impedance load to a high-impedance transmission line.

If the transmission line is so elevated that adjustment of a "fundamental" shorted stub cannot be accomplished easily from the ground, then the stub length may be increased by exactly one or two electrical half wavelengths, without appreciably affecting its operation.

While the correct position of the shorting bar and the point of attachment of the stub to the line can be determined entirely by experimental methods, the fact that the two adjustments are interdependent, or interlocking, makes such a cut-and-try procedure a tedious one. Much time can be saved by determining the approximate adjustments required by reference to a chart such as figure 12 and using them as a starter. Usually only a slight "touching up" will produce a perfect match and flat line.

In order to utilize figure 12, it is first necessary to locate accurately a voltage node or current node on the line in the vicinity that has been decided on for the stub, and also to determine the SWR.

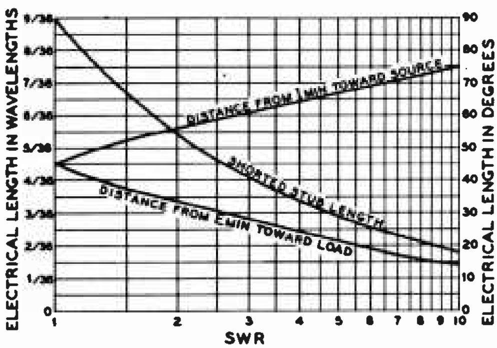


Figure 12

SHORTED-STUB LENGTH AND POSITION CHART

From the standing-wave ratio and current or voltage null position it is possible to determine the theoretically correct length and position of a shorted stub. In actual practice a slight discrepancy usually will be found between the theoretical and the experimentally optimized dimensions; therefore it may be necessary to "touch up" the dimensions after using the above data as a starting point.

Stub adjustment becomes more critical as the SWR increases, and under conditions of high SWR the current and voltage nulls are more sharply defined than the current and voltage maxima, or loops. Therefore, it is best to locate either a current null or voltage null, depending on whether a current-indicating device or a voltage-indicating device is used to check the standing-wave pattern.

The SWR is determined by means of a *directional coupler*, or by noting the ratio of E_{max} to E_{min} or I_{max} to I_{min} as read on an indicating device.

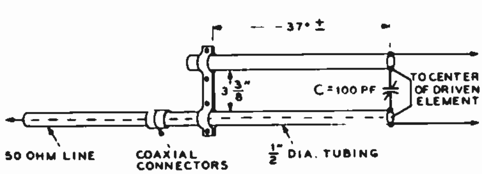


Figure 13

COAXIAL STUB BALUN FOR 14-MHz BEAM

Matching stub and balun are combined to provide balanced feed point for a 50-ohm transmission line to match low-impedance driven element. Balun is designed to be mounted on beam, at the center of driven element using short, heavy interconnecting leads.

It is assumed that the characteristic impedance of the section of line used as a stub is the same as that of the transmission line proper. It is preferable to have the stub section identical to the line physically as well as electrically.

A Stub Balun for a 14-MHz Yagi Beam A short, loaded matching stub may be combined with a balun transformer to provide a good match between a 50-ohm coaxial line and a low-impedance feedpoint of a typical 3-element 20-meter parasitic beam antenna (figure 13). The unit shown is designed to match a load impedance in the range of 17 to 25 ohms.

The stub balun is built of two sections of ½-inch diameter tubing. One section is about 40" long, and the other section is about 44" long and has a coaxial receptacle on one end. The tubes are separated about 3⅜". An adjustable shorting bar is placed at the transmission-line end of the assembly. A short length of RG-8/U coaxial line, with the outer jacket and braid removed is run from the coaxial receptacle, through the longer tube and out the free end. The wire is left long enough to cross-connect to the opposite balun tube. A variable capacitor is placed across the free end of the balun, as shown.

The balun is placed directly at the center of the driven element of the beam antenna. Length of the balun, the capacitance setting, and length of the driven element are the variables that determine the impedance match. Adjustment of these variables can provide a unity match at the resonant frequency of the array. The variable capacitor should be mounted in a waterproof box to protect it from moisture.

28-4 Unidirectional Driven Arrays

Three types of unidirectional driven arrays are illustrated in figure 14. The array shown in figure 14A is an end-fire system which may be used in place of a parasitic array of similar dimensions when greater frequency coverage than is available with the yagi type is desired. Figure 14B is a combination end-fire and collinear system

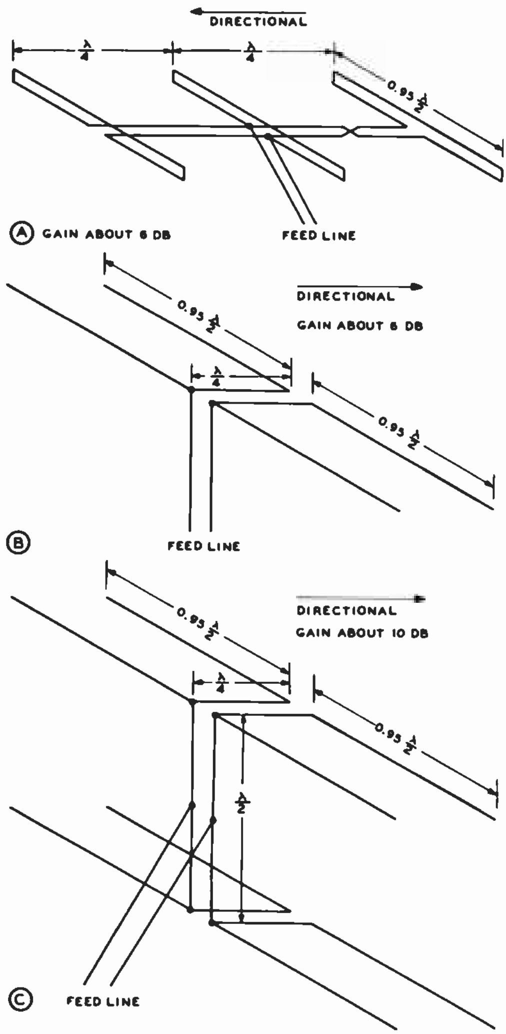


Figure 14

UNIDIRECTIONAL ALL-DRIVEN ARRAYS

A unidirectional all-driven end-fire array is shown at A. B shows an array with two half waves in phase with driven reflectors. A lazy-H array with driven reflectors is shown at C. Note that the directivity is through the elements with the greatest total feed-line length in arrays such as shown at B and C.

which will give approximately the same gain as the system of figure 14A, but which requires less boom length and greater total element length. Figure 14C illustrates the familiar lazy-H with driven reflectors (or directors, depending on the point of view) in a combination which will show wide bandwidth with a considerable amount of forward gain and good front-to-back ratio over the entire frequency coverage.

A simple driven array is the so-called *ZL Special,* which is one-half the array of figure 14B. The ZL Special is fed at the center point of the half-wave elements and provides a cardioid pattern with a gain of about 3 decibels.

Unidirectional Stacked Broadside Arrays Three practical types of unidirectional stacked broadside arrays are shown in figure 15. The first type, shown at figure 15A, is the simple lazy-H type of antenna with parasitic reflectors for each element. Figure 15B shows a simpler antenna array with a pair of folded dipoles spaced one-half wave vertically, operating with reflectors. In figure 15C is shown a more complex array with six half waves and six reflectors which will give a very worthwhile amount of gain.

In all three of the antenna arrays shown the spacing between the driven elements and the reflectors has been shown as one-quarter wavelength. This has been done to eliminate the requirement for tuning of the reflector, as a result of the fact that a half-wave element spaced exactly one-quarter wave from a driven element will make a unidirectional array when both elements are the same length. Using this procedure will give a gain of 3 db with the reflectors over the gain without the reflectors, with only a moderate decrease in the radiation resistance of the driven element. Actually, the radiation resistance of a half-wave dipole goes down from 73 ohms to 60 ohms when an identical half-wave element is placed one-quarter wave behind it.

A very slight increase in gain for the entire array (about 1 db) may be obtained at the expense of lowered radiation resistance, the necessity for tuning the reflectors, and decreased bandwidth by placing the reflectors 0.15 wavelength behind the driven elements and making them somewhat longer than the driven elements. The radiation resistance of each element will drop approximately to one-half the value obtained with untuned half-wave reflectors spaced one-quarter wave hehind the driven elements.

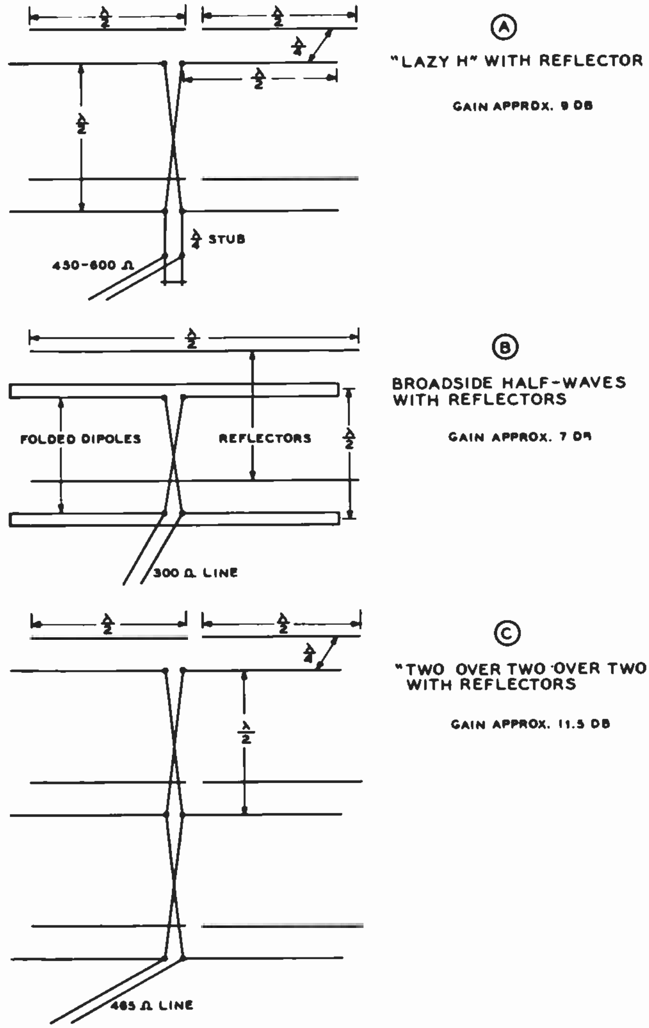


Figure 15

BROADSIDE ARRAYS WITH PARASITIC REFLECTORS

The apparent gain of the arrays illustrated will be greater than the values given due to concentration of the radiated signal at the lower elevation angles.

Antenna arrays of the type shown in figure 15 require the use of some sort of lattice work for the supporting structure since the arrays occupy appreciable distance in space in all three planes.

Feed Methods The requirements for the feed systems for antenna arrays of the type shown in figure 15 are less critical than those for the close-spaced parasitic arrays shown in the previous section. This is a natural result of the fact that a larger number of the radiating elements are directly fed with energy, and of the fact that the effective radiation resistance of each of the driven elements of the array is much higher than the feed-point resistance of a parasitic array. As a consequence of this fact, arrays of the type shown in figure 15 can be expected to cover a somewhat greater frequency band for a specified value of standing-wave ratio than the parasitic type of array.

In most cases a simple open-wire line may be coupled to the feed point of the array without any matching system. The standing-wave ratio with such a system of feed will often be less than 2-to-1. However, if a more accurate match between the antenna transmission line and the array is desired a conventional quarter-wave stub, or a quarter-wave matching transformer of appropriate impedance, may be used to obtain a low standing-wave ratio.

28-5 Construction of Rotatable Arrays

A considerable amount of ingenuity may be exercised in the construction of the supporting structure for a rotatable array. Every person has his own ideas as to the best method of construction. Often the most practical method of construction will be dictated by the availability of certain types of construction materials, but in any event be sure that sound mechanical engineering principles are used in the design of the supporting structure. There are few things quite as discouraging as the picking up of pieces, repairing of the roof, etc., when a newly constructed rotary comes down in the first strong wind. If the principles of mechanical engineering are understood it is wise to calculate the loads and torques which will exist in the various members of the structure with the highest wind velocity which may be expected in the locality of the installation. If this is not possible it will usually be worth the time and effort to look up a friend who understands these principles.

Radiating Elements One thing more or less standard about the construction of rotatable antenna arrays is the use of dural tubing for the self-supporting elements. Other materials may be used but an alloy known as 2024 has proven over a period of time to be quite satisfactory. Copper tubing is too heavy for a given strength, and steel tubing, unless copper plated, is likely to add an undesirably large loss resistance to the array. Also, steel tubing, even when plated, is not likely to withstand salt atmosphere (such as is encountered along the seashore) for a satisfactory period of time. Do not use a soft aluminum alloy for the elements unless they will be quite short; 2024 is a hard alloy and is noncorrosive. Alloy 2017 and 6061 are also satisfactory, cheaper, and easier to obtain. Do not use alloys 5052, 2014, or 3003 (EMT), as these signify alloys which have not been heat

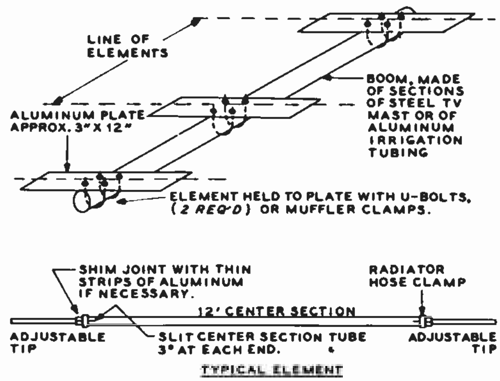


Figure 16

3-ELEMENT ALL-METAL ANTENNA ARRAY

All-metal configuration permits rugged, light assembly. Joints are made with U-bolts and metal plates for maximum rigidity.

treated for strength and rigidity. However, these softer alloys, and aluminum electrical conduit, may be used for short radiating elements such as would be used for the 50-MHz band or as interconnecting conductors in a stacked array.

All-Metal Construction It is characteristic of the conventional type of multielement parasitic array, such as discussed previously and outlined, that the centers of all the elements are at zero r-f potential with respect to ground. It is therefore possible to use a metallic structure without insulators for supporting the various elements of the array. A typical three-element array of this type is shown in figure 16. In this particular array, U-bolts and metal plates have been employed to fasten the elements to the boom. The elements are made of telescoping sections of aluminum tubing. The tips of the inner sections of tubing are split, and a tubing clamp is slipped over the joint, as shown in the drawing. Before assembly of the point, the mating pieces of aluminum are given a thin coat of *Penetrox-A* compound. (This antioxidizing paste is manufactured by *Burndy Co.*, Norwalk, Conn. and is distributed by the *General Electric Supply*

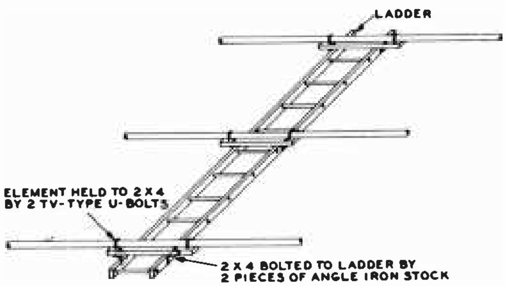


Figure 17

ALTERNATIVE WOODEN SUPPORTING ARRANGEMENT

A wooden ladder may be used to support a 10 or 15 meter array.

Co.) When the tubes are telescoped and the clamp is tightened, an airtight seal is produced, reducing corrosion to a minimum.

The boom of the parasitic array may be made from two or three sections of steel TV mast, or it may be made of a single section of aluminum irrigation pipe. This pipe is made by *Reynolds Aluminum Co.*, and others, and may often be purchased via the *Sears, Roebuck Co.* mail-order department. Three-inch pipe may be used for the 10- and 15-meter antennas, and the huskier four-inch pipe should be used for a 20-meter beam.

Automobile muffler clamps can often be used to affix the elements to the support plates. Larger clamps of this type will fasten the plates to the boom. In most cases, the muffler clamps are untreated, and they should be given one or two coats of rust-proof paint to protect them from inclement weather. All bolts, nuts, and washers used in the assembly of the array should be of the plated variety to reduce corrosion and rust.

If it is desired to use a split driven element for a balanced feed system, it is necessary to insulate the element from the supporting structure of the antenna. The element should be severed at the center, and the two halves driven onto a wooden dowel. The element may then be mounted on an aluminum support plate by means of four ceramic insulators. Metal-based insulators, such as the *Johnson 135-67* are recommended, since the all-ceramic types may break at the mounting holes when the array is subject to heavy winds.

28-6 Tuning the Array

Although satisfactory results may be obtained by precutting the antenna array to dimensions given earlier in this chapter, the occasion might arise when it is desired to make a check on the operation of the antenna before calling the job complete.

The process of tuning an array may satisfactorily be divided into two more or less distinct steps: the actual tuning of the array for best front-to-back ratio or for maximum forward gain, and the adjustment to obtain the best possible impedance match between the antenna transmission line and the feed point of the array.

Tuning the Array The actual tuning of the array for best front-to-back ratio or maximum forward gain may best be accomplished with the aid of a low-power transmitter feeding a dipole antenna (polarized the same as the array being tuned) at least four or five wavelengths away from the antenna being tuned and located at the same elevation as that of the antenna under test. A calibrated field-strength meter of the remote-indicating type is then coupled to the feed point of the antenna array being tuned. The transmissions from the portable transmitter should be made as short as possible and the call sign of the station making the test should be transmitted at least every ten minutes.

One satisfactory method of tuning the array proper, assuming that it is a system with several parasitic elements, is to set the directors to the dimensions given in figure 5 and then to adjust the reflector for maximum forward signal. Then the first director should be varied in length until maximum forward signal is obtained, and so on if additional directors are used. Then the array may be reversed in direction and the reflector adjusted for best front-to-back ratio. Subsequent small adjustments may then be made in both the directors and the reflector for best forward signal with a reasonable ratio of front-to-back signal. The adjustments in the directors and the reflector will be found to be interdependent to a certain degree, but if small adjustments are made after the preliminary tuning process a satisfactory set of adjustments for maximum performance will be obtained. It is usually best to make the end sections of the elements smaller in diameter so that they will slip inside the larger tubing sections. The smaller sliding sections may be clamped inside the larger main sections.

Matching to the Antenna Transmission Line The problem of matching the impedance of the antenna transmission line to the array is much simplified if the process of tuning the array is made a substantially separate process as just described. *After* the tuning operation is complete, the resonant frequency of the driven element of the antenna should be checked, directly at the center of the driven element if practical, with a grid-dip meter. It is important that the resonant frequency of the antenna be at the *center* of the frequency band to be covered. If the resonant frequency is found to be much different from the desired frequency, the length of the driven element of the array should be altered until this condition exists. A relatively small change in the length of the driven element will have only a second-order effect on the tuning of the parasitic elements of the array. Hence, a moderate change in the length of the driven element may be made without repeating the tuning process for the parasitic elements.

When the resonant frequency of the antenna system is correct, the antenna transmission line, with impedance-matching device or network between the line and antenna feed point, is then attached to the array and coupled to a low-power exciter unit or transmitter. Then, preferably, a standing-wave meter is connected in series with the antenna transmission line at a point relatively much closer to the transmitter than to the antenna.

If the standing-wave ratio is below 1.5 to 1 it is satisfactory to leave the installation as it is. If the ratio is greater than this range it will be best when twin line or coaxial line is being used, and advisable with open-wire line, to attempt to decrease the SWR.

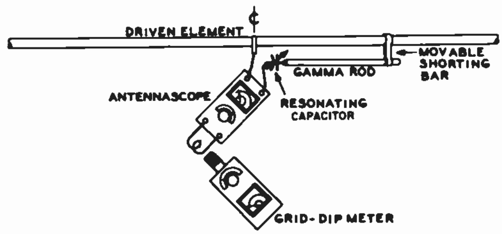


Figure 18

ADJUSTMENT OF GAMMA MATCH BY USE OF ANTENNASCOPE AND GRID-DIP METER

It must be remembered that no adjustments made at the *transmitter* end of the transmission line will alter the SWR on the line. All adjustments to better the SWR must be made at the *antenna* end of the line and to the device which performs the impedance transformation necessary to match the characteristic impedance of the antenna to that of the transmission line.

Before any adjustments to the matching system are made, the resonant frequency of the driven element must be ascertained, as explained previously. If all adjustments to correct impedance mismatch are made at this frequency, the problem of reactance termination of the transmission line is eliminated, greatly simplifying the problem. The following steps should be taken to adjust the impedance transformation:

1. The output impedance of the matching device should be measured. An Antennascope and a grid-dip oscillator are required for this step. The Antennascope is connected to the output terminals of the matching device. If the driven element is a folded dipole, the Antennascope connects directly to the split section of the dipole. If a gamma match or T-match is used, the Antennascope connects to the transmission-line end of the device. If a Q-section is used, the Antennascope connects to the bottom end of the section. The grid-dip oscillator is coupled to the input terminals of the Antennascope as shown in figure 18.
2. The grid-dip oscillator is tuned to the resonant frequency of the antenna,

which has been determined previously, and the Antennascope control is turned for a null reading on the meter of the Antennascope. The impedance presented to the Antennascope by the matching device may be read directly on the calibrated dial of the Antennascope.

3. Adjustments should be made to the matching device to present the desired impedance transformation to the Antennascope. If a folded dipole is used as the driven element, the transformation ratio of the dipole must be varied as explained previously in this chapter to provide a more exact match. If a T-match or gamma match system is used, the length of the matching rod may be changed to effect a proper match. If the Antennascope ohmic reading is *lower* than the desired reading, the length of the matching rod should be *increased*. If the Antennascope reading is *higher* than the desired reading, the length of the matching rod should be *decreased*. After each change in length of the matching rod, the series capacitor in the matching system should be re-resonated for best null on the meter of the Antennascope.

Raising and Lowering the Array A practical problem always present when tuning up and matching an array is the physical location of the structure. If the array is atop the mast it is inaccessible for adjustment, and if it is located on stepladders where it can be adjusted easily it cannot be rotated. One encouraging factor in this situation is the fact that experience has shown that if the array is placed 8 or 10 feet above ground on some stepladders for the preliminary tuning process, the raising of the system to its full height will not produce a serious change in the adjustments. So it is usually possible to make preliminary adjustments with the system located slightly greater than head height above ground, and then to raise the antenna to a position where it may be rotated for final adjustments. If the position of the matching device as determined near the ground is marked so that the adjustments will not be lost, the array may be raised to rotatable height and the fastening clamps left loose enough so that the elements may be slid in by means of a long bamboo pole. After a series of trials a satisfactory set of adjustments can be obtained.

The matching process does not require rotation, but it does require that the antenna proper be located at as nearly its normal operating position as possible. However, on a particular installation the standing-wave ratio on the transmission line near the transmitter may be checked with the array in the air, and then the array may be lowered to ascertain whether or not the SWR has changed. If it has not, and in most cases if the feeder line is strung out back and forth well above the ground as the antenna is lowered they will not change, the last adjustment may be determined, the standing-wave ratio again checked, and the antenna re-installed in its final location.

28-7 Indication of Direction

The most satisfactory method for indicating the direction of transmission of a rotatable array is that which uses *Selsyns* or *Synchros* for the transmission of the data from the rotating structure to the indicating pointer at the operating position. A number of Synchros and Selsyns of various types are available on the surplus market. Some of them are designed for operation on 115 volts at 60 *Hertz*, some are designed for operation on 60 Hertz but at a lowered voltage, and some are designed for operation from 400-Hertz or 800-*Hertz* energy. This latter type of high-frequency Selsyn is the most generally available type, and the high-frequency units are smaller and lighter than the 60-*Hertz* units. Since the indicating Selsyn must deliver an almost negligible amount of power to the pointer which it drives, the high-frequency types will operate quite satisfactorily from 60-Hertz power if the voltage on them is reduced to somewhere between 6.3 and 20 volts. In the case of many of the units available, a connection sheet is provided along with a recommendation in regard to the operating voltage when they are run on 60 Hertz. In any event the operating voltage should be held as low as it may be and still give satisfactory transmission of data from the antenna to the operat-

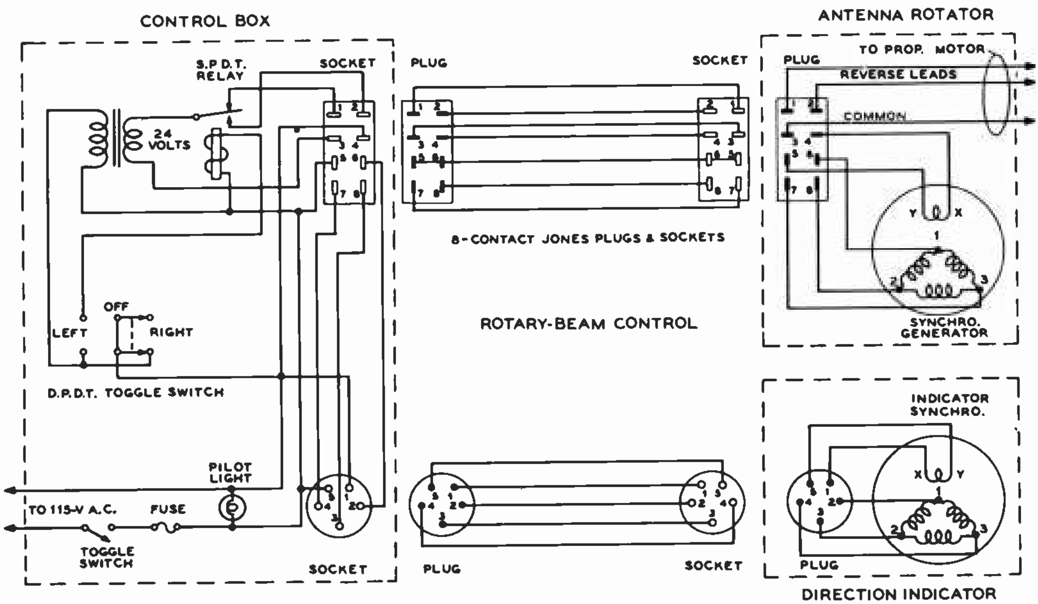


Figure 19

SCHEMATIC OF A COMPLETE ANTENNA CONTROL SYSTEM

ing position. Certainly it should not be necessary to run such a voltage on the units that they become overheated.

A suitable Selsyn indicating system is shown in figure 19.

Systems using a potentiometer capable of continuous rotation and a milliammeter, along with a battery or other source of direct current, may also be used for the indication of direction.

28-8 Three-Band Beams

A popular form of beam antenna introduced during the past few years is the so-called *three-band beam.* An array of this type is designed to operate on three adjacent amateur bands, such as the 10-, 15-, and 20-meter group. The principle of operation of this form of antenna is to employ parallel-tuned circuits placed at critical positions in the elements of the beam which serve to electrically connect and disconnect the outer sections of the elements as the frequency of excitation of the antenna is changed. A typical three-band element is shown in figure 20. At the lowest operating frequency, the tuned *traps* exert a minimum influence on the element which resonates at a frequency determined by the electrical length of the configuration, plus a slight degree of loading contributed by the traps. At some higher frequency (generally about 1.5 times the lowest operating frequency) the outer set of traps is in a parallel resonant condition, placing a high impedance between the element and the tips beyond the traps. Thus, the element resonates at a frequency 1.5 times higher than that deter-

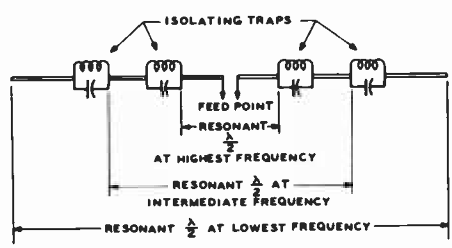


Figure 20

TRAP-TYPE "THREE BAND" ELEMENT

Isolating traps permit dipole to be self-resonant at three widely different frequencies.

Figure 21
HIGH-Q ISOLATING TRAP
This trap has a Q of nearly 300 and is well suited for multiband antennas. The coil is wound of No. 8 aluminum clothesline wire and is 3" in diameter and 3" long. The 15-meter trap has seven turns (illustrated) and the 10-meter trap has five turns. The capacitor is made from two lengths of aluminum tubing, coaxially aligned in a lucite dielectric. Capacitor length is about five inches and tubing sizes are 3/4 inch and 1-1/4 inch. Capacitance is about 25 pf. Lucite projects from end of capacitor to form 1/2-inch collar which is coated with epoxy to prevent deterioration of the dielectric under exposure to sunlight. Similar traps have been made using teflon as a dielectric material. Ends of aluminum tubes are slotted to facilitate assembly to antenna elements.

mined by the overall length of the element. As the frequency of operation is raised to approximately 2.0 times the lowest operating frequency, the inner set of traps becomes resonant, effectively disconnecting a larger portion of the element from the driven section. The length of the center section is resonant at the highest frequency of operation. The center section, plus the two adjacent inner sections are resonant at the intermediate frequency of operation, and the complete element is resonant at the lowest frequency of operation.

The efficiency of such a system is determined by the accuracy of tuning of both the element sections and the isolating traps. In addition the combined dielectric losses of the traps affect the overall antenna efficiency. As with all multipurpose devices, some compromise between operating convenience and efficiency must be made with antennas designed to operate over more than one narrow band of frequencies. Taking into account the theoretical difficulties that must be overcome it is a tribute to the designers of the better multiband beams that they perform as well as they do.

The Isolating Trap The parallel-tuned circuit which serves as an isolating trap for a multiband antenna should combine high circuit Q with good environmental protection. A highly satisfactory trap configuration based on the original design of W3DZZ is shown in figure 21. The trap capacitor, which has a value of about 25 pf, is made of two sections of aluminum tubing which form a portion of the antenna element. The capacitor dielectric is moulded lucite, or similar plastic material, given a coat of epoxy to help resist crazing and cracking caused by exposure to sunlight. The coil is wound of No. 8 aluminum wire and, with the capacitor placed within it, has a Q of nearly 300. The leads of the coil are bent around the tubing and a small aluminum block is used to form an inexpensive clamp. If desired, an aluminum cable clamp may be substituted for the homemade device.

The isolating trap is usually tuned to the lower edge of an amateur band, rather than to the center, to compensate for the length of the unit. In general, the 15-meter trap is tuned to approximately 20.8 MHz and the 10-meter trap is tuned near 27.8 MHz. The trap frequency is not critical within a few hundred kilohertz. Resonance is established by squeezing or expanding the turns of the coil while the trap is resonated on the bench with a grid-dip oscillator and a calibrated receiver.

A substitute for the moulded capacitor may be made up of two 40 pf, 5-kv ceramic capacitors connected in series (*Centralab 850S-50Z*) and mounted in a length of phenolic tubing of the proper diameter to slip within the aluminum antenna sections. The trap coil is then wound about the capacitor assembly in the manner shown in the photograph.

28-9 Lumped Baluns for Beam Antennas

A broadband coaxial balun was described in Chapter 20 of this Handbook. Baluns having similar broadband characteristics may be wound of wire on either air or ferrite cores, as shown in figure 22. The air-core design is rated for 2 kW PEP and maintains

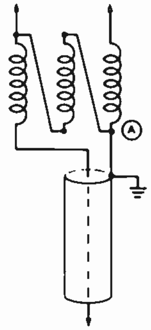


Figure 22

BROADBAND BALUN FOR BEAM ANTENNA

A trifilar balun may be used to match a 50-ohm coaxial line to a split driven element having an impedance of 15 to 50 ohms. Airwound balun consists of ten turns #14 Formvar insulated wire, wound on piece of 1-1/16" diameter plastic pipe, 4" long. Polyvinyl Chloride (PVC) pipe used for water pipe is satisfactory. Three windings are made, the wires placed in parallel, and wound side by side on the form as one, until ten trifilar turns are on the form.

Ferrite-core balun is composed of 6 turns #14 Formvar wire, trifilar wound on Q-1 material, ½" diameter. Use Indiana General CF-503 ferrite and break to proper length by nicking with file around the circumference and striking a hard blow. (Available Newark Electric Co., Chicago, part number 59F-1521).

a low value of SWR in the antenna system over a frequency range of 7 to 29.7 MHz. The balun may also be used at 80 meters, however, since it only exerts a slight detuning effect on the antenna.

The ferrite-core balun is designed for operation over the range of 3.5 to 29.7 MHz. While smaller in size than the air-core unit, the ferrite-core balun is power limited to about 200 watts at the high-frequency end of the operational range.

The baluns are trifilar wound, that is, three separate windings are placed on the form in parallel and then connected as shown in the illustration. The input terminals of the balun are nonsymmetrical; point A at the input end being taken as ground. Transposition of the input connections will degrade balancing action. Either end of the unit may be taken as the input, provided point A (the common connection of two windings) is ground.

When completed, the baluns should be protected from moisture by placing them within a waterproof, nonmetallic container. A plastic "squeeze bottle" may be used, with wooden discs cut for the ends and held in place with small screws.

CHAPTER TWENTY-NINE

Electronic Test Equipment

All amateur stations are required by law to have certain items of test equipment available within the station. A c-w station is required to have a frequency standard or other means, in addition to the transmitter frequency control, for insuring that the transmitted signal is on a frequency within one of the frequency bands assigned for such use. An SSB station is required in addition to have a means of determining that the transmitter is not being modulated in excess of its modulation capability, and in the case of an a-m transmitter, not more than 100 percent. Further, any station operating with a d-c power input greater than 900 watts is required to have a means of determining the exact input to the final stage of the transmitter, so as to insure that the d-c power input to the plate circuit of the output stage does not exceed 1000 watts.

The additional test and measurement equipment required by a station will be determined by the type of operation contemplated. It is desirable that all stations have an accurately calibrated voltohmmeter for routine transmitter and receiver checking and as an assistance in getting new pieces of equipment into operation. An oscilloscope and an audio oscillator make a very desirable adjunct to a phone station using a-m or f-m transmission, and are a necessity if single-sideband operation is contemplated. A calibrated signal generator is almost a necessity if much receiver work is contemplated, although a noise generator will serve in place of the signal general. Extensive antenna work invariably requires the use of some type of standing-wave meter. Lastly, if much construction work is to be done, a simple, grid-dip meter will be found to be one of the most used items of test equipment in the station.

Other modern pieces of test equipment such as digital voltmeters, counters and frequency synthesizers are becoming common items of station equipment as the amateur operator advances rapidly into today's world of solid-state equipment.

29-1 Voltage and Current

The measurement of *voltage* and *current* in radio circuits is very important in proper maintenance of equipment. Vacuum tubes and transistors of the types used in communications work must be operated within rather narrow limits in regard to filament or collector voltage, and they must be operated within certain maximum limits in regard to the voltage and current on other electrodes.

Both direct current and voltage are most commonly measured with the aid of an instrument consisting of a coil that is free to rotate in a constant magnetic field (*d'Arsonval* type instrument). If the instrument is to be used for the measurement of current it is called an *ammeter* or *milliammeter*. The current flowing through the circuit is caused to flow through the moving coil of this type of instrument. If the current to be measured is greater than 10 milliamperes or so, it is the usual practice to cause the majority of the current to flow through a bypass resistor called a *shunt*, only a specified portion of the current flowing through the moving coil of the instrument. The calculation of shunts for extending the range of d-c milliammeters and ammeters is discussed in Chapter Two.

A direct current *voltmeter* is merely a d-c milliammeter with a *multiplier* resistor in series with it. If it is desired to use a low-range milliammeter as a voltmeter the value of the multiplier resistor for any voltage range may be determined from the following formula:

$$R = \frac{1000\,E}{I}$$

where,

R equals multiplier resistor in ohms,
E equals desired full-scale voltage,
I equals full-scale current of meter in ma.

The sensitivity of a voltmeter is commonly expressed in *ohms per volt*. The higher the ohms per volt of a voltmeter the greater its sensitivity. When the full-scale current drain of a voltmeter is known, its sensitivity rating in ohms per volt may be determined by:

$$\textit{Ohms per volt} = \frac{1000}{I}$$

where,

I is the full-scale current drain of the indicating instrument in milliamperes.

Voltohmmeters An extremely useful piece of test equipment which should be found in every laboratory or radio station is the *voltohmmeter (v.o.m.)*. It consists of a multirange voltmeter with an additional fixed resistor, a variable resistor, and a battery. A typical example of such an instrument is diagrammed in figure 1. Tap 1 is used to permit use of the instrument as a 0-1 d-c milliammeter. Tap 2 permits accurate reading of resistors up to 100,000 ohms; taps 3, 4, 5, and 6 are for making voltage measurements, the full-scale voltages being 10, 50, 250, and 500 volts respectively.

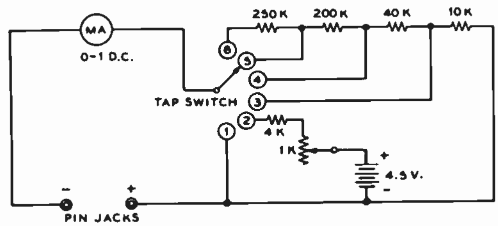


Figure 1

VOLTOHMMETER CIRCUIT

With the switch in position 1, the 0-1 milliammeter would be connected directly to the terminals. In position 2 the meter would read from 0-100,000 ohms, approximately, with a resistance value of 4500 ohms at half scale. (Note: The half-scale resistance value of an ohmmeter using this circuit is equal to the resistance in series with the battery inside the instrument.) The other four taps are voltage ranges with 10, 50, 250, and 500 volts full scale.

The 1000-ohm potentiometer is used to bring the needle to zero ohms when the terminals are shorted; this adjustment should always be made before a resistance measurement is taken. Higher voltages than 500 can be read if a higher value of multiplier resistor is added to an additional tap on the switch. The proper value for a given full-scale reading can be determined from Ohm's Law.

Resistances higher than 100,000 ohms cannot be measured accurately with the circuit constants shown; however, by increasing the ohmmeter battery to 45 volts and multiplying the 4000-ohm resistor and 1000-ohm potentiometer by 10, the ohms scale also will be multiplied by 10. This would permit accurate measurements up to 1 megohm.

For home-made voltohmmeters, good quality carbon resistors whose actual resistance has been checked may be used as multipliers where less accuracy is required.

Medium- and Low-Range Ohmmeter Most ohmmeters, including the one just described, are not adapted for accurate measurement of low-resistances — in the neighborhood of 100 ohms, for instance.

The ohmmeter diagrammed in figure 2 was especially designed for the reasonably accurate reading of resistances down to 1 ohm. Two scales are provided, one going in one direction and the other scale going in the other direction because of the different manner in which the milliammeter is used in each case. The low scale covers from 1 to 100 ohms and the high scale from 100 to 10,000 ohms. The high scale is in reality a medium-range scale. For accurate reading of resistances over 10,000 ohms, an ohmmeter of the type previously described should be used.

The calibration scale will depend on the internal resistance of the particular make of 1.5-ma meter used. The instrument can be calibrated by means of a Wheatstone bridge or a few resistors of known accuracy. The latter can be series-connected and parallel-connected to give sufficient calibration points

Measurement of Alternating Current and Voltage The measurement of alternating current and voltage is complicated by two factors; first, the frequency range covered in ordinary

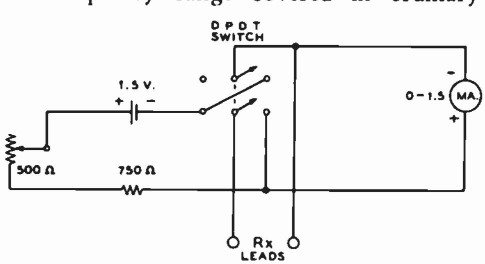


Figure 2

SCHEMATIC OF A LOW-RANGE OHMMETER

A description of the operation of this circuit is given in the text. With the switch in the left position the half-scale reading of the meter will occur with an external resistance of 1000 ohms. With the switch in the right position, half-scale deflection will be obtained with an external resistance equal to the d-c resistance of the milliammeter (20 to 50 ohms depending on the make of instrument).

communication channels is so great that calibration of an instrument becomes extremely difficult; second, there is no single type of instrument which is suitable for all a-c measurements—as the d'Arsonval type of movement is suitable for d-c. The *d'Arsonval* movement will not operate on alternating current since it indicates the average value of current flow, and the average value of an a-c wave is zero.

As a result of the inability of the reliable d'Arsonval type of movement to record an alternating current, either this current must be rectified and then fed to the movement. or a special type of movement which will operate from the *effective* value of the current can be used.

For the usual measurements of power-frequency alternating current (25-60 Hz), the *iron-vane* instrument is commonly used. For audio frequency alternating current (50-20,000 Hz) a d'Arsonval instrument having an integral copper-oxide, selenium, or (50-20,000 Hz), a d'Arsonval instrument having an integral diode rectifier is usually used. Radio-frequency voltage measurements are usually made with some type of vacuum-tube or solid-state voltmeter, while r-f current measurements are usually made with an instrument containing a thermocouple to convert the radio-frequency current into direct current for the meter movement.

Since an alternating-current wave can have an almost infinite variety of shapes, it can easily be seen that the ratios between the three fundamental quantities of the wave (peak, rms,effective, and average after rectification) can also vary widely. So it becomes necessary to know beforehand just which quality of the wave under measurement our instrument is going to indicate. For the purpose of simplicity we can list the usual types of alternating-current meters along with the characteristic of an alternating-current wave which they will indicate:

Iron-vane, thermocouple—rms.

Rectifier type (copper-oxide selenium, etc.)—average after rectification.

Vacuum-tube or solid-state voltmeter—rms, average, or peak, depending on design and calibration of the meter.

29-2 Electronic Voltmeters

An *electronic voltmeter* is essentially a detector in which a change in the input signal will produce a change in the indicating instrument (usually a d'Arsonval meter) placed in the output circuit. A *vacuum-tube voltmeter* (v.t.v.m.) may use a diode rectifier and several amplifying tubes, whereas a *solid-state voltmeter* makes use of transistors or ICs for the measurement of alternating or direct current.

When an electronic voltmeter is used in d-c measurement it is used primarily because of the very great input resistance of the device. Thus, the electronic voltmeter may be used for the measurement of agc, afc, and discriminator output voltages where no loading of the circuits can be tolerated.

The electronic voltmeter requires a closed d-c path for proper operation and—like the simple meter—can be overloaded and, thus, is limited in the amplitude of the voltage the input circuit can handle. Modern electronic voltmeters have an input resistance of 10 megohms, or more and usually incorporate a series resistance of 1 megohm, or more, to isolate the electronic voltmeter circuit from the circuit under test.

The Vacuum-Tube Voltmeter For the purpose of analysis, the operation of a modern v.t.v.m. will be described. The *Heath IM-13* is a fit instrument for such a description, since it is able to measure positive or negative d-c potentials, a-c rms values, peak-to-peak values, and resistance. The circuit of this unit is shown in figure 3. A sensitive 0 to 200 d-c microammeter is placed in the cathode circuit of a 12AU7 twin triode. The *zero-adjust* control sets up a balance between the two sections of the triode such that with zero input voltage applied to the first grid, the voltage drop across each portion of the zero-adjust control is the same. Under this condition of balance the meter will read zero. When a voltage is applied to the first grid, the balance in the cathode circuits is upset and the meter indicates the degree of unbalance. The relationship between the applied voltage on the first grid and the meter current is linear and therefore the meter can be calibrated with a linear scale. Since the tube is limited in the amount of current it can draw, the meter movement is electronically protected.

The maximum test voltage applied to the 12AU7 tube is about 3 volts. Higher applied voltages are reduced by a voltage divider which has a total resistance of about 10 megohms. An additional resistance of 1-megohm is located in the d-c test probe, thereby permitting measurements to be made in high-impedance circuits with minimum disturbance.

The rectifier portion of the vtvm is shown in figure 4. When a-c measurements are desired, a 6AL5 double diode is used as a full-wave rectifier to provide a d-c voltage proportional to the applied a-c voltage. This d-c voltage is applied through the voltage divider string to the 12AU7 tube causing the meter to indicate in the manner previously described. The a-c voltage scales of the meter are calibrated in both rms and peak-to-peak values. In the 1.5, 5, 15, 50, and 150 volt positions of the range switch, the full a-c voltage being measured is applied to the input of the 6AL5 full-wave rectifier. On the 500 and 1500 volt positions of the range switch, a divider network reduces the applied voltage in order to limit the voltage input to the 6AL5 to a safe recommended level.

The *a-c calibrate* control (figure 3) is used to obtain the proper meter deflection for the applied a-c voltage. Vacuum tubes develop a *contact potential* between tube elements. Such contact potential developed in the diode would cause a slight voltage to be present at all times. This voltage is cancelled out by proper application of a bucking voltage. The amount of bucking voltage is controlled by the *a-c balance* control. This eliminates zero shift of the meter when switching from a-c to d-c readings.

For resistance measurements, a 1.5-volt battery is connected through a string of multipliers and the external resistance to be measured, thus forming a voltage divider across the battery, and a resultant portion of the battery voltage is applied to the 12AU7 twin triode. The meter scale is calibrated in resistance (ohms) for this function.

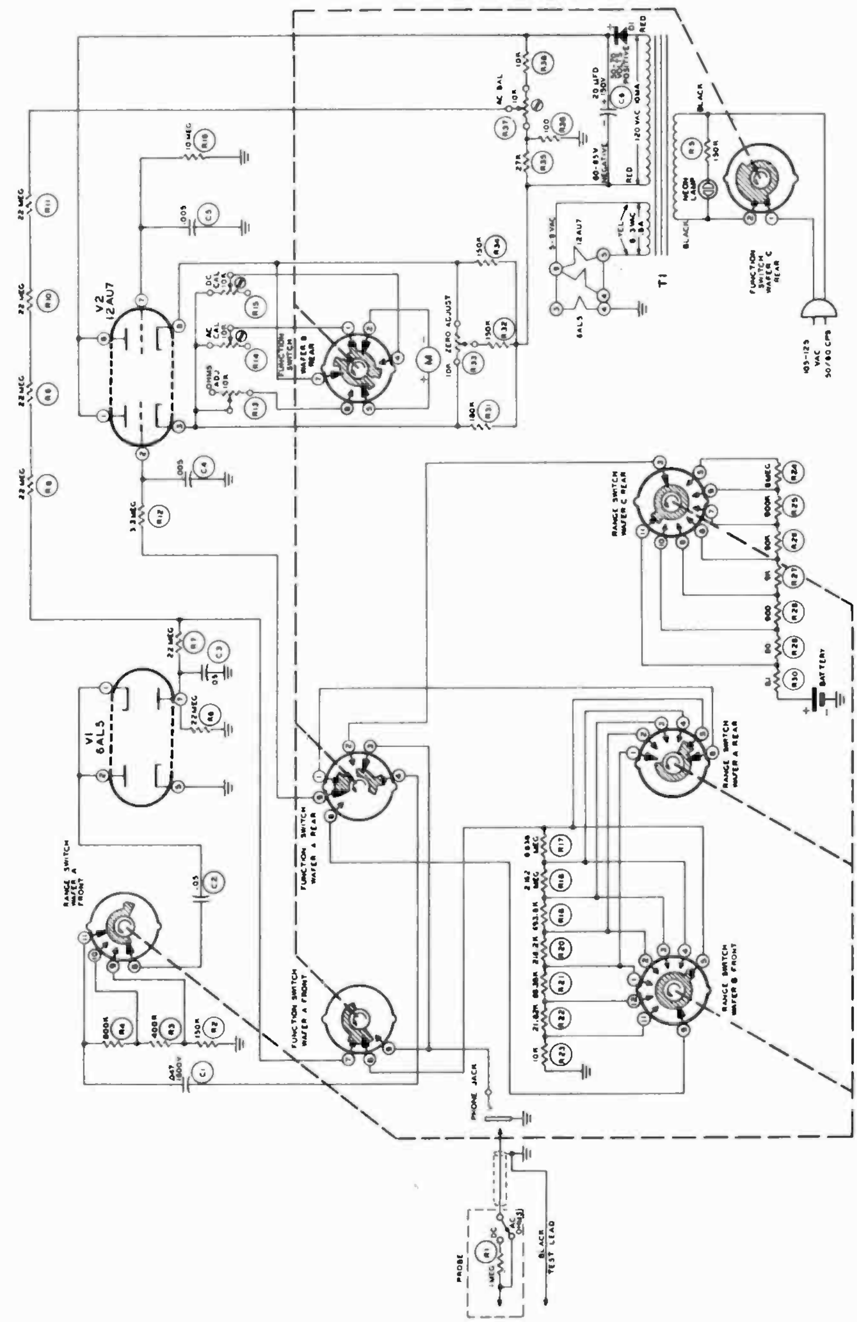


Figure 3

HEATH PEAK-TO-PEAK V.T.V.M.
MODEL IM-13

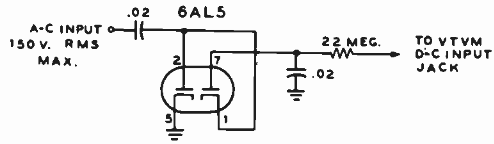


Figure 4

FULL-WAVE RECTIFIER FOR V.T.V.M.

Test Probes Auxiliary *test probes* may be used with the v.t.v.m. to extend the operating range, or to measure radio frequencies with high accuracy. Shown in figure 5 is a radio-frequency probe which provides linear response to over 150 MHz. A 1N270 is used as a rectifier, and d-c isolation is provided by a .005-μfd capacitor. The components of the detector are mounted within a shield at the end of a length of coaxial line, which terminates in the *d-c input* jack of the v.t.v.m. The readings obtained are rms, and should be multiplied by 1.414 to convert to peak readings.

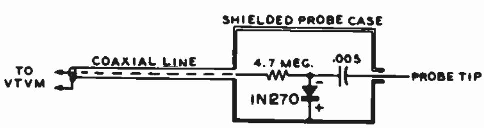


Figure 5

R-F PROBE SUITABLE FOR USE IN 1 kHz-150 MHz RANGE

The Solid-State Voltmeter The circuit of a solid-state voltmeter is shown in figure 6. The general operation of this circuit is similar to that of figure 3. The three input circuits (AC Volts, DC Volts and Ohms) are shown on the left-hand side of the schematic. These circuits perform the switching attenuation and rectification required to supply the correct voltage to the detecting and indicating circuits at the right-hand side of the schematic. Approximately 0.5 volt is required at the gate of FET input transistor Q_1 for full-scale deflection of the meter. Voltages greater than 0.5 are attenuated in the input circuits.

Input transistor Q_1 has a very high impedance gate circuit which keeps it from loading the input switching and attenuating circuits. A constant current source (Q_4), is used in place of a resistor in the source circuit of the FET. *Bias adjust* and *zero adjust* controls are provided to set the meter pointer to zero when no signal voltage is passed through the input circuits.

Transistors Q_2 and Q_3, together with a 3.3-megohm series input resistor are used to protect the input FET from accidental overload. The reverse-connected transistors perform like a 9-volt zener diode, short circuiting higher input voltages by virtue of the drop across the series input resistor.

The meter movement is driven by the voltage applied to the output circuit by Q_1. The source of Q_1 is directly coupled to the base of Q_5. Transistors Q_5 and Q_6 are used as emitter followers to provide the power to drive the meter. When the circuit is properly adjusted, no current flows through the meter without a signal being applied to Q_1.

Since the source current of Q_1 is constant and Q_5 is a direct-coupled emitter follower, voltage variations at the input of Q_1 are transferred to the meter circuit; a negative going input signal causing the meter pointer to move backwards. Meter polarity may be reversed so that negative going input voltages cause forward meter readings. The *zero adjust* control, moreover, varies the gate bias on Q_1 by introducing a positive voltage in series with the source which is returned to a "floating" negative return bus.

The Digital Voltmeter The *digital voltmeter* (d.v.m.) reads out a measurement in discrete numerals rather than as a pointer deflection on a continuous scale, as is commonly done in analog devices. The direct readout reduces reading error, eliminates meter parallax and increases reading speed. In common with the electronic voltmeter, the digital voltmeter features range changing, polarity changing, and overload protection. In addition the d.v.m. permits a permanent record to be made of measurements by the use of printout devices, card and tape punches, and magnetic-tape equipment. With data in digital form, it may be further processed with no loss of accuracy.

The heart of the electronic d.v.m. is the circuitry which converts analog voltage to a digital form, known as *analog-to-digital conversion* (ADC). Various forms of cir-

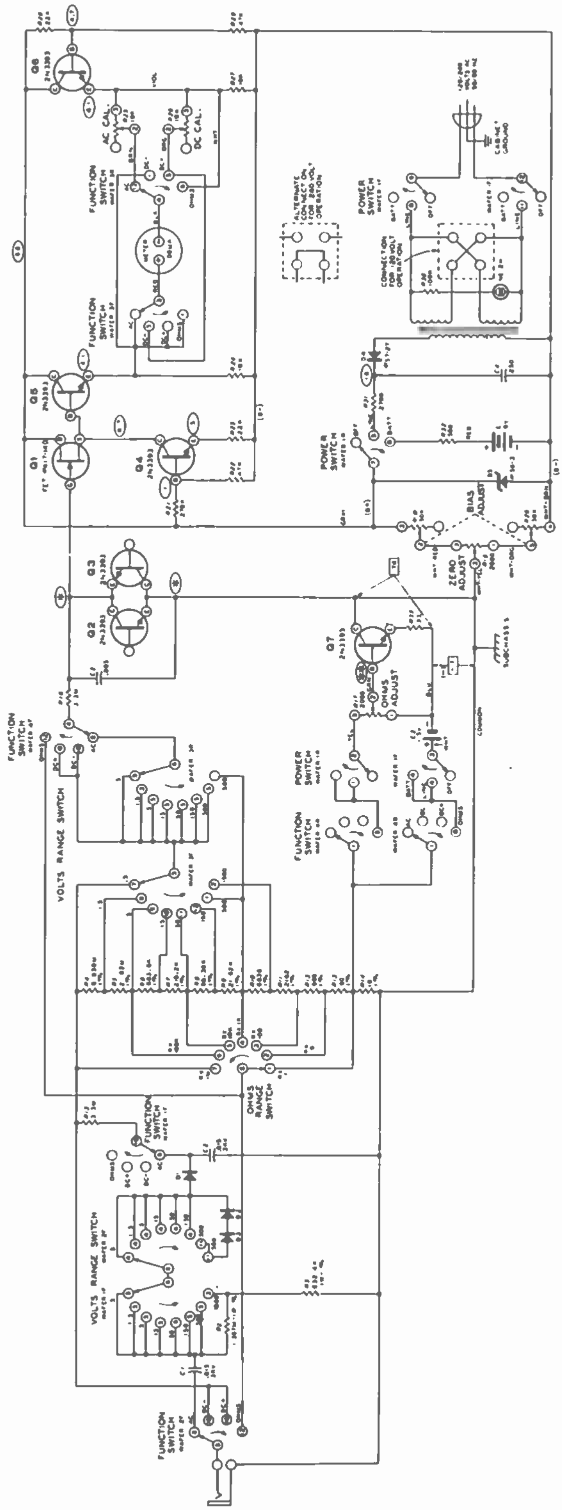


Figure 6
HEATH SOLID-STATE VOLTMETER IM-16

cuitry are in use to make this conversion. Among these circuit configurations are the *ramp*, *integrating*, and *dual-slope* variations.

The ramp circuit is a voltage-to-time conversion wherein the instrument measures the length of time it takes for a linear ramp of voltage to become equal to the unknown input voltage after starting from a known level. This time period is measured with an electronic time interval counter and displayed on an in-line indicating device.

The integrating circuit is a voltage-to-frequency conversion wherein the instrument measures the true average of the input voltage over a fixed encoding time instead of measuring the voltage at the end of the encoding time as do ramp type units, and others. The voltage is converted to a frequency by means of an integrating circuit. A feedback control governs a clock generator and the average voltage of the clock pulse train is equal to the d-c input voltage.

The dual-slope instrument makes a two-step measurement that combines integration in the first step with automatic comparison of its internal standard in the second. This technique rejects noise because of integration and achieves good stability from comparison with the standard. Direct numerical readout is accomplished with numerical display tubes or solid-state light-emitting devices.

A form of the dual-slope digital voltmeter is the *Heath Digital Multimeter IM-102.* This instrument measures a-c and d-c volts, a-c and d-c current, and resistance. All of the inputs are scaled to, or converted to, the basic measuring ranges of 200 millivolts or 2 volts, depending on the setting of the range switch. The measuring circuit is a high-impedance bipolar analog-to-digital converter. Resistance is measured by passing a scaled constant current through the unknown resistor and measuring the voltage drop across it. Alternating voltages are converted to d-c by an average-sensing, rms-calibrated, converter assembly. Current is measured by the voltage drop it establishes across a shunt network.

29-3 Power Measurements

Audio-frequency or radio-frequency power in a resistive circuit is most commonly and most easily determined by the indirect method, i.e., through the use of one of the following formulas:

$$P = EI, \quad P = E^2/R, \quad P = I^2R$$

These three formulas mean that if any two of the three factors determining power are known (resistance, current, voltage) the power being dissipated may be determined. In an ordinary 117-volt a-c line circuit the above formulas are not strictly true since the power factor of the load must be multiplied into the result—or a direct method of determining power such as a wattmeter may be used. But in a resistive a-f circuit and in a resonant r-f circuit the power factor of the load is taken as being unity.

For accurate measurement of a-f and r-f power, a *thermogalvanometer* or *thermocouple* ammeter in series with a noninductive resistor of known resistance can be used. The meter should have good accuracy, and the exact value of resistance should be known with accuracy. Suitable dummy-load resistors are available in various resistances in both 100-and 250-watt ratings. These are virtually noninductive, and may be considered as a pure resistance up to 30 MHz.

Sine-wave power measurements (r-f or single-frequency audio) may also be made through the use of a vtvm and a resistor of known value. In fact a vtvm of the type shown in figure 3 is particularly suited to this work. The formula, $P = E^2/R$ is used in this case. However, it must be remembered that a vtvm of the type shown in figure 3 indicates the *peak* value of the a-c wave. This reading must be converted to the rms or *heating* value of the wave by multiplying it by 0.707 before subsituting the voltage value in the formula. The same result can be obtained by using the formula $P = E^2/2R$. (*Note*: Some vtvm's are *peak reading* but are calibrated rms on the meter scale).

Power may also be measured through the use of a *calorimeter*, by actually measuring the amount of heat being dissipated. Through the use of a water-cooled dummy-load resistor this method of power output determination is being used by some of the most modern broadcast stations. But the method is too cumbersome for ordinary power determinations.

Power may also be determined *photometrically* through the use of a voltmeter, ammeter, incandescent lamp used as a load resistor, and a photographic exposure meter. With this method the exposure meter is used to determine the relative visual output of the lamp running as a dummy-load resistor and of the lamp running from the 117-volt a-c line. A rheostat in series with the lead from the a-c line to the lamp is used to vary its light intensity to the same value (as indicated by the exposure meter) as achieved as a dummy load. The a-c voltmeter in parallel with the lamp and ammeter in series with it is then used to determine lamp power input by: $P = EI$. This method of power determination is satisfactory for audio and low-frequency r.f. but is not satisfactory for vhf work because of variations in lamp efficiency due to uneven heating of the filament.

Finally, r-f power may be measured by means of a *directional coupler,* as discusssed later in this chapter.

Figure 7

2-KILOWATT DUMMY LOAD FOR 3-30 MHz

Load is built in case measuring 22" deep, 11" wide and 5" high. Meter is calibrated in watts against microampere scale as follows: (1), 22.3 μa. (5), 50 μa. (10), 70.5 μa. (15), 86.5 μa. (20), 100 μa. Scale may be marked off as shown in photograph. Calibration technique is discussed in text. Alternatively, a standing-wave bridge (calibrated in watts) such as "Micromatch" may be used to determine power input to load.

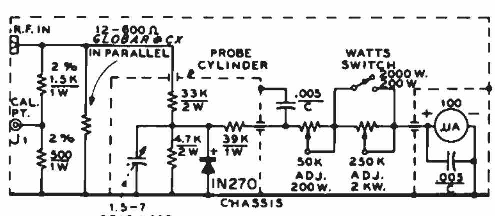
Vents in top of case, and 1/4-inch holes in chassis permit circulation of air about resistors. Unit should be fan-cooled for continuous dissipation.

Dummy Loads A suitable r-f load for power up to a few watts may be made by paralleling 2-watt composition resistors of suitable value to make a 50-ohm resistor of adequate dissipation.

A 2-kW dummy load having an SWR of less than 1.05 to 1 at 30 MHz is shown in figures 7, 8, and 9. The load consists of twelve 600-ohm, 120-watt *Globar type CX* noninductive resistors connected in parallel. A frequency-compensation circuit is used to balance out the slight capacitive reactance of the resistors. The compensation circuit is mounted in an aluminum tube 1" in diameter and 2 5/8" long. The tube is plugged at the ends by metal discs, and is mounted to the front panel of the box.

The resistors are mounted on aluminum T-bar stock and are grounded to the case at the rear of the assembly. Connection to the coaxial receptacle is made via copper strap.

The power meter is calibrated using a v.t.v.m. and r-f probe. Power is applied to the load at 3.5 MHz and the level is adjusted to provide 17.6 volts at "Calibration point." With the *Watts Switch* in the 200-watt position, the potentiometer is adjusted to provide a reading of 100 watts on the meter. In the 2000-watt position, the other potentiometer is adjusted for a meter reading of 200 watts. The excitation frequency is now changed to 29.7 MHz and the 17.6-volt level re-established. Adjust the frequency-compensating capacitor until meter again reads 100 watts. Recheck at 3.5 MHz and repeat until meter reads 100 watts at each frequency when 17.6-volt level is maintained.



NOTE: FIXED RESISTORS ARE *OHMITE "LITTLE DEVIL"* COMPOSITION UNITS.

Figure 8

SCHEMATIC, KILOWATT DUMMY LOAD

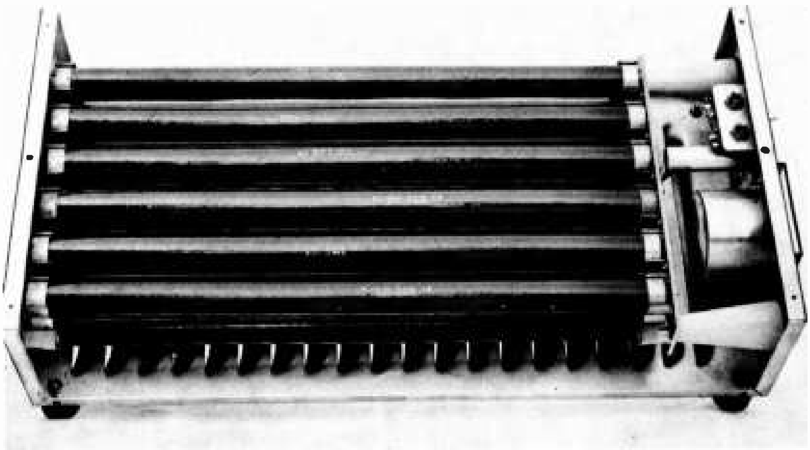


Figure 9

DUMMY-LOAD ASSEMBLY

Twelve Globar resistors (surplus) are mounted to aluminum "Tee" stock, six to a side, in fuse clips. Right end is supported by ceramic pillars from front panel. Probe, meter, and potentiometers are at right.

29-4 Measurement of Circuit Constants

The measurement of the resistance, capacitance, inductance, and Q (figure of merit) of the components used in communications work can be divided into three general methods: the impedance method, the substitution or resonance method, and the bridge method.

The Impedance Method The *impedance method* of measuring inductance and capacitance can be likened to the ohmmeter method for measuring resistance. An a-c voltmeter, or milliammeter in series with a resistor, is connected in series with the inductance or capacitance to be measured and the a-c line. The reading of the meter will be inversely proportional to the impedance of the component being measured. After the meter has been calibrated it will be possible to obtain the approximate value of the impedance directly from the scale of the meter. If the component is a capacitor, the value of impedance may be taken as its reactance at the measurement frequency and the capacitance determined accordingly. But the d-c resistance of an inductor must also be taken into consideration in determining its inductance. After the d-c resistance and the impedance have been determined, the reactance may be determined from the formula: $X_L = \sqrt{Z^2 - R^2}$. Then the inductance may be determined from: L equals $X_L/2\pi f$.

The Substitution Method The *substitution method* is a satisfactory system for obtaining the inductance or capacitance of high-frequency components. A large variable capacitor with a good dial having an accurate calibration curve is a necessity for making determinations by this method. If an unknown inductor is to be measured, it is connected in parallel with the standard capacitor and the combination tuned accurately to some known frequency. This tuning may be accomplished either by using the tuned circuit as a wavemeter and coupling it to the tuned circuit of a reference oscillator, or by using the tuned circuit in the controlling position of a two terminal oscillator such as a dynatron or transitron. The capacitance required to tune this first frequency is then noted as C_1. The circuit or the oscillator is then tuned to the *second harmonic* of this first frequency and the amount of capacitance again noted, this time as C_2. Then the distributed capacitance across the coil (including all stray capacitances) is equal to: $C_0 = (C_1 - 4C_2)/3$.

This value of distributed capacitance is then substituted in the following formula along with the value of the standard capacitance for either of the two frequencies of measurement:

$$L = \frac{1}{4\pi^2 f_1^2 (C_1 + C_0)}$$

The determination of an unknown capacitance is somewhat less complicated than the above. A tuned circuit including a coil, the unknown capacitor and the standard capacitor, all in parallel, is resonated to some convenient frequency. The capacitance of the standard capacitor is noted. Then the unknown capacitor is removed and the circuit re-resonated by means of the standard capacitor. The difference between the two readings of the standard capacitor is then equal to the capacitance of the unknown capacitor.

29-5 Measurements with a Bridge

The Wheatstone Bridge Experience has shown that one of the most satisfactory methods for measuring circuit constants (resistance, capacitance, and inductance) at audio frequencies is by

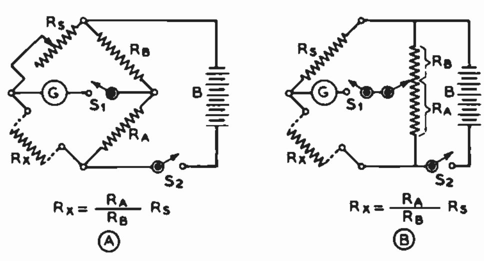


Figure 10

TWO WHEATSTONE BRIDGE CIRCUITS

These circuits are used for the measurement of d-c resistance. In A the "ratio arms" R_B and R_A are fixed and balancing of the bridge is accomplished by variation of the standard R_S. The standard in this case usually consists of a decade box giving resistance in 1-ohm steps from 0 to 1110 or to 11,110 ohms. In B a fixed standard is used for each range and the ratio arm is varied to obtain balance. A calibrated slide wire or potentiometer calibrated by resistance in terms of degrees is usually employed as R_A and R_B. It will be noticed that the formula for determining the unknown resistance from the known is the same in either case.

means of the a-c bridge. The *Wheatstone (d-c) bridge* is also one of the most accurate methods for the measurement of d-c resistance. With a simple bridge of the type shown in figure 10A it is entirely practical to obtain d-c resistance determinations accurate to four significant figures. With an a-c bridge operating within its normal rating as to frequency and range of measurement it is possible to obtain results accurate to three significant figures.

Both the a-c and the d-c bridges consist of a source of energy, a standard or reference of measurement, a means of balancing this standard against the unknown, and a means of indicating when this balance has been reached. The source of energy in the d-c bridge is a battery; the indicator is a sensitive galvanometer. In the a-c bridge the source of energy is an audio oscillator (usually in the vicinity of 1000 Hz), and the indicator is usually a pair of headphones. The standard for the d-c bridge is a resistance, usually in the form of a decade box. Standards for the a-c bridge can be resistance, capacitance, and inductance in varying forms.

Figure 10 shows two general types of the Wheatstone or d-c bridge. In A the so-called "ratio arms" (R_A and R_B) are fixed (usually in a ratio of 1-to-1, 1-to-10, 1-to-100, or 1-to-1000) and the standard resistor (R_S) is varied until the bridge is in balance. In commercially manufactured bridges there are usually two or more buttons on the galvanometer for progressively increasing its sensitivity as balance is approached. Figure 10B is the *slide-wire* type of bridge in which fixed standards are used and the ratio arm is continuously variable. The slide wire may actually consist of a moving contact along a length of wire of uniform cross section in which case the ratio of R_A to R_B may be read off directly in centimeters or inches, or in degrees of rotation if the slide wire is bent around a circular former. Alternatively, the slide wire may consist of a linear-wound potentiometer with its dial calibrated in degrees or in resistance from each end.

Figure 11A shows a simple type of a-c bridge for the measurement of capacitance and inductance. It can also, if desired, be used for the measurement of resistance. It

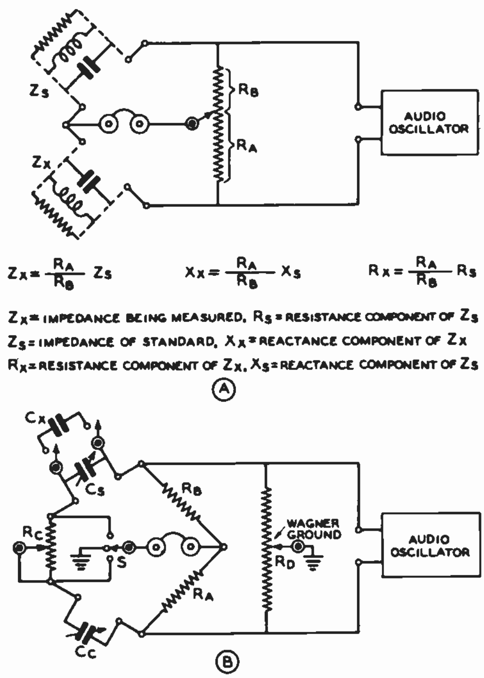


Figure 11
TWO A-C BRIDGE CIRCUITS

The operation of these bridges is essentially the same as those of figure 10 except that a-c is fed into the bridge instead of d-c and a pair of phones is used as the indicator instead of the galvanometer. The bridge shown at A can be used for the measurement of resistance, but it is usually used for the measurement of the impedance and reactance of coils and capacitors at frequencies from 200 to 1000 Hz. The bridge shown at B is used for the measurement of small values of capacitance by the substitution method. Full description of the operation of both bridges is given in the accompanying text.

is necessary with this type of bridge to use a standard which presents the same type of impedance as the unknown being measured: resistance standard for a resistance measurement, capacitance standard for capacitance, and inductance standard for inductance determination.

The Wagner Ground For measurement of capacitances from a few picofarads to about 0.001 μfd, a *Wagner-grounded substitution capacitance bridge* of the type shown in figure 11B will be found satisfactory. The ratio arms R_A and R_B should be of the same value within 1 percent; any value between 2500 and 10,000 ohms for both will be satisfactory. The two resistors R_C and R_D should be 1000-ohm wirewound potentiometers. C_S should be a straight-line capacitance capacitor with an accurate vernier dial; 500 to 1000 pf will be satisfactory. C_C can be a two- or three-gang broadcast capacitor from 700 to 1000 pf maximum capacitance.

The procedure for making a measurement is as follows: The unknown capacitor C_X is placed in parallel with the standard capacitor C_S. The *Wagner ground* (R_D) is varied back and forth a small amount from the center of its range until no signal is heard in the phones with the switch (S) in the center position. Then the switch (S) is placed in either of the two outside positions, C_C is adjusted to a capacitance somewhat greater than the assumed value of the unknown C_X, and the bridge is brought into balance by variation of the standard capacitor (C_S). It may be necessary to cut some resistance in at R_C and to switch to the other outside position of S before an exact balance can be obtained. The setting of C_S is then noted, C_X is removed from the circuit (but the leads which went to it are not changed in any way which would alter their mutual capacitance), and C_S is readjusted until balance is again obtained. The difference in the two settings of C_S is equal to the capacitance of the unknown capacitor C_X.

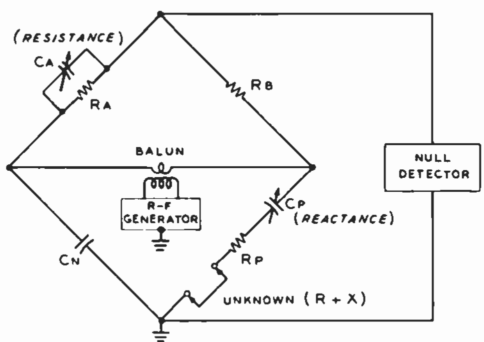
29-6 R-F Bridges

The basic bridge circuits are applicable to measurements at frequencies well up into the uhf band. While most of the null circuits used from d.c. to about 100 MHz are adaptations of the fundamental Wheatstone Bridge circuit, many other types of networks that can be adjusted to give zero transmission are employed at higher frequencies.

At very-high frequencies, where impedances can no longer be treated as lumped elements, null circuits based upon coaxial line techniques are used. The upper frequency limit of conventional bridge circuits using lumped parameters is determined by the magnitude of the residual impedance of the elements and the leads. The corrections for these usually become unmanageable at frequencies higher than 100 MHz or so.

The "General Radio" Bridge An *r-f bridge* suitable for use up to about 60 MHz is shown in figure 12. The bridge can measure resistances up to 1000 ohms and reactances over the range of plus or minus 5000 ohms at 1 MHz. The reactance range varies inversely as the frequency, and at other frequencies the reactance reading must be divided by the frequency in MHz. Measurements are made by a series-substitution method in which the bridge is first balanced by means of capacitors C_P and C_A with a short-circuit across the unknown terminals. The short is then removed, the unknown impedance connected in its place, and the bridge rebalanced. The unknown resistance and reactance values are then read from the difference between the initial and final balances.

A vhf variation of the r-f bridge provides direct measurements up to 500 MHz by sampling the electric and magnetic fields in a transmission line. Two attenuators are controlled simultaneously; one receives energy proportional to the electric field in the line, and the other receives energy proportional to the magnetic field. The magnitude of the unknown impedance is detemined by adjusting this combination for equal output from each attenuator. The two equal signals may also be applied to opposite ends of another transmission line, and phase angle can be determined from their point of cancellation.



$$\text{RESISTANCE} \quad R_X = R_B \times \frac{(C_{A2} - C_{A1})}{C_N}$$

$$\text{REACTANCE} \quad X_X = \frac{1}{\omega}\left(\frac{1}{C_{P2}} - \frac{1}{C_{P1}}\right)$$

Figure 12

THE "GENERAL RADIO" R-F BRIDGE

This bridge is suitable for r-f measurements up to 60 MHz or so. Calibrated reactance (C_P) and resistance (C_A) dials allow direct measurements at 1 MHz. At other frequencies reactance reading must be divided by the frequency in MHz. Wide-band balun input transformer allows bridge to be driven from signal generator via a coaxial line.

Above 500 MHz, impedance measurements are normally determined by inserting a detector probe in a slotted section of transmission line, as discussed in the next section of this chapter.

29-7 Antenna and Transmission-Line Instrumentation

The degree of adjustment of any amateur antenna can be judged by a study of the standing-wave ratio on the transmission line feeding the antenna. Various types of instruments have been designed to measure the ratio of forward to reflected power by sampling the r-f incident and reflected waves on the transmission line, or to measure the actual radiation resistance and reactance of the antenna in question. The most important of these instruments are the *slotted line*, the *directional coupler*, and the *r-f impedance bridge*.

The Slotted Line The relationship between the incident and the reflected power and standing wave present on a transmission line is expressed by:

$$K = \frac{1 + R}{1 - R}$$

where,

K = Standing-wave ratio,

R = Reflection coefficient, or ratio of relative amplitude of reflected signal to incident signal.

When measurements of a high degree of accuracy are required, it is necessary to insert an instrument into a section of line in order to ascertain the conditions existing within the shielded line. For most vhf measurements, wherein a wavelength is of manageable proportions, a *slotted line* is the instrument frequently used. Such an instrument, shown in figure 13, is an item of test equipment which could be constructed in a home workshop which includes a lathe and other metal-working tools. Commercially built slotted lines are very expensive since they are constructed with a high degree of accuracy for precise laboratory work. The slotted line consists essentially of a section of air-dielectric line having the same characteristic impedance as the transmission line into which it is inserted. Tapered fittings for the transmission line connectors at each end of the slotted line usually are required due to differences in the diameters of the slotted line

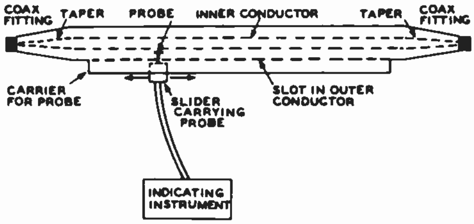


Figure 13

THE UHF SLOTTED LINE

The conductor ratios in the slotted line, including the tapered end sections should be such that the characteristic impedance of the equipment is the same as that of the transmission line with which the equipment is to be used. The indicating instrument may be operated by the d-c output of the rectifier coupled to the probe, or it may be operated by the a-c components of the rectified signal if the signal generator or transmitter is amplitude-modulated at a constant percentage.

and the line into which it is inserted. A narrow slot from ⅛-inch to ¼-inch in width is cut into the outer conductor of the line. A probe then is inserted into the slot so that it is coupled to the field inside the line. Some sort of accurately machined track or lead screw must be provided to ensure that the probe maintains a constant spacing from the inner conductor as it is moved from one end of the slotted line to the other. The probe usually includes some type of rectifying element whose output is fed to an indicating instrument alongside the slotted line.

The unfortunate part of the slotted-line system of measurement is that the line must be somewhat over one-half wavelength long at the test frequency, and for best results should be a full wavelength long. This requirement is easily met at frequencies of 420 MHz and above where a full wavelength is 28 inches or less. But for the lower frequencies such an instrument is mechanically impractical.

The Directional Coupler The r-f voltage on a transmission line may be considered to have two components. The *forward component* (incident *component*) and the *reverse component* (reflected component). The reverse component is brought about by operation of the line when terminated in a load that is unequal to the characteristic impedance of the line.

A *directional coupler* is an instrument that can sense either the forward or reflected components in a transmission line by taking advantage of the fact that the reflected components of voltage and current are 180 degrees out of phase while the forward components of voltage and current are in phase.

The directional coupler is inserted in the transmission line at an appropriate location. For a coaxial line, the instrument consists of a short section of line containing a small loop coplanar with the inner conductor (figure 14). The loop is connected through a resistor to the outer conductor, and this resistor is capacitively coupled to the inner conductor of the line. The voltage appearing across the series arrangement of loop and resistor is measured when the voltage across the resistor and the voltage induced in the loop are aiding and again when they are in opposition to each other. By rotating the loop through 180 degrees, the readings may be used to determine the amount of mismatch and the power carried by the line. Operation is substantially independent of load impedance and meter impedance at any frequency within the useful range of the instrument.

When the directional coupler is used to measure the SWR or the reflection coefficient on the line, the value obtained for these quantities depends only on the *ratio* of the two measured voltages. Power measurements are more stringent, since the absolute value of transmission line voltage must be determined and construction of a simple, compact r-f voltmeter that presents a linear reading over a wide frequency range and at various power levels is not simple.

In order to sample forward and reverse power, it is necessary to reverse the orientation of the directional coupler in the line, or to employ two couplers built in one unit but oriented oppositely. It is necessary, moreover, to have both couplers identical in *coupling factor* and *directivity.*

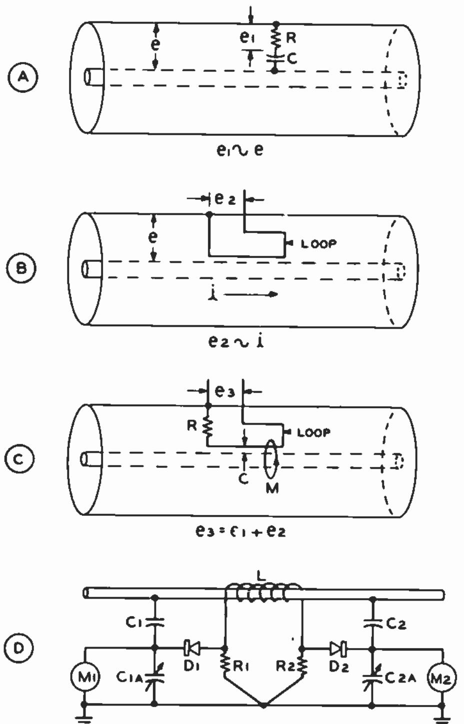


Figure 14

THE DIRECTIONAL COUPLER

The directional coupler (reflectometer) is a coaxial-line section containing an r-f voltmeter which reads the incident or reflected component of voltage, depending on the position of the pickup device in the line.
A—Voltage relationships for a series resistance-capacitance combination placed between the conductors of a coaxial line; e_1 is proportional to e.
B—Loop coupled to inner conductor will give voltage (e_2) proportional to current flowing in line (i).
C—Representation of reflectometer. Capacitance is provided by proximity of loop to inner conductor.
D—Double reflectometer provides simultaneous measurement of incident and reflected voltages. Ferrite core is placed around center conductor, with secondary winding acting as loop M.

The fraction of forward power that is sampled by the coupler is termed the *coupling factor,* and the directivity is the ability of the coupler to discriminate between opposite directions of current flow. If, for example one percent of the power is coupled out, the coupling factor is 20 decibels. If the coupler is now reversed to sample the power in the reverse direction, it may couple out, say 0.001 percent of the forward power even though there may be actually no reflected power. It is thus coupling out an amount of power 50 decibels below the power in the line. The discrimination between forward and reverse power is the difference between the coupled values, or 30 decibels. A *directivity* of 30 db is common for better types of reflectometers and SWR measurements derived from the measured reflection coefficient are sufficiently accurate for adjustment of simple beam antennas. It should be noted, however, that it is difficult to make measurements with any degree of accuracy at low SWR values with inexpensive directional couplers, because the directivity power ratio at SWR values below about 1.5/1 or so falls within the error limits of directivity capability of all but the best and most expensive reflectometers.

The SWR Bridge The SWR bridge is a useful device for determining the standing-wave ratio on, and the power transmitted along, a transmission line. When the SWR on a given line is unity, the line is terminated in a pure resistance equal to the characteristic impedance of the line. If the line and terminating load are made part of an r-f bridge circuit, the bridge will be in a balanced condition when the SWR is unity (figure 15). A sensitive r-f voltmeter connected across the bridge will indicate balance and the magnitude of bridge unbalance, and may be calibrated in terms of SWR, power, or both. It may be seen in figure 15A that the meter reading is proportional to bridge unbalance, and is thus proportional to the reflected power and is not influenced by the forward power in the circuit. The meter will read zero if, and only if, the transmission line is properly terminated in Z_l so that $Z_l = Z_o$ of

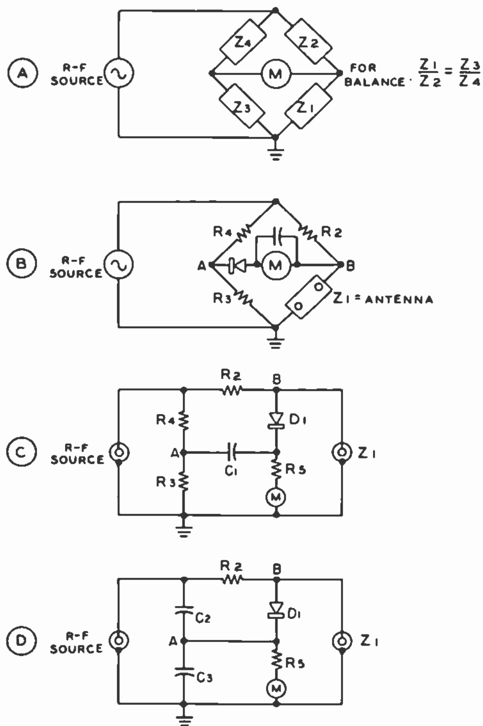


Figure 15

THE BRIDGE DIRECTIONAL COUPLER

A—When r-f bridge is balanced any change in load (Z_l) will result in bridge unbalance and cause a reading on meter M. Reading is due to reflected voltage. SWR may be derived from:

$$\text{SWR} = \frac{E_o - E_r}{E_o + E_r}$$

where,
E_o equals incident voltage,
E_r equals reflected voltage.

B—Equivalent bridge circuit. Bridge must be individually calibrated since performance differs from formula due to nonlinearity of voltmeter, circuit loading, and line discontinuity introduced by presence of bridge.

C, D—Practical bridge circuits having one side of meter grounded to line.

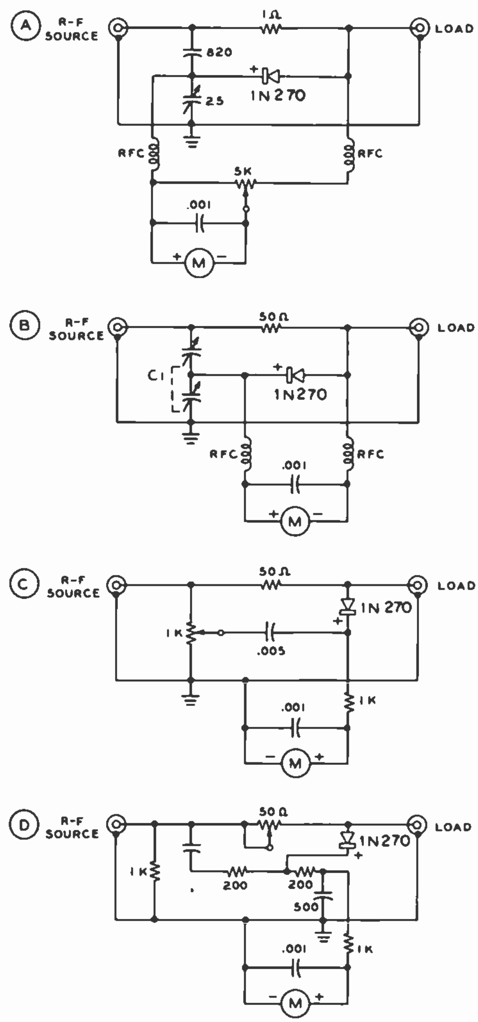


Figure 16

SWR BRIDGES

A—Micromatch bridge.
B—Capacitance ratio bridge.
C—Antennascope.
D—Antennascope with calibrating resistor in active leg of bridge.
Note: Meter M may be 0-500 d-c microammeter.

the line, so as to have unity standing-wave ratio.

Various forms of the SWR resistance bridge exist as shown in the illustration, but all of them are based on the principle of measurement of bridge balance by means of a null-indicating meter. Circuit B consists of two resistive voltage dividers across the r-f source, with an r-f voltmeter reading the difference of potential across the points A and B. Circuit C is identical, but redrawn so as to show a practical layout for measurement in a coaxial system with one side of the generator and the r-f voltmeter at ground potential. Circuit D is similar, except that one of the voltage dividers of the bridge is capacitive instead of resistive.

SWR Bridge Designs Various forms of the SWR bridge are shown in figure 16. Circuit A is the *Micromatch* capacitance bridge. In order to pass appreciable power through the bridge, the series resistor is reduced to one ohm, thus requiring the capacitance divided to maintain about the same ratio as set in the resistive arm. For a 50-ohm transmission line, the transformation ratio is 50/1, and the 25-pf variable capacitor must be set at a value corresponding to about fifty times the reactance of the 820-pf capacitor. The power-handling capability of the bridge is limited by the dissipation capability of the 1-ohm resistor.

Circuit B incorporates a differential capacitor to obtain an adjustable bridge ratio. The capacitor may be calibrated in terms of the unknown load and may be used to indicate resistive loads in the range of 10 to 500 ohms. The bridge has an advantage over the circuits of illustrations A and C in that it may be used in the manner of a simple impedance bridge to determine the radiation resistance of a *resonant* antenna. The bridge is placed at the antenna terminals, and the frequency of the driving source and the setting of the differential capacitor are varied to produce a null indication on the meter. The null occurs at the resonant frequency of the antenna, and the radiation resistance at that frequency may be read from the instrument.

A less-expensive variation of the variable r-f bridge is shown in illustrations C and D and is called the *Antennascope*. The Antennascope is a variable bridge making use of a (relatively) noninductive potentiometer in one leg. These simple instruments are useful in antenna adjustment as they indicate the resonant frequency of the antenna and the approximate radiation resistance of the driven element at this frequency. At other than the resonant frequency, the antenna exhibits a reactive component and the null of the instrument will not be complete. Even so, at the low values of impedance encountered in most amateur beam antennas, the readings obtained at frequencies off resonance approximate the resistive component of the radiation resistance of the antenna.

Construction information for a practical Antennascope and other SWR instruments will be described in the following section of this Handbook.

29-8 Practical SWR Instruments

Simple forms of the directional coupler and the SWR bridge are suited to home construction and will work well over the range of 1.8 to 148 MHz. No special tools are needed for construction and calibration may be accomplished with the aid of a handful of 1-watt composition resistors of known d-c value resistance.

The Antennascope The *Antennascope* is a modified SWR bridge in which one leg of the bridge is composed of a noninductive variable resistor (figure 16D). This resistor is calibrated in ohms, and when its setting is equal to the radiation resistance of a resonant antenna under test, the bridge is in a balanced state. If a sensitive voltmeter is connected across the bridge, it will indicate a voltage null at bridge balance. The radiation resistance of the antenna may then be read directly from the calibrated dial of the instrument.

When the test antenna is nonresonant, the null indication on the Antennascope will be incomplete. The frequency of the exciting signal must then be altered to the resonant frequency of the antenna to obtain accurate readings of radiation resistance. The resonant frequency of the antenna, of course, is also determined by this exercise.

The circuit of the Antennascope is shown in figure 18. A 100-ohm noninductive potentiometer (R_1) serves as the variable leg of the bridge. The other legs are composed of the 200-ohm composition resistors and the radiation resistance of the antenna. If the radiation resistance of the antenna or external load under test is 50 ohms, and the potentiometer is set at midscale, the bridge is balanced and the diode voltmeter will read zero. If the radiation resistance of the antenna is any other value between about 10 and 100 ohms, the bridge may be balanced to this new value by varying the setting on the potentiometer, which is calibrated in ohms.

Building the Antennascope—The Antennascope is constructed within an aluminum

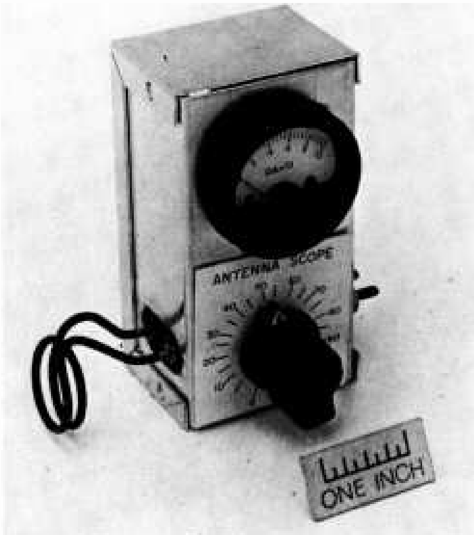


Figure 17
THE ANTENNASCOPE

The antennascope may be used to measure the resonant radiation resistance of antennas at frequencies up to 150 MHz. Grid-dip oscillator is coupled to input loop of antennascope and antenna under test is connected to output terminals with short, heavy leads.

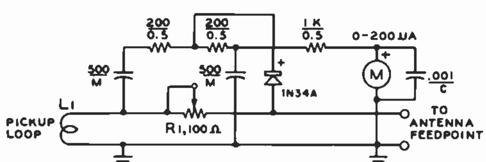


Figure 18
SCHEMATIC, ANTENNASCOPE

R_1—100-ohm composition potentiometer. Ohmite AB or Allen-Bradley type J linear taper
L_1—2 turns brass wire to fit gdo coil. See photos
M—0-100 μa d-c meter

box chassis measuring about 4″ × 2″ × 1½″, and placement of the major components may be seen in the photographs. A 1¼-inch diameter hole is drilled in the lower portion of the panel and the variable potentiometer is mounted in this hole on a thin piece of insulating material such as micarta or bakelite. The terminals of the potentiometer and the case are at r-f potential, so it is essential for proper bridge operation to have a minimum of capacitance between the potentiometer and ground.

The two 200-ohm, ½-watt resistors should be matched on an ohmmeter, and a number of the 500-pf capacitors should be checked on a bridge to find two units of equal capacitance. The exact value of resistance and capacitance in either case is not critical, it is only necessary that the companion units be equal in value. Care should be taken when soldering the small resistors in the circuit to see that they do not become overheated, causing the resistance value to permanently change. In like manner, the germanium diode should be soldered in the circuit using a pair of long-nose pliers as a heat sink to remove the soldering heat from the unit as rapidly as possible.

As shown in the photographs, copper strap cut from flashing stock is used for wiring the important r-f leads. The output leads terminate in an insulated terminal strip on one side of the box and the input coupling loop is made of a section of brass rod,

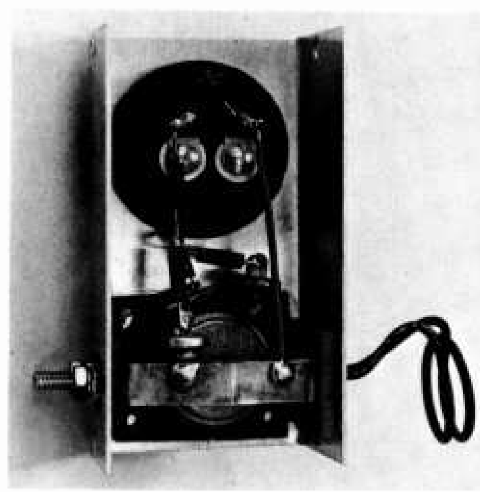


Figure 19
INTERIOR OF ANTENNA SCOPE

Strap connection is made between common input and output terminals. Grid-dip oscillator coupling loop is at right.

which is tapped at each end for 6-32 machine nuts. The loop is bent and positioned so as to slip over the coil of a grid-dip oscillator used as the driving source.

Testing the Antennascope—When the instrument is completed, a grid-dip oscillator may be coupled to the input link. The oscillator should be set somewhere in the 10-MHz to 20-MHz range and coupling is adjusted to obtain a half-scale reading on the meter of the Antennascope. Various values of precalibrated 1-watt composition resistors ranging from 10 to 90 ohms should be placed across the output terminals of the Antennascope and the potentiometer adjusted for nulls on the indicating meter. The settings of the potentiometer may then be marked on a temporary paper dial and, by interpolation, 5-ohm points can be marked on the scale for the complete rotation of the control. The dial may then be removed and inked.

This calibration will hold to frequencies well above the 2-meter band, but as the internal lead inductance of the Antennascope starts to become a factor, it will no longer be possible to obtain a complete null on the indicating meter. Wired as shown, the meter null begins to rise off zero in the region of 150 MHz.

Using the Antennascope—The Antennascope is coupled to a grid-dip oscillator by means of the input link. Additional turns may need to be added to the link to obtain sufficient pickup below 7 MHz or so. Enough coupling should be obtained to allow at least ¾-scale reading on the meter with no load connected to the measuring terminals. For general use, the measuring terminals of the instrument are connected across the antenna terminals at the feedpoint. Either a balanced or unbalanced antenna system may be measured, the "hot" lead of the unbalanced antenna connection to the ungrounded terminal of the Antennascope. Excitation is supplied from the grid-dip oscillator and the frequency of excitation and the Antennascope control dial are varied until a complete meter null is obtained. The frequency of the source of excitation now indicates the resonant frequency of the antenna under test, and the approximate radiation resistance of the antenna may be read upon the dial of the Antennascope.

On measurements made on 40- and 80-meter antennas it may be found impossible to obtain a complete null on the Antennascope. This is usually caused by pickup of a nearby broadcast station, in which case the rectified signal of the station will obscure the null action of the Antennascope. This action is only noticed when antennas of large size are being checked.

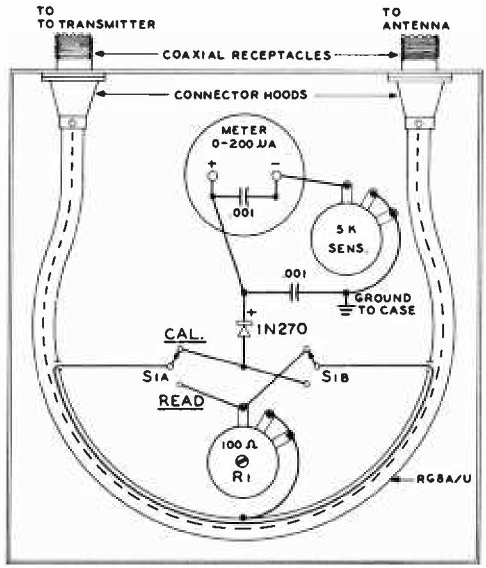


Figure 20

MONIMATCH

R_1—100-ohm composition potentiometer. Ohmite AB, or Allen Bradley type J, linear taper
S_1—Dpdt rotary switch. Centralab 1462
Case—5" × 7" × 2" chassis with back plate.

The Antennascope is designed to be used directly at the antenna terminals without an intervening feedline. It is convenient to mount the instrument and the grid-dip oscillator as a single package on a strip of wood. This unit may then be carried up the tower and attached to the terminals of the beam antenna. It is also possible to make remote measurements on an antenna with the use of an electrical half-wavelength of transmission line placed between the Antennascope and the antenna terminals.

The Monimatch The *Monimatch* is a dual reflectometer constructed from a length of flexible coaxial transmission line (figure 20). The heart of the Monimatch is a pickup line made from a 14-inch length of RG-8A/U coaxial cable. The coupling loop of this special section is a piece of No. 22 enamel or *formvar* covered wire slid under the flexible outer shield of the coaxial line for a distance of about eight inches. The coaxial pickup line is then conveniently wound around the inside walls of the mounting box so that the protruding ends of the coupling loop fall adjacent to the simple switching circuit. The coupling loop and center conductor of the coaxial line form a simple reflectometer terminated at either end by a noninductive potentiometer. Choice of termination is determined by the panel switch. When the potentiometer is adjusted to the balance point, the bridge is calibrated and ready for use. The selector switch permits reading forward or reverse power in the coaxial line and an SWR of unity is indicated by a null reading on the meter of the instrument.

The special coaxial pickup loop is easily made. A 14-inch length of RG-8A/U cable is trimmed square at the ends and the outer vinyl jacket is carefully removed. Two holes to pass the pickup wire are carefully made in the outer braid of the section with the aid of an awl or needle. Be careful not to break the fine wires of the braid. The holes are made 8 inches apart, and centered on the section. The outer shield is next bunched up a bit to loosen it and a length of No. 22 wire is threaded under the braid, in and out of the holes. A stiff copper wire may be threaded through the holes and used as a needle to pass the flexible copper wire under the braid. Finally, the braid is smoothed out to its original length and the pickup wire checked with an ohmmeter to make sure that no short exists between the braid and the wire. The braid is then wrapped with vinyl tape at the two holes. The last step is to solder connector hoods and coaxial receptacles on each end of the line, making the assembly "r-f tight."

The special line may now be mounted in the instrument case, along with the various other components, as shown in the illustration. The calibrating potentiometer is mounted on an insulating plate in the center of a one-inch hole to reduce the capacity of the unit to ground. The coaxial line should be grounded only at the coaxial receptacles, and should otherwise be wrapped with vinyl tape to prevent it from shorting to the case or other components.

A noninductive 52-ohm dummy load is attached to the output of the Monimatch and it is driven from an r-f source. Place the panel switch in the *Calibrate* position and adjust the *sensitivity* control for a half-scale reading of the meter. Now switch to the *Read* position and adjust the sensitivity control for full-scale reading. Adjust the *Calibrate* potentiometer in the back of the Monimatch for a null in the meter reading —it should be very close to zero on the scale. Switch back to *Calibrate* again and once again adjust the sensitivity control for full-scale meter reading. Finally, switch once again to *Read* and re-null the meter with the *Calibrate* potentiometer. The Monimatch is now ready for use.

Using the Monimatch—The Monimatch is inserted in the coaxial line to the antenna, power is applied and the switch set to *Calibrate* position. The sensitvity control is adjusted for full-scale reading and the switch is thrown to the *Read* position. Adjustments to the antenna may now be made to reach an SWR of unity, at which point the meter reading will be at maximum null, or close to zero. If desired, the Monimatch may be calibrated in terms of SWR by observing the reading when various values of noninductive composition resistors of known value are measured with the device.

A Practical Reflectometer The *reflectometer* is an accurate, inexpensive and easily constructed instrument for the experimenter. Shown in this section is a practical reflectometer made from a short section of coaxial transmission line. It is designed for use with output power of up to 2000 watts and at frequencies up to 150 MHz. An easily wound toroid transformer is used for a pickup element, in conjunction with two reverse-connected diode voltmeters, affording quick indication of forward and reverse conditions within the transmission line. The instrument is of the type shown in figure 14D. One voltmeter reads the incident component of voltage and the other reads the reflected component. The magnitude of standing-wave ratio on the transmission line is the ratio of these two components.

The upper frequency limit of the reflectometer is determined by the dimensions of the pickup loop which should be a small fraction of a wavelength in size. When used

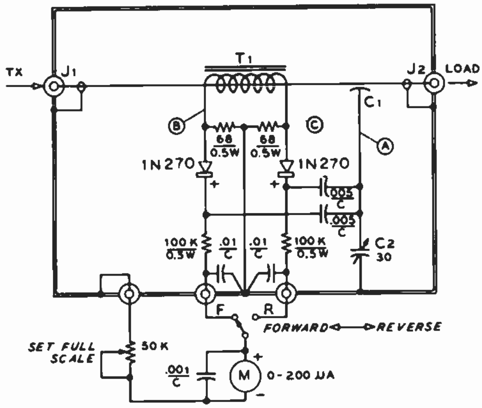


Figure 21

REFLECTOMETER

C_1—Sleeve formed of #28 tinned wire wrapped around inner dielectric of line for 3/8-inch length. See text
T_1—40 turns #28 insulated wire equally spaced around toroid core, Q-1 material. Indiana-General CF-114, 1.25" diameter × 0.38" thick. See figure 23 for assembly

to measure SWR, the resultant figure depends on the ratio of two measured voltages which are usually valid figures regardless of variations in load impedance and frequency. When used as a wattmeter, the absolute transmisssion-line voltage must be measured and the detection devices must have a flat frequency response with diodes operating in the square-law region for widest frequency coverage.

When used for SWR measurements, calibration of the reflectometer is not required since relative readings indicate the degree of mismatch and all system adjustments are conducted so as to make this ratio as high as possible, regardless of the absolute values. Power measurements may be made if the instrument is calibrated against a known dummy load in both the forward and reverse directions. The reflectometer may be left in the transmission line to indicate SWR and relative power output of the transmitter.

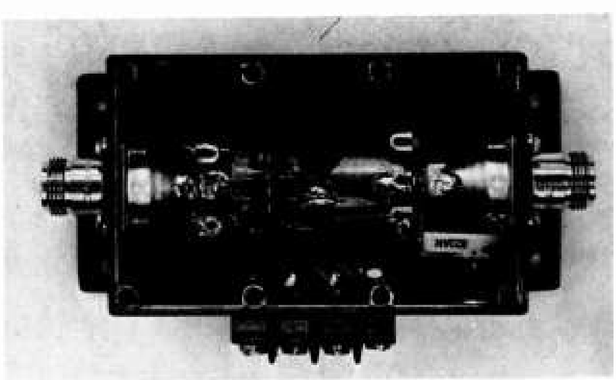
Building the Reflectometer—Assembly of the reflectometer is shown in figure 24. A short length of coaxial line of the chosen impedance is trimmed to length. The outer insulation and outer braid are cut with a sharp knife for a distance of about 3/4 of an inch at the center of the line, exposing a section of the inner dielectric. Around the dielectric a length of No. 28 tinned wire is wound to form a sleeve about 3/8-inch long for 50-ohm cable. If 70-ohm cable is used, the sleeve should be about 5/8-inch long. The sleeve is tinned and forms capacitor C_1 to the inner conductor. A short length of insulated wire is soldered to the sleeve (*lead A*). The capacitor is now wrapped with vinyl tape. Next, a short section of thin copper shim stock is wrapped over the tape to form a simple Faraday shield which ensures that the coupling between the primary of T_1 (the inner conductor of the coaxial line) and the secondary (the winding on the ferrite core) is inductive and not capacitive. *One end* of the shield is carefully soldered to the outer braid of the coaxial line and the other end is left free.

The ferrite core is now wrapped with vinyl tape and 40 turns of No. 28 insulated wire are evenly wrapped around the core. The core is then slipped over the cable sec-

Figure 22

INTERIOR VIEW OF REFLECTOMETER

Complete assembly including accessory components is placed in cast aluminum box, 4 × 2½" × 1½" (Pomona Electronic #2904). Calibrating capacitor is adjustable through small hole drilled in box.



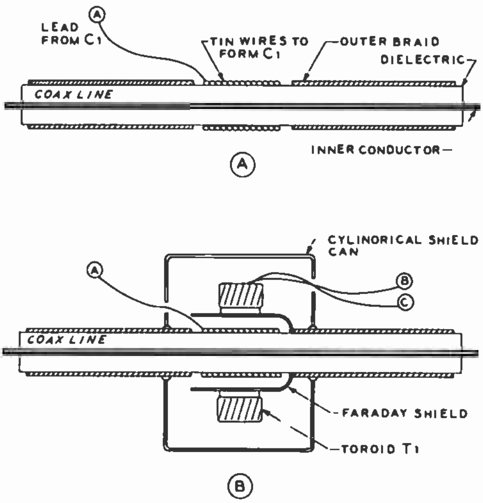


Figure 23

A—Assembly of coaxial capacitor C_1.
B—Assembly of capacitor, Faraday shield and toroid transformer T_1. Leads A, B, and C connect as shown in figure 21.

tion and positioned directly above capacitor C_1. The reflectometer section is then completed by forming a copper shield around the toroid assembly. In this case, the shield is made up of two copper discs soldered to the cable braid, over which is slipped a copper cylinder made of thin shim stock. The cylinder and end rings are soldered into an inclusive shield, as shown in the photograph, with the three pickup leads passing through small holes placed in the cylindrical end sections.

The reflectometer and associated components are placed in an aluminum box (figure 22) having a terminal strip attached for connection to an external reversal switch and meter. Final adjustment is accomplished by feeding power through the reflectometer into a dummy load having a low value of SWR and adjusting capacitor C_2 for minimum meter indication when the instrument is set for a reflected-power reading.

29-9 Frequency and Time Measurements

All frequency and time measurements within the United States are based on data transmitted from the *National Bureau of Standards*. Several time scales are used for time measurement: (1)—*Universal Time* (*UT*). Universal time, or *Greenwich Mean Time* (GMT), is a system of mean solar time based on the rotation of the earth about its axis relative to the position of the sun. Several UT scales are used: uncorrected astronomical observations are denoted Uϕ; the UT time scale corrected for the earth's polar variation is denoted UT1; the UT1 scale corrected for annual variation in the rotation of the earth is denoted UT2. Time signals tansmitted by standard stations are generally based on the UT2 time scale. Although UT is in common use, it is nonuniform because of changes in the earth's speed of rotation. (2)—*Ephemeris Time* (ET). Scientific measurements of precise time intervals require a uniform time scale.

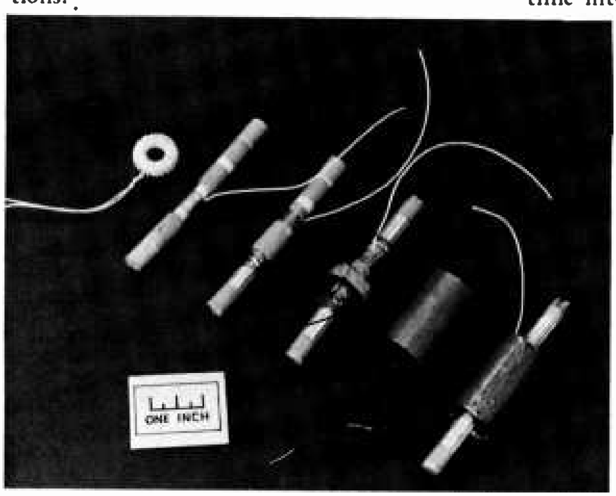


Figure 24

ASSEMBLY SEQUENCE OF REFLECTOMETER UNIT

Left-to-Right—Toroid-core transformer T_1, coaxial capacitor assembly, Faraday shield, completed unit, outer shield, transformer with attached leads.

The fundamental standard of constant time is defined by the orbital motion of the earth about the sun and is called Ephemeris time, and is determined from lunar observations. (3)—*Atomic Time* (AT). Molecular and atomic resonance characteristics can be used to provide time scales which are apparently constant and nearly equivalent to ET. The designation A.1 has been given to the time scale derived from the zero-field resonance of *cesium*. The U. S. Frequency Standard at Boulder, Colorado, is maintained by reference to the A.1 time scale.

Standard Radio Frequency and Time Signals High- and low-frequency time signals are broadcast on standard frequencies in the United States by the *National Bureau of Standards* over radio stations WWV, WWVB, and WWVL (located near Fort Collins, Colorado) and WWVH (located near Kekaha, Kaui, Hawaii). The broadcasts of WWV may also be heard by telephone by dialing (303) 499-7111, Boulder, Colorado.

Stations WWV, WWVH, and WWVL broadcast nominal frequencies and time consistent with the internationally agreed upon time scale, *Universal Coordinated Time* (UTC). WWV broadcasts on 2.5, 5, 10, 15, 20, and 25 MHz; while WWVH broadcasts on all these frequencies except 25 MHz. Transmissions are continuous. WWVB broadcasts *Stepped Atomic Time* (SAT) on the standard frequency of 60 kHz and WWVL on 20 kHz. These two stations broadcast continuously except for scheduled maintenance periods. WWVL also transmits experimental, multiple frequencies, on occasion, at 19.9, 20.0 and 20.9 kHz.

Frequency accuracy, offset and effects of the propagation medium are covered in a technical bulletin *NBS Frequency and Time Broadcast Services*, NBS Special Publication 236, available for 25¢ from the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402.

Standard audio frequencies of 440 Hz, 500 Hz and 600 Hz are broadcast by WWV and WWVH as well as one second markers. In addition, short term forecasts of radio propagation along paths in the North Atlantic area, such as Washington, D.C. to London or New York to Berlin are broadcast in voice during part of every 15th minute of each hour from WWV. Geophysical alerts are broadcast in voice during the 19th minute of each hour from WWV and during the 46th minute of each hour from WWVH. These broadcasts tell of geophysical and solar events affecting radio propagation.

In addition to these broadcasts, storm warning broadcasts for the North Atlantic and North Pacific areas prepared by the National Weather Bureau are broadcast over WWV and WWVH. A summary of WWV/WWVH broadcasts is shown in figure 25.

In addition to the NBS broadcasts, the Dominion Observatory of Canada transmits time ticks and voice announcements in English and French on 3.330, 7.335, and 14.670 MHz. Many other countries of the world also transmit standard frequency and time signals, particularly on 5, 10 and 15 MHz.

The standard-frequency transmissions may be used for accurately determining the limits of the various amateur bands with the aid of the station receiver and a *secondary frequency standard* which utilizes an accurate low-frequency crystal oscillator. The crystal is zero-beat with WWV by means of its harmonics and then left with only an occasional check to see that the frequency has not drifted off with time. Accurate signals at smaller frequency intervals may be derived from the secondary frequency standard by the use of multivibrator or divider circuits to produce markers at intervals of 25, 10, 5, or 1 kHz. In addition, a variable-frequency *interpolation oscillator* may be used in conjunction with the secondary standard to measure frequencies at any point in the radio spectrum.

Shown in figure 26 is a simple 100-kHz calibration oscillator which provides marker signals up to 30 mHz or so.

29-10 A Precision Crystal Calibrator

Modern direct-reading h-f receivers require a high order of calibrator accuracy. Shown in this section is a versatile crystal-controlled secondary frequency standard utilizing a 1 MHz AT-cut crystal of excellent

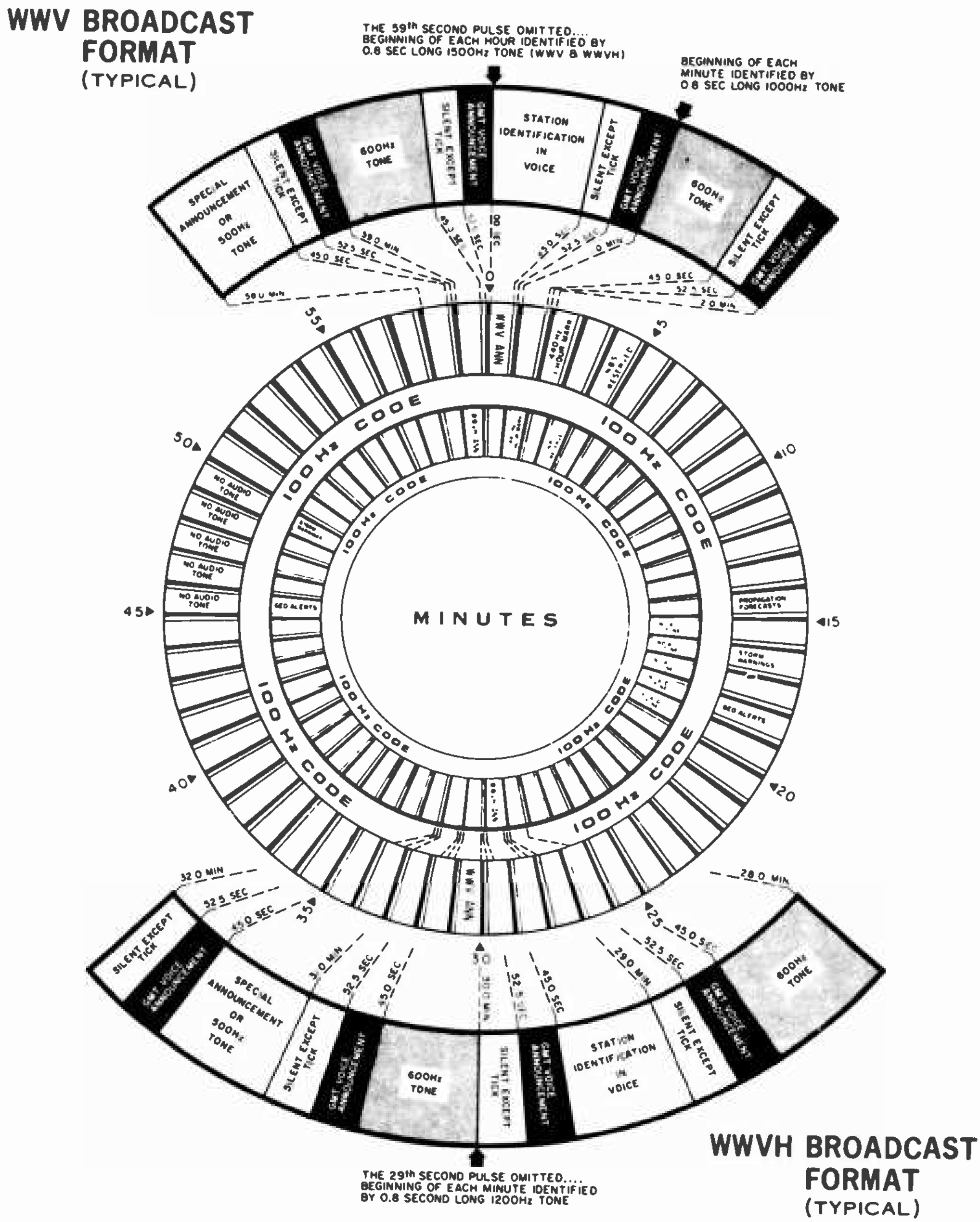


Figure 25

HOURLY BROADCAST SCHEDULE OF WWV AND WWVH

temperature stability. The circuit of this instrument is shown in figure 27.

The crystal is used in an FET oscillator (Q_1) having a high input impedance coupled to an amplifier (Q_2), followed by an impedance transformer (Q_3) to the logic circuit level. Integrated circuit U_1 is a quadruple TTL-type gate used as a Schmitt

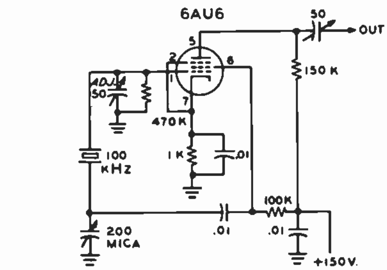


Figure 26
100-kHz MARKER OSCILLATOR

trigger to provide fast rise and fall time for the decade divider (U_2) and the dual flip-flop (U_3). The available outputs are: 1 MHz, 500 kHz, 100 kHz, 50 kHz, and 25 kHz. The IC (U_2) is configured as a divide-by-two and a divide-by-five combination to provide the 500-kHz and 100-kHz markers. A dual-voltage, regulated power supply provides plus fourteen and plus five volts with very low ripple and good regulation.

Frequency of the 1-MHz crystal is set by adjusting capacitor C_1 while zero-beating one of the 1-MHz harmonics with a transmission of WWV, or the frequency may be set with the aid of a frequency counter connected to the 1-MHz output.

For receiver calibration, a 5-pfd capacitor at the receiver end of a short length of low capacitance coaxial cable (93 ohm) will permit maximum harmonic signal to be delivered at the antenna terminals.

29-11 Instruments for Shop and Station

A Silicon Diode Noise Generator The limiting factor in signal reception above 25 MHz is usually the thermal noise generated in the receiver. At any frequency, however, the tuned circuits of the receiver must be accurately aligned for best signal-to-noise ratio. Circuit changes (and even

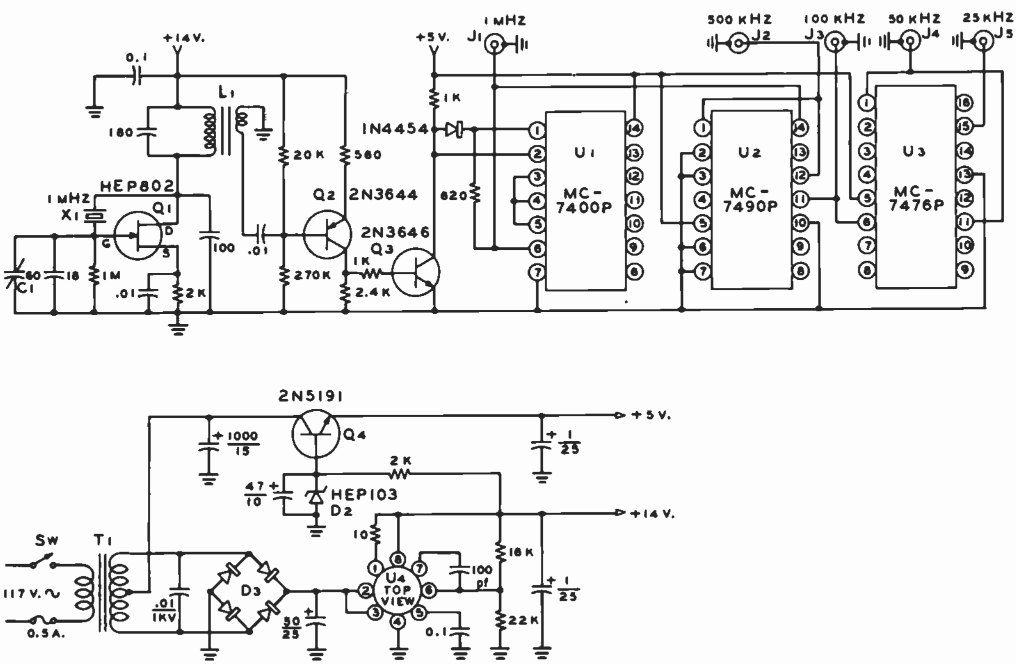


Figure 27

SCHEMATIC, PRECISION CRYSTAL CALIBRATOR

D_3—HEP 176
U_4—LM 300, SG 305T or CA 3055
L_1—120-240 µH. CTC 2060-8. Secondary is 10 turns # 24 insulated wire
T_1—16-volt, center-tapped. Triad F-90X

alignment changes) in the r-f stages of a receiver may do much to either enhance or degrade the noise figure of the receiver. It is exceedingly hard to determine whether changes of either alignment or circuitry are really providing a boost in signal-to-noise ratio of the receiver, or are merely increasing the gain (and noise) of the unit.

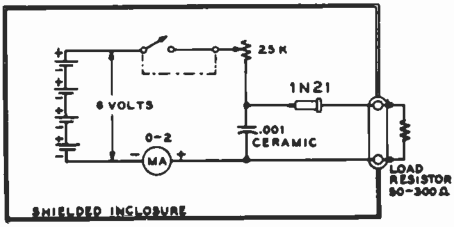


Figure 28

A SILICON DIODE NOISE GENERATOR

A simple means of determining the degree of actual sensitivity of a receiver is to inject a minute signal in the input circuit and then measure the amount of this signal that is needed to overcome the inherent receiver noise. The less injected signal needed to override the receiver noise by a certain, fixed amount, the more sensitive the receiver is.

A simple source of minute signal may be obtained from a silicon crystal diode. If a small d-c current is passed through a silicon crystal in the direction of higher resistance, a small but constant r-f noise (or hiss) is generated. The voltage necessary to generate this noise may be obtained from a few flashlight cells. The *noise generator* is a broadband device and requires no tuning. If built with short leads, it may be employed for receiver measurements well above 150 MHz. The noise generator should be used for comparative measurements only, since calibration against a high-quality commercial noise generator is necessary for absolute measurements.

A Practical Noise Generator Described in this section is a simple silicon crystal noise generator. The schematic of this unit is illustrated in figure 28. The 1N21 crystal and .001-μfd ceramic capacitor are connected in series directly across the output terminals of the instrument. Three small flashlight batteries are wired in series and mounted inside the case, along with the 0-2 d-c milliammeter and the noise-level potentiometer.

To prevent heat damage to the 1N21 crystal during the soldering process, the crystal should be held with a damp rag, and the connections soldered to it quickly with a very hot iron. Across the terminals (and in parallel with the equipment to be attached to the generator) is a 1-watt carbon resistor whose resistance is equal to the impedance level at which measurements are to be made. This will usually be either 50 or 300 ohms. If the noise generator is to be used at one impedance level only, this resistor may be mounted permanently inside of the case.

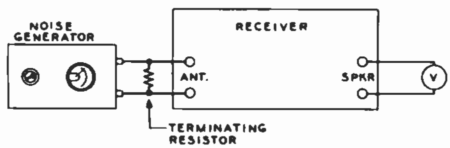


Figure 29
TEST SETUP FOR NOISE GENERATOR

Using the Noise Generator The test setup for use of the noise generator is shown in figure 29. The noise generator is connected to the antenna terminals of the receiver under test. The receiver is turned on, the avc turned off, and the r-f gain control placed full on. The audio volume control is adjusted until the output meter advances to one-quarter scale. This reading is the basic receiver noise. The noise generator is turned on, and the noise-level potentiometer adjusted until the noise output voltage of the receiver is doubled. The more resistance in the diode circuit, the better is the signal-to-noise ratio of the receiver under test. The r-f circuit of the receiver may be aligned for maximum signal-to-noise ratio with the noise generator by aligning for a 2/1 noise ratio at minimum diode current.

An Inexpensive Transistor Tester This inexpensive and compact transistor checker will measure the d-c parameters of most common transistors. Either NPN or PNP transistors may be checked. A six-position test switch permits the following parameters to be measured: (1) I_{CO}—D-c collector current when collector junction is reverse-biased and emitter is open circuited; (2) I_{CO-20}—collector current when base current is 20 microamperes; (3) I_{CO-100}—collector current when base current is 100 microamperes; (4) I_{CEO}—collector current when collector junction is reverse-biased and base is open circuited; (5) I_{CES}—collector current when collector junction is reverse-biased and base is shorted to emitter; (6) I_{EO}—emitter current when emitter junction is reverse-biased and collector is open circuited.

Using the data derived from these tests, the *static* and *a-c forward-current transfer ratios* (h_{FE} and h_{fe} respectively) may be computed as shown in figure 31. This data may be compared with the information

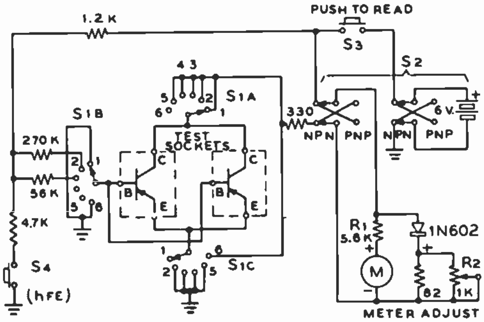
Figure 30

TRANSISTOR CHECKER

An expanded-scale meter provides accurate measurement of transistor parameters in this easily built instrument. Six d-c parameters may be measured and with the data derived from these tests, the a-c forward-current transfer ratios may be computed. Two transistor sockets are mounted at the left of the tester, with the three selector switches to the right. Six-position test switch is mounted to bottom side of box. Tip jacks are placed in parallel with transistor socket terminals to permit test of transistors having unorthodox bases.

listed in the transistor data sheet to determine the condition of the transistor under test.

The transistor parameters are read on a 0-100 d-c microammeter placed in a diode



TO TEST	WHEN	ADJUST S_1 TO	RESULT
I_{CO}	V_{CB} = 6 V.	1	READ METER DIRECT
I_C	I_B = 20 µA	2	"
I_C	I_B = 100 µA	3	"
I_{CEO}	V_{CE} = 6 V.	4	"
I_{CES}	V_{CE} = 6 V.	5	"
I_{EO}	V_{EO} = 6 V.	6	"
h_{FE}	I_B = 20 µA	2	CALCULATE: $h_{FE} = \frac{I_C}{I_B} = \frac{\text{METER READING}}{20\ \mu A}$
h_{FE}	I_B = 100 µA	3	CALCULATE: $h_{FE} = \frac{I_C}{I_B} = \frac{\text{METER READING}}{100\ \mu A}$
h_{fe}	I_B = 20 µA	2	CALCULATE: $h_{fe} = \frac{I_{C1}-I_{C2}}{4\times10^{-6}}$ WHERE: I_{C1} = METER READING
h_{fe}	I_B = 100 µA	3	CALCULATE: $h_{fe} = \frac{I_{C1}-I_{C2}}{20\times10^{-6}}$ I_{C2} = METER READING WITH S_4 CLOSED
6 V. BATTERY	—	4	WITH 150Ω RESISTOR CONNECTED TO C-E OF TEST SOCKET, FULL-SCALE METER DEFLECTION WILL RESULT WHEN S3 IS PRESSED.

Figure 31

SCHEMATIC OF TRANSISTOR CHECKER

S_1A, B, C—Three-pole, 6-position. Centralab 1021
S_2, S_3, S_4—Centralab type 1400 nonshorting lever switch
M—0-200 d-c microammeter. General Electric or Simpson (4½")

network which provides a nearly linear scale to 20 microamperes, a highly compressed scale from 20 microamperes to one milliampere, and a nearly linear scale to full scale at 10 milliamperes. Transistor parameters may be read to within 10 percent on all transistor types from mesas to power alloys without switching meter ranges and without damage to the meter movement or transistor.

By making the sum of the internal resistance of the meter plus series resistor R_1 equal to about 6K, the meter scale is compressed only one microampere at 20 microamperes. *Meter adjust* potentiometer R_2 is set to give 10 milliamperes full-scale meter

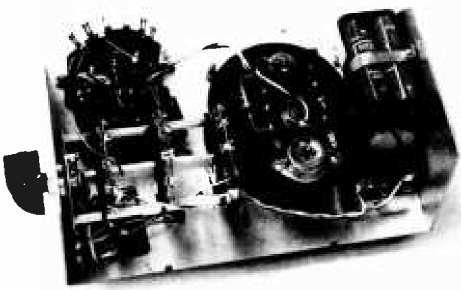


Figure 32

INTERIOR VIEW OF TRANSISTOR CHECKER

Components of meter diode circuit are mounted to phenolic board attached to meter terminals. Other small resistors may be wired directly to switch lugs. The four 1½-volt batteries are held in a small clamp at the rear of the case. Chassis is cut out for lever-action switches and opening is covered with three-position switch plate.

deflection. The scale may then be calibrated by comparison with a conventional meter.

If the *NPN-PNP switch* (S_2) is in the wrong position, the collector and emitter junctions will be forward biased during the I_{CO} and I_{EO} tests (switch positions 1 and 6). The high resulting current may be used as a check for open or intermittent connections within the transistor.

The transistor checker also measures h_{FE} with 20 microamperes and 100 microamperes base current. Depressing the h_{fe} switch (S_1) decreases the base drive about 20 percent, permitting h_{fe} to be estimated from the corresponding change in collector current (formulas 1 and 2). All tests are conducted with a 330-ohm resistor limiting the collector current to about 12 milliamperes and the maximum transistor dissipation to about 20 milliwatts. The checker therefore cannot harm a transistor regardless of how it is plugged in or how the test switches are set.

The *battery test* provides full-scale meter deflection of 10 milliamperes when the battery potential is 6 volts. This is achieved by connecting a 150-ohm resistor from collector to emitter of a test socket.

Test Set Construction The transistor checker is built in an aluminum box measuring 3" × 5" × 7", as shown in the photographs. Test switch S_1 is mounted on the end of the box; and the transistor sockets, microammeter, and the various other switches are placed on the top of the box. Three insulated tip jacks are wired to the leads of one transistor test socket so that transistors having unorthodox bases or leads may be clipped to the tester by means of short test leads. Four 1½-volt flashlight cells are mounted to the rear of the case by an aluminum clamp. Potentiometer R_2, the meter diode, and associated components are fastened to a phenolic board attached to the meter terminals. Switch S_1 has an indicator scale made of heavy white cardboard, lettered with India ink and a lettering pen.

A Transistorized Capacitance Meter Described in this section is a simple and inexpensive transistorized capacitance meter using a single unijunction tran-

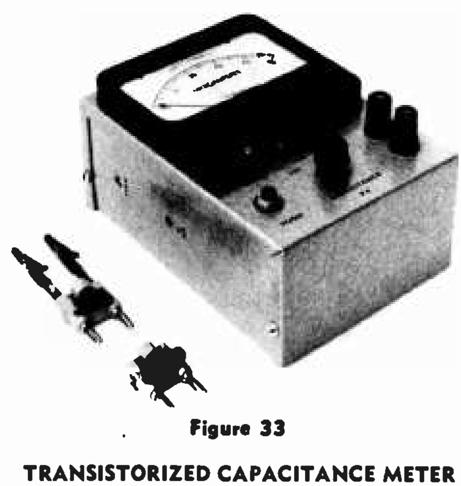


Figure 33

TRANSISTORIZED CAPACITANCE METER

This small, inexpensive test instrument measures capacitance directly up to 0.1 μfd. Using a small self-contained battery, the tester employs a single unijunction transistor in a simple oscillator counter circuit. The "unknown" terminals are at the right of the panel, with the range switch and the push-to-test button to the left. Two jack plugs are made up with "standard" capacitors. The top plug has two alligator clips soldered to jack tips which may be inserted in the tester. Calibration potentiometers are adjusted through the small holes in the side of the case.

sistor (figure 33). The instrument measures capacitance values ranging in size from a few pf up to 0.1 μfd in four ranges.

The capacitance meter uses a simple RC relaxation oscillator to generate square audio-frequency pulses (figure 34). The unknown capacitor is pulse-charged through a diode (D_1) and is discharged through the indicating meter and its series resistance. The discharge current is directly proportional to the value of capacitance under test provided the frequency and amplitude of the charging pulses are held constant.

The frequency of the RC oscillator is switched to provide four capacitance ranges: 100 pf, 1000 pf, .01 μfd and .1 μfd. A 0 to 50 d-c microammeter serves as a read-out device so the reading of the meter must be multiplied by two to obtain the value of measured capacitance. The base resistance of the unijunction transistor is switched in order to achieve full-scale meter deflection on the 100-pf range.

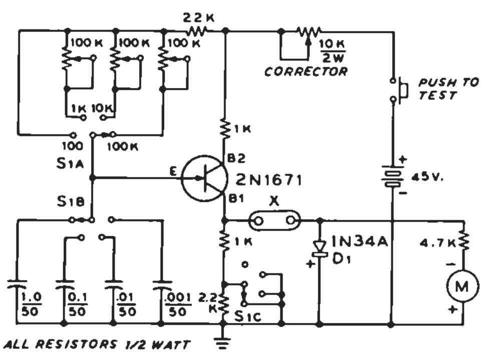


Figure 34

SCHEMATIC OF CAPACITANCE METER

S_1A, B, C—Three-pole, 4-position. Centralab PA-1007
M—0-50 d-c microamperes. Simpson model 49 (4½")

Capacitance Meter Construction The instrument is built in an aluminum box measuring 3″ × 5″ × 7″ (figures 33 and 35). Small components are mounted on two phenolic boards which are supported on either side of the meter by small metal angle brackets. The three 100K calibration potentiometers and the 10K corrector potentiometer are mounted on these boards so that the slotted shafts may

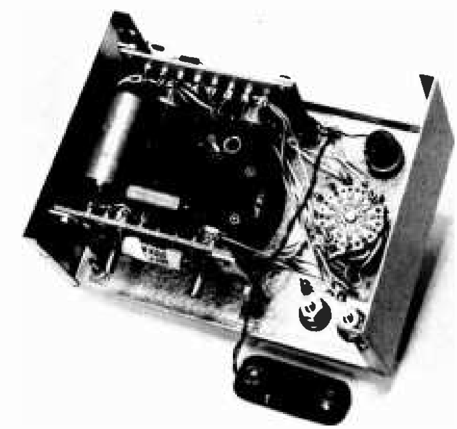


Figure 35

INTERIOR VIEW OF CAPACITANCE METER

The smaller components are mounted to phenolic terminal strips held in place by angle brackets fastened to the meter bolts. The battery is bolted to the rear of the box and connections to the instrument are made via the battery clip in the foreground.

be adjusted through small holes drilled in the sides of the case. The unijunction transistor is mounted in place by its leads. The battery is clamped to the rear half of the case with a small aluminum bracket.

Meter Calibration When the wiring has been completed and checked, the capacitance meter may be calibrated with the aid of capacitors of known value. Ten-percent tolerance paper or mica capacitors that have been checked on a capacitance bridge of good accuracy may be used, or a set of one-percent tolerance capacitors may be used as "standards." A 100-pf standard capacitor is placed between the "unknown" terminals of the capacitance meter (marked X on the schematic) and the meter switch is set to the 100-pf range. The *press to test* button is depressed and the *corrector* potentiometer is adjusted for full-scale meter deflection. The 1000-pf capacitor is now used on the next range to achieve full-scale deflection when the 100K range-calibration potentiometer is properly adjusted. The two higher ranges are adjusted in a like manner with standard capacitors of .01μfd and .1

µfd. The *corrector* potentiometer should be adjusted only on the 100-pf range and should not be retouched until recalibration is necessary as a result of low battery voltage. Normal battery drain is about 5 milliamperes.

A Two-Tone Audio Generator To examine linearity of an amplifier by observation of the output signal some means must be provided to vary the output signal level from zero to maximum with a regular pattern that is easily interpreted. A simple means is to use two audio tones of equal amplitude to modulate the SSB transmitter. This is termed a *two-tone test.* This procedure causes the transmitter to emit two steady signals separated by the frequency difference of the two audio tones. The resultant, or beat, between the two r-f signals produces a pattern which, when observed on an oscilloscope has the appearance of a carrier 100-percent modulated by a series of half sine waves, as previously shown in chapter 9, figure 6B.

With a two-equal-tone test signal, the following equations approximate the relationships between two-tone meter readings, peak envelope power, and average power for class-AB or class-B operation:

D-c plate current:

$$I_b = \frac{2 \times i_{pm}}{\pi^2}$$

Plate Power Input (watts):

$$P_{in} = \frac{2 \times i_{pm} \times E_b}{\pi^2}$$

Average Power Output (watts):

$$P_o = \frac{i_{pm} \times e_p}{8}$$

Plate efficiency:

$$N_p = \left(\frac{\pi}{4}\right)^2 \times \frac{e_p}{E_b}$$

where,

i_{pm} equals peak of the plate current pulse,
e_p equals peak value of plate voltage swing,
E_b equals d-c plate voltage,
π equals 3.14

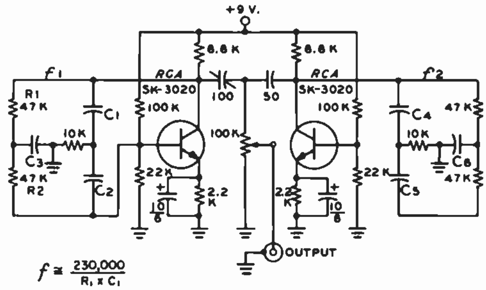


Figure 36

TWO-TONE AUDIO OSCILLATOR

Relative amplitude of oscillators may be leveled by adjusting 100-pf capacitor. For f_1 = 900 Hz, C_1 = C_2 = .005 µfd, and C_3 = .01 µfd. For f_2 = 1300 Hz, C_4 = C_5 = .003 µfd and C_6 = .006 µfd.

Finally, peak-envelope-power *output* under these conditions is twice the average-power output. Thus, using a two-tone test signal, a linear amplifier may be tuned up at a power-output level of half that normally achieved at the so-called "two kilowatt PEP" input level. Power-*input* level, on the other hand, of the two-tone test condition is about two-thirds that of the single-tone condition.

Shown in figure 36 is a transistor two-tone generator that may be used in conjunc-

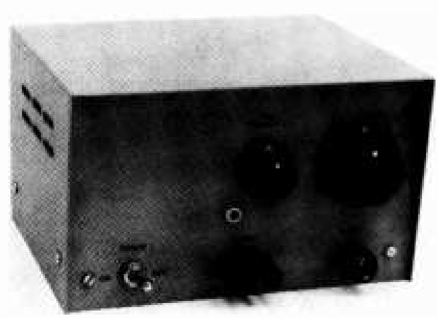


Figure 37

VARIABLE-FREQUENCY AUDIO GENERATOR

This compact, solid-state audio generator covers the range of 20 Hz to 20 kHz with a distortion level of 0.05 percent or less. The frequency-control potentiometer is near center, with the frequency-range switch at the right. Unit is built in a small aluminum utility cabinet.

tion with SSB equipment for appropriate tests. Two high-beta silicon NPN planar transistors are used in a twin-T dual oscillator circuit. With the values specified, frequency f_1 is about 900 Hz and frequency f_2 is about 1300 Hz. Increasing the capacitors C_1, C_2, and C_3 will lower the oscillator frequency. Capacitor C_1 should always equal C_2 and C_3 should have a value twice that of C_1 or C_2. Resistor R_3 normally falls between 8K and 12K and is adjusted for best oscillator waveform. The oscillators should be tested separately and their waveform viewed on an oscilloscope.

The Two-tone Test—The test oscillator is connected to the audio system of the SSB transmitter which is tuned up into a dummy load with an oscilloscope coupled to the load to show a typical test pattern. The transmitter is adjusted for maximum power output without waveform flattopping. Under these conditions, the power input is:

$$\text{PEP Input (watts)} = I_b \times E_b \left(1.57 - 0.57 \frac{I_b}{I_o}\right)$$

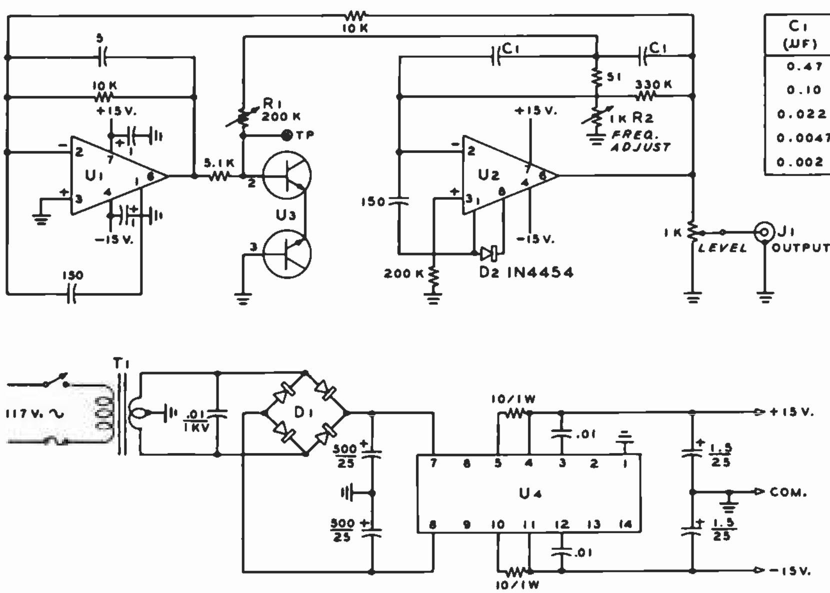
where,

E_b equals d-c plate voltage,
I_b equals two-tone d-c plate current,
I_o equals idling plate current with no test signal.

29-12 A Variable-Frequency Audio Generator

Described in this section is a high-quality, variable-frequency audio generator that covers the range of 20 Hz to 20 kHz, with a distortion level of 0.05% or less (figure 37).

Unlike the expensive laboratory oscillators which require dual (tracking) variable



C_1 (µF)	MIN. FREQ.	MAX. FREQ.
0.47	18 Hz	80 Hz
0.10	80 Hz	380 Hz
0.022	380 Hz	1.7 kHz
0.0047	1.7 kHz	8 kHz
0.002	4.4 kHz	20 kHz

Figure 38

SCHEMATIC, AUDIO GENERATOR

U_1, U_2—LM 301A (National Semiconductor)
U_3—LM709C used as dual zener (pins 2 and 3)
U_4—SG 3501D (Silicon General)
D_1—HEP 176
T_1—32-volt, center-tapped, Triad F-90X

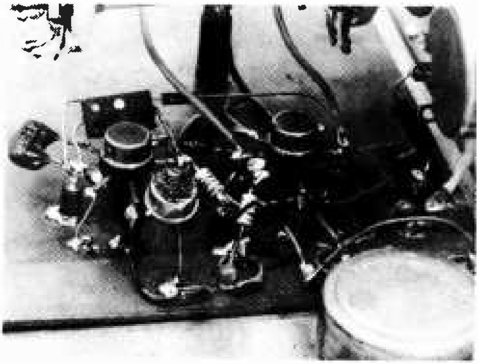


Figure 39

COMPONENTS OF AUDIO GENERATOR ARE MOUNTED ON P.C. BOARD

U_3, the inexpensive IC used as a dual zener diode, is in the foreground at left. The two op-amps are placed in sockets supported on small terminals soldered to the board.

resistors or capacitors, this compact oscillator uses a single variable resistor for tuning. The circuit is shown in figure 38.

Three operational IC amplifiers are used. Op-amp U_2 functions as an active bandpass filter, U_1 serves as a broadband amplifier, and U_3 is used as a dual zener diode. The feedback loop that sustains oscillation involves 180 degrees of phase shift around U_1 and 180 degrees of phase shift around U_2. To permit oscillation, sufficient circuit gain occurs only at the maximum response frequency of the active bandpass filter that is designed around U_2. The frequency of oscillation is thus controlled by varying the center frequency of the bandpass filter. Level stabilization is obtained by clipping the sine wave by means of U_3, the Q of the active filter circuit removing the harmonics created by the clipping. Only the base-emitter diodes of the two input transistors of U_3 are used (figure 39), the other leads are left floating. The LM 709C was used because of its very low price in comparison to the cost of a good seven-volt zener diode.

A test point is provided for the builder to monitor the percentage of sine-wave clipping in use, the level being set by potentiometer R_1. This is normally set so that about 20 percent of the sine wave total amplitude is clipped when the frequency control potentiometer (R_2) is at the low-frequency (maximum resistance) position.

To power the audio oscillator, a simple dual-voltage regulated supply providing plus and minus 15 volts is included.

CHAPTER THIRTY

The Oscilloscope

The *cathode-ray oscilloscope* is an instrument which permits visual examination of various electrical phenomena of interest to the electronic engineer. Instantaneous changes in voltage, current and phase are observable if they take place slowly enough for the eye to follow, or if they are periodic for a long enough time so that the eye can obtain an impression from the screen of the cathode-ray tube. In addition, the cathode-ray oscilloscope may be used to study any variable (within the limits of its frequency-response characteristic) which can be converted into electrical potentials. This conversion is made possible by the use of some type of *transducer,* such as a vibration pickup unit, pressure pickup unit, photoelectric cell, microphone, or a variable impedance. The use of such a transducer makes the cscilloscope a valuable tool in fields other than electronics.

30-1 A Modern Oscilloscope

For the purpose of analysis, the operation of a modern oscilloscope will be described. The 'scope is completely solid state except for the cathode-ray tube. The simplified block diagram of the instrument is shown in figure 1. This oscilloscope (the *Heath* model IO-102) is capable of reproducing sine waves up to 5 MHz and has a rise time of 80 nanoseconds. The sweep speed is continuously variable from 10 Hz to 500 kHz in five ranges, and the electron beam of the cathode-ray tube can be moved vertically or horizontally, or the movements may be combined to produce composite patterns on the screen. As shown in the diagram, the cathode-ray tube receives signals from two sources: the *vertical* (Y-axis) and the *horizontal* (X-axis) *amplifiers,* and also receives *blanking pulses* that remove unwanted return trace signals from the screen. The operation of the cathode-ray tube has been covered in an earlier chapter and the auxiliary circuits pertaining to signal presentation will be discussed here.

The Vertical Amplifier The incoming signal to be displayed is coupled through a frequency-compensated attenuator network (figure 2). The gain may thus be controlled in calibrated steps. A capacitor blocks the d-c component of the signal when a-c signals are applied to the circuit. A portion of the input signal is applied through a voltage-limiting resistor and two limiting diodes (D_1, D_2) to a FET connected as a source follower amplifier (Q_1). This device provides the high input impedance necessary to prevent circuit loading. Transistor Q_2 is a constant-current source for the FET and diodes D_4 and D_5 hold the base of Q_2 at a constant voltage. Since Q_2 is a form of emitter follower, the emitter

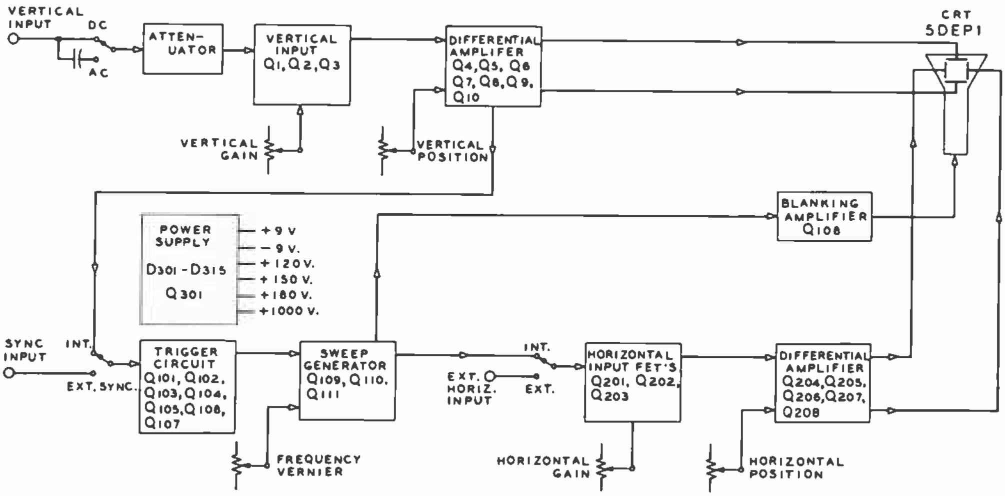


Figure 1

BLOCK DIAGRAM OF A MODERN OSCILLOSCOPE

This simplified diagram of the Heath IO-102 solid-state oscilloscope features triggered sweep and a blanking circuit that permits observation of extremely short pulses. The cathode-ray tube is the only vacuum tube in the instrument.

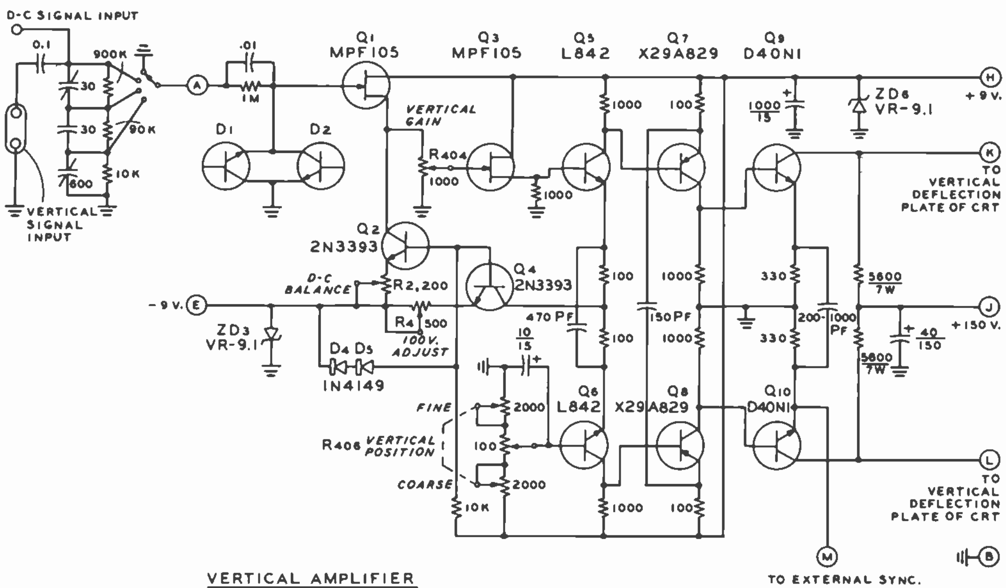


Figure 2

VERTICAL AMPLIFIER

The vertical amplifier is capable of passing sine waves up to 5 MHz. The compensated input attenuator and peaking circuits provide gain that is essentially independent of frequency. Emitter-follower Q_5 is coupled to amplifier Q_6 to provide push-pull signal necessary for the deflection plates of the cathode-ray tube. The input signal is limited in amplitude by diodes D_1 and D_2 (the junction of inexpensive bipolar transistors).

voltage is a function of the base voltage, and the emitter voltage also remains constant. This voltage appears across the *d-c balance* control which is adjusted so that the source voltage of the FET is zero when an input signal is not present. Thus, a signal applied to the gate of Q_1 causes only voltage changes at the source because the current through Q_1 is constant .The voltage variations are applied across the *vertical gain* control and a portion of this signal is applied to the gate of source follower Q_3. Transistor Q_4 forms a constant-current source for transistors Q_5 and Q_6. Since the emitter of each device is connected to this source, the source serves as a common-emitter resistance and sets the operating point for the following stages.

Transistors Q_5 and Q_6 have a common-emitter resistance and any signal present at the Q_5 emitter is coupled to the emitter of Q_6, which functions as a common-base amplifier whose base is held constant by the *vertical position* potentiometer. The signal at the collector of transistor Q_6 is 180° out of phase with the signal at the collector of Q_5, thus forming a push-pull configuration required to drive the deflection plates of the cathode-ray tube.

Drive transistors Q_7 and Q_8 are common-emitter amplifiers which drive output amplifier transistors Q_9 and Q_{10} which have their collector potential derived from the +150 volt supply.

The Sweep Circuit Investigation of electrical waveforms by the use of a cathode-ray tube requires that some means be readily available to determine the variation in the waveforms with respect to time. An X-axis *time base* on the screen of the cathode-ray tube shows the variation in amplitude of the input signal with respect to time (figure 3). This display is made possible by a *time-base generator (sweep generator)* which moves the spot across the screen at a constant rate from left to right between selected points, returns the spot almost instantaeously to its original position, and repeats this procedure at a specified rate (referred to as the *sweep frequency*).

The Sweep-Trigger Circuit—An external *synchronizing impulse* which may be either a portion of the amplified signal or a signal applied to the *external sync* terminals is coupled to the gate of source follower Q_{101}. Two limiting diodes protect the transistor from high voltage surges. Constant-current source Q_{102} is adjusted by the *sync level* control to provide proper bias for the synchronizing circuits. This ensures that even

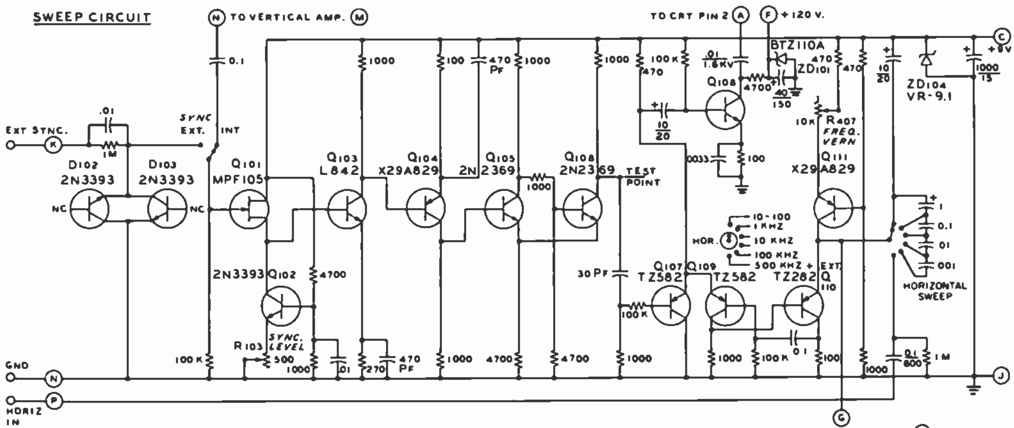


Figure 3

THE SWEEP CIRCUIT

The sweep may be triggered either by the input signal or by an external source. Schmitt trigger circuit (Q_{105} and Q_{106}) produces a regular pulse each time it is triggered, driving the astable multivibrator (Q_{109} and Q_{110}). Timing capacitors and the frequency vernier potentiometer determine sweep speed. During the wait period between trigger pulses, the CRT is cut off so that the blanking waveform is not seen. Negative pulse from blanking amplifier Q_{108} is applied to pin #2 of the cathode-ray tube to perform this function.

a small signal can synchronize the sweep generator.

Transistors Q_{103} and Q_{104} amplify the signal and apply it to the *Schmitt trigger circuit* consisting of Q_{105} and Q_{106}. This trigger circuit is a regenerative bistable circuit which produces a regular pulse output each time it is triggered and reset. Devices Q_{109} and Q_{110} form an astable multivibrator whose frequency is controlled by the switchable timing capacitors. The capacitors are charged through Q_{110} and discharged through the constant-current source circuit of Q_{111}. The *frequency vernier* potentiometer determines the current flowing through Q_{111} which, in turn, determines the discharge current and discharge time of the timing capacitor. As the capacitor discharges, a positive-going sawtooth voltage is generated and coupled to the horizontal amplifier. The frequency of the horizontal sweep is determined by the particular timing capacitor and the discharge current.

The Blanking Circuit—During the wait period between trigger pulses, the cathode-ray tube is completely cut off so that the blanking waveform is not seen. Since transistors Q_{107} and Q_{109} have a common emitter resistor, a signal applied to the base of Q_{107} is emitter-coupled to transistor Q_{108}. The pulse output of the Schmitt trigger (Q_{106}) is coupled to Q_{109}. This causes this transistor to turn on and Q_{110} to cut off and start the sweep just prior to the time it would normally begin. When the signal at the emitter of Q_{109} goes positive, a positive pulse is applied to the base of blanking amplifier Q_{108}. A negative-going output pulse is coupled to the grid of the cathode-ray tube which turns off the electron beam during retrace.

The Horizontal Amplifier — Since the amplitude of the sweep waveform at the output of the sweep generator is not large enough to drive the horizontal deflection plates of the cathode-ray tube, further amplification is needed. The signal from the sweep generator is applied to the horizontal amplifier, whose circuitry is similar to that of the vertical amplifier (figure 4). The major difference is that the horizontal amplifier does not have a PNP amplifier stage corresponding to Q_7 and Q_8 in the vertical amplifier. The positive-going sawtooth wave

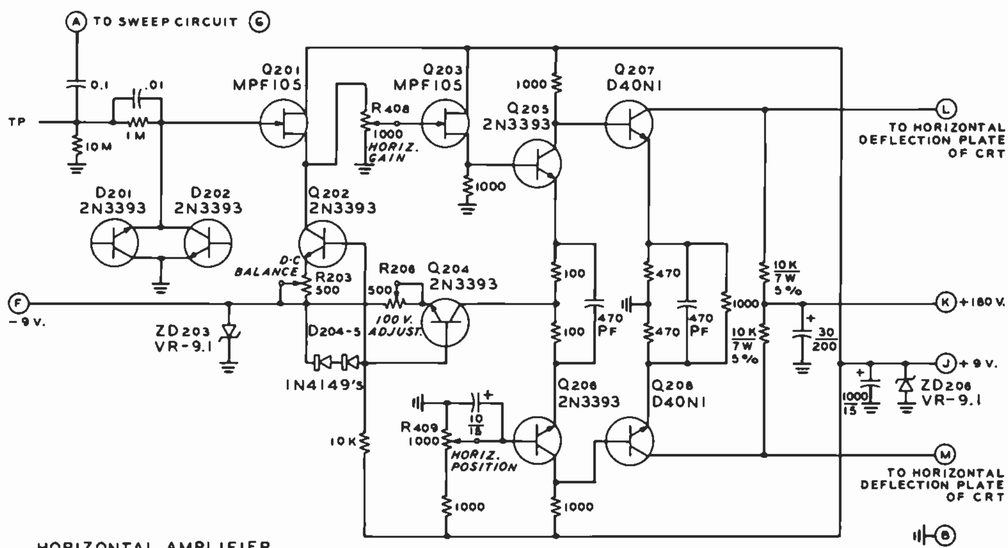


Figure 4

HORIZONTAL AMPLIFIER

The horizontal amplifier is similar to the vertical amplifier except it does not have PNP stage Q_7-Q_8 shown in figure 3. Amplified sweep waveform is applied to the horizontal-deflection plates of the CRT causing the electron beam to sweep across the face of the tube producing a visible trace. Transistor Q_{205} serves as an emitter follower to produce push-pull driving signal for Q_{207} and Q_{208}. Horizontal positioning of signal on screen of CRT is determined by the base bias of Q_{206}.

from the sweep generator is amplified and applied to the horizontal plates of the cathode-ray tube. This increasing voltage causes the electron beam to sweep across the face of the tube producing a visible trace. The sweep rate of the electron beam is determined by the sawtooth frequency.

The Power Supply The power supply provides positive and negative voltages for the various stages of the oscilloscope, as shown in figure 5. A high-voltage winding of the power transformer is connected to a voltage-doubler circuit to provide —1500 volts to the cathode-ray tube. *Intensity* and *focus* voltages are also supplied from a voltage-divider network. A separate 6.3-volt winding supplies the filament voltage for the cathode-ray tube. Optimum focus is obtained when the deflection plates of the cathode-ray tube and the *astigmatism* grid are at the same potential. Since the vertical-deflection plate voltages (collectors of Q_9 and Q_{10}) are adjusted to 100 volts d-c by the constant-current source Q_4, the astigmatism potential is also adjusted to 100 volts. A low-voltage regulated supply provides +9 and —9 volts and a third supply provides the various other voltages required by the oscilloscope circuits.

30-2 Display of Waveforms

Together with a working knowledge of the controls of the oscilloscope, an understanding of how the patterns are traced on the screen must be obtained for a thorough knowledge of oscilloscope operation. With this in mind a careful analysis of two fundamental waveform patterns is discussed under the following headings:

1. Patterns plotted against time (using the sweep generator for horizontal deflection).
2. *Lissajous figures* (using a sine wave for horizontal deflection).

Patterns Plotted Against Time A sine wave is typical of such a pattern and is convenient for this study. This wave is amplified by the vertical amplifier and impressed on the vertical (Y-axis) deflection plates of the cathode-ray tube. Simultaneously the sawtooth wave from the time-base generator is amplified and impressed on the horizontal (X-axis) deflection plates.

The electron beam moves in accordance with the resultant of the sine and sawtooth signals. The effect is shown in figure 6 where the sine and sawtooth waves are graphically represented on time and voltage axes. Points on the two waves that occur simultaneously are numbered similarly. For example, point 2 on the sine wave and point 2 on the sawtooth wave occur at the same instant. Therefore the position of the beam at instant 2 is the resultant of the voltages on the horizontal and vertical deflection plates at instant 2. Referring to figure 8, by projecting lines from the two point-2 positions, the position of the electron beam at instant 2 can be located. If projections were drawn from every other instantaneous position of each wave to intersect on the circle representing the tube screen, the intersections of similarly timed projects would trace out a sine wave.

In summation, figure 6 illustrates the principles involved in producing a sine-wave trace on the screen of a cathode-ray tube. Each intersection of similarly timed projections represents the position of the electron beam acting under the influence of the varying voltage waveforms on each pair of deflection plates. Figure 7 shows the effect on the pattern of decreasing the frequency of the sawtooth wave. Any recurrent waveform plotted against time can be displayed and analyzed by the same procedure as used in these examples.

The sine-wave problem just illustrated is typical of the method by which any waveform can be displayed on the screen of the cathode-ray tube. Such waveforms as square wave, sawtooth wave, and many more irregular recurrent waveforms can be observed by the same method explained in the preceding paragraphs.

30-3 Lissajous Figures

Another fundamental pattern is the *Lissajous figures*, named after the 19th-century French scientist. This type of pattern is of particular use in determining the frequency

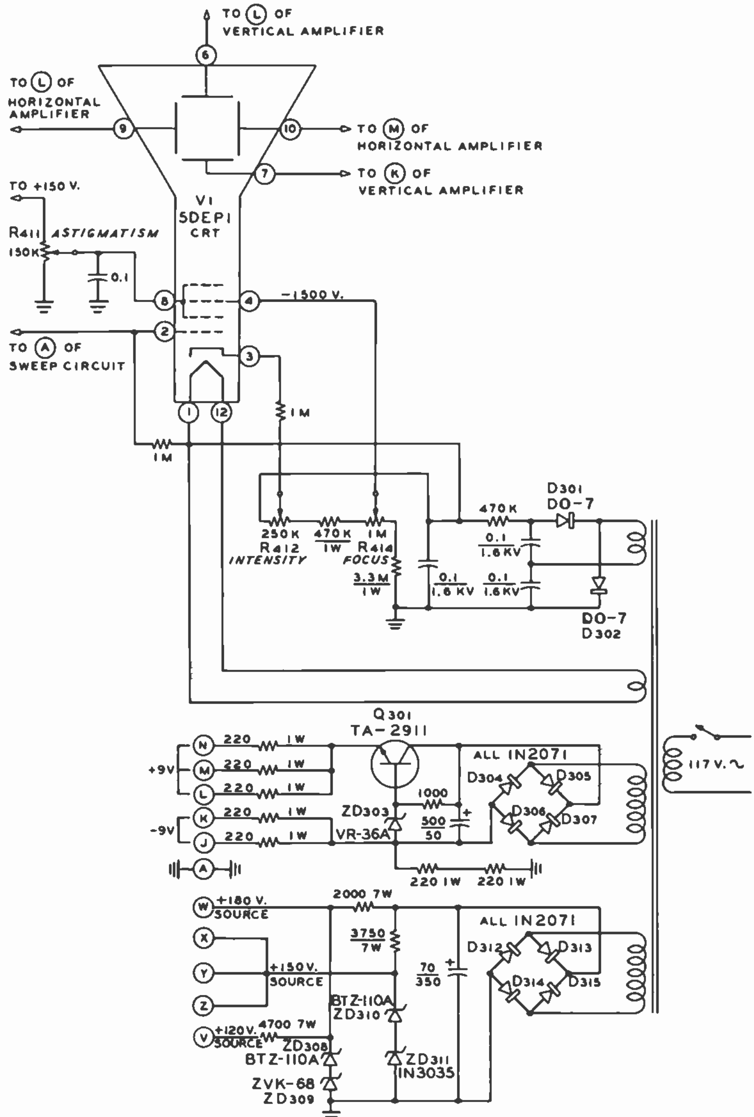


Figure 5

POWER SUPPLY

Power supply provides −1500 volts for CRT and various low voltages for solid-state circuitry of the oscilloscope. Intensity and focus voltages are supplied from a voltage-divider network. Optimum focus is obtained when the deflection plates of CRT and the astigmatism grid are at the same potential.

ratio between two sine-wave signals. If one of these signals is known, the other can be easily calculated from the pattern made by the two signals on the screen of the cathode-ray tube. Common practice is to connect the known signal to the horizontal channel and the unknown signal to the vertical channel.

The presentation of Lissajous figures can

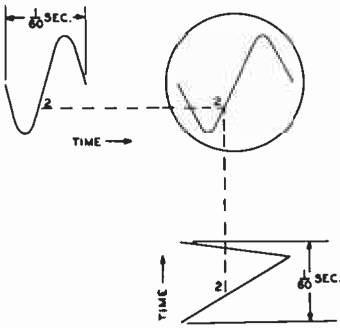


Figure 6

PROJECTION DRAWING OF A SINE WAVE APPLIED TO THE VERTICAL AXIS AND A SAWTOOTH WAVE OF THE SAME FREQUENCY APPLIED SIMULTANEOUSLY ON THE HORIZONTAL AXIS

be analyzed by the same method as previously used for sine-wave presentation. A simple example is shown in figure 8. The frequency ratio of the signal on the horizontal axis to the signal on the vertical axis is 3 to 1. If the known signal on the horizontal axis is 180 Hertz, the signal on the vertical axis is 60 Hertz.

Obtaining a Lissajous Pattern on the Screen; Oscilloscope Settings 1. The horizontal amplifier should be disconnected from the sweep oscillator. The

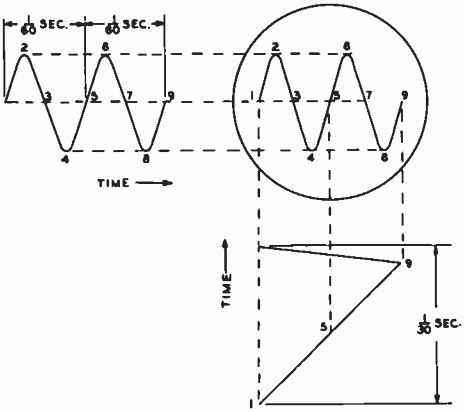


Figure 7

PROJECTION DRAWING SHOWING THE RESULTANT PATTERN WHEN THE FREQUENCY OF THE SAWTOOTH IS ONE-HALF OF THAT EMPLOYED IN FIGURE 6

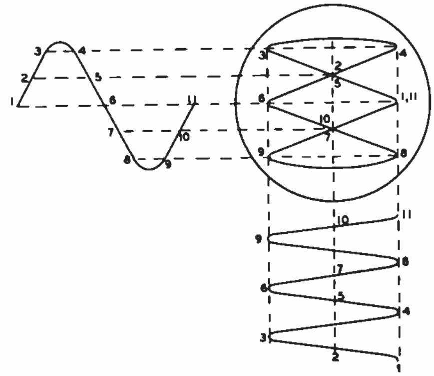


Figure 8

PROJECTION DRAWING SHOWING THE RESULTANT LISSAJOUS PATTERN WHEN A SINE WAVE APPLIED TO THE HORIZONTAL AXIS IS THREE TIMES THAT APPLIED TO THE VERTICAL AXIS

signal to be examined should be connected to the horizontal amplifier of the oscilloscope.

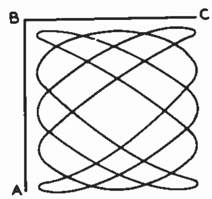


Figure 9

METHOD OF CALCULATING FREQUENCY RATIO OF LISSAJOUS FIGURES

2. An audio oscillator signal should be connected to the vertical amplifier of the oscilloscope.

3. By adjusting the frequency of the audio oscillator a stationary pattern should be obtained on the screen of the oscilloscope. It is not necessary to stop the pattern, but merely to slow it up enough to count the loops at the side of the pattern.

4. Count the number of loops which intersect an imaginary vertical line *AB* and the number of loops which intersect the imaginary horizontal line *BC* as shown in figure 9. The ratio of the number of loops which intersect *AB* is to the number of loops which intersect *BC* as the frequency of the horizontal signal is to the frequency of the vertical signal.

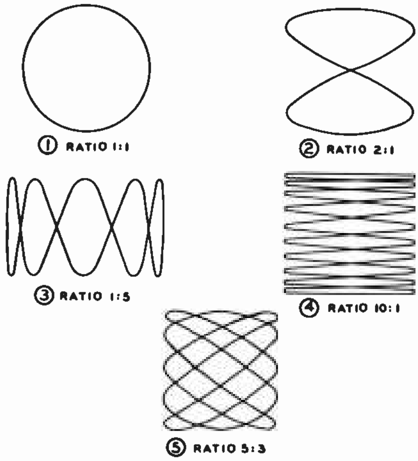


Figure 10

OTHER LISSAJOUS PATTERNS

Figure 10 shows other examples of Lissajous figures. In each case the frequency ratio shown is the frequency ratio of the signal on the horizontal axis to that on the vertical axis.

Phase Difference Patterns Coming under the heading of Lissajous figures is the method used to determine the phase difference between signals of the same frequency. The patterns involved take on the form of ellipses with different degrees of eccentricity.

The following steps should be taken to obtain a phase-difference pattern:

1. With no signal input to the oscilloscope, the spot should be centered on the screen of the tube.
2. Connect one signal to the vertical amplifier of the oscilloscope, and the other signal to the horizontal amplifier.
3. Connect a common ground between the two frequencies under investigation and the oscilloscope.
4. Adjust the vertical amplifier gain so as to give about 3 inches of deflection on a 5-inch tube, and adjust the calibrated scale of the oscilloscope so that the vertical axis of the scale coincides precisely with the vertical deflection of the spot.
5. Remove the signal from the vertical amplifier, being careful not to change the setting of the vertical gain control.
6. Increase the gain of the horizontal amplifier to give a deflection exactly the same as that to which the vertical amplifier control is adjusted (3 inches). Reconnect the signal to the vertical amplifier.

The resulting pattern will give an accurate picture of the exact phase difference between the two waves. If these two patterns are exactly the same frequency but different in phase and maintain that difference, the pattern on the screen will remain stationary. If, however, one of these frequencies is drifting slightly, the pattern will drift slcwly through 360°. The phase angles of 0°, 45°, 90°, 135°, 180°, 225°, 270°, and 315° are shown in figure 11.

Each of the eight patterns in figure 11 can be analyzed separately by the previously used projection method. Figure 14 shows two sine waves which differ in phase being projected on to the screen of the cathode-ray tube. These signals represent a phase difference of 45°. It is extremely important that (1) the spot has been centered on the screen of the cathode-ray tube, (2) that both the horizontal and vertical amplifiers have been adjusted to give exactly the same gain, and (3) that the calibrated scale be originally set to coincide with the displacement of the signal along the vertical axis. If the amplifiers of the oscilloscope are not used for conveying the signal to the deflection plates of the cathode-ray tube, the coarse frequency switch should be set to *horizontal input direct* and the vertical input switch to *direct* and the outputs of the two signals must be adjusted to result in exactly the same vertical deflection as horizontal deflection. Once this deflection has been set by either the oscillator output controls or the amplifier gain controls in the oscillograph, it should not be changed for the duration of the measurement.

Determinination of the Phase Angle The relation commonly used in determining the phase angle between signals is:

$$\text{Sine } \theta = \frac{Y \ intercept}{Y \ maximum}$$

where,

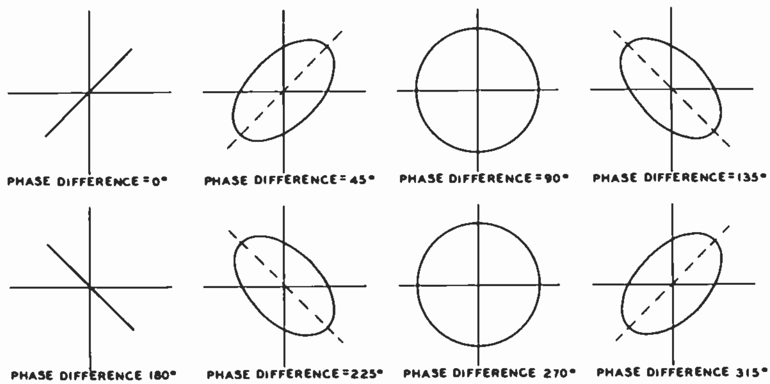


Figure 11

LISSAJOUS PATTERNS OBTAINED FROM THE MAJOR PHASE DIFFERENCE ANGLES

θ equals phase angle between signals,
Y intercept equals point where ellipse crosses vertical axis measured in tenths of inches (calibrations on the calibrated screen),
Y maximum equals highest vertical point on ellipse in tenths of inches.

Several examples of the use of the formula are given in figure 13. In each case the *Y intercept* and *Y maximum* are indicated together with the sine of the angle and the angle itself. For the operator to observe these various patterns with a single signal source such as the test signal, there are many types of phase shifters which can be used. Circuits can be obtained from a number of radio textbooks. The procedure is to connect the original signal to the horizontal channel of the oscilloscope and the signal which has passed through the phase shifter to the vertical channel of the oscilloscope, and follow the procedure set forth in this discussion to observe the various phase-shift patterns.

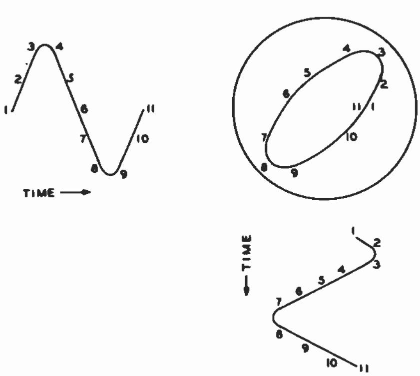


Figure 12

PROJECTION DRAWING SHOWING THE RESULTANT PHASE-DIFFERENCE PATTERN OF TWO SINE WAVES 45° OUT OF PHASE

30-4 Receiver I-F Alignment with an Oscilloscope

The alignment of the i-f amplifiers of a receiver consists of adjusting all the tuned circuits to resonance at the intermediate frequency and at the same time permitting passage of a predetermined number of sidebands. The best indication of this adjustment is a resonance curve representing the response of the i-f circuit to its particular range of frequencies.

As a rule medium- and low-priced receivers use i-f transformers whose bandwidth is about 5 kHz on each side of the fundamental frequency. The response curve of these i-f transformers is shown in figure 14. High-fidelity receivers usually contain i-f transformers which have a broader bandwidth which is usually 10 kHz on each side of the fundamental. The response curve for this type transformer is shown in figure 15.

Resonance curves such as these can be displayed on the screen of an oscilloscope. For a complete understanding of the procedure

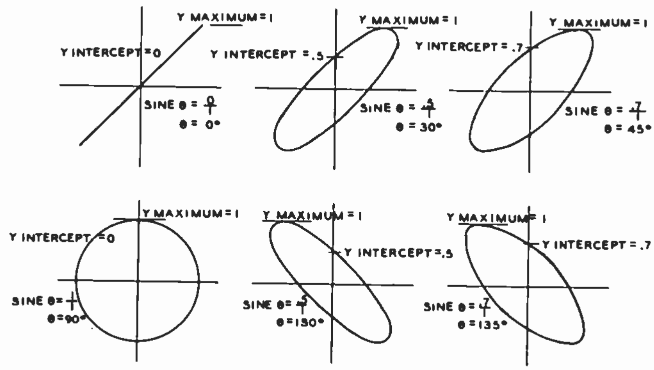


Figure 13

EXAMPLES SHOWING THE USE OF THE INTERCEPT FORMULA FOR DETERMINATION OF PHASE DIFFERENCE

it is important to know how the resonance curve is traced.

The Resonance Curve on the Screen To present a resonance curve on the screen, a frequency-modulated signal source must be available. This signal source is a signal generator whose output is the fundamental i-f frequency which is frequency-modulated 5 to 10 kHz each side of the fundamental frequency. A signal generator of this type generally takes the form of an ordinary signal generator with a rotating motor-driven tuned-circuit capacitor, called a *wobbulator,* or its electronic equivalent, which is a reactance tube.

The method of presenting a resonance curve on the screen is to connect the vertical channel of the oscilloscope across the detector load of the receiver as shown in the detectors of figure 16 (between point A and ground) and the time-base generator output to the horizontal channel. In this way the d-c voltage across the detector load varies with the frequencies which are passed by the i-f system. Thus, if the time-base generator is set at the frequency of rotation of the motor-driven capacitor, or the reactance tube, a pattern resembling figure 17 (a double resonance curve) appears on the screen.

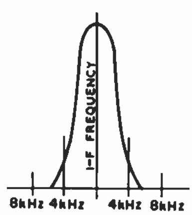


Figure 14

I-F FREQUENCY RESPONSE CURVE OF A LOW PRICED RECEIVER

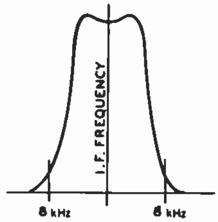


Figure 15

FREQUENCY RESPONSE OF HIGH-FIDELITY I-F SYSTEM

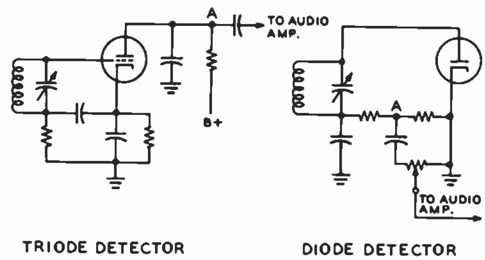


Figure 16

CONNECTION OF THE OSCILLOSCOPE ACROSS THE DETECTOR LOAD

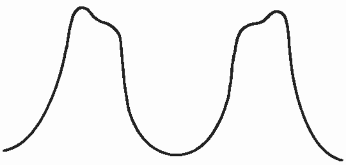


Figure 17

DOUBLE-RESONANCE CURVE

Figure 17 is explained by considering figure 18. In half a rotation of the motor-driven capacitor the frequency increases from 445 kHz to 465 kHz, more than covering the range of frequencies passed by the i-f system. Therefore, a full resonance curve is presented on the screen during this half rotation since only *half* a cycle of the voltage producing horizontal deflection has transpired. In the second half of the rotation the motor-driven capacitor takes the frequency of the signal in the reverse order through the range of frequencies passed by the i-f system. In this interval the time-base generator sawtooth waveform completes its cycle, drawing the electron beam further across the screen and then returning it to the starting point. Subsequent cycles of the motor-driven capacitor and the sawtooth voltage merely retrace the same pattern. Since the signal being viewed is applied through the vertical amplifier, the sweep can be synchronized internally.

Some signal generators, particularly those employing a reactance tube, provide a sweep output in the form of a sine wave which is synchronized to the frequency with which the reactance tube is swinging the fundamental frequency through its limits, (usually 60 hertz). If such a signal is used for horizontal deflection, it is already synchronized.

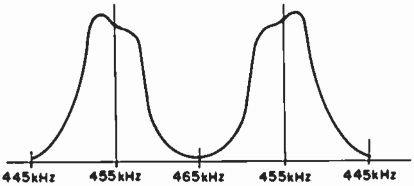


Figure 18

DOUBLE-RESONANCE ACHIEVED BY COMPLETE ROTATION OF THE MOTOR-DRIVEN CAPACITOR

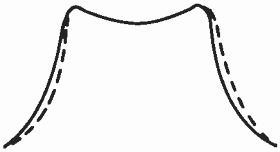


Figure 19

SUPERPOSITION OF RESONANCE CURVES

Since this signal is a sine wave, the response curve is observed as it sweeps the spot across the screen from left to right; and it is observed again as the sine wave sweeps the spot back again from right to left. Under these conditions the two response curves are superimposed on each other and the high-frequency responses of both curves are at one end and the low-frequency response of both curves is at the other end. The i-f trimmer capacitors are adjusted to produce a response curve which is symmetrical on each side of the fundamental frequency.

When using sawtooth sweep, the two response curves can also be superimposed. If the sawtooth signal is generated at exactly twice the frequency of rotation of the motor-driven capacitor, the two resonance curves will be superimposed (figure 19) if the i-f transformers are properly tuned. If the two curves do not coincide the i-f trimmer capacitors should be adjusted. At the point of coincidence the tuning is correct. It should be pointed out that rarely do the two curves agree perfectly. As a result, optimum adjustment is made by making the peaks coincide. This latter procedure is the one generally used in i-f adjustment. When the two curves coincide, it is evident that the i-f system responds equally to signals higher and lower than the fundamental i-f frequency.

30-5 Single-Sideband Applications

Measurement of power output and distortion are of particular importance in SSB transmitter adjustment. These measurements are related to the extent that distortion rises rapidly when the power amplifier is overloaded. The usable power output of an SSB transmitter is often defined as the maximum peak envelope power obtainable with a specified *signal-to-distortion* ratio. The oscilloscope is a useful instrument for measuring and studying distortion of all types that may be generated in single-sideband equipment.

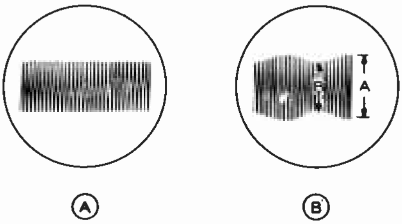


Figure 20

SINGLE-TONE PRESENTATION

Oscilloscope trace of SSB signal modulated by single tone (A). Incomplete carrier supression or spurious products will show modulated envelope of (B). The ratio of supression is:

$$S = 20 \log \frac{A+B}{A-B}$$

Single-Tone Observations When an SSB transmitter is modulated with a single audio tone, the r-f output should be a single radio frequency. If the vertical plates of the oscilloscope are coupled to the output of the transmitter, and the horizontal amplifier sweep is set to a slow rate, the scope presentation will be as shown in figure 20. If unwanted distortion products or carrier are present, the top and bottom of the pattern will develop a "ripple" proportional to the degree of spurious products.

The Linearity Tracer The *linearity tracer* is an auxiliary detector to be used with an oscilloscope for quick observation of amplifier adjustments and parameter variations. This instrument consists of two SSB *envelope detectors* the outputs of which connect to the horizontal and vertical inputs of an oscilloscope. Figure 21 shows a block diagram of a typical linearity test setup. A two-tone test signal is normally employed to supply an SSB modulation envelope, but any modulating signal that provides an envelope that varies from zero to full amplitude may be used. Speech modulation gives a satisfactory trace, so that this instrument may be used as a visual monitor of transmitter linearity. It is particularly useful for monitoring the signal level and clearly shows when the amplifier under observation is overloaded. The linearity trace will be a straight line regardless of the envelope shape if the

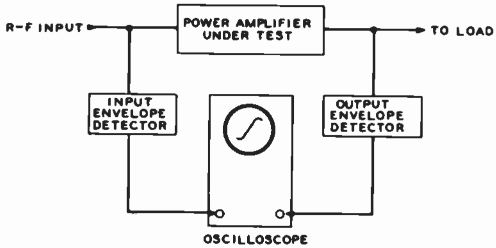


Figure 21

BLOCK DIAGRAM OF LINEARITY TRACER

amplifier has no distortion. Overloading causes a sharp break in the linearity curve. Distortion due to too much bias is also easily observed and the adjustment for low distortion can easily be made.

Another feature of the linearity detector is that the distortion of each individual stage can be observed. This is helpful in troubleshooting. By connecting the input envelope detector to the output of the SSB generator, the over-all distortion of the entire r-f circuit beyond this point is observed. The unit can also serve as a voltage indicator which is useful in making tuning adjustments.

The circuit of a typical envelope detector is shown in figure 22. Two matched germanium diodes are used as detectors. The detectors are not linear at low signal levels, but if the nonlinearity of the two detectors is matched, the effect of their nonlinearity on

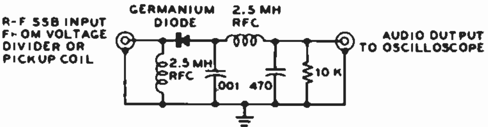


Figure 22

SCHEMATIC OF ENVELOPE DETECTOR

the oscilloscope trace is cancelled. The effect of diode differences is minimized by using a diode load of 5000 to 10,000 ohms, as shown. It is important that both detectors operate at approximately the same signal level so that their differences will cancel more exactly. The operating level should be 1 volt or higher.

It is convenient to build the detector in a small shielded enclosure such as an i-f transformer can fitted with coaxial input and output connectors. Voltage dividers can be similarly constructed so that it is easy to insert the desired amount of voltage attenuation from the various sources. In some cases it is convenient to use a pickup loop on the end of a short length of coaxial cable.

The phase shift of the amplifiers in the oscilloscope should be the same and their frequency response should be flat out to at least twenty times the frequency difference of the two test tones. Excellent high-frequency characteristics are necessary because the rectified SSB envelope contains harmonics extending to the limit of the envelope detector's response. Inadequate frequency response of the vertical amplifier may cause a little "foot" to appear on the lower end of the trace, as shown in figure 23. If it is small, it may be safely neglected.

Another spurious effect often encountered is a double trace, as shown in figure 24. This can usually be corrected with an RC network placed between one detector and the oscilloscope. The best method of testing the detectors and the amplifiers is to connect the input of the envelope detectors in parallel. A perfectly straight line trace will result when everything is working properly. One detector is then connected to the other r-f source through a voltage divider adjusted so that no appreciable change in the setting of

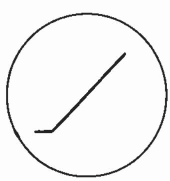


Figure 23

EFFECT OF INADEQUATE RESPONSE OF VERTICAL AMPLIFIER

the oscilloscope amplifier controls is required. Figure 25 illustrates some typical linearity traces. *Trace A* is caused by inadequate static plate current in class-A or class-B amplifiers or a mixer stage. To regain linearity, the grid bias of the stage should be reduced, the screen voltage should be raised,

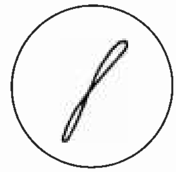


Figure 24

DOUBLE TRACE CAUSED BY PHASE SHIFT

or the signal level should be decreased. *Trace B* is a result of poor grid-circuit regulation when grid current is drawn, or a result of nonlinear plate characteristics of the amplifier tube at large plate swings. More grid swamping should be used, or the exciting signal should be reduced. A combination of the effects of A and B are shown in *Trace C*. *Trace D* illustrates amplifier overloading. The exciting signal should be reduced.

A means of estimating the distortion level observed is quite useful. The first- and third-order distortion components may be derived by an equation that will give the approximate signal-to-distortion level ratio of a *two-tone* test signal, operating on a given linearity curve. Figure 26 shows a linearity curve with two ordinates erected at half and full peak input signal level. The length of

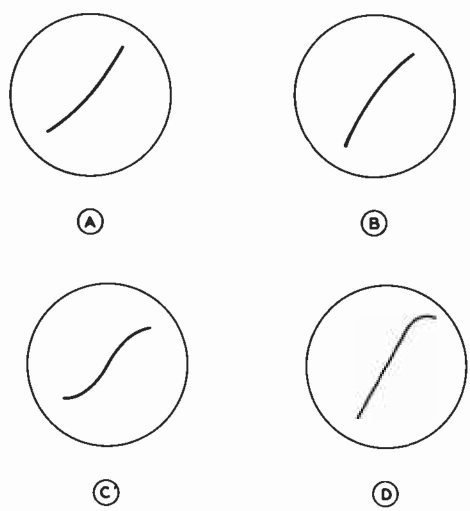


Figure 25

TYPICAL LINEARITY TRACES

the ordinates e_1 and e_2 may be scaled and used in the following equation:

Signal-to-distortion ratio in db =

$$20 \log \frac{3\, e_1 - e_2}{2\, e_1 - e_2}$$

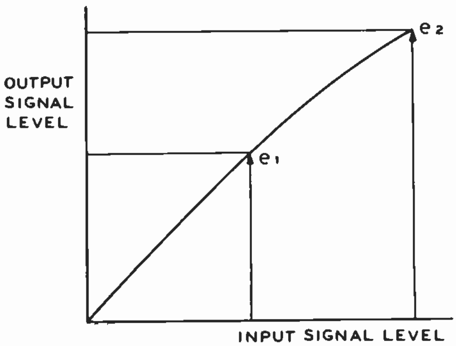


Figure 26

ORDINATES ON LINEARITY CURVE FOR 3RD-ORDER DISTORTION EQUATION

30-6 A-M Applications

The oscilloscope may be used as an aid for the proper operation of an a-m transmitter, and may be used as an indicator of the over-all performance of the transmitter output signal, and as a modulation monitor.

Waveforms There are two types of patterns that can serve as indicators, the *trapezoidal pattern* (figure 27) and the *modulated-wave pattern* (figure 28). The trapezoidal pattern is presented on the screen by impressing a modulated carrier-wave signal on the vertical deflection plates and the signal that modulates the carrier-wave signal (the modulating signal) on the horizontal deflection plates. The trapezoidal pattern can be analyzed by the method used previously in analyzing waveforms. Figure 29 shows how the signals cause the electron beam to trace out the pattern.

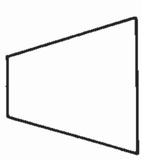


Figure 27

TRAPEZOIDAL MODULATION PATTERN

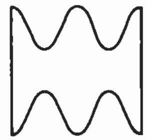


Figure 28

MODULATED CARRIER-WAVE PATTERN

The modulated-wave pattern is accomplished by presenting a modulated carrier wave on the vertical deflection plates and by using the time-base generator for horizontal deflection. The modulated-wave pattern also can be used for analyzing waveforms. Fig-

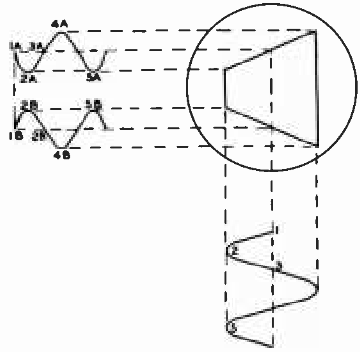


Figure 29

PROJECTION DRAWING SHOWING TRAPEZOIDAL PATTERN

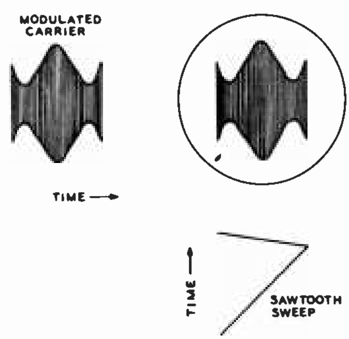


Figure 30

PROJECTION DRAWING SHOWING MODULATED-CARRIER WAVE PATTERN

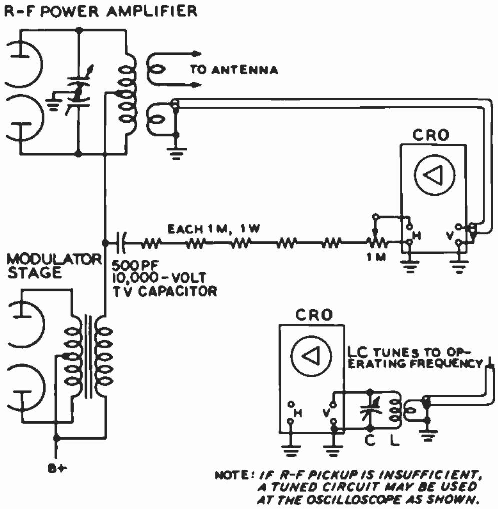


Figure 31

MONITORING CIRCUIT FOR TRAPEZOIDAL MODULATION PATTERN

ure 30 shows how the two signals cause the electron beam to trace out the pattern.

The Trapezoidal Pattern The oscilloscope connections for obtaining a trapezoidal pattern are shown in figure 31. A portion of the audio output of the transmitter modulator is applied to the horizontal input of the oscilloscope. The vertical amplifier of the oscilloscope is disconnected, and a small amount of modulated r-f energy is coupled directly to the vertical deflection plates of the oscilloscope. A small pickup loop, loosely coupled to the final amplifier tank circuit and connected to the vertical deflection plates by a short length of coaxial line will suffice. The amount of excitation to

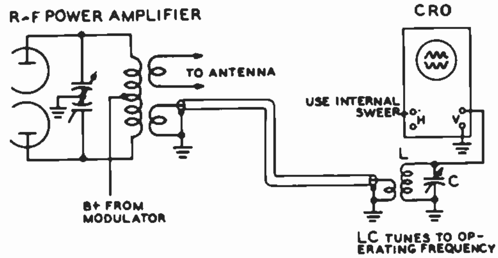


Figure 35

MONITORING CIRCUIT FOR MODULATED-WAVE PATTERN

the plates of the oscilloscope may be adjusted to provide a pattern of convenient size. On modulation of the transmitter, the trapezoidal pattern will appear. By changing the degree of modulation of the carrier wave the shape of the pattern will change. Figures 32 and 33 show the trapezoidal pattern for various degrees of modulation. The percentage of modulation may be determined by the following formula:

$$\text{Modulation percentage} = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \times 100$$

where,

E_{max} and E_{min} are defined as in figure 32.

An overmodulated signal is shown in figure 34.

The Modulated-Wave Pattern The oscilloscope connections for obtaining a modulated-wave pattern are shown in figure 35. The internal sweep circuit of the oscilloscope is applied to the horizontal

TRAPEZOIDAL PATTERNS

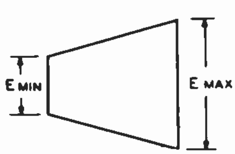


Figure 32

(LESS THAN 100% MODULATION)

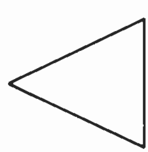


Figure 33

(100% MODULATION)

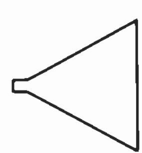


Figure 34

(OVERMODULATION)

CARRIER-WAVE PATTERN

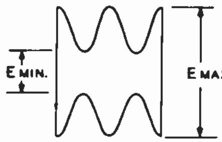


Figure 36

(LESS THAN 100% MODULATION)

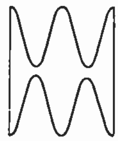


Figure 37

(100% MODULATION)

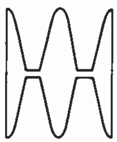


Figure 38

(OVERMODULATION)

plates, and the modulated r-f signal is applied to the vertical plates, as described before. If desired, the internal sweep circuit may be synchronized with the modulating signal of the transmitter by applying a small portion of the modulator output signal to the *external sync* post of the oscilloscope. The percentage of modulation may be determined in the same fashion as with a trapezoidal pattern. Figures 36, 37, and 38 show the modulated-wave pattern for various degrees of modulation.

CHAPTER THIRTY-ONE

Construction Practices

With a few possible exceptions, such as cabinets, brackets, neutralizing capacitors and transmitting coils, it hardly pays one to attempt to build the components required for the construction of an amateur transmitter. This is especially true when the parts are of the type used in construction and replacement work on receivers and TV, as mass production has made these parts very inexpensive.

Those who have and wish to spend the necessary time can effect considerable monetary saving in their equipment by building them from the component parts. The necessary data is given in the construction chapter of this handbook.

To many builders, the construction is as fascinating as the operation of the finished transmitter; in fact, many amateurs get so much satisfaction out of building a well-performing piece of equipment that they spend more time constructing and rebuilding equipment than they do operating the equipment on the air.

31-1 Tools

Beautiful work can be done with metal chassis and panels with the help of only a few inexpensive tools. The time required for construction, however, will be greatly reduced if a fairly complete assortment of metal-working tools is available. Thus, while an array of tools will speed up the work, excellent results may be accomplished with few tools, if one has the time and patience.

The investment one is justified in making in tools is dependent upon several factors. If you like to tinker, there are many tools useful in radio construction that you would probably buy anyway, or perhaps already have, such as screwdrivers, hammer, saws, square, vise, files, etc. This means that the money taken for tools from your radio budget can be used to buy the more specialized tools, such as socket punches or hole saws, taps and dies, etc.

The amount of construction work one does determines whether buying a large assortment of tools is an economical move. It also determines if one should buy the less expensive type offered at surprisingly low prices by the familiar mail order houses, "five and ten" stores, and chain auto-supply stores, or whether one should spend more money and get first-grade tools. The latter cost considerably more and work but little better when new, but will outlast several sets of the cheaper tools. Therefore they are a wise investment for the experimenter who does lots of construction work (if he can afford the initial cash outlay). The amateur who constructs only an occasional piece of apparatus need not be so concerned with tool life, as even the cheaper grade tools will last him several years, if they are given proper care.

The hand tools and materials in the accompanying lists will be found very useful around the home workshop. Materials not listed but ordinarily used, such as paint, can best be purchased as required for each individual job.

ESSENTIAL HAND TOOLS AND MATERIALS

1 *Good* electric soldering iron, about 100 watts; or soldering gun
1 Spool rosin-core wire solder
1 Each large, medium, small, and midget screwdrivers
1 *Good* hand drill (eggbeater type), preferably two-speed
1 Pair regular pliers, 6 inch
1 Pair long-nose pliers, 6 inch
1 Pair cutting pliers (diagonals), 5 inch or 6 inch
1 ⅝-inch socket punch
1 "Boy Scout" knife
1 Combination square and steel rule, 1 foot
1 Yardstick or steel pushrule
1 Scratch awl
1 Center punch
1 Dozen or more assorted round shank drills (as many as you can afford between No. 50 and ¼ or ⅜ inch, depending upon size of hand drill chuck)
1 Set Allen and spline-head wrenches
1 Set Phillips screwdrivers
1 Hacksaw and blades
1 Medium file and handle
1 Cold chisel (½ inch tip)
1 Wrench for socket punch
1 Hammer
Light machine oil (in squirt can)
Vinyl electrical tape

HIGHLY DESIRABLE HAND TOOLS AND MATERIALS

1 Bench vise (jaws at least 3 inch)
1 Carpenter's brace, ratchet type
1 Square-shank countersink bit
1 Square-shank taper reamer, small
1 Square-shank taper reamer, large (the two reamers should overlap; ½ inch and ⅞ inch size will usually be suitable)
1 ¾-inch socket punch
1 ⅞-inch socket punch
1 Adjustable circle cutter for holes to 3 inch
1 Set small, inexpensive, open-end wrenches
1 Set jewelers screwdrivers
1 Soldering pencil, 30 watt
1 Wood chisel (½ inch tip)
1 Pair wing dividers
1 Coarse mill file, flat 12 inch
1 Coarse bastard file, round, ½ or ¾ inch
1 Set nutdrivers (¼, 5/16, 11/32)
4 Small "C" clamps
6 or 8 Assorted small files; round, half-round or triangular, flat, square, rat-tail
Sandpaper and emery cloth, coarse, medium, and fine
Epoxy cement
File brush

USEFUL TOOLS AND MATERIALS

1 Jig or scroll saw (small) with metal-cutting blades
1 Aerosol can, contact cleaner
1 Wiss metal snips
1 Wire stripper
1 "Pop" rivet gun
1 Tap and die set for 4-40, 6-32, 8-32, 10-32 and 10-24 machine screw threads
4 Medium size "C" clamps
1 Metal "nibbling" tool
1 Set alignment tools
1 Electric drill, ¼-inch, variable speed
1 DYMO label embosser
1 Can paint thinner
1 Drill press
1 Shop vacuum cleaner
Aerosol spray paints.
Dusting brush
Paint brushes
Sheet Lucite, or polystyrene

Not listed in the table are several special-purpose radio tools which are somewhat of a luxury, but are nevertheless quite handy, such as various around-the-corner screwdrivers and wrenches, special soldering iron tips, etc. These can be found in the larger radio parts stores and are usually listed in their mail order catalogs.

Metal Chassis Though quite a few more tools and considerably more time will be required for metal-chassis con-

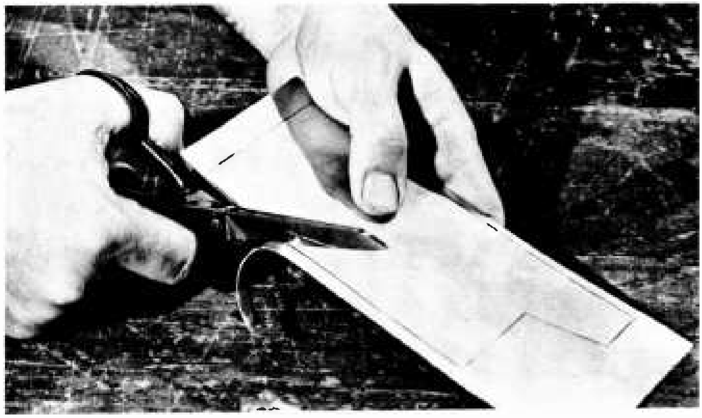


Figure 1
SOFT ALUMINUM SHEET MAY BE CUT WITH HEAVY KITCHEN SHEARS

struction, much neater and more satisfactory equipment can be built by mounting the parts on sheet metal chassis or circuit boards instead of "breadboards." This type of construction is necessary when shielding of the apparatus is required. A front panel and a back shield minimize the danger of shock and complete the shielding of the inclosure.

31-2 The Material

Electronic equipment may be built on a foundation of circuit board, steel, or aluminum. The choice of foundation material is governed by the requirements of the electrical circuit, the weight of the components of the assembly, and the financial cost of the project when balanced against the pocketbook contents of the constructor.

Breadboard and Brassboard Experimental circuits may be built up in a temporary fashion termed *breadboarding*, a term reflecting the old practice of the "twenties" when circuits were built on wooden boards. Modern breadboards may be built upon circuit board material or upon prepunched phenolic boards. The prepunched boards contain a grid of small holes into which the component leads may be anchored for soldering.

A *brassboard* is an advanced form of assembly in which the experimental circuit is

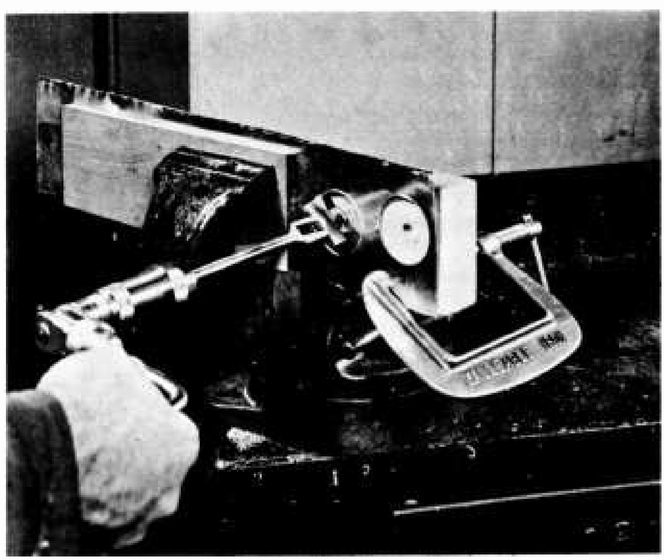


Figure 2
CONVENTIONAL WOOD EXPANSION BIT IS EFFECTIVE IN DRILLING SOCKET HOLES IN SOFT ALUMINUM

Figure 3
SOFT ALUMINUM TUBING MAY BE BENT AROUND WOODEN FORM BLOCKS. TO PREVENT THE TUBE FROM COLLAPSING ON SHARP BENDS, IT IS PACKED WITH WET SAND

built up in semipermanent form on a metal chassis or copper-plated circuit board. Manufacture and use of printed-circuit boards is covered later in this chapter.

Figure 4
A WOODWORKING PLANE MAY BE USED TO SMOOTH OR TRIM THE EDGES OF ALUMINUM STOCK.

Special Frameworks For high-powered r-f stages, many amateur constructors prefer to discard the more conventional types of construction and employ instead special metal frameworks and brackets which they design specially for the parts which they intend to use. These are usually arranged to give the shortest possible r-f leads and to fasten directly behind a panel by means of a few bolts, with the control shafts projecting through corresponding holes in the panel.

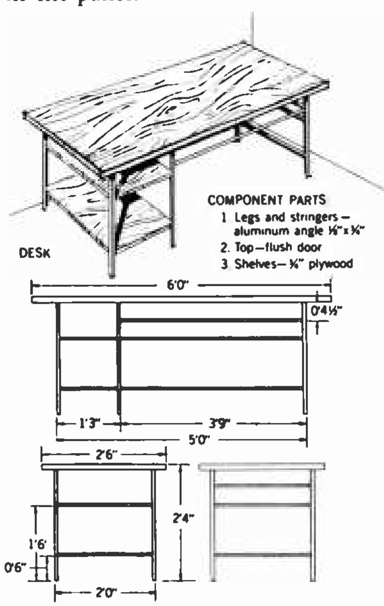


Figure 5
INEXPENSIVE OPERATING DESK MADE FROM ALUMINUM ANGLE STOCK, PLYWOOD AND A FLUSH-TYPE DOOR

Working with Aluminum The necessity of employing "electrically tight inclosures" for the containment of TVI-producing harmonics has led to the general use of aluminum for chassis, panel, and inclosure construction. If the proper type of aluminum material is used, it may be cut and worked with the usual woodworking tools found in the home shop. Hard, brittle aluminum alloys such as 2024 and 6061 should be avoided, and the softer materials such as 1100 or 3003 should be employed.

Reynold's *Do-it-Yourself* aluminum, which is being distributed on a nationwide basis through hardware stores, lumber yards, and building material outlets, is an alloy which is temper selected for easy working with ordinary tools. Aluminum sheet, bar, and angle stock may be obtained, as well as perforated sheets for ventilated inclosures.

Figures 1 through 4 illustrate how this soft material may be cut and worked with ordinary shop tools, and figure 5 shows a simple operating desk that may be made from aluminum angle stock, plywood, and a flush-type six-foot door.

31-3 TVI-Proof Inclosures

Armed with a right-angle square, tin-snips and a straight edge, the home constructor will find the assembly of aluminum inclosures an easy task. This section will show simple construction methods, and short cuts in producing inclosures.

The simplest type of aluminum inclosure is that formed from a single sheet of perforated material as shown in figure 6. The top, sides, and back of the inclosure are of one piece, complete with folds that permit the formed inclosure to be bolted together along the edges. The top area of the inclosure should match the area of the chassis to ensure a close fit. The front edge of the inclosure is attached to aluminum angle strips that are bolted to the front panel of the unit; the sides and back can either be bolted to matching angle strips affixed to the chassis, or may simply be attached to the edge of the chassis with self-tapping sheet-metal screws.

A more sophisticated inclosure is shown in figure 7. In this assembly aluminum angle stock is cut to length to form a framework on which the individual sides, back, and top of the inclosure are bolted. For greatest strength, small aluminum gusset plates should be affixed in each corner of the inclosure. The complete assembly may be held together by No. 6 sheet-metal screws or "pop" rivets.

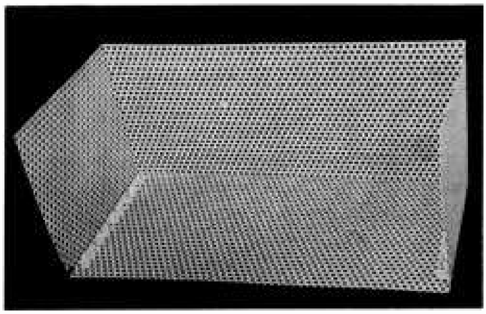


Figure 6
TVI INCLOSURE MADE FROM SINGLE SHEET OF PERFORATED ALUMINUM
Reynolds Metal Co. "Do-it-yourself" aluminum sheet may be cut and folded to form TVI-proof inclosure. One-half inch lip on edges is bolted to center section with 6-32 machine screws.

Regardless of the type of inclosure to be made, care should be taken to ensure that all joints are square. Do not assume that all prefabricated chassis and panels are absolutely true and square. Check them before you start to form your shield because any dimensional errors in the foundation will cause endless patching and cutting *after* your inclosure is bolted together. Finally, be sure that paint is removed from the panel and chassis at the point the inclosure attaches to the foundation. A clean, metallic contact along the seam is required for maximum harmonic suppression.

31-4 Inclosure Openings

Openings into shielded inclosures may be made simply by covering them by a piece of shielding held in place by sheet-metal screws.

Openings through vertical panels, however, usually require a bit more attention to prevent leakage of harmonic energy through the crack of the door which is supposed to seal the opening. Hinged door openings, however, do not seal tightly enough to be called TVI-proof. In areas of high TV signal strength where a minimum of operation on

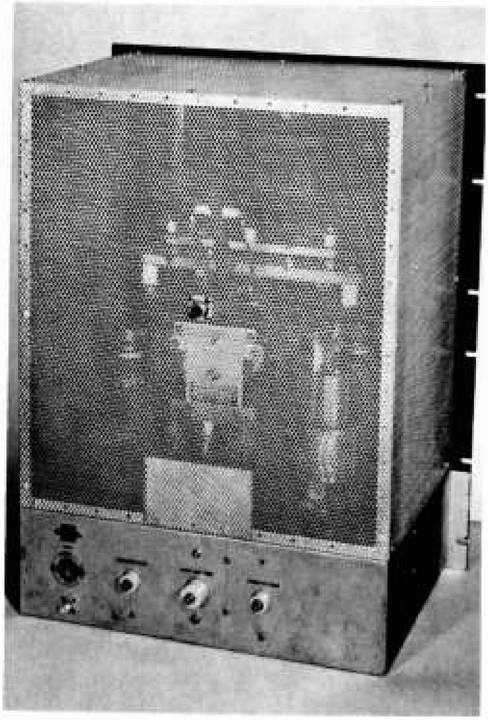


Figure 7
TVI-PROOF INCLOSURE BUILT OF PERFORATED ALUMINUM SHEET AND ANGLE STOCK

28 MHz is contemplated, the door is satisfactory as-is. To accomplish more complete harmonic suppression the edges of the opening should be lined with preformed contact finger stock. *Eimac* finger stock is an excellent means of providing good contact continuity when using components with adjustable or moving contact surfaces, or in acting as electrical "weatherstrip" around access doors in inclosures. Harmonic leakage through such a sealed opening is reduced to a negligible level. The mating surface to the finger stock should be free of paint, and should provide a good electrical connection to the stock.

31-5 Construction Practice

Chassis Layout The chassis first should be covered with a layer of wrapping paper, which is drawn tightly down on all sides and fastened with scotch tape. This allows any number of measurement lines and hole centers to be spotted in the correct positions without making any marks on the chassis itself. Place on it the parts to be mounted and play a game of chess with them, trying different arrangements until all the grid and plate leads are made as short as possible, tubes are clear of coil fields, r-f chokes are in safe positions, etc. Remember, especially if you are going to use a panel, that a good mechanical layout often can accompany sound electrical design, but that the electrical design should be given first consideration.

All too often parts are grouped to give a symmetrical panel, irrespective of the arrangement behind. When a satisfactory arrangement has been reached, the mounting holes may be marked. The same procedure now must be followed for the underside, always being careful to see that there are no clashes between the two (that no top mounting screws come down into the middle of a paper capacitor on the underside, that the variable capacitor rotors do not hit anything when turned, etc.).

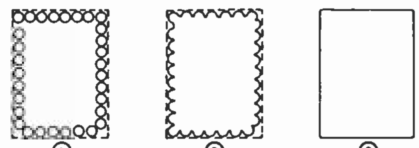


Figure 8
MAKING A RECTANGULAR CUTOUT

When all the holes have been spotted, they should be center-punched *through* the paper into the chassis. Don't forget to spot holes for leads which must also come through the chassis.

For transformers which have lugs on the bottoms, the clearance holes may be spotted by pressing the transformer on a piece of paper to obtain impressions, which may then be transferred to the chassis.

Punching In cutting socket holes, one can use either a fly-cutter or socket punches. These punches are easy to operate and only a few precautions are necessary. The guide pin should fit snugly in the guide

Figure 9
NUMBERED DRILL SIZES

DRILL NUMBER	Diameter (in.)	Clears Screw	Correct for Tapping Steel or Brass†
1	.228	—	—
2	.221	12-24	—
3	.213	—	14-24
4	.209	12-20	—
5	.205	—	—
6	.204	—	—
7	.201	—	—
8	.199	—	—
9	.196	—	—
10*	.193	10-32	—
11	.191	10-24	—
12*	.189	—	—
13	.185	—	—
14	.182	—	—
15	.180	—	—
16	.177	—	12-24
17	.173	—	—
18*	.169	8-32	—
19	.166	—	12-20
20	.161	—	—
21*	.159	—	10-32
22	.157	—	—
23	.154	—	—
24	.152	—	—
25*	.149	—	10-24
26	.147	—	—
27	.144	—	—
28*	.140	6-32	—
29*	.136	—	8-32
30	.128	—	—
31	.120	—	—
32	.116	—	—
33*	.113	4-36 4-40	—
34	.111	—	—
35*	.110	—	6-32
36	.106	—	—
37	.104	—	—
38	.102	—	—
39*	.100	3-48	—
40	.098	—	—
41	.096	—	—
42*	.093	—	4-36 4-40
43	.089	2-56	—
44	.086	—	—
45*	.082	—	3-48

*Sizes most commonly used in radio construction.
†Use next size larger for tapping bakelite and similar composition materials (plastics, etc.).

hole. This increases the accuracy of location of the socket. If this is not of great importance, one may well use a drill of 1/32 inch larger diameter than the guide pin.

The male part of the punch should be placed in the vise, cutting edge up and the female portion forced against the metal with a wrench. These punches can be obtained in sizes to accommodate all tube sockets and even large enough to be used for meter holes. In the large socket sizes they require the use of a 3/8-inch center hole to accommodate the bolt.

Transformer Cutouts Cutouts for transformers and chokes are not so simply handled. After marking off the part to be cut, drill about a 1/4-inch hole on each of the inside corners and tangential to the edges. After burring the holes, clamp the piece and a block of cast iron or steel in the vise. Then, take your burring chisel and insert it in one of the corner holes. Cut out the metal by hitting the chisel with a hammer. The blows should be light and numerous. The chisel acts against the block in the same way that the two blades of a pair of scissors work against each other. This same process is repeated for the other sides. A file is used to trim up the completed cutout.

Another method is to drill the four corner holes large enough to take a hack saw blade, then saw instead of chisel. The four holes permit nice looking corners.

Still another method is shown in figure 8. When heavy panel steel is used and a drill press or electric drill is available, this is the most satisfactory method.

Removing Burrs In both drilling and punching, a burr is usually left on the work. There are three simple ways of removing these. Perhaps the best is to take a chisel (be sure it is one for use on metal) and set it so that its bottom face is parallel to the piece. Then gently tap it with a hammer. This usually will make a clean job with a little practice. If one has access to a counterbore, this will also do a nice job. A countersink will work, although it bevels the edges. A drill of several sizes larger is a much used arrangement. The third method is by filing off the burr, which does a good job but scratches the adjacent metal surfaces badly.

Mounting Components There are two methods in general use for the fastening of transformers, chokes, and similar pieces of apparatus to chassis or breadboards. The first, using nuts and machine screws, is slow, and the manufacturing practice of using self-tapping screws or rivets is gaining favor. For the mounting of small parts such as resistors and capacitors, "tie points" are very useful to gain rigidity. They also contribute materially to the appearance of finished apparatus.

Rubber grommets of the proper size placed in all chassis holes through which wires are to be passed, will give a neater appearing job and also will reduce the possibility of short circuits.

Soldering Making a strong, low-resistance solder joint does not mean just dropping a blob of solder on the two parts to be joined and then hoping that they'll stick. There are several definite rules that *must* be observed.

All parts to be soldered must be absolutely clean. To clean a wire, lug, or whatever it may be, take your pocket knife and scrape it thoroughly, until fresh metal is laid bare. It is not enough to make a few streaks; scrape until the part to be soldered is bright.

Make a good mechanical joint before applying any solder. Solder is intended primarily to make a good *electrical* connection; mechanical rigidity should be obtained by bending the wire into a small hook at the end and nipping it firmly around the other part, so that it will hold well even before the solder is applied.

Keep your iron properly tinned. It is impossible to get the work hot enough to take the solder properly if the iron is dirty. To tin your iron, file it, while hot, on one side until a full surface of clean metal is exposed. Immediately apply rosin core solder until a thin layer flows completely over the exposed, surface. Repeat for the other faces. Then take a clean rag and wipe off all excess solder and rosin. The iron should also be wiped frequently while the actual construction is going on; it helps prevent pitting the tip.

Apply the solder to the work, not to the iron. The iron should be held against the parts to be joined until they are thoroughly heated. The solder should then be applied against the parts and the iron should be held in place until the solder flows smoothly and envelops the work. If it acts like water on a greasy plate, and forms a ball, the work is not sufficiently clean.

The completed joint must be held perfectly still until the solder has had time to solidify. If the work is moved before the solder has become *completely* solid, a "cold" joint will result. This can be identified immediately, because the solder will have a dull "white" appearance rather than one of shiny "silver." Such joints tend to be of high resistance and will very likely have a bad effect on a circuit. The cure is simple, merely reheat the joint and do the job correctly.

For general construction work, 60-40 solder (60% tin, 40% lead) is generally used. It melts at 370°F.

Finishes If the apparatus is constructed on a painted chassis (commonly available in black and gray wrinkle and "hammertone"), there is no need for application of a protective coating when the equipment is finished, assuming that you are careful not to scratch or mar the finish while drilling holes and mounting parts. However, many amateurs prefer to use unpainted (zinc or cadmium plated) chassis, because it is much simpler to make a chassis ground connection with this type of chassis. In localities near the sea coast it is a good idea to lacquer the edges of the various chassis cutouts even on a painted chassis, as rust will get a good start at these points unless the metal is protected where the drill or saw has exposed it. If too thick a coat is applied, the lacquer will tend to peel. It may be thinned with lacquer thinner to permit application of a light coat. A thin coat will adhere to any *clean* metal surface that is not too shiny.

An attractive dull gloss finish, almost velvety can be put on aluminum by sand-blasting it with a very weak blast and fine particles and then lacquering it. Soaking the aluminum in a solution of lye produces somewhat the same effect as a fine-grain sand blast.

Metal panels and inclosures may be painted an attractive color with the aid of aerosol spray paint, available in many colors. After the panel is spray-painted, press-on *decals* may be used to letter the panel. Once the decals have dried, the panel may then be given a spray coat of clear plastic or lacquer to hold the decals in position and to protect the surface.

To protect brass from tarnish, thoroughly cleanse and remove the last trace of grease by the use of paint thinner. The brass must be carefully rinsed with water and dried; but in doing it, care must be taken not to handle any portion with the bare hands or anything else that is greasy. Then apply an acrylic aerosol spray.

31-6 Printed Circuits

Etched or *printed circuits* were developed to apply mass-production techniques to electronic assemblies, utilizing the processes of the graphic arts industry. On a large-volume basis, the etched-circuit technique provides uniformity of layout and freedom from wiring errors at a substantial reduction in assembly time and cost. In this assembly scheme, the methods of the photoengraving process are used to print photographic patterns representing electronic circuitry on copper-foil clad insulating board. By using an *etch-resistant* material (impervious to acid) for the pattern of conductors, the unmasked areas of the foil may be etched away, leaving the desired conducting pattern, conforming to the wiring harness of the electronic assembly.

The etched board is drilled at appropriate places to accept lead wires, thus permitting small components such as resistors and capacitors to be affixed to the board by inserting the leads in the matching holes. Larger components, such as sockets, inductors, and small transformers, are fitted with tabs which pass through matching holes in the board. The various components are interconnected by the foil conductors on one or both sides of the board. All joints are soldered at one time by immersing one side of the board in molten solder.

The foil-clad circuit board is usually made of laminated material such as phenolic, silicon, *teflon*, or *fiberglas*, impregnated with resin and having a copper foil of 0.0007- to 0.009-inch thickness affixed to the board under heat and pressure. Boards are available in thicknesses of 1/64 to 1/4 inch.

While large production runs of etched-circuit boards are made by a photographic process utilizing a master negative and photosensitive board, a simpler process may be used by experimenters to produce circuit boards in the home workshop through the use of *tape* or *ink resist*, plus a chemical solution which etches away all unmasked copper, without affecting the circuit board.

Homemade Circuit Boards Circuit boards may be easily constructed for electronic assemblies without the need of photographic equipment. The method is simple and fast and requires few special materials. The circuit board is made from a full-scale template of the circuit. Precut board is available from large radio supply houses as are the etchant and resist used in this process. This is how the board is prepared:

Step 1—A full-scale template of the desired circuit is drawn. Lead placement must be arranged so that the conductors do not cross each other except at interconnection points. Holes for component leads and terminals are surrounded by a foil area for the soldered connection. It is suggested that a trial layout be drawn on a piece of graph paper, making the conductors about 1/16-inch wide and the terminal circles about 1/8-inch in diameter. When conductors must cross, a point is selected where a component may be used to bridge one conductor; or a wire jumper may be added to the circuit.

Special layout paper marked with the same pattern as on perforated boards may also be used.

Step 2—The template is transferred to the foil-clad board. The board should be unsensitized and cut somewhat oversize. Either single-clad or dual-clad board may be used. For simple circuits, the complete layout can be traced on the board by eye, using a ruler and a pencil. For more complicated circuits, the template should be applied directly to the copper foil by the use of rubber cement. The circuit is traced and the board lightly centerpunched at all drill points for reference. The template and cement are now removed.

Step 3—Once the board has been punched, the board is cleaned to remove copper oxide. A bright, uniform finish is required to ensure proper adhesion of the resist and complete etching. Kitchen cleaning powder may be used for this operation, followed by a thorough washing of the board in water. Care should be taken to avoid touching the copper foil from this point on. Now, to etch out the circuit on the copper foil, the resist material is applied to areas where the copper will remain, and the areas that are not covered with resist will be etched away.

Step 4—The conductors and interconnecting points are laid down on the copper laminate using resist material (figure 10). One form of resist is liquid and is applied from a resist marking pen. A second form

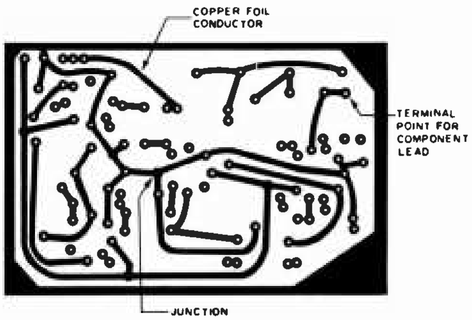


Figure 10
LIQUID RESIST MAKES PRACTICAL PRINTED CIRCUIT

Liquid resist is applied to copper foil of circuit board to protect conductor areas from etchant. Each lead hole is circled, the circle being about four times the diameter of the hole. After the holes have been circled, lines are drawn between them in accordance with the circuit sketch. Junctions are marked with a solid circle. Connecting path should be about 1/16-inch wide, ample to carry a current of about 10 amperes, if required. Tape or a "transfer" resist material provides professional appearance to board. Placement of components may be marked on reverse of board in India Ink.

of resist is thin vinyl tape having adhesive backing. In an emergency, India Ink or nail polish may be used for resist. Using the original templates as a visual guide, the resist is applied to the clean foil and allowed to dry.

Suitable etchants are *ferric chloride* or *ammonium persulfate*. The etchant may be liquid or a powder which is mixed with hot water according to directions. Ready-made etchant kits using these chemicals are available from several manufacturers.

The board is now ready to be immersed in an *etchant bath,* or tank. A quick and effective etching technique makes use of a *froth etching* bath (figure 11), described as follows.

The Froth Etching Technique The froth etcher is designed for fast etching of both single and double faced boards on which fine resolution is also important. It produces uniformly etched boards in about four minutes with very little undercutting of the foil. As a bonus, the process automatically aerates the etchant, greatly extending its life.

Figure 11

FROTH ETCHING IS QUICK AND EASY

The continuous air flow through the aerators creates a surface froth that "scrubs" the circuit board with constantly agitated etchant. The sliding clamp holder which is attached to the dish cover permits rapid insertion or reversal of the printed circuit board. Sample board is clamped to cover holder in foreground.

Constructing the froth etcher tank is quite simple. A heat-resistant glass dish (*Pyrex*, or equivalent) with cover serves as the tank. Also required are a tungsten-carbide hacksaw blade to notch the dish cover, some two-part epoxy adhesive, some rubber air tubes and a thermometer. To provide the continuous air flow, three inexpensive ceramic aquarium aerators and an aquarium pump are used. Finally, a plexiglass holder for the boards is required.

The small ceramic aerators are cemented to the bottom of the glass dish, as shown in figure 12. The quick-change printed circuit board holder is cemented to the glass cover as shown in figure 13. The thermometer and short lengths of plastic tubing which serve as holders for the air hoses are cemented to

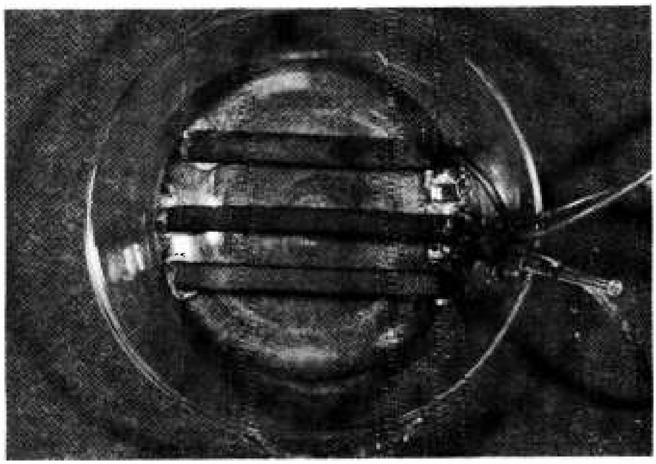


Figure 12

INTERIOR OF ETCHING BATH

Aquarium aerators are cemented into the bottom of the heat-resistant glass dish, along with sections of plastic tubing to support the rubber air tubes and the thermometer. The tubes are connected by T-fittings to a single tube running to the main air supply, which is an aquarium air pump.

the side of the dish and the cover is notched to provide egress for them. The complete froth bath assembly is shown in figure 13.

The continuous air flow through the aerators creates a surface froth that "scrubs" the circuit board with constantly agitated etchant. The board is held in position in the bath by the plexiglass holder shown in figure 14. The etchant used consists of ferric chloride in the proportion of 1¾ pounds of *FeCl* to every quart of water, mixed at a temperature of between 100°F and 110°F.

The froth bath is placed on an electric hot plate and filled with etchant to a level that just reaches the bottom of the copper-clad board when it is mounted in the lid holder. The etchant is heated to its lower operating temperature (100°F) and the hot plate is turned off. The board is now placed in the holder, the cover placed on the dish and the air supply is turned on, adjusting it to create a continuous, vigorous froth over the total surface of the etchant. After a few minutes—anything from three to eight minutes, depending on the freshness of the solution—inspect the board by raising the cover. The air supply must be turned off first to prevent splattering of the etchant. When the process is observed to be complete, the board is removed and washed in clean water.

The resist material can be left on the board to protect the conductors until the board is cut to final size, clamped between wood blocks in a vise and trimmed with a fine hacksaw blade. The resist is then removed with soft steel wool or a solvent. The

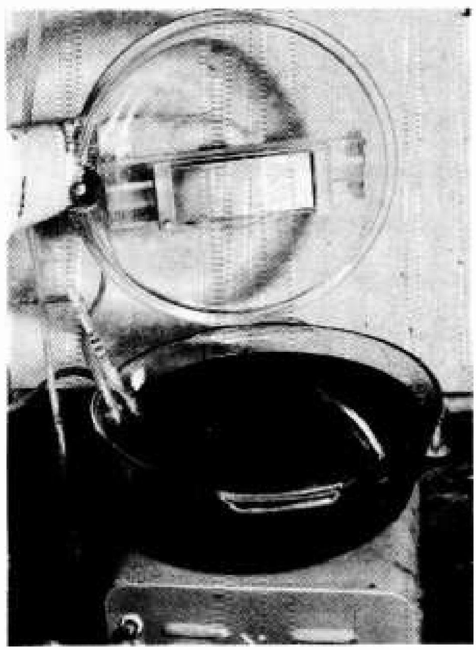


Figure 13

BATH LID AND CIRCUIT BOARD HOLDER

Plexiglass holder grips the edge of the printed circuit board, assuring uniform etch of the entire surface. One clamp is threaded and fitted with a nylon screw to accomodate boards of various sizes. A rubber band around the clamp provides tension. Observing the etching process is easily done by lifting the heat-resistant glass etcher cover to which the printed circuit board is attached. Before the cover is removed, air supply must be turned off to prevent any splattering of the etchant.

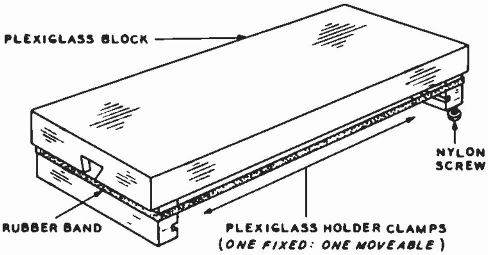


Figure 14

PLEXIGLASS HOLDER FOR ETCHANT TANK

complete board is then given a final cleaning with soft steel wool and the center-punched points drilled with a #54 pilot drill. The holes are then drilled out to a larger size as required for component assembly.

The components are mounted to the board on the side opposite the conductors. The leads are passed through the appropriate holes, bent slightly to hold the component in place, and then clipped close to the conductor surface. After checking placement and observing polarity where necessary, the leads may be individually soldered to the conductor with a small pencil-tip iron. Use small diameter (0.032-inch diameter or smaller) solder and take care not to overheat the board or components during this operation. The last step is to wash the circuit side of the board with solvent to remove any soldering flux and then to give the board a coating of clear acrylic (*Krylon*) plastic spray from an aerosol can.

(The Froth etching technique is reprinted from *Electronics*, July 3, 1972; copyright McGraw-Hill, Inc. 1972).

31-7 Coaxial Cable Terminations

Commercial electronics equipment usually employs *series N* and *series BNC* coaxial connectors, whereas the majority of amateur equipment employs the older *UHF series* coaxial connectors. Shown in figure 15 is a simplified and quick method of placing the UHF plug (PL-259) on RG-8A/U or RG-11/U coaxial line. The only special tools needed are a *Stanley 99A* (or equivalent) shop knife and a *General Hardware* 123 (or equivalent) midget tubing cutter.

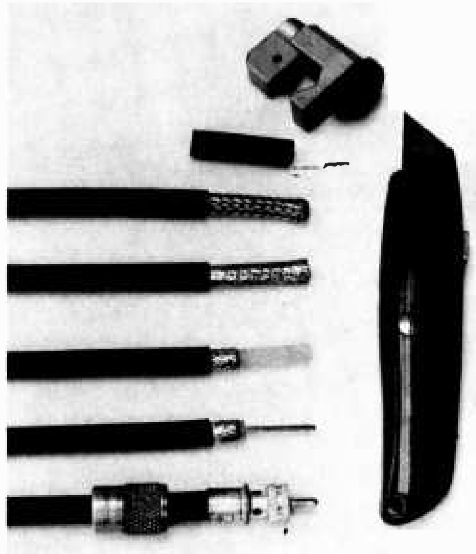


Figure 15

CABLE PREPARATION FOR PL-259 COAXIAL PLUG

Midget tubing cutter and utility knife are used to prepare RG-8/U or RG-11/U cable for uhf-type plug. Cable jacket is removed and outer braid tinned with hot iron. Braid is then cut with tubing cutter and inner insulation trimmed with knife. PL-259 shell is twisted on cable and soldered in position through holes in shank.

The first step is to slide the coupling ring of the PL-259 plug over the coaxial line. Next, the utility knife is used to circumscribe a cut in the outer, black vinyl jacket of the cable 1¼ inches back from the end. The cut should be square, and the free jacket piece is slit and removed from the cable.

Next, using a hot iron or soldering gun, quickly tin the exposed braid of the cable. Do this quickly so the inner polyethylene insulation does not soften. Clean the flux from the braid with paint thinner after the solder cools.

The next step is to cut the solid, tinned braid with the tubing cutter so that 7/16 inch remains. Mark the cutting line with a pencil and place the cutting wheel over the mark. Tighten the wheel and revolve the cutter about the cable. The unwanted braid end may be removed, using wire cutters as snips.

Next, trim the inner polyethylene insulation with the utility knife so that 1/16 inch remains exposed beyond the braid. Using a circular cut, slice the insulation and pull the slug free with a twisting motion. Tin the inner conductor. The last step is to push the shell of the PL-259 plug on the prepared cable end. Screw it on with your fingers until the tinned braid is fully visible through the solder holes of the plug. Using an iron with a small point, solder the plug to the braid through the four holes, using care that the solder does not run over the outer threads of the plug. Lastly, run the coupling ring down over the plug and solder the inner conductor to the plug tip.

31-8 Workshop Layout

The *size* of your workshop is relatively unimportant since the shop *layout* will determine its efficiency and the ease with which you may complete your work.

Shown in figure 16 is a workshop built into a 10′ × 10′ area in the corner of a garage. The workbench is 32″ wide, made up of four strips of 2″ × 8″ lumber supported on a solid framework made of 2″ × 4″ lumber. The top of the workbench is covered with hard-surface *Masonite.* The edge of the surface is protected with aluminum "counter edging" strip, obtainable at large hardware stores. Two wooden shelves 12″ wide are placed above the bench to hold the various items of test equipment. The shelves are bolted to the wall studs with large angle brackets and have wooden end pieces. Along the edge of the lower shelf a metal "outlet strip" is placed that has a 117-volt outlet every six inches along its length. A similar strip is run along the *back* of the lower shelf. The front strip is used for equipment that is being bench-tested, and the rear strip powers the various items of test equipment placed on the shelves.

At the left of the bench is a storage bin for small components. A file cabinet can be placed at the right of the bench. This neccessary item holds schematics, transformer data sheets, and other papers that normally are lost in the usual clutter and confusion.

The area below the workbench has two storage shelves which are concealed by sliding doors made of 1/4-inch *Masonite.* Heavier tools, and large components are stored in this area. On the floor and not shown in the

Figure 16

GOOD SHOP LAYOUT AIDS CAREFUL WORKMANSHIP

Built in a corner of a garage, this shop has all features necessary for electronic work. Test instruments are arranged on shelves above bench. Numerous outlets reduce "haywire" produced by tangled line cords. Not shown in picture are drill press and sander at end of left bench.

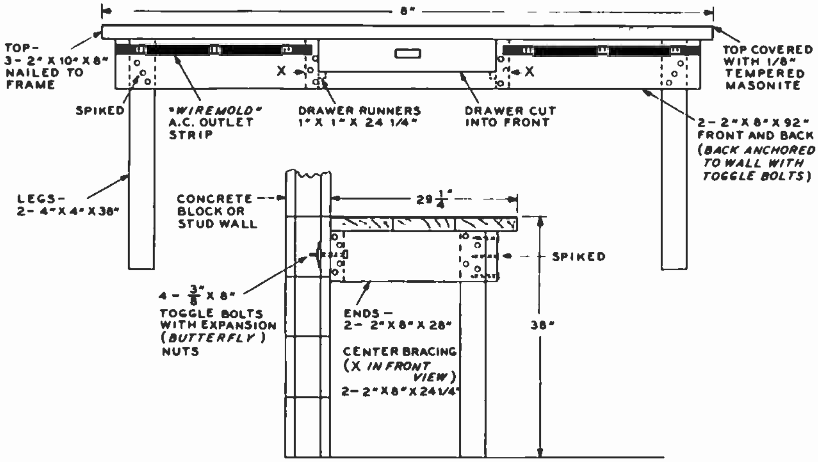


Figure 17

HEAVY-DUTY WORKBENCH TO BOLT TO CEMENT BLOCK OR STUD WALL

photograph is a very necessary item of shop equipment: a large trash receptacle.

A heavy duty workbench that may be bolted to a cement block or stud wall is shown in figure 17.

31-9 Components and Hardware

Procurement of components and hardware for a construction project can often be a time consuming and vexing task as smaller radio parts stores often have limited or incomplete stocks of only the most fast-moving items. Larger distributors carrying industrial stocks, however, maintain warehouse inventories of components or have facilities for obtaining them at short notice. It is recommended, therefore, that the experimenter have at hand catalogs from some of the larger supply houses which distribute to the electronics industry. The following industrial catalogs of large mail-order distributors are suggested as part of your technical library:

Allied Electronics Co., 100 No. Western Ave., Chicago, Ill. 60680; *Lafayette Radio Electronics*, 111 Jericho Turnpike, Syosset, N.Y. 11791; *Newark Electronics*, 500 No. Pulaski Rd., Chicago, Ill. 60624.

A complete 1700-page catalog of electronic parts and components (*The Radio Electronic Master Catalog*) may be obtained from United Technical Publications, 645 Stewart Ave., Garden City, N.Y. 11530. Copies of this master catalog are often available at large radio supply houses.

CHAPTER THIRTY-TWO

Radio Mathematics and Calculations

Radiomen often have occasion to calculate sizes and values of required parts. This requires some knowledge of mathematics. The following pages contain a review of those parts of mathematics necessary to understand and apply the information contained in this book. It is assumed that the reader has i.ad some mathematical training; this chapter is not intended to teach those who have never learned anything of the subject.

Fortunately only a knowledge of fundamentals is necessary, although this knowledge must include several branches of the subject. Fortunately, too, the majority of practical applications in radio work reduce to the solution of equations or formulas or the interpretation of graphs.

Arithmetic

Notation of Numbers In writing numbers in the Arabic system we employ ten different symbols, digits, or figures: 1, 2, 3, 4, 5, 6, 7, 8, 9, and 0, and place them in a definite sequence. If there is more than one figure in the number the *position* of each figure or digit is as important in determining its value as is the digit itself. When we deal with whole numbers the righthandmost digit represents units, the next to the left represents tens, the next hundreds, the next thousands, from which we derive the rule that every time a digit is placed one space further to the left its value is multiplied by ten.

8	**1**	**4**	**3**
thousands	hundreds	tens	units

It will be seen that any number is actually a sum. In the example given above it is the sum of eight thousands, plus one hundred, plus four tens, plus three units, which could be written as follows:

8	thousands (10 x 10 x 10)
1	hundreds (10 x 10)
4	tens
3	units
8143	

The number in the units position is sometimes referred to as a *first order* number, that in the tens position is of the *second order*, that in the hundreds position the *third order*, etc.

The idea of letting the position of the symbol denote its value is an outcome of the abacus. The abacus had only a limited number of wires with beads, but it soon became apparent that the quantity of symbols might be continued indefinitely towards the left, each further space multiplying the digit's value by ten. Thus any quantity, however large, may readily be indicated.

It has become customary for ease of reading to divide large numbers into groups of three digits, separating them by commas.

6,000,000 rather than 6000000

Our system of notation then is characterized by two things: the use of positions to indicate the value of each symbol, and the use of ten symbols, from which we derive the name *decimal system*.

Retaining the same use of positions, we might have used a different number of symbols, and displacing a symbol one place to the left might multiply its value by any other factor such as 2, 6 or 12. Such other systems have been in use in history, but will not be discussed here. There are also systems in which displacing a symbol to the left multiplies its value by

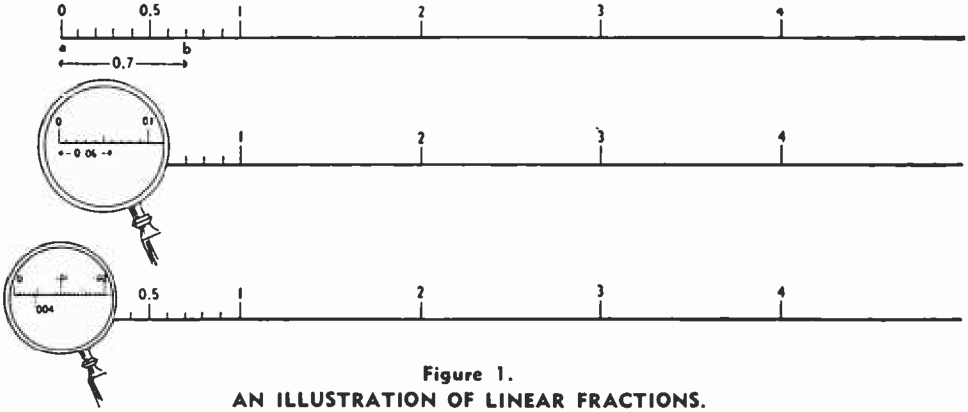


Figure 1.
AN ILLUSTRATION OF LINEAR FRACTIONS.

varying factors in accordance with complicated rules. The English system of measurements is such an inconsistent and inferior system.

Decimal Fractions Since we can extend a number indefinitely to the left to make it bigger, it is a logical step to extend it towards the right to make it smaller. Numbers smaller than unity are *fractions* and if a displacement one position to the right divides its value by ten, then the number is referred to as a *decimal fraction*. Thus a digit to the right of the units column indicates the number of *tenths*, the second digit to the right represents the number of *hundredths*, the third, the number of *thousandths*, etc. Some distinguishing mark must be used to divide unit from tenths so that one may properly evaluate each symbol. This mark is the *decimal point*.

A decimal fraction like *four-tenths* may be written .4 or 0.4 as desired, the latter probably being the clearer. Every time a digit is placed one space further to the right it represents a ten times smaller part. This is illustrated in Figure 1, where each large division represents a unit; each unit may be divided into ten parts although in the drawing we have only so divided the first part. The length *ab* is equal to seven of these tenth parts and is written as 0.7.

The next smaller divisions, which should be written in the second column to the right of the decimal point, are each one-tenth of the small division, or one one-hundredth each. They are so small that we can only show them by imagining a magnifying glass to look at them, as in Figure 1. Six of these divisions is to be written as 0.06 (six hundredths). We need a microscope to see the next smaller division, that is those in the third place, which will be a tenth of one one-hundredth, or a thousandth; four such divisions would be written as 0.004 (four thousandths).

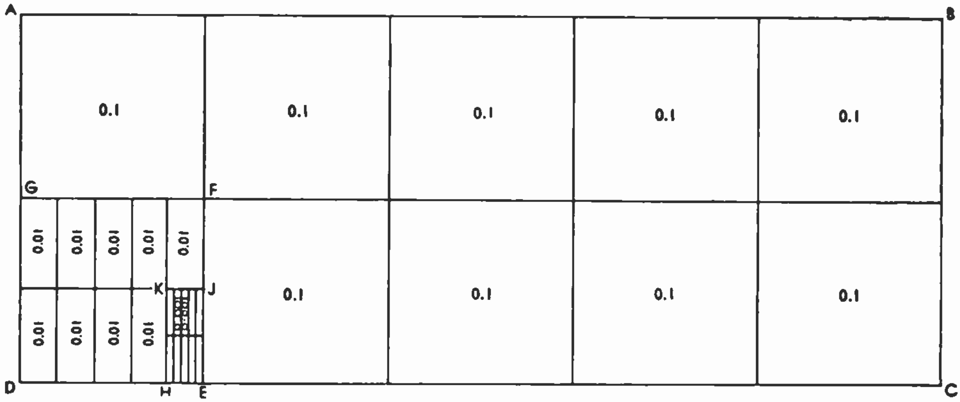


Figure 2.
IN THIS ILLUSTRATION FRACTIONAL PORTIONS ARE REPRESENTED IN THE FORM OF RECTANGLES RATHER THAN LINEARLY.
ABCD = 1.0; GFED = 0.1; KJEH = 0.01; each small section within KJEH equals 0.001

It should not be thought that such numbers are merely of academic interest for very small quantities are common in radio work.

Possibly the conception of fractions may be clearer to some students by representing it in the form of rectangles rather than linearly (see Figure 2).

Addition When two or more numbers are to be added we sometimes write them horizontally with the plus sign between them. + is the sign or *operator* indicating addition. Thus if 7 and 12 are to be added together we may write $7+12=19$.

But if larger or more numbers are to be added together they are almost invariably written one under another in such a position that the decimal points fall in a vertical line. If a number has no decimal point, it is still considered as being just to the right of the units figure; such a number is a whole number or *integer*. Examples:

```
  654      0.654      654
   32      3.2         32
53041     53.041    5304.1
-----     ------    ------
53727     56.895    5990.1
```

The result obtained by adding numbers is called the *sum*.

Subtraction Subtraction is the reverse of addition. Its operator is — (the *minus* sign). The number to be subtracted is called the *subtrahend*, the number from which it is subtracted is the *minuend*, and the result is called the *remainder*.

```
 minuend
—subtrahend
-----------
 remainder
```

Examples:

```
 65.4     65.4
—32      —32.21
-----    ------
 33.4     33.19
```

Multiplication When numbers are to be multiplied together we use the ×, which is known as the *multiplication* or the *times* sign. The number to be multiplied is known as the *multiplicand* and that by which it is to be multiplied is the *multiplier*, which may be written in words as follows:

```
  multiplicand
  × multiplier
-----------------
 partial product
partial product
-----------------
p r o d u c t
```

The result of the operation is called the *product*.

From the examples to follow it will be obvious that there are as many partial products as there are digits in the multiplier. In the following examples note that the righthandmost digit of each partial product is placed one space farther to the left than the previous one.

```
  834        834
× 26       × 206
-----      ------
 5004       5004
1668        000
-----     1668
21684     ------
          171804
```

In the second example above it will be seen that the inclusion of the second partial product was unnecessary; whenever the multiplier contains a cipher (zero) the next partial product should be moved an *additional* space to the left.

Numbers containing decimal fractions may first be multiplied exactly as if the decimal point did not occur in the numbers at all; the position of the decimal point in the product is determined after all operations have been completed. It must be so positioned in the product that the number of digits to its right is equal to the number of decimal places in the multiplicand plus the number of decimal places in the multiplier.

This rule should be well understood since many radio calculations contain quantities which involve very small decimal fractions. In the examples which follow the explanatory notations "2 places," etc., are not actually written down since it is comparatively easy to determine the decimal point's proper location mentally.

```
  5.43    2 places
 ×0.72    2 places
 ------
  1086
 3 801
 ------
 3.9096   2+2=4 places

   0.04   2 places
 ×0.003   3 places
 -------
0.00012   2+3=5 places
```

Division Division is the reverse of multiplication. Its operator is the ÷, which is called the *division sign*. It is also common to indicate division by the use of the fraction bar (/) or by writing one number over the other. The number which is to be divided is called the *dividend* and is written before the division sign or fraction bar or over the horizontal line indicating a fraction. The number by which the dividend is to be divided is called the *divisor* and follows the division sign or fraction bar or comes under the horizontal line of the fraction. The answer or result is called the *quotient.*

```
           quotient
  divisor ) dividend
```

or

$$\text{dividend} \div \text{divisor} = \text{quotient}$$

or

$$\frac{\text{dividend}}{\text{divisor}} = \text{quotient}$$

Examples:

```
          126               49
  834 ) 105084       49 ) 2436
        834                196
        ----               ---
        2168               476
        1668               441
        ----               ---
         5004               35 remainder
         5004
         ----
```

Note that one number often fails to divide into another evenly. Hence there is often a quantity left over called the *remainder.*

The rules for placing the decimal point are the reverse of those for multiplication. The number of decimal places in the quotient is equal to the difference between the number of decimal places in the dividend and that in the divisor. It is often simpler and clearer to remove the decimal point entirely from the divisor by multiplying both dividend and divisor by the necessary factor; that is we move the decimal point in the divisor as many places to the right as is necessary to make it a whole number and then we move the decimal point in the dividend exactly the same number of places to the right regardless of whether this makes the dividend a whole number or not. When this has been done the decimal point in the quotient will automatically come directly above that in the dividend as shown in the following example.

Example: Divide 10.5084 by 8.34. Move the decimal point of both dividend and divisor two places to the right.

```
           1.26
  834 ) 1050.84
         834
         ----
         2168
         1668
         ----
          5004
          5004
          ----
```

Another example: Divide 0.000325 by 0.017. Here we must move the decimal point three places to the right in both dividend and divisor.

```
       0.019
  17 ) 0.325
         17
         --
         155
         153
         ---
           2
```

In a case where the dividend has fewer decimals than the divisor the same rules still may be applied by adding ciphers. For example to divide 0.49 by 0.006 we must move the decimal point three places to the right. The 0.49 now becomes 490 and we write:

```
      81
  6 ) 490
      48
      ---
       10
        6
       --
        4
```

When the division shows a remainder it is sometimes necessary to continue the work so as to obtain more figures. In that case ciphers may be annexed to the dividend, brought down to the remainder, and the division continued as long as may be necessary; be sure to place a decimal point in the dividend before the ciphers are annexed if the dividend does not already contain a decimal point. For example:

```
      80.33
  6 ) 482.00
      48
      --
        20
        18
        --
         20
         18
         --
          2
```

This operation is not very often required in radio work since the accuracy of the measurements from which our problems start seldom justifies the use of more than three significant figures. This point will be covered further later in this chapter.

Fractions Quantities of less than one (unity) are called *fractions.* They may be expressed by decimal notation as we have seen, or they may be expressed as *vulgar fractions.* Examples of vulgar fractions:

$$\frac{\textbf{numerator}}{\textbf{denominator}} \quad \frac{3}{4} \quad \frac{6}{7} \quad \frac{1}{5}$$

The upper position of a vulgar fraction is called the *numerator* and the lower position the *denominator*. When the numerator is the smaller of the two, the fraction is called a *proper fraction;* the examples of vulgar fractions given above are proper vulgar fractions. When the numerator is the larger, the expression is an *improper fraction,* which can be reduced to an integer or whole number with a proper fraction, the whole being called a mixed number. In the following examples improper fractions have been reduced to their corresponding mixed numbers.

$$\frac{7}{4} = 1\frac{3}{4} \qquad \frac{5}{3} = 1\frac{2}{3}$$

Adding or Subtracting Fractions Except when the fractions are very simple it will usually be found much easier to add and subtract fractions in the form of decimals. This rule likewise applies for practically all other operations with fractions. However, it is occasionally necessary to perform various operations with vulgar fractions and the rules should be understood.

When adding or subtracting such fractions the denominators must be made equal. This may be done by multiplying both numerator and denominator of the first fraction by the denominator of the other fraction, after which we multiply the numerator and denominator of the second fraction by the denominator of the first fraction. This sounds more complicated than it usually proves in practice, as the following examples will show.

$$\frac{1}{2} + \frac{1}{3} = \left[\frac{1 \times 3}{2 \times 3} + \frac{1 \times 2}{3 \times 2}\right] = \frac{3}{6} + \frac{2}{6} = \frac{5}{6}$$

$$\frac{3}{4} - \frac{2}{5} = \left[\frac{3 \times 5}{4 \times 5} - \frac{2 \times 4}{5 \times 4}\right] = \frac{15}{20} - \frac{8}{20} = \frac{7}{20}$$

Except in problems involving large numbers the step shown in brackets above is usually done in the head and is not written down.

Although in the examples shown above we have used proper fractions, it is obvious that the same procedure applies with improper fractions. In the case of problems involving mixed numbers it is necessary first to convert them into improper fractions. Example:

$$2\frac{3}{7} = \frac{2 \times 7 + 3}{7} = \frac{17}{7}$$

The numerator of the improper fraction is equal to the whole number multiplied by the denominator of the original fraction, to which the numerator is added. That is in the above example we multiply 2 by 7 and then add 3 to obtain 17 for the numerator. The denominator is the same as is the denominator of the original fraction. In the following example we have added two mixed numbers.

$$2\frac{3}{7} + 3\frac{3}{4} = \frac{17}{7} + \frac{15}{4} = \left[\frac{17 \times 4}{7 \times 4} + \frac{15 \times 7}{4 \times 7}\right]$$

$$= \frac{68}{28} + \frac{105}{28} = \frac{173}{28} = 6\frac{5}{28}$$

Multiplying Fractions All vulgar fractions are multiplied by multiplying the numerators together and the denominators together, as shown in the following example:

$$\frac{3}{4} \times \frac{2}{5} = \left[\frac{3 \times 2}{4 \times 5}\right] = \frac{6}{20} = \frac{3}{10}$$

As above, the step indicated in brackets is usually not written down since it may easily be performed mentally. As with addition and subtraction any mixed numbers should be first reduced to improper fractions as shown in the following example:

$$\frac{3}{23} \times 4\frac{1}{3} = \frac{3}{23} \times \frac{13}{3} = \frac{39}{69} = \frac{13}{23}$$

Division of Fractions Fractions may be most easily divided by inverting the divisor and then multiplying. Example:

$$\frac{2}{5} \div \frac{3}{4} = \frac{2}{5} \times \frac{4}{3} = \frac{8}{15}$$

In the above example it will be seen that to divide by ¾ is exactly the same thing as to multiply by 4/3. Actual division of fractions is a rather rare operation and if necessary is usually postponed until the final answer is secured when it is often desired to reduce the resulting vulgar fraction to a decimal fraction by division. It is more common and usually results in least overall work to reduce vulgar fractions to decimals at the beginning of a problem. Examples:

$$\frac{3}{8} = 0.375 \qquad \frac{5}{32} = 0.15625$$

```
      0.15625
32 ) 5.00000
     3 2
     1 80
     1 60
       200
       192
         80
         64
         160
         160
```

It will be obvious that many vulgar fractions cannot be reduced to exact decimal equivalents. This fact need not worry us, however, since the degree of equivalence can always be as much as the data warrants. For instance, if we know that one-third of an ampere is flowing in a given circuit, this can be written as 0.333 amperes. This is not the exact equivalent of 1/3 but is close enough since it shows the value to the nearest thousandth of an ampere and it is probable that the meter from which we secured our original data was not accurate to the nearest thousandth of an ampere.

Thus in converting vulgar fractions to a decimal we unhesitatingly stop when we have reached the number of significant figures warranted by our original data, which is very seldom more than three places (see section *Significant Figures* later in this chapter).

When the denominator of a vulgar fraction contains only the factors 2 or 5, division can be brought to a finish and there will be no remainder, as shown in the examples above.

When the denominator has other factors such as 3, 7, 11, etc., the division will seldom come out even no matter how long it is continued but, as previously stated, this is of no consequence in practical work since it may be carried to whatever degree of accuracy is necessary. The digits in the quotient will usually repeat either singly or in groups, although there may first occur one or more digits which do not repeat. Such fractions are known as *repeating fractions*. They are sometimes indicated by an oblique line (fraction bar) through the digit which repeats, or through the first and last digits of a repeating group. Example:

$$\frac{1}{3} = 0.3333\ldots = 0.\cancel{3}$$

$$\frac{1}{7} = 0.142857142857\ldots = 0.\cancel{1}4285\cancel{7}$$

The foregoing examples contained only repeating digits. In the following example a non-repeating digit precedes the repeating digit:

$$\frac{7}{30} = 0.2333\ldots = 0.2\cancel{3}$$

While repeating decimal fractions can be converted into their vulgar fraction equivalents, this is seldom necessary in practical work and the rules will be omitted here.

Powers and Roots When a number is to be multiplied by itself we say that it is to be *squared* or to be *raised to the second power*. When it is to be multipled by itself once again, we say that it is *cubed* or *raised to the third power*.

In general terms, when a number is to be multipled by itself we speak of *raising to a power* or *involution;* the number of times which the number is to be multiplied by itself is called the *order of the power*. The standard notation requires that the order of the power be indicated by a small number written after the number and above the line, called the *exponent*. Examples:

$2^2 = 2 \times 2$, or 2 **squared**, or the **second power of 2**

$2^3 = 2 \times 2 \times 2$, or 2 **cubed**, or the **third power of 2**

$2^4 = 2 \times 2 \times 2 \times 2$, or the **fourth power of 2**

Sometimes it is necessary to perform the reverse of this operation, that is, it may be necessary, for instance, to find that number which multiplied by itself will give a product of nine. The answer is of course 3. This process is known as *extracting the root* or *evolution*. The particular example which is cited would be written:

$$\sqrt{9} = 3$$

The sign for extracting the root is $\sqrt{\ }$, which is known as the *radical sign;* the order of the root is indicated by a small number above the radical as in $\sqrt[4]{\ }$, which would mean the fourth root; this number is called the *index*. When the radical bears no index, the square or second root is intended.

Restricting our attention for the moment to square root, we know that 2 is the square root of 4, and 3 is the square root of 9. If we want the square root of a number between 3 and 9, such as the square root of 5, it is obvious that it must lie between 2 and 3. In general the square root of such a number cannot be *exactly* expressed either by a vulgar fraction or a decimal fraction. However, the square root can be carried out decimally as far as may be necessary for sufficient accuracy. In general such a decimal fraction will contain a never-ending series of digits without repeating groups. Such a number is an *irrational number*, such as

$$\sqrt{5} = 2.2361\ldots$$

The extraction of roots is usually done by tables or logarithms the use of which will be described later. There are longhand methods of extracting various roots, but we shall give only that for extracting the square root since the others become so tedious as to make other methods almost invariably preferable. Even the longhand method for extracting the square root will usually be used only if logarithm tables, slide rule, or table of roots are not handy.

Extracting the Square Root First divide the number the root of which is to be extracted into groups of two digits starting at the decimal point and going in both directions. If the lefthandmost group proves to have only one digit instead of two, no harm will be done. The righthandmost group may be made to have two digits by annexing a zero if necessary. For example, let it be required to find the square root of 5678.91. This is to be divided off as follows:

$$\sqrt{56'\,78.91}$$

The mark used to divide the groups may be anything convenient, although the prime-sign (′) is most commonly used for the purpose.

Next find the largest square which is contained in the first group, in this case 56. The largest square is obviously 49, the square of 7. Place the 7 above the first group of the number whose root is to be extracted, which is sometimes called the *dividend* from analogy to ordinary division. Place the square of this figure, that is 49, under the first group, 56, and subtract leaving a remainder of 7.

```
  7
√56' 78.91
 49
 --
  7
```

Bring down the next group and annex it to the remainder so that we have 778. Now to the left of this quantity write down twice the root so far found (2 × 7 or 14 in this example), annex a cipher as a trial divisor, and see how many times the result is contained in 778. In our example 140 will go into 778 5 times. Replace the cipher with a 5, and multiply the resulting 145 by 5 to give 725. Place the 5 directly above the second group in the dividend and then subtract the 725 from 778.

```
               7  5
              √56' 78.91
               49
               --
140             7 78
145 × 5 =       7 25
                ----
                  53
```

The next step is an exact repetition of the previous step. Bring down the third group and annex it to the remainder of 53, giving 5391. Write down twice the root already found and annex the cipher (2 x 75 or 150 plus the cipher, which will give 1500). 1500 will go into 5391 3 times. Replace the last cipher with a three and multiply 1503 by 3 to give 4509. Place 3 above the third group. Subtract to find the remainder of 882. The quotient 75.3 which has been found so far is not the exact square root which was desired; in most cases it will be sufficiently accurate. However, if greater accuracy is desired groups of two ciphers can be brought down and the process carried on as long as necessary.

```
                7  5. 3
              √56' 78.91
               49
               --
 140            7 78
 145 × 5 =      7 25
                ----
1500              53 91
1503 × 3 =        45 09
                  -----
                   8 82
```

Each digit of the root should be placed directly above the group of the dividend from which it was derived; if this is done the decimal point of the root will come directly above the decimal point of the dividend.

Sometimes the remainder after a square has been subtracted (such as the 1 in the following example) will not be sufficiently large to contain twice the root already found even after the next group of figures has been brought down. In this case we write a cipher above the group just brought down and bring down another group.

```
                 7. 0  8  2
               √50.16' 00' 00
                49
                --
 1400            1 16 00
 1408 × 8 =      1 12 64
                 -------
14160               3 36 00
14162 × 2 =         2 83 24
                    -------
                      52 76
```

In the above example the amount 116 was not sufficient to contain twice the root already found with a cipher annexed to it; that is, it was not sufficient to contain 140. Therefore we write a zero above 16 and bring down the next group, which in this example is a pair of ciphers.

Order of Operations One frequently encounters problems in which several of the fundamental operations of arithmetic which have been described are to be performed. The order in which these operations must be performed is important. First all powers and roots should be calculated; multiplication and division come next; adding and subtraction come last. In the example

$$2 + 3 \times 4^2$$

we must first square the 4 to get 16; then we multiply 16 by 3, making 48, and to the product we add 2, giving a result of 50.

If a different order of operations were followed, a different result would be obtained. For instance, if we add 2 to 3 we would obtain 5, and then multiplying this by the square of 4 or 16, we would obtain a result of 80, which is incorrect.

In more complicated forms such as fractions whose numerators and denominators may both be in complicated forms, the numerator and denominator are first found separately before the division is made, such as in the following example:

$$\frac{3 \times 4 + 5 \times 2}{2 \times 3 + 2 + 3} = \frac{12 + 10}{6 + 2 + 3} = \frac{22}{11} = 2$$

Problems of this type are very common in dealing with circuits containing several inductances, capacities, or resistances.

The order of operations specified above does not always meet all possible conditions; if a series of operations should be performed in a different order, this is always indicated by *parentheses* or *brackets*, for example:

$$2 + 3 \times 4^2 = 2 + 3 \times 16 = 2 + 48 = 50$$

$$(2 + 3) \times 4^2 = 5 \times 4^2 = 5 \times 16 = 80$$

$$2 + (3 \times 4)^2 = 2 + 12^2 = 2 + 144 = 146$$

In connection with the radical sign, brackets may be used or the "hat" of the radical may be extended over the entire quantity whose root is to be extracted. Example:

$$\sqrt{}4 + 5 = \sqrt{4} + 5 = 2 + 5 = 7$$

$$\sqrt{}(4 + 5) = \sqrt{4 + 5} = \sqrt{9} = 3$$

It is recommended that the radical always be extended over the quantity whose root is to be extracted to avoid any ambiguity.

Cancellation In a fraction in which the numerator and denominator consist of several factors to be multiplied, considerable labor can often be saved if it is found that the same factor occurs in both numerator and denominator. These factors cancel each other and can be removed. Example:

$$\frac{\cancel{2} \times \cancel{3} \times \overset{5}{\cancel{25}}}{\underset{\cancel{2}}{\cancel{6}} \times \cancel{5} \times 7} = \frac{5}{7}$$

In the foregoing example it is obvious that the 3 in the numerator goes into the 6 in the denominator twice. We may thus cross out the three and replace the 6 by a 2. The 2 which we have just placed in the denominator cancels the 2 in the numerator. Next the 5 in the denominator will go into the 25 in the numerator leaving a result of 5. Now we have left only a 5 in the numerator and a 7 in the denominator, so our final result is 5/7. If we had multiplied $2 \times 3 \times 25$ to obtain 150 and then had divided this by $6 \times 5 \times 7$ or 210, we would have obtained the same result but, with considerably more work.

Algebra

Algebra is not a separate branch of mathematics but is merely a form of *generalized arithmetic* in which letters of the alphabet and occasional other symbols are substituted for numbers, from which it is often referred to as *literal notation.* It is simply a shorthand method of writing operations which could be spelled out.

The laws of most common electrical phenomena and circuits (including of course radio phenomena and circuits) lend themselves particularly well to representation by literal notation and solution by algebraic equations or formulas.

While we may write a particular problem in Ohm's Law as an ordinary division or multiplication, the general statement of all such problems calls for the replacement of the numbers by symbols. We might be explicit and write out the names of the units and use these names as symbols:

$$\text{volts} = \text{amperes} \times \text{ohms}$$

Such a procedure becomes too clumsy when the expression is more involved and would be unusually cumbersome if any operations like multiplication were required. Therefore as a short way of writing these generalized relations the numbers are represented by letters. Ohm's Law then becomes

$$E = I \times R$$

In the statement of any particular problem the significance of the letters is usually indicated directly below the equation or formula using them unless there can be no ambiguity. Thus the above form of Ohm's Law would be more completely written as:

$$E = I \times R$$

where E = e.m.f. in volts
I = current in amperes
R = resistance in ohms

Letters therefore represent numbers, and for any letter we can read "any number." When the same letter occurs again in the same expression we would mentally read "the same number," and for another letter "another number of any value."

These letters are connected by the usual operational symbols of arithmetic, +, −, ×, ÷, and so forth. In algebra, the sign for division is seldom used, a division being usually written as a fraction. The multiplication sign, ×, is usually omitted or one may write a period only. Examples:

$$2 \times a \times b = 2ab$$
$$2.3.4.5a = 2 \times 3 \times 4 \times 5 \times a$$

In practical applications of algebra, an expression usually states some physical law and each letter represents a variable quantity which is therefore called a *variable*. A fixed number in front of such a quantity (by which it is to be multiplied) is known as the *coefficient*. Sometimes the coefficient may be unknown, yet to be determined; it is then also written as a letter; *k* is most commonly used for this purpose.

The Negative Sign In ordinary arithmetic we seldom work with negative numbers, although we may be "short" in a subtraction. In algebra, however, a number may be either negative or positive. Such a thing may seem *academic* but a negative quantity can have a real existence. We need only refer to a *debt* being considered a negative possession. In electrical work, however, a result of a problem might be a negative number of amperes or volts, indicating that the direction of the current is opposite to the direction chosen as positive. We shall have illustrations of this shortly.

Having established the existence of negative quantities, we must now learn how to work with these negative quantities in addition, subtraction, multiplication and so forth.

In addition, a negative number added to a positive number is the same as subtracting a positive number from it.

$$\begin{array}{r} 7 \\ -3 \\ \hline 4 \end{array} \text{ (add) is the same as } \begin{array}{r} 7 \\ 3 \\ \hline 4 \end{array} \text{ (subtract)}$$

or we might write it

$$7 + (-3) = 7 - 3 = 4$$

Similarly, we have:

$$a + (-b) = a - b$$

When a minus sign is in front of an expression in brackets, this minus sign has the effect of reversing the signs of every term within the brackets:

$$-(a - b) = -a + b$$
$$-(2a + 3b - 5c) = -2a - 3b + 5c$$

Multiplication. When both the multiplicand and the multiplier are negative, the product is positive. When only one (either one) is negative the product is negative. The four possible cases are illustrated below:

$$+ \times + = + \qquad + \times - = -$$
$$- \times + = - \qquad - \times - = +$$

Division. Since division is but the reverse of multiplication, similar rules apply for the sign of the quotient. When both the dividend and the divisor have the same sign (both negative or both positive) the quotient is positive. If they have unlike signs (one positive and one negative) the quotient is negative.

$$\frac{+}{+} = + \qquad \frac{+}{-} = -$$
$$\frac{-}{+} = - \qquad \frac{-}{-} = +$$

Powers. Even powers of negative numbers are positive and *odd* powers are negative. Powers of positive numbers are always positive. Examples:

$$-2^2 = -2 \times -2 = +4$$
$$-2^3 = -2 \times -2 \times -2 = +4 \times -2 = -8$$

Roots. Since the square of a negative number is positive and the square of a positive number is also positive, it follows that a positive number has two square roots. The square root of 4 can be either +2 or −2 for $(+2) \times (+2) = +4$ and $(-2) \times (-2) = +4$.

Addition and Subtraction *Polynomials* are quantities like $3ab^2 + 4ab^3 - 7a^2b^4$ which have several terms of different *names*. When adding polynomials, only terms of the same name can be taken together.

$$\begin{array}{r} 7a^3 + 8\,ab^2 + 3\,a^2b \phantom{{}- b^3} + 3 \\ a^3 - 5\,ab^2 \phantom{{}+ 3\,a^2b} - b^3 \phantom{{}+ 3} \\ \hline 8a^3 + 3\,ab^2 + 3\,a^2b - b^3 + 3 \end{array}$$

Collecting terms. When an expression contains more than one term of the same name, these can be added together and the expression made simpler:

$$5x^2 + 2xy + 3xy^2 - 3x^2 + 7xy =$$

$$5x^2 - 3x^2 + 2xy + 7xy + 3xy^2 =$$

$$2x^2 + 9xy + 3xy^2$$

Multiplication Multiplication of single terms is indicated simply by writing them together.

$$a \times b \text{ is written as } ab$$

$$a \times b^2 \text{ is written as } ab^2$$

Bracketed quantities are multiplied by a single term by multiplying each term:

$$a(b + c + d) = ab + ac + ad$$

When two bracketed quantities are multiplied, each term of the first bracketed quantity is to be multiplied by each term of the second bracketed quantity, thereby making every possible combination.

$$(a + b)(c + d) = ac + ad + bc + bd$$

In this work particular care must be taken to get the signs correct. Examples:

$$(a + b)(a - b) = a^2 + ab - ab - b^2 = a^2 - b^2$$

$$(a + b)(a + b) = a^2 + ab + ab + b^2 = a^2 + 2ab + b^2$$

$$(a - b)(a - b) = a^2 - ab - ab + b^2 = a^2 - 2ab + b^2$$

Division It is possible to do longhand division in algebra, although it is somewhat more complicated than in arithmetic. However, the division will seldom come out even, and is not often done in this form. The method is as follows: Write the terms of the dividend in the order of descending powers of one variable and do likewise with the divisor. Example:

Divide $5a^2b + 21b^3 + 2a^3 - 26ab^2$ by $2a - 3b$

Write the dividend in the order of descending powers of a and divide in the same way as in arithmetic.

$$\begin{array}{r|llll}
 & a^2 & + 4ab & - 7b^2 & \\
2a - 3b & 2a^3 & + 5a^2b & - 26ab^2 & + 21b^3 \\
 & 2a^3 & - 3a^2b & & \\ \hline
 & & + 8a^2b & - 26ab^2 & \\
 & & + 8a^2b & - 12ab^2 & \\ \hline
 & & & - 14ab^2 & + 21b^3 \\
 & & & - 14ab^2 & + 21b^3 \\ \hline
\end{array}$$

Another example: Divide $x^3 - y^3$ by $x - y$

$$\begin{array}{r|llll}
x - y & x^3 & + 0 & + 0 & - y^3 \;(\, x^2 + xy + y^2 \\
 & x^3 & - x^2y & & \\ \hline
 & & + x^2y & & \\
 & & x^2y & - xy^2 & \\ \hline
 & & & + xy^2 & - y^3 \\
 & & & xy^2 & - y^3 \\ \hline
\end{array}$$

Factoring Very often it is necessary to simplify expressions by finding a factor. This is done by collecting two or more terms having the same factor and bringing the factor outside the brackets:

$$6ab + 3ac = 3a(2b + c)$$

In a four term expression one can take together two terms at a time; the intention is to try getting the terms within the brackets the same after the factor has been removed:

$$\begin{gathered}
30ac - 18bc + 10ad - 6bd = \\
6c(5a - 3b) + 2d(5a - 3b) = \\
(5a - 3b)(6c + 2d)
\end{gathered}$$

Of course, this is not always possible and the expression may not have any factors. A similar process can of course be followed when the expression has six or eight or any even number of terms.

A special case is a three-term polynomial, which can sometimes be factored by writing the middle term as the sum of two terms:

$$\begin{gathered}
x^2 - 7xy + 12y^2 \text{ may be rewritten as} \\
x^2 - 3xy - 4xy + 12y^2 = \\
x(x - 3y) - 4y(x - 3y) = \\
(x - 4y)(x - 3y)
\end{gathered}$$

The middle term should be split into two in such a way that the sum of the two new terms equals the original middle term and that their product equals the product of the two outer terms. In the above example these conditions are fulfilled for $-3xy - 4xy = -7xy$ and $(-3xy)(-4xy) = 12x^2y^2$. It is not always possible to do this and there are then no simple factors.

Working with Powers and Roots When two powers of the same number are to be multiplied, the exponents are *added.*

$$a^2 \times a^3 = aa \times aaa = aaaaa = a^5 \text{ or}$$
$$a^2 \times a^3 = a^{(2+3)} = a^5$$

$$b^3 \times b = b^4$$

$$c^5 \times c^7 = c^{12}$$

Similarly, dividing of powers is done by subtracting the exponents.

$$\frac{a^3}{a^2} = \frac{aaa}{aa} = a \text{ or } \frac{a^3}{a^2} = a^{(3-2)} = a^1 = a$$

$$\frac{b^5}{b^3} = \frac{bbbbb}{bbb} = b^2 \text{ or } \frac{b^5}{b^3} = b^{(5-3)} = b^2$$

Now we are logically led into some important new ways of notation. We have seen that when dividing, the exponents are subtracted. This can be continued into negative exponents. In the following series, we successively divide by *a* and since this can now be done in two ways, the two ways of notation must have the same meaning and be identical.

a^5	a^2	$a^{-1} = \frac{1}{a}$
a^4	$a^1 = a$	$a^{-2} = \frac{1}{a^2}$
a^3	$a^0 = 1$	$a^{-3} = \frac{1}{a^3}$

These examples illustrate two rules: (1) any number raised to "zero' power equals one or unity; (2) any quantity raised to a negative power is the inverse or reciprocal of the same quantity raised to the same positive power.

$$n^0 = 1 \qquad a^{-n} = \frac{1}{a^n}$$

Roots. The product of the square root of two quantities equals the square root of their product.

$$\sqrt{a} \times \sqrt{b} = \sqrt{ab}$$

Also, the quotient of two roots is equal to the root of the quotient.

$$\frac{\sqrt{a}}{\sqrt{b}} = \sqrt{\frac{a}{b}}$$

Note, however, that in addition or subtraction the square root of the sum or difference is *not* the same as the sum or difference of the square roots.

$$\text{Thus, } \sqrt{9} - \sqrt{4} = 3 - 2 = 1$$
$$\text{but } \sqrt{9 - 4} = \sqrt{5} = 2.2361$$
$$\text{Likewise } \sqrt{a} + \sqrt{b} \text{ is not the same as } \sqrt{a + b}$$

Roots may be written as fractional powers. Thus $\sqrt{a}$ may be written as $a^{1/2}$ because

$$\sqrt{a} \times \sqrt{a} = a$$
$$\text{and, } a^{1/2} \times a^{1/2} = a^{1/2+1/2} = a^1 = a$$

Any root may be written in this form

$$\sqrt{b} = b^{1/2} \quad \sqrt[3]{b} = b^{1/3} \quad \sqrt[4]{b^3} = b^{3/4}$$

The same notation is also extended in the negative direction:

$$b^{-1/2} = \frac{1}{b^{1/2}} = \frac{1}{\sqrt{b}} \qquad c^{-1/3} = \frac{1}{c^{1/3}} = \frac{1}{\sqrt[3]{c}}$$

Following the previous rules that exponents add when powers are multiplied,

$$\sqrt[3]{a} \times \sqrt[3]{a} = \sqrt[3]{a^2}$$
$$\text{but also } a^{1/3} \times a^{1/3} = a^{2/3}$$
$$\text{therefore } a^{2/3} = \sqrt[3]{a^2}$$

Powers of powers. When a power is again raised to a power, the exponents are multiplied;

$$(a^2)^3 = a^6 \qquad (b^{-1})^3 = b^{-3}$$
$$(a^3)^4 = a^{12} \qquad (b^{-2})^{-4} = b^8$$

This same rule also applies to roots of roots and also powers of roots and roots of powers because a root can always be written as a fractional power.

$$\sqrt[3]{\sqrt{a}} = \sqrt[6]{a} \text{ for } (a^{1/2})^{1/3} = a^{1/6}$$

Removing radicals. A root or radical in the denominator of a fraction makes the expression difficult to handle. If there must be a radical it should be located in the numerator rather than in the denominator. The removal of the radical from the denominator is done by multiplying both numerator and denominator by a quantity which will remove the radical from the denominator, thus *rationalizing* it:

$$\frac{1}{\sqrt{a}} = \frac{\sqrt{a}}{\sqrt{a} \times \sqrt{a}} = \frac{1}{a}\sqrt{a}$$

Suppose we have to rationalize

$\dfrac{3a}{\sqrt{a} + \sqrt{b}}$ In this case we must multiply numerator and denominator by $\sqrt{a} - \sqrt{b}$, the same terms but with the second having the opposite sign, so that their product will not contain a root.

$$\frac{3a}{\sqrt{a} + \sqrt{b}} = \frac{3a(\sqrt{a} - \sqrt{b})}{(\sqrt{a}+\sqrt{b})(\sqrt{a} - \sqrt{b})} = \frac{3a(\sqrt{a} - \sqrt{b})}{a - b}$$

Imaginary Numbers Since the square of a negative number is positive and the square of a positive number is also positive, the square root of a negative number can be neither positive nor negative. Such a number is said to be *imaginary;* the most common such number ($\sqrt{-1}$) is often represented by the letter *i* in mathematical work or *j* in electrical work.

$$\sqrt{-1} = i \text{ or } j \text{ and } i^2 \text{ or } j^2 = -1$$

Imaginary numbers do not exactly correspond to anything in our experience and it is best not to try to visualize them. Despite this fact, their interest is much more than academic, for they are extremely useful in many calculations involving alternating currents.

The square root of any other negative number may be reduced to a product of two roots, one positive and one negative. For instance:

$$\sqrt{-57} = \sqrt{-1}\ \sqrt{57} = i\sqrt{57}$$

or, in general

$$\sqrt{-a} = i\sqrt{a}$$

Since $i = \sqrt{-1}$, the powers of *i* have the following values:

$$i^2 = -1$$

$$i^3 = -1 \times i = -i$$

$$i^4 = +1$$

$$i^5 = +1 \times i = i$$

Imaginary numbers are different from either positive or negative numbers; so in addition or subtraction they must always be accounted for separately. Numbers which consist of both real and imaginary parts are called *complex* numbers. Examples of complex numbers:

$$3 + 4i = 3 + 4\sqrt{-1}$$
$$a + bi = a + b\sqrt{-1}$$

Since an imaginary number can never be equal to a real number, it follows that in an equality like

$$a + bi = c + di$$

a must equal **c** and **bi** must equal **di**

Complex numbers are handled in algebra just like any other expression, considering *i* as a known quantity. Whenever powers of *i* occur, they can be replaced by the equivalents given above. This idea of having in one equation two separate sets of quantities which must be accounted for separately, has found a symbolic application in vector notation. These are covered later in this chapter.

Equations of the First Degree Algebraic expressions usually come in the form of equations, that is, one set of terms equals another set of terms. The simplest example of this is Ohm's Law:

$$E = IR$$

One of the three quantities may be unknown but if the other two are known, the third can be found readily by substituting the known values in the equation. This is very easy if it is *E* in the above example that is to be found; but suppose we wish to find *I* while *E* and *R* are given. We must then rearrange the equation so that *I* comes to stand alone to the left of the equality sign. This is known as *solving the equation for I.*

Solution of the equation in this case is done simply by transposing. If two things are equal then they must still be equal if both are multiplied or divided by the same number. Dividing both sides of the equation by R:

$$\frac{E}{R} = \frac{IR}{R} = I \text{ or } I = \frac{E}{R}$$

If it were required to solve the equation for *R*, we should divide both sides of the equation by *I*.

$$\frac{E}{I} = R \text{ or } R = \frac{E}{I}$$

A little more complicated example is the equation for the reactance of a condenser:

$$X = \frac{1}{2\pi f C}$$

To solve this equation for *C*, we may multiply both sides of the equation by *C* and divide both sides by *X*

$$X \times \frac{C}{X} = \frac{1}{2\pi f C} \times \frac{C}{X}, \text{ or}$$
$$C = \frac{1}{2\pi f X}$$

This equation is one of those which requires a good knowledge of the placing of the decimal point when solving. Therefore we give a few examples: What is the reactance of a 25 $\mu\mu$fd. capacitor at 1000 kc.? In filling in the given values in the equation we must remember that the units used are farads, cycles, and ohms. Hence, we must write 25 $\mu\mu$fd. as 25 millionths of a millionth of a farad or 25×10^{-12} farad; similarly, 1000 kc. must be converted to 1,000,000 cycles. Substituting these values in the original equation, we have

$$X = \frac{1}{2 \times 3.14 \times 1{,}000{,}000 \times 25 \times 10^{-12}}$$

$$X = \frac{1}{6.28 \times 10^{6} \times 25 \times 10^{-12}} = \frac{10^{6}}{6.28 \times 25}$$
$$= 6360 \text{ ohms}$$

A bias resistor of 1000 ohms should be by-passed, so that at the lowest frequency the reactance of the condenser is 1/10th of that of the resistor. Assume the lowest frequency to be 50 cycles, then the required capacity should have a reactance of 100 ohms, at 50 cycles:

$$C = \frac{1}{2 \times 3.14 \times 50 \times 100} \text{ farads}$$

$$C = \frac{10^{6}}{6.28 \times 5000} \text{ microfarads}$$

$$C = 32\ \mu\text{fd.}$$

In the third possible case, it may be that the frequency is the unknown. This happens for instance in some tone control problems. Suppose it is required to find the frequency which makes the reactance of a 0.03 μfd. condenser equal to 100,000 ohms.

First we must solve the equation for f. This is done by transposition.

$$X = \frac{1}{2\pi f C} \qquad f = \frac{1}{2\pi C X}$$

Substituting known values

$$f = \frac{1}{2 \times 3.14 \times 0.03 \times 10^{-6} \times 100{,}000} \text{ cycles}$$

$$f = \frac{1}{0.01884} \text{ cycles} = 53 \text{ cycles}$$

These equations are known as first degree equations with one unknown. First degree, because the unknown occurs only as a first power. Such an equation always has one possible solution or *root* if all the other values are known.

If there are two unknowns, a single equation will not suffice, for there are then an infinite number of possible solutions. In the case of two unknowns we need *two independent* simultaneous equations. An example of this is:

$$3x + 5y = 7 \qquad 4x - 10y = 3$$

Required, to find x and y.

This type of work is done either by the *substitution method* or by the *elimination method.* In the substitution method we might write for the first equation:

$$3x = 7 - 5y \therefore x = \frac{7 - 5y}{3}$$

(The symbol $\therefore$ means *therefore* or *hence).*

This value of x can then be substituted for x in the second equation making it a single equation with but one unknown, y.

It is, however, simpler in this case to use the elimination method. Multiply both sides of the first equation by two and add it to the second equation:

$$\begin{aligned} 6x + 10y &= 14 \\ 4x - 10y &= 3 \\ \hline 10x \quad &= 17 \quad \text{add} \qquad x = 1.7 \end{aligned}$$

Substituting this value of x in the first equation, we have

$$5.1 + 5y = 7 \therefore 5y = 7 - 5.1 = 1.9 \therefore$$
$$y = 0.38$$

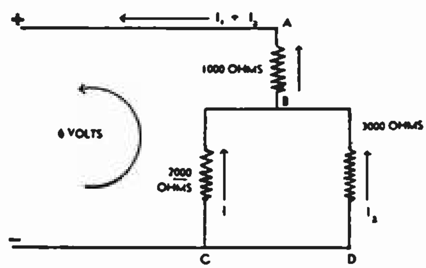


Figure 3.

In this simple network the current divides through the 2000-ohm and 3000-ohm resistors. The current through each may be found by using two simultaneous linear equations. Note that the arrows indicate the direction of electron flow as explained on page 18.

An application of two simultaneous linear equations will now be given. In Figure 3 a simple network is shown consisting of three resistances; let it be required to find the currents I_1 and I_2 in the two branches.

The general way in which all such problems can be solved is to assign directions to the currents through the various resistances. When these are chosen wrong it will do no harm for the result of the equations will then be negative, showing up the error. In this simple illustration there is, of course, no such difficulty.

Next we write the equations for the meshes, in accordance with Kirchhoff's second law. All voltage drops in the direction of the curved arrow are considered positive, the reverse ones negative. Since there are two unknowns we write two equations.

$$1000\ (I_1 + I_2) + 2000\ I_1 = 6$$

$$-2000\ I_1 + 3000\ I_2 = 0$$

Expand the first equation

$$3000\ I_1 + 1000\ I_2 = 6$$

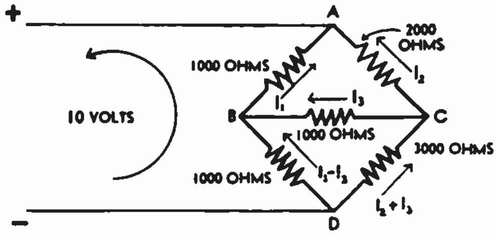


Figure 4.

A MORE COMPLICATED PROBLEM REQUIRING THE SOLUTION OF CURRENTS IN A NETWORK.

This problem is similar to that in Figure 3 but requires the use of three simultaneous linear equations.

Multiply this equation by 3

$$9000\ I_1 + 3000\ I_2 = 18$$

Subtracting the second equation from the first

$$11000\ I_1 = 18$$
$$I_1 = 18/11000 = 0.00164 \text{ amp.}$$

Filling in this value in the second equation

$$3000\ I_2 = 3.28 \qquad I_2 = 0.00109 \text{ amp.}$$

A similar problem but requiring three equations is shown in Figure 4. This consists of an unbalanced bridge and the problem is to find the current in the bridge-branch, I_3. We again assign directions to the different currents, guessing at the one marked I_3. The voltages around closed loops ABC [eq. (1)] and BDC [eq. (2)] equal zero and are assumed to be positive in a counterclockwise direction; that from D to A equals 10 volts [eq. (3)].

(1)
$$-1000\ I_1 + 2000\ I_2 - 1000\ I_3 = 0$$

(2)
$$-1000\ (I_1 - I_3) + 1000\ I_3 + 3000\ (I_2 + I_3) = 0$$

(3)
$$1000\ I_1 + 1000\ (I_1 - I_3) - 10 = 0$$

Expand equations (2) and (3)

(2)
$$-1000\ I_1 + 3000\ I_2 + 5000\ I_3 = 0$$

(3)
$$2000\ I_1 - 1000\ I_3 - 10 = 0$$

Subtract equation (2) from equation (1)

(a)
$$-1000\ I_2 - 6000\ I_3 = 0$$

Multiply the second equation by 2 and add it to the third equation

(b)
$$6000\ I_2 + 9000\ I_3 - 10 = 0$$

Now we have but two equations with two unknowns.

Multiplying equation (a) by 6 and adding to equation (b) we have

$$-27000\ I_3 - 10 = 0$$
$$I_3 = -10/27000 = -0.00037 \text{ amp.}$$

Note that now the solution is negative which means that we have drawn the arrow for I_3 in Figure 4 in the wrong direction. The current is 0.37 ma. in the other direction.

Second Degree or Quadratic Equations A somewhat similar problem in radio would be, if power in watts and resistance in ohms of a circuit are given, to find the voltage and the current. Example: When lighted to normal brilliancy, a 100 watt lamp has a resistance of 49 ohms; for what line voltage was the lamp designed and what current would it take.

Here we have to use the simultaneous equations:

$$P = EI \quad \text{and} \quad E = IR$$

Filling in the known values:

$$P = EI = 100 \text{ and } E = IR = I \times 49$$

Substitute the second equation into the first equation

$$P = EI = (I) \times I \times 49 = 49\ I^2 = 100$$

$$\therefore\ I = \sqrt{\frac{100}{49}} = \frac{10}{7} = 1.43 \text{ amp.}$$

Substituting the found value of 1.43 amp. for I in the first equation, we obtain the value of the line voltage, 70 volts.

Note that this is a *second degree* equation for we finally had the second power of I. Also, since the current in this problem could only be positive, the negative square root of 100/49 or $-10/7$ was not used. Strictly speaking, however, there are two more values that satisfy both equations, these are -1.43 and -70.

In general, a second degree equation in one unknown has two roots, a third degree equation three roots, etc.

The Quadratic Equation Quadratic or second degree equations with but one unknown can be reduced to the general form

$$ax^2 + bx + c = 0$$

where x is the unknown and a, b, and c are constants.

This type of equation can sometimes be solved by the method of factoring a three-term expression as follows:

$$2x^2 + 7x + 6 = 0$$

$$2x^2 + 4x + 3x + 6 = 0$$

factoring:

$$2x\,(x + 2) + 3\,(x + 2) = 0$$

$$(2x + 3)\,(x + 2) = 0$$

There are two possibilities when a product is zero. Either the one or the other factor equals zero. Therefore there are two solutions.

$$2x_1 + 3 = 0 \qquad x_2 + 2 = 0$$

$$2x_1 = -3 \qquad x_2 = -2$$

$$x_1 = -1\tfrac{1}{2}$$

Since factoring is not always easy, the following general solution can usually be employed; in this equation a, b, and c are the coefficients referred to above.

$$X = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

Applying this method of solution to the previous example:

$$X = \frac{-7 \pm \sqrt{49 - 8 \times 6}}{4} = \frac{-7 \pm \sqrt{1}}{4} = \frac{-7 \pm 1}{4}$$

$$X_1 = \frac{-7 + 1}{4} = -1\tfrac{1}{2}$$

$$X_2 = \frac{-7 - 1}{4} = -2$$

A practical example involving quadratics is the law of impedance in a.c. circuits. However, this is a simple kind of quadratic equation which can be solved readily without the use of the special formula given above.

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

This equation can always be solved for R, by squaring both sides of the equation. It should now be understood that squaring both sides of an equation as well as multiplying both sides with a term containing the unknown *may add a new root.* Since we know here that Z and R are positive, when we square the expression there is no ambiguity.

$$Z^2 = R^2 + (X_L - X_C)^2$$

$$\text{and } R^2 = Z^2 - (X_L - X_C)^2$$

$$\text{or } R = \sqrt{Z^2 - (X_L - X_C)^2}$$

$$\text{Also: } (X_L - X_C)^2 = Z^2 - R^2$$

$$\text{and } \pm\,(X_L - X_C) = \sqrt{Z^2 - R^2}$$

But here we do not know the sign of the solution unless there are other facts which indicate it. To find either X_L or X_C alone it would have to be known whether the one or the other is the larger.

Logarithms

Definition and Use A logarithm is the power (or exponent) to which we must raise one number to obtain another. Although the large numbers used in logarithmic work may make them seem difficult or complicated, in reality the principal use of logarithms is to *simplify* calculations which would otherwise be extremely laborious.

We have seen so far that every operation in arithmetic can be reversed. If we have the addition:

$$a + b = c$$

we can reverse this operation in two ways. It may be that b is the unknown, and then we reverse the equation so that it becomes

$$c - a = b$$

It is also possible that we wish to know a, and that b and c are given. The equation then becomes

$$c - b = a$$

We call both of these reversed operations *subtraction,* and we make no distinction between the two possible reverses.

Multiplication can also be reversed in two manners. In the multiplication

$$ab = c$$

we may wish to know a, when b and c are given, or we may wish to know b when a and c are given. In both cases we speak of *division,* and we make again no distinction between the two.

In the case of powers we can also reverse the operation in two manners, but now they are not equivalent. Suppose we have the equation

$$a^b = c$$

If a is the unknown, and b and c are given, we may reverse the operation by writing

$$\sqrt[b]{c} = a$$

This operation we call *taking the root.* But there is a third possibility: that a and c are given, and that we wish to know b. In other words, the question is "to which power must we raise a so as to obtain c?". This operation is known as *taking the logarithm*, and b is the logarithm of c to the base a. We write this operation as follows:

$$\log_a c = b$$

Consider a numerical example. We know $2^3=8$. We can reverse this operation by asking "to which power must we raise *2* so as to obtain *8?*" Therefore, the logarithm of *8* to the base *2* is *3*, or

$$\log_2 8 = 3$$

Taking any single base, such as *2*, we might write a series of all the powers of the base next to the series of their logarithms:

Number:	2	4	8	16	32	64	128	256	512	1024
Logarithm:	1	2	3	4	5	6	7	8	9	10

We can expand this table by finding terms between the terms listed above. For instance, if we let the logarithms increase with ½ each time, successive terms in the upper series would have to be multiplied by the square root of 2. Similarly, if we wish to increase the logarithm by 1/10 at each term, the ratio between two consecutive terms in the upper series would be the tenth root of 2. Now this short list of numbers constitutes a small logarithm table. It should be clear that one could find the logarithm of any number to the base 2. This logarithm will usually be a number with many decimals.

Logarithmic Bases The fact that we chose 2 as a base for the illustration is purely arbitrary. Any base could be used, and therefore there are many possible systems of logarithms. In practice we use only two bases: The most frequently used base is 10, and the system using this base is known as the system of *common* logarithms, or Briggs' logarithms. The second system employs as a base an odd number, designated by the letter e; e = 2.71828. . . . This is known as the *natural* logarithmic system, also as the Napierian system, and the hyperbolic system. Although different writers may vary on the subject, the usual notation is simply *log a* for the common logarithm of a, and $\log_e a$ (or sometimes *ln a*) for the natural logarithm of a. We shall use the common logarithmic system in most cases, and therefore we shall examine this system more closely.

Common Logarithms In the system wherein *10* is the base, the logarithm of *10* equals *1*; the logarithm of *100* equals *2*, etc., as shown in the following table:

$$\log 10 = \log 10^1 = 1$$
$$\log 100 = \log 10^2 = 2$$
$$\log 1{,}000 = \log 10^3 = 3$$
$$\log 10{,}000 = \log 10^4 = 4$$
$$\log 100{,}000 = \log 10^5 = 5$$
$$\log 1{,}000{,}000 = \log 10^6 = 6$$

This table can be extended for numbers less than 10 when we remember the rules of powers discussed under the subject of algebra. Numbers less than unity, too, can be written as powers of ten.

$$\log 1 = \log 10^0 = 0$$
$$\log 0.1 = \log 10^{-1} = -1$$
$$\log 0.01 = \log 10^{-2} = -2$$
$$\log 0.001 = \log 10^{-3} = -3$$
$$\log 0.0001 = \log 10^{-4} = -4$$

From these examples follow several rules: The logarithm of any number between zero and + 1 is negative; the logarithm of zero is minus infinity; the logarithm of a number greater than + 1 is positive. *Negative numbers have no logarithm.* These rules are true of common logarithms and of logarithms to any base.

The logarithm of a number between the powers of ten is an irrational number, that is, it has a never ending series of decimals. For instance, the logarithm of 20 must be between 1 and 2 because 20 is between 10 and 100; the value of the logarithm of 20 is 1.30103. . . . The part of the logarithm to the left of the decimal point is called the *characteristic*, while the decimals are called the *mantissa*. In the case of 1.30103 . ., the logarithm of 20, the characteristic is 1 and the mantissa is .30103 . .

Properties of Logarithms If the base of our system is ten, then, by definition of a logarithm:

$$10^{\log a} = a$$

or, if the base is raised to the power having an exponent equal to the logarithm of a number, the result is that number.

The logarithm of a *product* is equal to the *sum* of the logarithms of the two factors.

$$\log ab = \log a + \log b$$

This is easily proved to be true because, it

Figure 5. FOUR PLACE LOGARITHM TABLES.

N	0	1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396

N	0	1	2	3	4	5	6	7	8	9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7527
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122
85	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996

was shown before that when multiplying to powers, the exponents are added; therefore,

$$a \times b = 10^{\log a} \times 10^{\log b} = 10^{(\log a + \log b)}$$

Similarly, the logarithm of a quotient is the difference between the logarithm of the dividend and the logarithm of the divisor.

$$\log \frac{a}{b} = \log a - \log b$$

This is so because by the same rules of exponents:

$$\frac{a}{b} = \frac{10^{\log a}}{10^{\log b}} = 10^{(\log a - \log b)}$$

We have thus established an easier way of multiplication and division since these operations have been reduced to adding and subtracting.

The logarithm of a power of a number is equal to the logarithm of that number, multiplied by the exponent of the power.

$$\log a^2 = 2 \log a \text{ and } \log a^3 = 3 \log a$$

or, in general:

$$\log a^n = n \log a$$

Also, the logarithm of a root of a number is equal to the logarithm of that number divided by the index of the root:

$$\log \sqrt[n]{a} = \frac{1}{n} \log a$$

It follows from the rules of multiplication, that numbers having the same digits but different locations for the decimal point, have logarithms with the same mantissa:

$$\log 829 = 2.918555$$

$$\log 82.9 = 1.918555$$

$$\log 8.29 = 0.918555$$

$$\log 0.829 = -1.918555$$

$$\log 0.0829 = -2.918555$$

$$\log 829 = \log (8.29 \times 100) = \log 8.29 + \log 100 = 0.918555 + 2$$

Logarithm tables give the mantissas of logarithms only. The characteristic has to be determined by inspection. The characteristic is equal to the number of digits to the left of the decimal point *minus one*. In the case of logarithms of numbers less than unity, the characteristic is negative and is equal to the number of ciphers to the right of the decimal point *plus one*.

For reasons of convenience in making up logarithm tables, it has become the rule that the mantissa should always be positive. Such notations above as -1.918555 really mean $(+0.918555 \ -1)$; and -2.981555 means $(+0.918555 - 2)$. There are also some other notations in use such as

$$\bar{1}.918555 \text{ and } \bar{2}.918555$$

$$\text{also } 9.918555 - 10 \quad 8.918555 - 10 \quad 7.918555 - 10, \text{ etc.}$$

When, after some addition and subtraction of logarithms a mantissa should come out negative, one cannot look up its equivalent number or *anti-logarithm* in the table. The mantissa must first be made positive by adding and subtracting an appropriate integral number. Example: Suppose we find that the logarithm of a number is -0.34569, then we can transform it into the proper form by adding and subtracting 1

$$\begin{array}{ll} 1 & -1 \\ -0.34569 & \\ \hline 0.65431 & -1 \text{ or } -1.65431 \end{array}$$

Using Logarithm Tables

Logarithms are used for calculations involving multiplication, division, powers, and roots. Especially when the numbers are large and for higher, or fractional powers and roots, this becomes the most convenient way.

Logarithm tables are available giving the logarithms to three places, some to four places, others to five and six places. The table to use depends on the accuracy required in the result of our calculations. The four place table, printed in this chapter, permits the finding of answers to problems to four significant figures which is good enough for most constructional purposes. If greater accuracy is required a five place table should be consulted. The five place table is perhaps the most popular of all.

Referring now to the four place table, to find a common logarithm of a number, proceed as follows. Suppose the number is 5576. First, determine the characteristic. An inspection will show that the characteristic should be 3. This figure is placed to the left of the decimal point. The mantissa is now found by reference to the logarithm table. The first two numbers are 55; glance down the *N* column until coming to these figures. Advance to the right until coming in line with the column headed 7; the mantissa will be 7459. (Note that the column headed 7 corresponds to the *third figure* in the number 5576.) Place the mantissa 7459 to the right of the decimal point, making the logarithm of 5576 now read 3.7459. *Important:* do not consider the last figure 6 in the

N	L.	0*	1	2	3	4	5	6	7	8	9	P.P.
250	39	794	811	829	846	863	881	898	915	933	950	
251		967	985	*002	*019	*037	*054	*071	*088	*106	*123	18
252	40	140	157	175	192	209	226	243	261	278	295	1 1.8
253		312	329	346	364	381	398	415	432	449	466	2 3.6
254		483	500	518	535	552	569	586	603	620	637	3 5.4
												4 7.2
256		654	671	688	705	722	739	756	773	790	807	etc.

Figure 6.

A SMALL SECTION OF A FIVE PLACE LOGARITHM TABLE.

Logarithms may be found with greater accuracy with such tables, but they are only of use when the accuracy of the original data warrants greater precision in the figure work. Slightly greater accuracy may be obtained for intermediate points by interpolation, as explained in the text.

number 5576 when looking for the mantissa in the accompanying four place tables; in fact, one may usually disregard all digits beyond the first three when determining the mantissa. (*Interpolation*, sometimes used to find a logarithm more accurately, is unnecessary unless warranted by unusual accuracy in the available data.) However, be doubly sure to include all figures when ascertaining the magnitude of the characteristic.

To find the anti-logarithm, the table is used in reverse. As an example, let us find the anti-logarithm of 1.272 or, in other words, find the number of which 1.272 is the logarithm. Look in the table for the mantissa closest to 272. This is found in the first half of the table and the nearest value is 2718. Write down the first two significant figures of the anti-logarithm by taking the figures at the beginning of the line on which 2718 was found. This is 18; add to this, the digit above the column in which 2718 was found; this is 7. The anti-logarithm is 187 but we have not yet placed the decimal point. The characteristic is 1, which means that there should be two digits to the left of the decimal point. Hence, 18.7 is the anti-logarithm of 1.272.

For the sake of completeness we shall also describe the same operation with a five-place table where interpolation is done by means of tables of proportional parts (P.P. tables). Therefore we are reproducing here a small part of one page of a five-place table.

Finding the logarithm of 0.025013 is done as follows: We can begin with the characteristic, which is -2. Next find the first three digits in the column, headed by N and immediately after this we see 39, the first two digits of the mantissa. Then look among the headings of the other columns for the next digit of the number, in this case *1*. In the column, headed by *1* and on the line headed *250*, we find the next three digits of the logarithm, 811. So far, the logarithm is -2.39811 but this is the logarithm of 0.025010 and we want the logarithm of 0.025013. Here we can interpolate by observing that the difference between the log of 0.02501 and 0.02502 is $829 - 811$ or 18, in the last two significant figures. Looking in the P.P. table marked *18* we find after 3 the number 5.4 which is to be added to the logarithm.

$$\begin{array}{r} -2.39811 \\ 5.4 \\ \hline \end{array}$$

-2.39816, the logarithm of 0.025013

Since our table is only good to five places, we must eliminate the last figure given in the P.P. table if it is less than 5, otherwise we must add one to the next to the last figure, rounding off to a whole number in the P.P. table.

Finding the anti-logarithm is done the same way but with the procedure reversed. Suppose it is required to find the anti-logarithm of 0.40100. Find the first two digits in the column headed by L. Then one must look for the next three digits or the ones nearest to it, in the columns after 40 and on the lines from 40 to 41. Now here we find that numbers in the neighborhood of 100 occur only with an asterisk on the line just before 40 and still after 39. The asterisk means that instead of the 39 as the first two digits, these mantissas should have 40 as the first two digits. The logarithm 0.40100 is between the logs 0.40088 and 0.40106; the anti-logarithm is between 2517 and 2518. The difference between the two logarithms in the table is again 18 in the last two figures and our logarithm 0.40100 differs with the lower one 12 in the last figures. Look in the P.P. table of 18 which number comes closest to 12. This is found to be 12.6 for $7 \times 1.8 = 12.6$. Therefore we may add the digit 7 to the anti-logarithm already found; so we have 25177. Next, place the decimal point according to the rules: There are as many digits to the left of the decimal point as indicated in the characteristic plus one. The anti-logarithm of 0.40100 is 2.5177.

In the following examples of the use of logarithms we shall use only three places from the tables printed in this chapter since a greater degree of precision in our calculations would not be warranted by the accuracy of the data given.

In a 375 ohm bias resistor flows a current of 41.5 milliamperes; how many watts are dissipated by the resistor?

We write the equation for power in watts:

$$P = I^2R$$

and filling in the quantities in question, we have:

$$P = 0.0415^2 \times 375$$

Taking logarithms,

$$\log P = 2 \log 0.0415 + \log 375$$

$$\log 0.0415 = -2.618$$

$$\text{So } 2 \times \log 0.0415 = -3.236$$

$$\log 375 = 2.574$$

$$\log P = -1.810$$

antilog = 0.646. Answer = 0.646 watts

Caution: Do not forget that the negative sign before the characteristic belongs to the characteristic only and that mantissas are *always* positive. Therefore we recommend the other notation, for it is less likely to lead to errors. The work is then written:

$$\log 0.0415 = 8.618 - 10$$

$$2 \times \log 0.0415 = 17.236 - 20 = 7.236 - 10$$

$$\log 375 = 2.574$$

$$\log P = 9.810 - 10$$

Another example follows which demonstrates the ease in handling powers and roots. Assume an all-wave receiver is to be built, covering from 550 kc. to 60 mc. Can this be done in five ranges and what will be the required tuning ratio for each range if no overlapping is required? Call the tuning ratio of one band, x. Then the total tuning ratio for five such bands is x^5. But the total tuning ratio for all bands is 60/0.55. Therefore:

$$X^5 = \frac{60}{0.55} \text{ or : } x = \sqrt[5]{\frac{60}{0.55}}$$

Taking logarithms:

$$\log x = \frac{\log 60 - \log 0.55}{5}$$

log 60	1.778
log 0.55	−1.740
subtract	2.038

Remember again that the mantissas are positive and the characteristic alone can be negative. Subtracting −1 is the same as adding +1.

$$\log x = \frac{2.038}{5} = 0.408$$

$$x = \text{antilog } 0.408 = 2.56$$

The tuning ratio should be 2.56.

db	Power Ratio
0	1.00
1	1.26
2	1.58
3	2.00
4	2.51
5	3.16
6	3.98
7	5.01
8	6.31
9	7.94
10	10.00
20	100
30	1,000
40	10,000
50	100,000
60	1,000,000
70	10,000,000
80	100,000,000

Figure 7.
A TABLE OF DECIBEL GAINS VERSUS POWER RATIOS.

The Decibel

The decibel is a unit for the comparison of power or voltage levels in sound and electrical work. The sensation of our ears due to sound waves in the surrounding air is roughly proportional to the logarithm of the energy of the sound-wave and not proportional to the energy itself. For this reason a logarithmic unit is used so as to approach the reaction of the ear.

The decibel represents a *ratio* of two power levels, usually connected with gains or loss due to an amplifier or other network. The decibel is defined

$$N_{db} = 10 \log \frac{P_o}{P_i}$$

where P_o stands for the output power, P_i for the input power and N_{db} for the number of decibels. When the answer is positive, there is a gain; when the answer is negative, there is a loss.

The gain of amplifiers is usually given in decibels. For this purpose both the input power and output power should be measured. Example: Suppose that an intermediate amplifier is being driven by an input power of 0.2 watt and after amplification, the output is found to be 6 watts.

$$\frac{P_o}{P_i} = \frac{6}{0.2} = 30$$

$$\log 30 = 1.48$$

Therefore the gain is 10 × 1.48 = 14.8 decibels. The decibel is a logarithmic unit; when the power was multiplied by 30, the power level in decibels was increased by 14.8 decibels, or 14.8 decibels *added*.

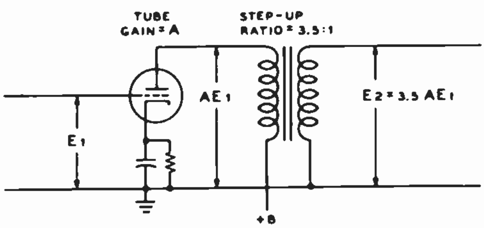


Figure 8.
STAGE GAIN.
The voltage gain in decibels in this stage is equal to the amplification in the tube plus the step-up ratio of the transformer, both expressed in decibels.

When one amplifier is to be followed by another amplifier, power gains are multiplied but the decibel gains are added. If a main amplifier having a gain of 1,000,000 (power ratio is 1,000,000) is preceded by a pre-amplifier with a gain of 1000, the total gain is 1,000,000,000. But in decibels, the first amplifier has a gain of 60 decibels, the second a gain of 30 decibels and the two of them will have a gain of 90 decibels when connected in cascade. (This is true only if the two amplifiers are properly matched at the junction as otherwise there will be a reflection loss at this point which must be subtracted from the total.)

Conversion of power ratios to decibels or vice versa is easy with the small table shown on these pages. In any case, an ordinary logarithm table will do. Find the logarithm of the power ratio and multiply by ten to find decibels.

Sometimes it is more convenient to figure decibels from voltage or current ratios or gains rather than from power ratios. This applies especially to voltage amplifiers. The equation for this is

$$N_{db} = 20 \log \frac{E_o}{E_i} \text{ or } 20 \log \frac{I_o}{I_i}$$

where the subscript, $_o$, denotes the output voltage or current and $_i$ the input voltage or current. Remember, this equation is true only if the voltage or current gain in question represents a power gain which is the square of it and not if the power gain which results from this is some other quantity due to impedance changes. This should be quite clear when we consider that a matching transformer to connect a speaker to a line or output tube does not represent a gain or loss; there is a voltage change and a current change yet the power remains the same for the impedance has changed.

On the other hand, when dealing with voltage amplifiers, we can figure the gain in a stage by finding the voltage ratio from the grid of the first tube to the grid of the next tube.

Example: In the circuit of Figure 8, the gain in the stage is equal to the amplification in the tube and the step-up ratio of the transformer. If the amplification in the tube is 10 and the step-up in the transformer is 3.5, the voltage gain is 35 and the gain in decibels is:

$$20 \times \log 35 = 20 \times 1.54 = 30.8 \text{ db}$$

Decibels as Power Level The original use of the decibel was only as a *ratio* of power levels—not as an absolute measure of power. However, one may use the decibel as such an absolute unit by fixing an arbitrary "zero" level, and to indicate any power level by its number of decibels above or below this arbitrary zero level. This is all very good so long as we agree on the zero level.

Any power level may then be converted to decibels by the equation:

$$N_{db} = 10 \log \frac{P_o}{P_{ref.}}$$

where N_{db} is the desired power level in decibels, P_o the output of the amplifier, $P_{ref.}$ the arbitrary reference level.

The zero level most frequently used (but not always) is 6 milliwatts or 0.006 watts. For this zero level, the equation reduces to

$$N_{db} = 10 \log \frac{P_o}{0.006}$$

Example: An amplifier using a 6F6 tube should be able to deliver an undistorted output of 3 watts. How much is this in decibels?

$$\frac{P_o}{P_{ref.}} = \frac{3}{.006} = 500$$

$$10 \times \log 500 = 10 \times 2.70 = 27.0$$

Therefore the power level at the output of the 6F6 is 27.0 decibels. When the power level to be converted is less than 6 milliwatts, the level is noted as negative. Here we must remember all that has been said regarding logarithms of numbers less than unity and the fact that the characteristic is negative but not the mantissa.

A preamplifier for a microphone is feeding 1.5 milliwatts into the line going to the regular speech amplifier. What is this power level expressed in decibels?

$$\text{decibels} = 10 \log \frac{P_o}{0.006} = 10 \log \frac{0.0015}{0.006} = 10 \log 0.25$$

Log 0.25 = $\bar{1}.398$ (from table). Therefore, $10 \times \bar{1}.398 = (10 \times -1 = -10) + (10 \times .398 = 3.98)$; adding the products algebraically, gives -6.02 db.

The conversion chart reproduced in this chapter will be of use in converting decibels to watts and vice versa.

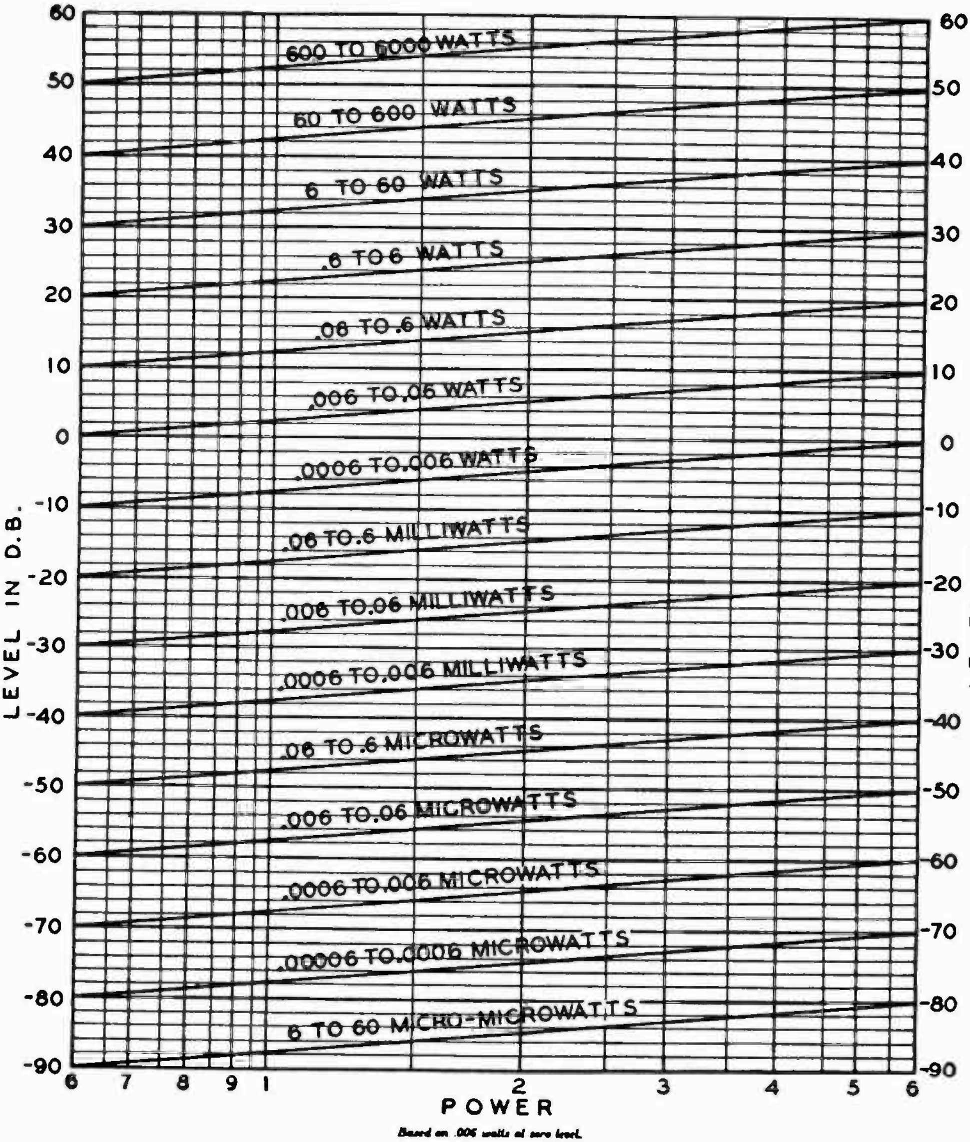


Figure 9.

CONVERSION CHART: POWER TO DECIBELS

Power levels between 6 micromicrowatts and 6000 watts may be referred to corresponding decibel levels between -90 and 60 db, and vice versa, by means of the above chart. Fifteen ranges are provided. Each curve begins at the same point where the preceding one ends, enabling uninterrupted coverage of the wide db and power ranges with condensed chart. For example: the lowermost curve ends at -80 db or 60 micromicrowatts and the next range starts at the same level. Zero db level is taken as 6 milliwatts (.006 watt).

Converting Decibels to Power It is often convenient to be able to convert a decibel value to a power equivalent. The formula used for this operation is

$$P = 0.006 \times \text{antilog} \frac{N_{db}}{10}$$

where P is the desired level in watts and N_{db} the decibels to be converted.

To determine the power level *P* from a decibel equivalent, simply divide the decibel value by 10; then take the number comprising the antilog and multiply it by 0.006; the product gives the level in watts.

Note: In problems dealing with the conversion of *minus* decibels to power, it often happens that the decibel value $-N_{db}$ is not divisible by 10. When this is the case, the numerator in the factor $-\frac{N_{db}}{10}$ must be made evenly divisible by 10, the negative signs must be observed, and the quotient labeled accordingly.

To make the numerator evenly divisible by 10 proceed as follows: Assume, for example, that $-N_{db}$ is some such value as -38; to make this figure evenly divisible by 10, we must add -2 to it, and, since we have added a negative 2 to it, we must also add a positive 2 so as to keep the net result the same.

Our decibel value now stands, $-40 + 2$. Dividing both of these figures by 10, as in the equation above, we have -4 and $+0.2$. Putting the two together we have the logarithm -4.2 with the negative characteristic and the positive mantissa as required.

The following examples will show the technique to be followed in practical problems.

(a) The output of a certain device is rated at -74 db. What is the power equivalent? Solution:

$$\frac{N_{db}}{10} = \frac{-74}{10} \text{ (not evenly divisible by 10)}$$

Routine:

$$\begin{array}{rr} -74 & \\ -\ 6 & +6 \\ \hline -80 & +6 \end{array}$$

$$\frac{N_{db}}{10} = \frac{-80+6}{10} = -8.6$$

$$\text{antilog } -8.6 = 0.000\ 000\ 04$$

$$.006 \times 0.000\ 000\ 04 = 0.000\ 000\ 000\ 24 \text{ watt or } 240 \text{ micro-microwatt}$$

(b) This example differs somewhat from that of the foregoing one in that the mantissas are added differently. A low-powered amplifier has an input signal level of -17.3 db. How many milliwatts does this value represent?

Solution:

$$\begin{array}{rr} -17.3 & \\ -\ 2.7 & +2.7 \\ \hline -20 & +2.7 \end{array}$$

$$\frac{N_{db}}{10} = \frac{-20+2.7}{10} = -2.27$$

$$\text{Antilog } -2.27 = 0.0186$$

$$0.006 \times 0.0186 = 0.000\ 1116 \text{ watt or } 0.1116 \text{ milliwatt}$$

Input voltages: To determine the required input voltage, take the peak voltage necessary to drive the last class A amplifier tube to maximum output, and divide this figure by the total overall voltage gain of the preceding stages.

Computing Specifications: From the preceding explanations the following data can be computed with any degree of accuracy warranted by the circumstances:

(1) Voltage amplification
(2) Overall gain in db
(3) Output signal level in db
(4) Input signal level in db
(5) Input signal level in watts
(6) Input signal voltage

When a power level is available which must be brought up to a new power level, the gain required in the intervening amplifier is equal to the difference between the two levels in decibels. If the required input of an amplifier for full output is -30 decibels and the output from a device to be used is but -45 decibels, the pre-amplifier required should have a gain of the difference, or 15 decibels. Again this is true only if the two amplifiers are properly matched and no losses are introduced due to mismatching.

Push-Pull Amplifiers To double the output of any cascade amplifier, it is only necessary to connect in push-pull the last amplifying stage, and replace the interstage and output transformers with push-pull types.

To determine the voltage gain (voltage ratio) of a push-pull amplifier, take the ratio of one *half* of the secondary winding of the push-pull transformer and multiply it by the μ of one of the output tubes in the push-pull stage; the product, *when doubled*, will be the voltage amplification, or step-up.

Other Units and Zero Levels When working with decibels one should not immediately take for granted that the zero level is 6 milliwatts for there are other zero levels in use.

In broadcast stations an entirely new system is now employed. Measurements made in acoustics are now made with the standard zero level of 10^{-16} watts per square cm.

Microphones are often rated with reference to the following zero level: *one volt at open circuit when the sound pressure is one millibar.* In any case, the rating of the microphone must include the loudness of the sound. It is obvious that this zero level does not lend itself readily for the calculation of required gain in an amplifier.

The VU: So far, the decibel has always referred to a type of signal which can readily be measured, that is, a steady signal of a single frequency. But what would be the power level of a signal which is constantly varying in volume and frequency? The measurement of voltage would depend on the type of instrument employed, whether it is measured with a thermal square law meter or one that shows average values; also, the inertia of the movement will change its indications at the peaks and valleys.

After considerable consultation, the broadcast chains and the Bell System have agreed on the *VU*. The level in VU is the level in decibels above *1 milliwatt* zero level and measured with a carefully defined type of instrument across a 600 ohm line. So long as we deal with an unvarying sound, the level in VU is equal to decibels above 1 milliwatt; but when the sound level varies, the unit is the VU and the special meter must be used. There is then no equivalent in decibels.

The Neper: We might have used the natural logarithm instead of the common logarithm when defining our logarithmic unit of sound. This was done in Europe and the unit obtained is known as the *neper* or *napier*. It is still found in some American literature on filters.

1 neper = 8.686 decibels
1 decibel = 0.1151 neper

AC Meters With Decibel Scales Many test instruments are now equipped with scales calibrated in decibels which is very handy when making measurements of frequency characteristics and gain. These meters are generally calibrated for connection across a 500 ohm line and for a zero level of 6 milliwatts. When they are connected across another impedance, the reading on the meter is no longer correct for the zero level of 6 milliwatts. A correction factor should be applied consisting in the addition or subtraction of a steady figure to all readings on the meter. This figure is given by the equation:

$$\text{db to be added} = 10 \log \frac{500}{Z}$$

where Z is the impedance of the circuit under measurement.

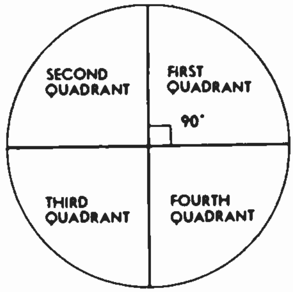


Figure 10.
THE CIRCLE IS DIVIDED INTO FOUR QUADRANTS BY TWO PERPENDICULAR LINES AT RIGHT ANGLES TO EACH OTHER.
The "northeast" quadrant thus formed is known as the first quadrant; the others are numbered consecutively in a counterclockwise direction.

Trigonometry

Definition and Use Trigonometry is the science of mensuration of *triangles*. At first glance triangles may seem to have little to do with electrical phenomena; however, in a.c. work most currents and voltages follow laws equivalent to those of the various trigonometric relations which we are about to examine briefly. Examples of their application to a.c. work will be given in the section on *Vectors*.

Angles are measured in *degrees* or in *radians*. The circle has been divided into 360 degrees, each degree into 60 minutes, and each minute into 60 seconds. A decimal division of the degree is also in use because it makes calculation easier. Degrees, minutes and seconds are indicated by the following signs: °, ′ and ″ Example: 6° 5′ 23″ means six degrees, five minutes, twenty-three seconds. In the decimal notation we simply write 8.47°, eight and forty-seven hundredths of a degree.

When a circle is divided into four quadrants by two perpendicular lines passing through the center (Figure 10) the angle made by the two lines is 90 degrees, known as a *right angle*. Two right angles, or 180° equals a *straight angle*.

The radian: If we take the radius of a circle and bend it so it can cover a part of the circumference, the arc it covers subtends an angle called a *radian* (Figure 11). Since the diameter of a circle equals 2 times the radius, there are 2π radians in 360°. So we have the following relations:

1 radian = 57° 17′ 45″ = 57.2958° $\quad \pi = 3.14159$
1 degree = 0.01745 radians
π radians = 180° $\quad \pi/2$ radians = 90°
$\pi/3$ radians = 60°

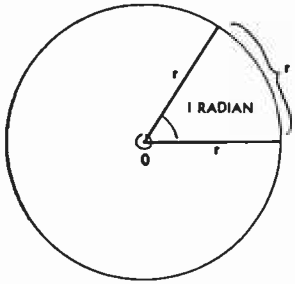


Figure 11.
THE RADIAN.
A radian is an angle whose arc is exactly equal to the length of either side. Note that the angle is constant regardless of the length of the side and the arc so long as they are equal. A radian equals 57.2958°.

In trigonometry we consider an angle *generated* by two lines, one stationary and the other rotating as if it were hinged at 0, Figure 12. Angles can be greater than 180 degrees and even greater than 360 degrees as illustrated in this figure.

Two angles are complements of each other when their sum is 90°, or a right angle. *A* is the complement of *B* and *B* is the complement of *A* when

$$A=(90°-B)$$

and when

$$B=(90°-A)$$

Two angles are supplements of each other when their sum is equal to a straight angle, or 180°. A is the supplement of B and B is the supplement of A when

$$A=(180°-B)$$

and

$$B=(180°-A)$$

In the angle *A*, Figure 13A, a line is drawn from *P*, perpendicular to *b*. Regardless of the point selected for *P*, the *ratio a/c* will always be the same for any given angle, *A*. So will all the other proportions between *a, b,* and *c* remain constant regardless of the position of point *P* on *c*. The six possible ratios each are named and defined as follows:

$$\text{sine A} = \frac{a}{c} \qquad \text{cosine A} = \frac{b}{c}$$

$$\text{tangent A} = \frac{a}{b} \qquad \text{cotangent A} = \frac{b}{a}$$

$$\text{secant A} = \frac{c}{b} \qquad \text{cosecant A} = \frac{c}{a}$$

Let us take a special angle as an example. For instance, let the angle *A* be 60 degrees as in Figure 13B. Then the relations between the sides are as in the figure and the six functions become:

$$\sin 60° = \frac{a}{c} = \frac{\frac{1}{2}\sqrt{3}}{1} = \tfrac{1}{2}\sqrt{3}$$

$$\cos 60° = \frac{b}{c} = \frac{\frac{1}{2}}{1} = \tfrac{1}{2}$$

$$\tan 60° = \frac{a}{b} = \frac{\frac{1}{2}\sqrt{3}}{\frac{1}{2}} = \sqrt{3}$$

$$\cot 60° = \frac{\frac{1}{2}}{\frac{1}{2}\sqrt{3}} = \frac{1}{\sqrt{3}} = \tfrac{1}{3}\sqrt{3}$$

$$\sec 60° = \frac{c}{b} = \frac{1}{\frac{1}{2}} = 2$$

$$\csc 60° = \frac{c}{a} = \frac{1}{\frac{1}{2}\sqrt{3}} = \tfrac{2}{3}\sqrt{3}$$

Another example: Let the angle be 45°, then the relations between the lengths of *a, b,* and *c* are as shown in Figure 13C, and the six functions are:

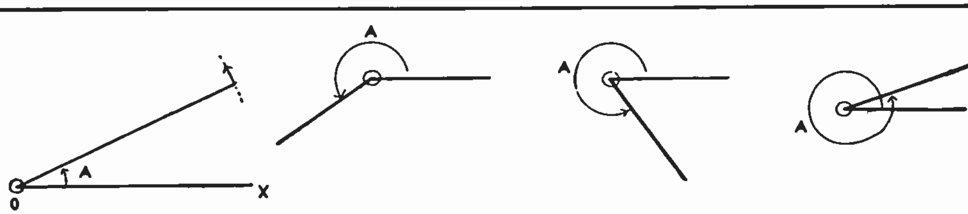


Figure 12.
AN ANGLE IS GENERATED BY TWO LINES, ONE STATIONARY AND THE OTHER ROTATING.
The line OX is stationary; the line with the small arrow at the far end rotates in a counterclockwise direction. At the position illustrated in the lefthandmost section of the drawing it makes an angle, A, which is less than 90° and is therefore in the first quadrant. In the position shown in the second portion of the drawing the angle A has increased to such a value that it now lies in the third quadrant; note that an angle can be greater than 180°. In the third illustration the angle A is in the fourth quadrant. In the fourth position the rotating vector has made more than one complete revolution and is hence in the fifth quadrant; since the fifth quadrant is an exact repetition of the first quadrant, its values will be the same as in the lefthandmost portion of the illustration.

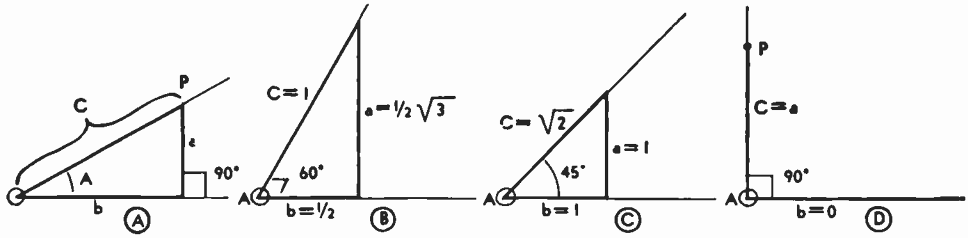


Figure 13.

THE TRIGONOMETRIC FUNCTIONS.

In the right triangle shown in (A) the side opposite the angle A is a, while the adjoining sides are b and c; the trigonometric functions of the angle A are completely defined by the ratios of the sides a, b and c. In (B) are shown the lengths of the sides a and b when angle A is 60° and side c is 1. In (C) angle A is 45°; a and b equal 1, while c equals $\sqrt{2}$. In (D) note that c equals a for a right angle while b equals 0.

$$\sin 45^\circ = \frac{1}{\sqrt{2}} = \tfrac{1}{2}\sqrt{2} \qquad \cos 45^\circ = \frac{1}{\sqrt{2}} = \tfrac{1}{2}\sqrt{2}$$

$$\tan 45^\circ = \frac{1}{1} = 1 \qquad \cot 45^\circ = \frac{1}{1} = 1$$

$$\sec 45^\circ = \frac{\sqrt{2}}{1} = \sqrt{2} \qquad \text{cosec } 45^\circ = \frac{\sqrt{2}}{1} = \sqrt{2}$$

There are some special difficulties when the angle is zero or 90 degrees. In Figure 13D an angle of 90 degrees is shown; drawing a line perpendicular to b from point P makes it fall on top of c. Therefore in this case $a = c$ and $b = 0$. The six ratios are now:

$$\sin 90^\circ = \frac{a}{c} = 1 \qquad \cos 90^\circ = \frac{b}{c} = \frac{0}{c} = 0$$

$$\tan 90^\circ = \frac{a}{b} = \frac{a}{0} = \infty \qquad \cot 90^\circ = \frac{0}{a} = 0$$

$$\sec 90^\circ = \frac{c}{b} = \frac{c}{0} = \infty \qquad \text{cosec } 90^\circ = \frac{c}{a} = 1$$

When the angle is zero, $a=0$ and $b=c$. The values are then:

$$\sin 0^\circ = \frac{a}{c} = \frac{0}{c} = 0 \qquad \cos 0^\circ = \frac{b}{c} = 1$$

$$\tan 0^\circ = \frac{a}{b} = \frac{0}{b} = 0 \qquad \cot 0^\circ = \frac{b}{a} = \frac{b}{0} = \infty$$

$$\sec 0^\circ = \frac{c}{b} = 1 \qquad \text{cosec } 0^\circ = \frac{c}{a} = \frac{c}{0} = \infty$$

In general, for every angle, there will be definite values of the six functions. Conversely, when any of the six functions is known, the angle is defined. Tables have been calculated giving the value of the functions for angles.

From the foregoing we can make up a small table of our own (Figure 14), giving values of the functions for some common angles.

Relations Between Functions It follows from the definitions that

$$\sin A = \frac{1}{\text{cosec } A} \qquad \cos A = \frac{1}{\sec A}$$

$$\text{and } \tan A = \frac{1}{\cot A}$$

From the definitions also follows the relation

$$\cos A = \sin\,(\text{complement of } A) = \sin\,(90^\circ - A)$$

because in the right triangle of Figure 15, $\cos A = b/c = \sin B$ and $B = 90^\circ - A$ or the complement of A. For the same reason:

$$\cot A = \tan\ (90^\circ - A)$$

$$\csc A = \sec\ (90^\circ - A)$$

Relations in Right Triangles In the right triangle of Figure 15, $\sin A = a/c$ and by transposition

$$a = c \sin A$$

For the same reason we have the following identities:

$$\tan A = a/b \qquad a = b \tan A$$
$$\cot A = b/a \qquad b = a \cot A$$

In the same triangle we can do the same for functions of the angle B

Angle	Sin	Cos.	Tan	Cot	Sec.	Cosec.
0	0	1	0	∞	1	∞
30°	½	$\frac{1}{2}\sqrt{3}$	$\frac{1}{3}\sqrt{3}$	$\sqrt{3}$	$\frac{2}{3}\sqrt{3}$	2
45°	$\frac{1}{2}\sqrt{2}$	$\frac{1}{2}\sqrt{2}$	1	1	$\sqrt{2}$	$\sqrt{2}$
60°	$\frac{1}{2}\sqrt{3}$	½	$\sqrt{3}$	$\frac{1}{3}\sqrt{3}$	2	$\frac{2}{3}\sqrt{3}$
90°	1	0	∞	0	∞	1

Figure 14.

Values of trigonometric functions for common angles in the first quadrant.

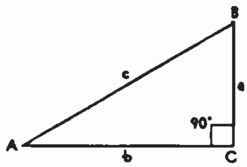


Figure 15.

In this figure the sides a, b, and c are used to define the trigonometric functions of angle B as well as angle A.

sin B = b/c	b = c sin B
cos B = a/c	a = c cos B
tan B = b/a	b = a tan B
cot B = a/b	a = b cot B

Functions of Angles Greater than 90 Degrees In angles greater than 90 degrees, the values of *a* and *b* become negative on occasion in accordance with the rules of Cartesian coordinates. When *b* is measured from 0 towards the left it is considered negative and similarly, when *a* is measured from 0 downwards, it is negative. Referring to Figure 16, an angle in the *second quadrant* (between 90° and 180°) has some of its functions negative:

$$\sin A = \frac{a}{c} = \text{pos.} \qquad \cos A = \frac{-b}{c} = \text{neg.}$$

$$\tan A = \frac{a}{-b} = \text{neg.} \qquad \cot A = \frac{-b}{a} = \text{neg.}$$

$$\sec A = \frac{c}{-b} = \text{neg.} \qquad \text{cosec } A = \frac{c}{a} = \text{pos.}$$

For an angle in the third quadrant (180° to 270°), the functions are

$$\sin A = \frac{-a}{c} = \text{neg.} \qquad \cos A = \frac{-b}{c} = \text{neg.}$$

$$\tan A = \frac{-a}{-b} = \text{pos.} \qquad \cot A = \frac{-b}{-a} = \text{pos.}$$

$$\sec A = \frac{c}{-b} = \text{neg.} \qquad \text{cosec } A = \frac{c}{-a} = \text{neg.}$$

And in the fourth quadrant (270° to 360°):

$$\sin A = \frac{-a}{c} = \text{neg.} \qquad \cos A = \frac{b}{c} = \text{pos.}$$

$$\tan A = \frac{-a}{b} = \text{neg.} \qquad \cot A = \frac{b}{-a} = \text{neg.}$$

$$\sec A = \frac{c}{b} = \text{pos.} \qquad \text{cosec } A = \frac{c}{-a} = \text{neg.}$$

Summarizing, the sign of the functions in each quadrant can be seen at a glance from Figure 17, where in each quadrant are written the names of functions which are positive; those not mentioned are negative.

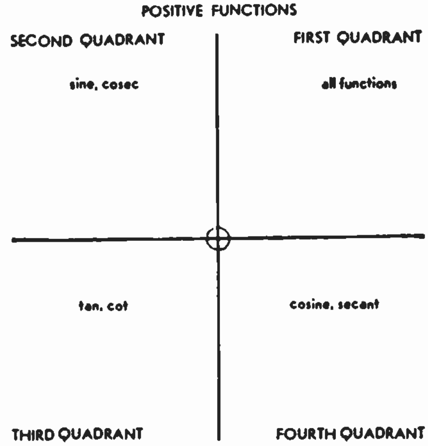


Figure 17.

SIGNS OF THE TRIGONOMETRIC FUNCTIONS.

The functions listed in this diagram are positive; all other functions are negative.

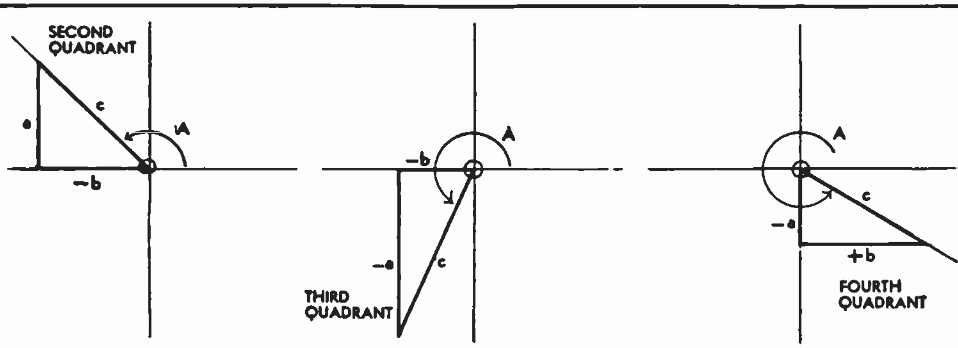


Figure 16.

TRIGONOMETRIC FUNCTIONS IN THE SECOND, THIRD, AND FOURTH QUADRANTS.

The trigonometric functions in these quadrants are similar to first quadrant values, but the signs of the functions vary as listed in the text and in Figure 17.

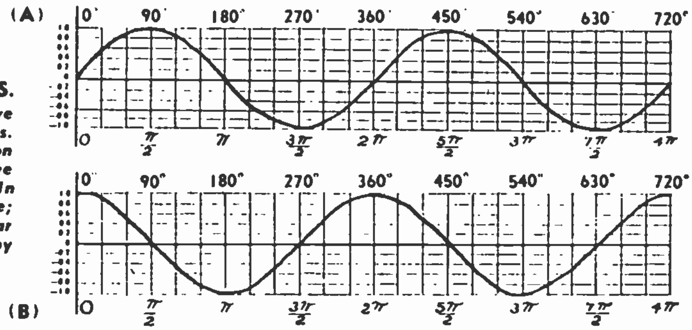


Figure 18.
SINE AND COSINE CURVES.
In (A) we have a sine curve drawn in Cartesian coordinates. This is the usual representation of an alternating current wave without substantial harmonics. In (B) we have a cosine wave; note that it is exactly similar to a sine wave displaced by 90° or π/2 radians.

Graphs of Trigonometric Functions

The sine wave. When we have the relation $y = \sin x$, where x is an angle measured in radians or degrees, we can draw a curve of y versus x for all values of the independent variable, and thus get a good conception how the sine varies with the magnitude of the angle. This has been done in Figure 18A. We can learn from this curve the following facts.

1. The sine varies between +1 and −1
2. It is a periodic curve, repeating itself after every multiple of 2π or 360°
3. $\sin x = \sin (180° - x)$ or $\sin (\pi - x)$
4. $\sin x = -\sin (180° + x)$, or $-\sin (\pi + x)$

The cosine wave. Making a curve for the function $y = \cos x$, we obtain a curve similar to that for $y = \sin x$ except that it is displaced by 90° or $\pi/2$ radians with respect to the Y-axis. This curve (Figure 18B) is also periodic but it does not start with zero. We read from the curve:

1. The value of the cosine never goes beyond +1 or −1
2. The curve repeats, after every multiple of 2π radians or 360°
3. $\cos x = -\cos (180° - x)$ or $-\cos (\pi - x)$
4. $\cos x = \cos (360° - x)$ or $\cos (2\pi - x)$

The graph of the tangent is illustrated in Figure 19. This is a discontinuous curve and illustrates well how the tangent increases from zero to infinity when the angle increases from zero to 90 degrees. Then when the angle is further increased, the tangent starts from minus infinity going to zero in the second quadrant, and to infinity again in the third quadrant.

1. The tangent can have any value between $+\infty$ and $-\infty$
2. The curve repeats and the period is π radians or 180°, not 2π radians
3. $\tan x = \tan (180° + x)$ or $\tan (\pi + x)$
4. $\tan x = -\tan (180° - x)$ or $-\tan (\pi - x)$

The graph of the cotangent is the inverse of that of the tangent, see Figure 20. It leads us to the following conclusions:

1. The cotangent can have any value between $+\infty$ and $-\infty$
2. It is a periodic curve, the period being π radians or 180°
3. $\cot x = \cot (180° + x)$ or $\cot (\pi + x)$
4. $\cot x = -\cot (180° - x)$ or $-\cot (\pi - x)$

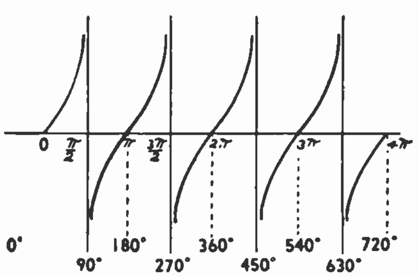


Figure 19.
TANGENT CURVES.
The tangent curve increases from 0 to ∞ with an angular increase of 90°. In the next 180° it increases from −∞ to +∞.

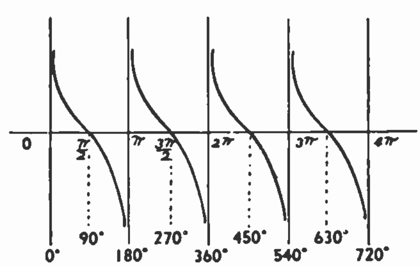


Figure 20.
COTANGENT CURVES.
Cotangent curves are the inverse of the tangent curves. They vary from +∞ to −∞ in each pair of quadrants.

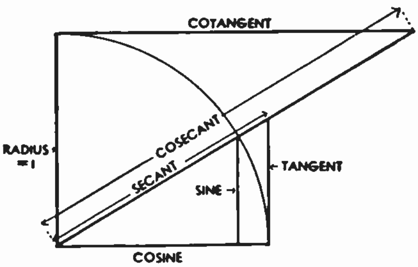


Figure 21.
ANOTHER REPRESENTATION OF TRIGONOMETRIC FUNCTIONS.
If the radius of a circle is considered as the unit of measurement, then the lengths of the various lines shown in this diagram are numerically equal to the functions marked adjacent to them.

The graphs of the secant and cosecant are of lesser importance and will not be shown here. They are the inverse, respectively, of the cosine and the sine, and therefore they vary from +1 to infinity and from −1 to −infinity.

Perhaps another useful way of visualizing the values of the functions is by considering Figure 21. If the radius of the circle is the unit of measurement then the lengths of the lines are equal to the functions marked on them.

Trigonometric Tables There are two kinds of trigonometric tables. The first type gives the functions of the angles, the second the logarithms of the functions. The first kind is also known as the table of *natural* trigonometric functions.

These tables give the functions of all angles between 0 and 45°. This is all that is necessary for the function of an angle between 45° and 90° can always be written as the co-function of an angle below 45°. Example: If we had to find the sine of 48°, we might write

$$\sin 48° = \cos (90° - 48°) = \cos 42°$$

Tables of the logarithms of trigonometric functions give the common logarithms ($\log_{10}$) of these functions. Since many of these logarithms have negative characteristics, one should add −10 to all logarithms in the table which have a characteristic of 6 or higher. For instance, the log sin 24° = 9.60931 −10. Log tan 1° = 8.24192 −10 but log cot 1° = 1.75808. When the characteristic shown is less than 6, it is supposed to be positive and one should not add −10.

Vectors

A *scalar* quantity has *magnitude* only; a *vector* quantity has both *magnitude* and *direction*. When we speak of a speed of 50 miles per hour, we are using a scalar quantity, but when we say the wind is Northeast and has a velocity of 50 miles per hour, we speak of a vector quantity.

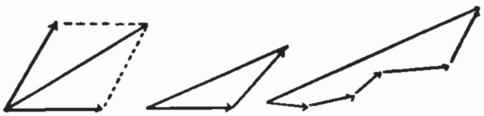


Figure 22.
Vectors may be added as shown in these sketches. In each case the long vector represents the vector sum of the smaller vectors. For many engineering applications sufficient accuracy can be obtained by this method which avoids long and laborious calculations.

Vectors, representing forces, speeds, displacements, etc., are represented by arrows. They can be added graphically by well known methods illustrated in Figure 22. We can make the parallelogram of forces or we can simply draw a triangle. The addition of many vectors can be accomplished graphically as in the same figure.

In order that we may define vectors algebraically and add, subtract, multiply, or divide them, we must have a logical notation system that lends itself to these operations. For this purpose vectors can be defined by coordinate systems. Both the Cartesian and the polar coordinates are in use.

Vectors Defined by Cartesian Coordinates Since we have seen how the sum of two vectors is obtained, it follows from Figure 23, that the vector $\dot{Z}$ equals the sum of the two vectors $\dot{x}$ and $\dot{y}$. In fact, any vector can be resolved into vectors along the X- and Y-axis. For convenience in working with these quantities we need to dis-

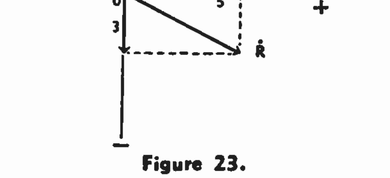


Figure 23.
RESOLUTION OF VECTORS.
Any vector such as Ż may be resolved into two vectors, x and y, along the X- and Y-axes. If vectors are to be added, their respective x and y components may be added to find the x and y components of the resultant vector.

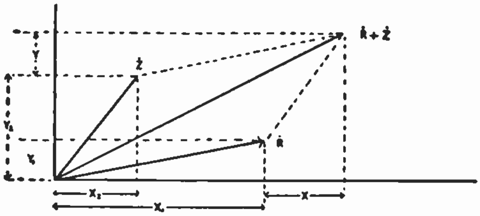


Figure 24.
ADDITION OR SUBTRACTION OF VECTORS.
Vectors may be added or subtracted by adding or subtracting their x or y components separately.

tinguish between the *X*- and *Y*-component, and so it has been agreed that the *Y*-component alone shall be marked with the letter *j*. Example (Figure 23):

$$\dot{Z} = 3 + 4j$$

Note again that the sign of components along the *X*-axis is positive when measured from 0 to the right and negative when measured from 0 towards the left. Also, the component along the *Y*-axis is positive when measured from 0 upwards, and negative when measured from 0 downwards. So the vector, $\dot{R}$, is described as

$$\dot{R} = 5 - 3j$$

Vector quantities are usually indicated by some special typography, especially by using a point over the letter indicating the vector, as $\dot{R}$.

Absolute Value of a Vector The absolute or scalar value of vectors such as $\dot{Z}$ or $\dot{R}$ in Figure 23 is easily found by the theorem of Pythagoras, which states that in any right-angled triangle the square of the side opposite the right angle is equal to the sum of the squares of the sides adjoining the right angle. In Figure 23, *OAB* is a right-angled triangle; therefore, the square of *OB* (or *Z*) is equal to the square of *OA* (or *x*) plus the square of *AB* (or *y*). Thus the absolute values of *Z* and *R* may be determined as follows:

$$|Z| = \sqrt{x^2 + y^2}$$

$$|Z| = \sqrt{3^2 + 4^2} = 5$$

$$|R| = \sqrt{5^2 + 3^2} = \sqrt{34} = 5.83$$

The vertical lines indicate that the absolute or scalar value is meant without regard to sign or direction.

Addition of Vectors An examination of Figure 24 will show that the two vectors

$$\dot{R} = x_1 + j\,y_1$$

$$\dot{Z} = x_2 + j\,y_2$$

can be added, if we add the *X*-components and the *Y*-components separately.

$$\dot{R} + \dot{Z} = x_1 + x_2 + j\,(y_1 + y_2)$$

For the same reason we can carry out subtraction by subtracting the horizontal components and subtracting the vertical components

$$\dot{R} - \dot{Z} = x_1 - x_2 + j\,(y_1 - y_2)$$

Let us consider the operator *j*. If we have a vector *a* along the X-axis and add a *j* in front of it (multiplying by *j*) the result is that the direction of the vector is rotated forward 90 degrees. If we do this twice (multiplying by j^2) the vector is rotated forward by 180 degrees and now has the value $-a$. Therefore multiplying by j^2 is equivalent to multiplying by -1. Then

$$j^2 = -1 \text{ and } j = \sqrt{-1}$$

This is the imaginary number discussed before under algebra. In electrical engineering the letter *j* is used rather than *i*, because *i* is already known as the symbol for current.

Multiplying Vectors When two vectors are to be multiplied we can perform the operation just as in algebra, remembering that $j^2 = -1$.

$$\dot{R}\dot{Z} = (x_1 + jy_1)\,(x_2 + jy_2)$$

$$= x_1\,x_2 + jx_1\,y_2 + j\,x_2\,y_1 + j^2\,y_1\,y_2$$

$$= x_1\,x_2 - y_1\,y_2 + j\,(x_1\,y_2 + x_2\,y_1)$$

Division has to be carried out so as to remove the *j*-term from the denominator. This can be done by multiplying both denominator and numerator by a quantity which will eliminate *j* from the denominator. Example:

$$\frac{\dot{R}}{\dot{Z}} = \frac{x_1 + jy_1}{x_2 + jy_2} = \frac{(x_1 + jy_1)\,(x_2 - jy_2)}{(x_2 + jy_2)\,(x_2 - jy_2)}$$

$$= \frac{x_1x_2 + y_1y_2 + j\,(x_2y_1 - x_1y_2)}{x_2^2 + y_2^2}$$

Polar Coordinates A vector can also be defined in polar coordinates by its magnitude and its vectorial angle with an arbitrary reference axis. In Figure 25

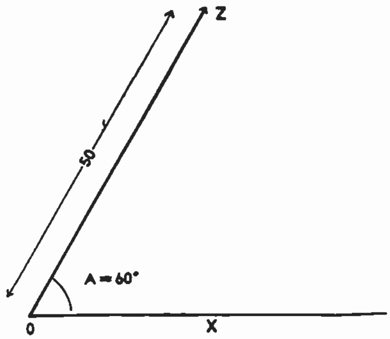


Figure 25.
IN THIS FIGURE A VECTOR HAS BEEN REPRESENTED IN POLAR INSTEAD OF CARTESIAN CO-ORDINATES.
In polar coordinates a vector is defined by a magnitude and an angle, called the vectorial angle, instead of by two magnitudes as in Cartesian coordinates.

the vector $\dot{Z}$ has a magnitude 50 and a vectorial angle of 60 degrees. This will then be written

$$\dot{Z} = 50\angle 60°$$

A vector $a + jb$ can be transformed into polar notation very simply (see Figure 26)

$$\dot{Z} = a + jb = \sqrt{a^2 + b^2}\,\angle \tan^{-1}\frac{b}{a}$$

In this connection tan^{-1} means *the angle of which the tangent is*. Sometimes the notation *arc tan b/a* is used. Both have the same meaning.

A polar notation of a vector can be transformed into a Cartesian coordinate notation in the following manner (Figure 27)

$$\dot{Z} = p\angle A = p \cos A + jp \sin A$$

A sinusoidally alternating voltage or current is symbolically represented by a rotating vector, having a magnitude equal to the peak voltage or current and rotating with an angular velocity of $2\pi f$ radians per second or as many revolutions per second as there are cycles per second.

The instantaneous voltage, *e*, is always equal to the sine of the vectorial angle of this rotating vector, multiplied by its magnitude.

$$e = E \sin 2\pi ft$$

The alternating voltage therefore varies with time as the sine varies with the angle. If we plot time horizontally and instantaneous voltage vertically we will get a curve like those in Figure 18.

In alternating current circuits, the current

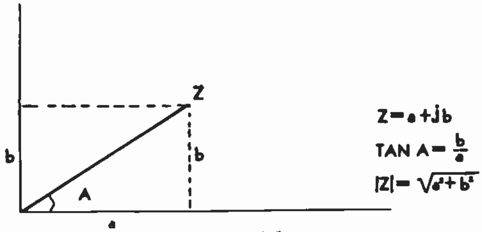


Figure 26.
Vectors can be transformed from Cartesian into polar notation as shown in this figure.

which flows due to the alternating voltage is not necessarily in step with it. The rotating current vector may be ahead or behind the voltage vector, having a *phase difference* with it. For convenience we draw these vectors as if they were standing still, so that we can indicate the difference in phase or the *phase angle*. In Figure 28 the current lags behind the voltage by the angle θ, or we might say that the voltage leads the current by the angle θ.

Vector diagrams show the phase relations between two or more vectors (voltages and currents) in a circuit. They may be added and subtracted as described; one may add a voltage vector to another voltage vector or a current vector to a current vector but not a current vector to a voltage vector (for the same reason that one cannot add a force to a speed). Figure 28 illustrates the relations in the simple series circuit of a coil and resistor. We know that the current passing through coil and resistor must be the same and in the same phase, so we draw this current *I* along the *X*-axis. We know also that the voltage drop *IR* across the resistor is in phase with the current, so the vector *IR* representing the voltage drop is also along the *X*-axis.

The voltage across the coil is 90 degrees ahead of the current through it; *IX* must therefore be drawn along the *Y*-axis. $\dot{E}$ the applied voltage must be equal to the vectorial sum of the two voltage drops, *IR* and *IX*, and we have so constructed it in the drawing. Now expressing the same in algebraic notation, we have

$$\dot{E} = IR + jIX$$

$$I\dot{Z} = IR + jIX$$

Dividing by *I*

$$\dot{Z} = R + jX$$

Due to the fact that a *reactance* rotates the voltage vector ahead or behind the current vector by 90 degrees, we must mark it with a *j* in vector notation. Inductive reactance will have a plus sign because it shifts the voltage vector forwards; a capacitive reactance is neg-

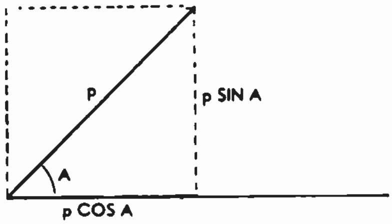


Figure 27.
Vectors can be transformed from polar into Cartesian notation as shown in this figure.

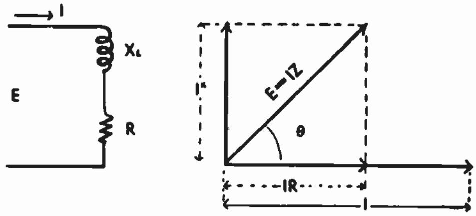


Figure 28.
VECTOR REPRESENTATION OF A SIMPLE SERIES CIRCUIT.
The righthand portion of the illustration shows the vectors representing the voltage drops in the coil and resistance illustrated at the left. Note that the voltage drop across the coil X_L leads that across the resistance by 90°.

ative because the voltage will lag behind the current. Therefore:

$$X_L = +j\,2\pi fL$$

$$X_C = -j\,\frac{1}{2\pi fC}$$

In Figure 28 the angle θ is known as the phase angle between E and I. When calculating power, only the real components count. The power in the circuit is then

$$P = I\,(IR)$$

$$\text{but } IR = E\cos\theta$$

$$\therefore P = EI\cos\theta$$

This *cos θ* is known as the power factor of the circuit. In many circuits we strive to keep the angle θ as small as possible, making *cos θ* as near to unity as possible. In tuned circuits, we use reactances which should have as low a power factor as possible. The merit of a coil or condenser, its *Q*, is defined by the tangent of this phase angle:

$$Q = \tan\theta = X/R$$

For an efficient coil or condenser, *Q* should be as large as possible; the phase-angle should then be as close to 90 degrees as possible, making the power factor nearly zero. *Q* is almost but not quite the inverse of *cos θ*. Note that in Figure 29

$$Q = X/R \qquad \text{and } \cos\theta = R/Z$$

When Q is more than 5, the power factor is less than 20%; we can then safely say $Q = 1/\cos\theta$ with a maximum error of about 2½ percent, for in the worst case, when $\cos\theta = 0.2$, Q will equal $\tan\theta = 4.89$. For higher values of Q, the error becomes less.

Note that from Figure 29 can be seen the simple relation:

$$\dot{Z} = R + jX_L$$

$$|Z| = \sqrt{R^2 + X_L^2}$$

Graphical Representation

Formulas and physical laws are often presented in graphical form; this gives us a "bird's eye view" of various possible conditions due to the variations of the quantities involved. In some cases graphs permit us to solve equations with greater ease than ordinary algebra.

Coordinate Systems All of us have used coordinate systems without realizing it. For instance, in modern cities we have numbered streets and numbered avenues. By this means we can define the location of any spot in the city if the nearest street crossings are named. This is nothing but an application of Cartesian coordinates.

In the Cartesian coordinate system (named after Descartes), we define the location of any point in a plane by giving its distance from each of two perpendicular lines or *axes*. Figure 30 illustrates this idea. The vertical axis is called the *Y-axis*, the horizontal axis is the *X-axis*. The intersection of these two axes is called the *origin, O*. The location of a point, P, (Figure 30) is defined by measuring the respective distances, *x* and *y* along the X-axis and the Y-axis. In this example the distance along the X-axis is 2 units and along the Y-axis is 3 units. Thus we define the point as

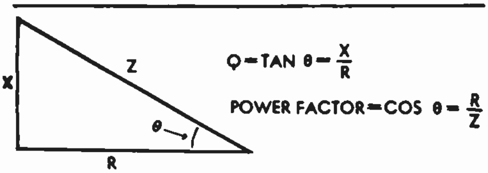


Figure 29.
The figure of merit of a coil and its resistance is represented by the ratio of the inductive reactance to the resistance, which as shown in this diagram is equal to $\frac{X_L}{R}$ which equals tan θ. For large values of θ (the phase angle) this is approximately equal to the reciprocal of the cos θ.

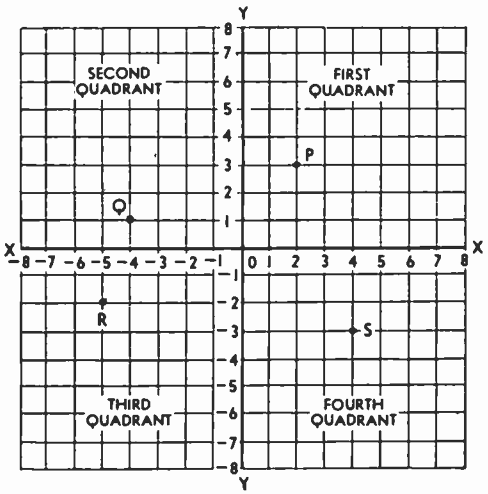


Figure 30.

CARTESIAN COORDINATES.

The location of any point can be defined by its distance from the X and Y axes.

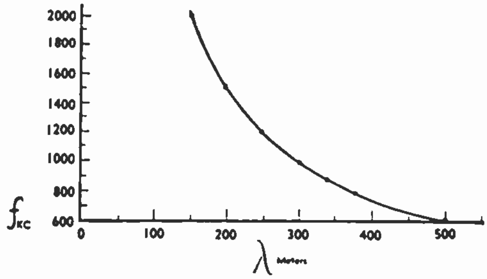


Figure 31.

REPRESENTATION OF A SIMPLE FUNCTION IN CARTESIAN CO-ORDINATES.

In this chart of the function $f_{kc} = \frac{300{,}000}{\lambda_{meters}}$ distances along the X axis represent wave-length in meters, while those along the Y axis represent frequency in kilocycles. A curve such as this helps to find values between those calculated with sufficient accuracy for most purposes.

P 2, 3 or we might say $x = 2$ and $y = 3$. The measurement x is called the *abscissa* of the point and the distance y is called its *ordinate.* It is arbitrarily agreed that distances measured from 0 to the right along the X-axis shall be reckoned positive and to the left negative. Distances measured along the Y-axis are positive when measured upwards from 0 and negative when measured downwards from 0. This is illustrated in Figure 30. The two axes divide the plane area into four parts called quadrants. These four quadrants are numbered as shown in the figure.

It follows from the foregoing statements, that points lying within the first quadrant have both x and y positive, as is the case with the point P. A point in the second quadrant has a negative abscissa, x, and a positive ordinate, y. This is illustrated by the point Q, which has the coordinates $x = -4$ and $y = +1$. Points in the third quadrant have both x and y negative. $x = -5$ and $y = -2$ illustrates such a point, R. The point S, in the fourth quadrant has a negative ordinate, y and a positive abscissa or x.

In practical applications we might draw only as much of this plane as needed to illustrate our equation and therefore, the scales along the X-axis and Y-axis might not start with zero and may show only that part of the scale which interests us.

Representation of Functions In the equation:

$$f = \frac{300{,}000}{\lambda}$$

f is said to be a function of λ. For every value of f there is a definite value of λ. A variable is said to be a function of another variable when for every possible value of the latter, or *independent* variable, there is a definite value of the first or *dependent* variable. For instance, if $y = 5x^2$, y is a function of x and x is called the independent variable. When $a = 3b^3 + 5b^2 - 25b + 6$ then a is a function of b.

A function can be illustrated in our coordinate system as follows. Let us take the equation for frequency versus wavelength as an example. Given different values to the independent variable find the corresponding values of the dependent variable. Then plot the *points* represented by the different sets of two values.

$f_{kc.}$	λ_{meters}
600	500
800	375
1000	300
1200	250
1400	214
1600	187
1800	167
2000	150

Plotting these points in Figure 31 and drawing a smooth curve through them gives us the *curve* or *graph* of the equation. This curve will help us find values of f for other values of λ (those in between the points calculated) and so a curve of an often-used equation may serve better than a table which always has gaps.

When using the coordinate system described so far and when measuring linearly along both axes, there are some definite rules regarding

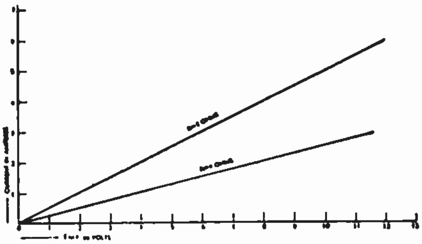


Figure 32.

Only two points are needed to define functions which result in a straight line as shown in this diagram representing Ohm's Law.

the kind of curve we get for any type of equation. In fact, an expert can draw the curve with but a very few plotted points since the equation has told him what kind of curve to expect.

First, when the equation can be reduced to the form y = mx + b, where *x* and *y* are the variables, it is known as a *linear* or *first degree* function and the curve becomes a straight line. (Mathematicians still speak of a "curve" when it has become a straight line.)

When the equation is of the second degree, that is, when it contains terms like x^2 or y^2 or *xy*. the graph belongs to a group of curves, called *conic sections*. These include the circle, the ellipse, the parabola and the hyperbola. In the example given above, our equation is of the form

xy = c, c being equal to 300,000

which is a second degree equation and in this case, the graph is a hyperbola.

This type of curve does not lend itself readily for the purpose of calculation except near the middle, because at the ends a very large change in λ represents a small change in *f* and vice versa. Before discussing what can be done about this let us look at some other types of curves.

Suppose we have a resistance of 2 ohms and we plot the function represented by Ohm's Law: $E = 2I$. Measuring E along the X-axis and amperes along the Y-axis, we plot the necessary points. Since this is a first degree equation, of the form $y = mx + b$ (for $E = y$, $m = 2$ and $I = x$ and $b = 0$) it will be a straight line so we need only two points to plot it.

	I	E
(line passes through origin)	0	0
	5	10

The line is shown in Figure 32. It is seen to be a straight line passing through the origin.

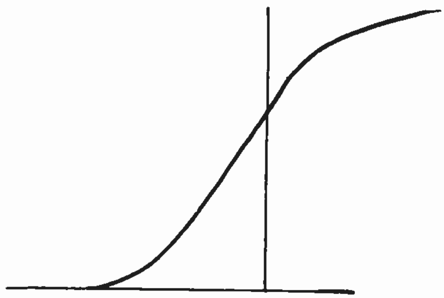


Figure 33.

A TYPICAL GRID-VOLTAGE PLATE-CURRENT CHARACTERISTIC CURVE.

The equation represented by such a curve is so complicated that we do not use it. Data for such a curve is obtained experimentally, and intermediate values can be found with sufficient accuracy from the curve.

If the resistance were 4 ohms, we should get the equation E = 4I and this also represents a line which we can plot in the same figure. As we see, this line also passes through the origin but has a different slope. In this illustration the slope defines the resistance and we could make a protractor which would convert the angle into ohms. This fact may seem inconsequential now, but use of this is made in the drawing of loadlines on tube curves.

Figure 33 shows a typical, grid-voltage, plate-current static characteristic of a triode. The equation represented by this curve is rather complicated so that we prefer to deal with the curve. Note that this curve extends through the first and second quadrant.

Families of curves. It has been explained that curves in a plane can be made to illustrate the relation between *two* variables when one of them varies independently. However, what are we going to do when there are *three* variables and *two* of them vary independently. It is possible to use three dimensions and three axes but this is not conveniently done. Instead of this we may use a *family of curves*. We have already illustrated this partly with Ohm's Law. If we wish to make a chart which will show the current through *any* resistance with *any* voltage applied across it, we must take the equation E = IR, having three variables.

We can now draw one line representing a resistance of 1 ohm, another line representing 2 ohms, another representing 3 ohms, etc., or as many as we wish and the size of our paper will allow. The whole set of lines is then applicable to any case of Ohm's Law falling within the range of the chart. If any two of the three quantities are given, the third can be found.

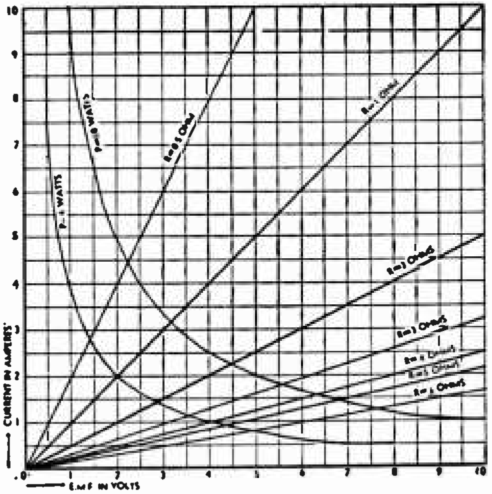


Figure 34.

A FAMILY OF CURVES.

An equation such as Ohm's Law has three variables, but can be represented in Cartesian coordinates by a family of curves such as shown here. If any two quantities are given, the third can be found. Any point in the chart represents a definite value each of E, I, and R, which will satisfy the equation of Ohm's Law. Values of R not situated on an R line can be found by interpolation.

AVERAGE PLATE CHARACTERISTICS

$E_f = 6.3$ v.

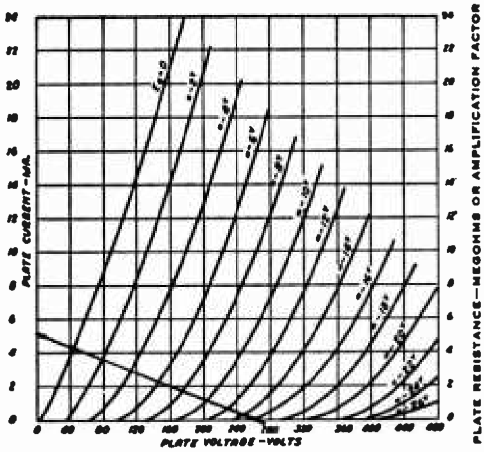


Figure 35.

"PLATE" CURVES FOR A TYPICAL VACUUM TUBE.

In such curves we have three variables, plate voltage, plate current, and grid bias. Each point on a grid bias line corresponds to the plate voltage and plate current represented by its position with respect to the X and Y axes. Those for other values of grid bias may be found by interpolation. The loadline shown in the lower left portion of the chart is explained in the text.

Figure 34 shows such a family of curves to solve Ohm's Law. Any point in the chart represents a definite value each of E, I, and R which will satisfy the equation. The value of R represented by a point that is not situated on an R line can be found by interpolation.

It is even possible to draw on the same chart a second family of curves, representing a fourth variable. But this is not always possible, for among the four variables there should be no more than *two independent variables.* In our example such a set of lines could represent power in watts; we have drawn only two of these but there could of course be as many as desired. A single point in the plane now indicates the four values of E, I, R, and P which belong together and the knowledge of any two of them will give us the other two by reference to the chart.

Another example of a family of curves is the dynamic transfer characteristic or *plate family* of a tube. Such a chart consists of several curves showing the relation between plate voltage, plate current, and grid bias of a tube. Since we have again three variables, we must show several curves, each curve for a fixed value of one of the variables. It is customary to plot plate voltage along the X-axis, plate current along the Y-axis, and to make different curves for various values of grid bias. Such a set of curves is illustrated in Figure 35. Each point in the plane is defined by three values, which belong together, plate voltage, plate current, and grid voltage.

Now consider the diagram of a resistance-coupled amplifier in Figure 36. Starting with the B-supply voltage, we know that whatever plate current flows must pass through the resistor and will conform to Ohm's Law. The voltage drop across the resistor is subtracted from the plate supply voltage and the remainder is the actual voltage at the plate, the kind that is plotted along the X-axis in Figure 35. We can now plot on the plate family of the

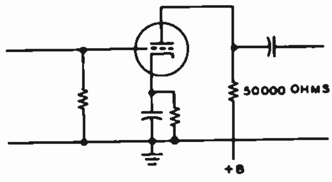


Figure 36.

PARTIAL DIAGRAM OF A RESISTANCE COUPLED AMPLIFIER.

The portion of the supply voltage wasted across the 50,000-ohm resistor is represented in Figure 35 as the loadline.

tube the *loadline*, that is the line showing which part of the plate supply voltage is across the resistor and which part across the tube for any value of plate current. In our example, let us suppose the plate resistor is 50,000 ohms. Then, if the plate current were zero, the voltage drop across the resistor would be zero and the full plate supply voltage is across the tube. Our first point of the loadline is $E = 250$, $I = 0$. Next, suppose, the plate current were 1 ma., then the voltage drop across the resistor would be 50 volts, which would leave for the tube 200 volts. The second point of the loadline is then $E = 200$, $I = 1$. We can continue like this but it is unnecessary for we shall find that it is a straight line and two points are sufficient to determine it.

This loadline shows at a glance what happens when the grid-bias is changed. Although there are many possible combinations of plate voltage, plate current, and grid bias, we are now restricted to points along this line as long as the 50,000 ohm plate resistor is in use. This line therefore shows the voltage drop across the tube as well as the voltage drop across the load for every value of grid bias. Therefore, if we know how much the grid bias varies, we can calculate the amount of variation in the plate voltage and plate current, the amplification, the power output, and the distortion.

Logarithmic Scales Sometimes it is convenient to measure along the axes the *logarithms* of our variable quantities. Instead of actually calculating the logarithm, special paper is available with logarithmic scales, that is, the distances measured along the axes are proportional to the logarithms of the numbers marked on them rather than to the numbers themselves.

There is semi-logarithmic paper, having logarithmic scales along one axis only, the other scale being linear. We also have full logarithmic paper where both axes carry logarithmic scales. Many curves are greatly simplified and some become straight lines when plotted on this paper.

As an example let us take the wavelength-frequency relation, charted before on straight cross-section paper.

$$f = \frac{300,000}{\lambda}$$

Taking logarithms:

$$\log f = \log 300,000 - \log \lambda$$

If we plot $\log f$ along the Y-axis and $\log \lambda$ along the X-axis, the curve becomes a straight line. Figure 37 illustrates this graph on full logarithmic paper. The graph may be read with the same accuracy at any point in contrast to the graph made with linear coordinates.

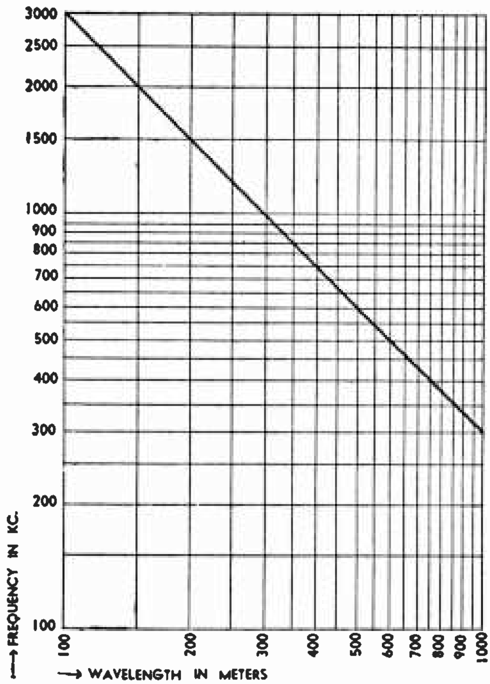


Figure 37.

A LOGARITHMIC CURVE.

Many functions become greatly simplified and some become straight lines when plotted to logarithmic scales such as shown in this diagram. Here the frequency versus wavelength curve of Figure 31 has been replotted to conform with logarithmic axes. Note that it is only necessary to calculate two points in order to determine the "curve" since this type of function results in a straight line.

This last fact is a great advantage of logarithmic scales in general. It should be clear that if we have a linear scale with 100 small divisions numbered from 1 to 100, and if we are able to read to one tenth of a division, the possible error we can make near 100, way up the scale, is only 1/10th of a percent. But near the beginning of the scale, near 1, one tenth of a division amounts to 10 percent of 1 and we are making a 10 percent error.

In any logarithmic scale, our possible error in measurement or reading might be, say 1/32 of an inch which represents a fixed amount of the log depending on the scale used. The net result of adding to the logarithm a fixed quantity, as 0.01, is that the anti-logarithm is multiplied by 1.025, or the error is 2½%. No matter at what part of the scale the 0.01 is added, the error is always 2½%.

An example of the advantage due to the use

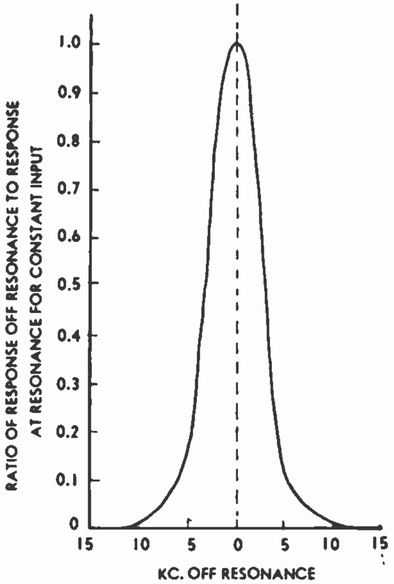


Figure 38.
A RECEIVER RESONANCE CURVE.
This curve represents the output of a receiver versus frequency when plotted to linear coordinates.

of semi-logarithmic paper is shown in Figures 38 and 39. A resonance curve, when plotted on linear coordinate paper will look like the curve in Figure 38. Here we have plotted the output of a receiver against frequency while the applied voltage is kept constant. It is the kind of curve a "wobbulator" will show. The curve does not give enough information in this form for one might think that a signal 10 kc. off resonance would not cause any current at all and is tuned out. However, we frequently have off resonance signals which are 1000 times as strong as the desired signal and one cannot read on the graph of Figure 38 how much any signal is attenuated if it is reduced more than about 20 times.

In comparison look at the curve of Figure 39. Here the response (the current) is plotted in logarithmic proportion, which allows us to plot clearly how far off resonance a signal has to be to be reduced 100, 1,000, or even 10,000 times.

Note that this curve is now "upside down"; it is therefore called a *selectivity* curve. The reason that it appears upside down is that the method of measurement is different. In a selectivity curve we plot the increase in signal voltage necessary to cause a standard output off resonance. It is also possible to plot this increase along the Y-axis in decibels; the curve then looks the same although linear paper can

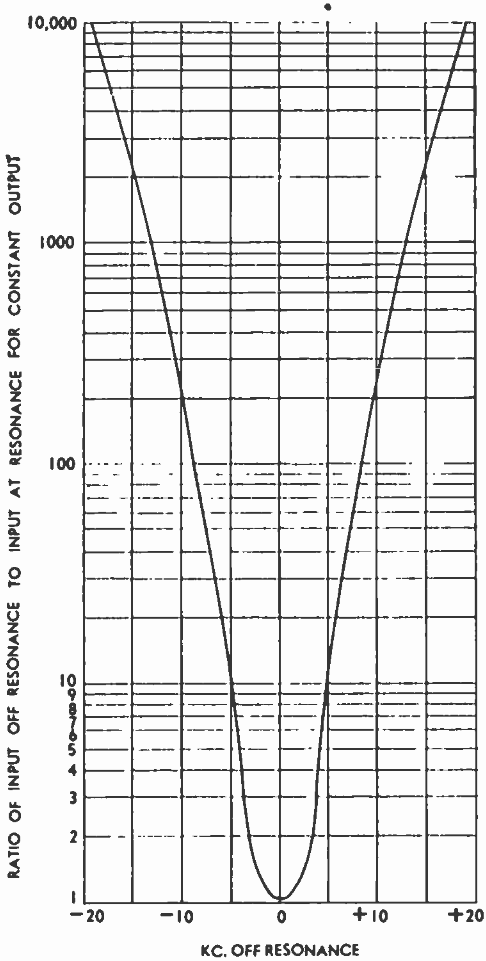


Figure 39.
A RECEIVER SELECTIVITY CURVE.
This curve represents the selectivity of a receiver plotted to logarithmic coordinates for the output, but linear coordinates for frequency. The reason that this curve appears inverted from that of Figure 38 is explained in the text.

be used because now our unit is logarithmic.

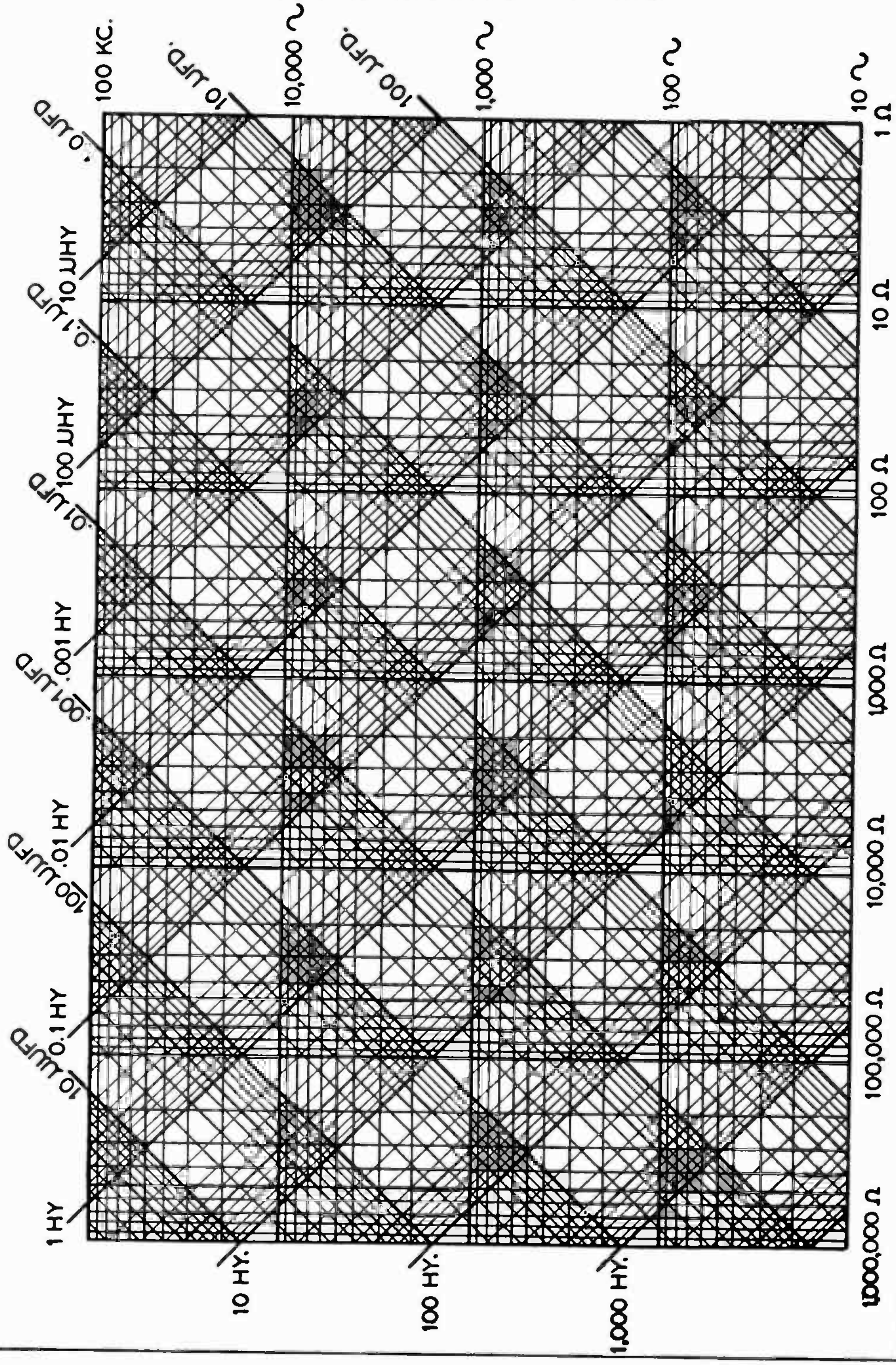
An example of full logarithmic paper being used for families of curves is shown in the reactance charts of Figures 40 and 41.

Nomograms or Alignment Charts An alignment chart consists of three or more sets of scales which have been so laid out that to solve the formula for which the chart was made, we have but to lay a straight edge along the two given values on any two of the scales, to find the third and unknown value on the third scale. In its sim-

Figure 40.

REACTANCE-FREQUENCY CHART FOR AUDIO FREQUENCIES

See text for applications and instructions for use.



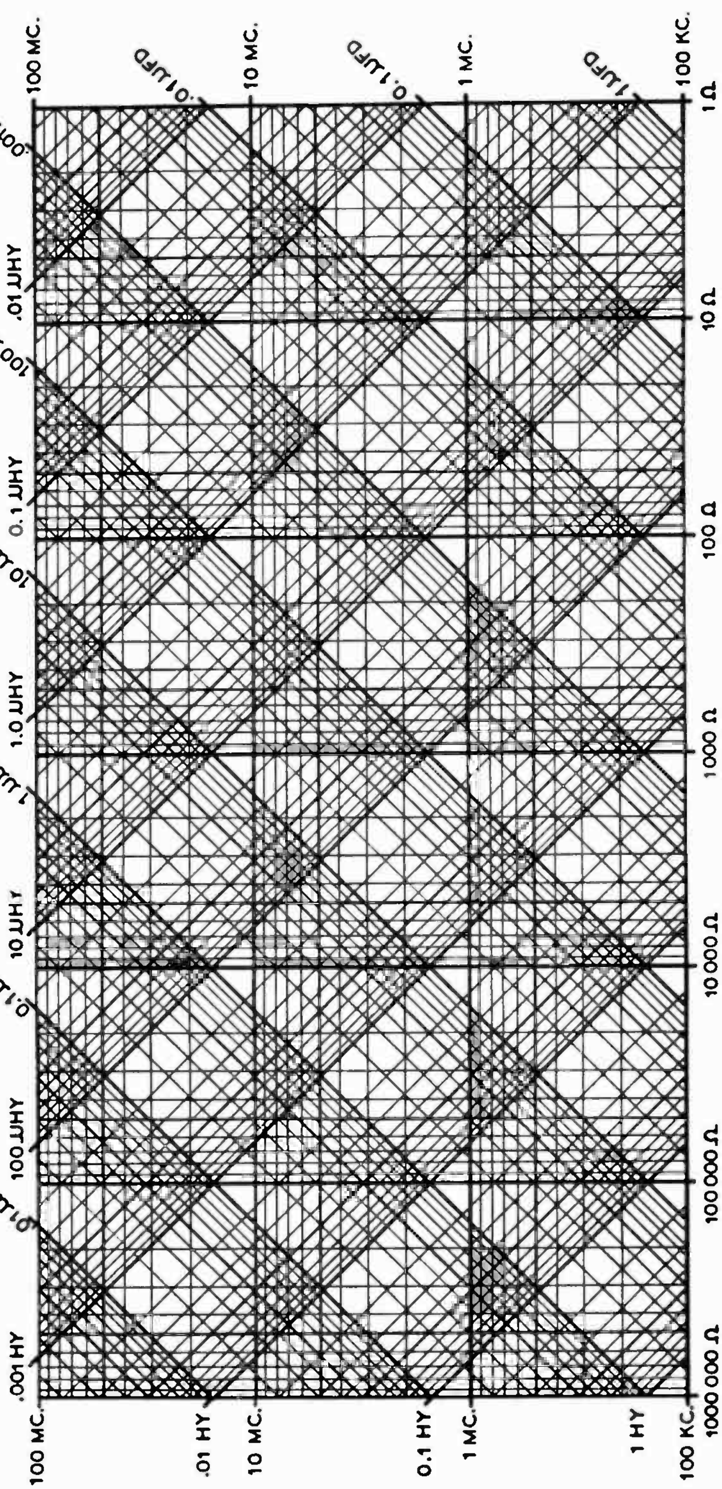


Figure 41.

REACTANCE-FREQUENCY CHART FOR 100 KC. TO 100 MC.

This chart is used in conjunction with the nomograph (Figure 45) for radio frequency tank coil computations.

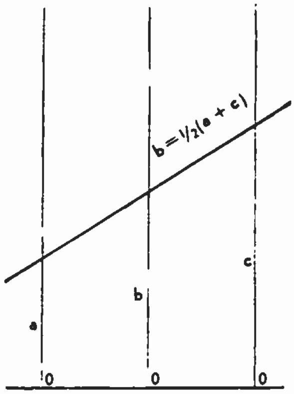


Figure 42.
THE SIMPLEST FORM OF NOMOGRAM.

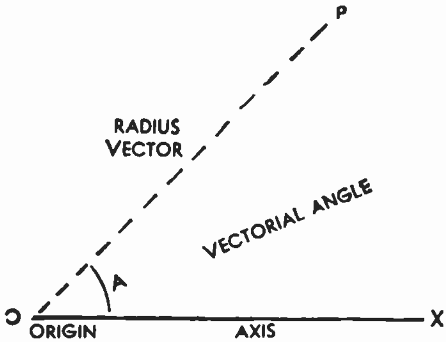


Figure 43.
THE LOCATION OF A POINT BY POLAR COORDINATES.

In the polar coordinate system any point is determined by its distance from the origin and the angle formed by a line drawn from it to the origin and the 0-X axis.

plest form, it is somewhat like the lines in Figure 42. If the lines *a*, *b*, and *c* are parallel and equidistant, we know from ordinary geometry, that b = ½ (a + c). Therefore, if we draw a scale of the same units on all three lines, starting with zero at the bottom, we know that by laying a straight-edge across the chart at any place, it will connect values of *a*, *b*, and *c*, which satisfy the above equation. When any two quantities are known, the third can be found.

If, in the same configuration we used logarithmic scales instead of linear scales, the relation of the quantities would become

$$\log b = \tfrac{1}{2} (\log a + \log c) \text{ or } b = \sqrt{ac}$$

By using different kinds of scales, different units, and different spacings between the scales, charts can be made to solve many kinds of equations.

If there are more than three variables it is generally necessary to make a double chart, that is, to make the result from the first chart serve as the given quantity of the second one. Such an example is the chart for the design of coils illustrated in Figure 45. This nomogram is used to convert the inductance in microhenries to physical dimensions of the coil and vice versa. A pin and a straight edge are required. The method is shown under "R. F. Tank Circuit Calculations" later in this chapter.

Polar Coordinates Instead of the Cartesian coordinate system there is also another system for defining algebraically the location of a point or line in a plane. In this, the polar coordinate system, a point is determined by its distance from the origin, *O*, and by the angle it makes with the axis *O-X*. In Figure 43 the point *P* is defined by the length of *OP*, known as the radius vector and by the angle *A* the vectorial angle. We give these data in the following form

$$P = 3 \angle 60^\circ$$

Polar coordinates are used in radio chiefly for the plotting of directional properties of microphones and antennas. A typical example of such a directional characteristic is shown in Figure 44. The radiation of the antenna represented here is proportional to the distance of the characteristic from the origin for every possible direction.

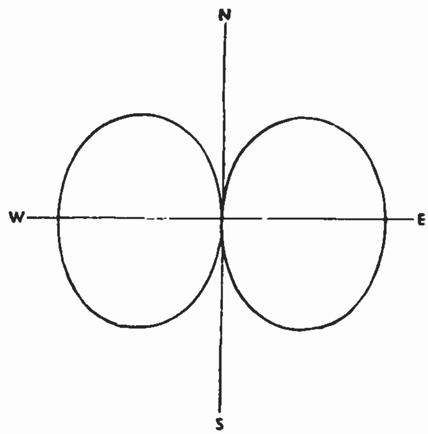


Figure 44.
THE RADIATION CURVE OF AN ANTENNA.

Polar coordinates are used principally in radio work for plotting the directional characteristics of an antenna where the radiation is represented by the distance of the curve from the origin for every possible direction.

Reactance Calculations

In audio frequency calculations, an accuracy to better than a few per cent is seldom required, and when dealing with calculations involving inductance, capacitance, resonant frequency, etc., it is much simpler to make use of reactance-frequency charts such as those in figures 40 and 41 rather than to wrestle with a combination of unwieldy formulas. From these charts it is possible to determine the reactance of a condenser or coil if the capacitance or inductance is known, and vice versa. It follows from this that resonance calculations can be made directly from the chart, because resonance simply means that the inductive and capacitive reactances are equal. The capacity required to resonate with a given inductance, or the inductance required to resonate with a given capacity, can be taken directly from the chart.

While the chart may look somewhat formidable to one not familiar with charts of this type, its application is really quite simple, and can be learned in a short while. The following example should clarify its interpretation.

For instance, following the lines to their intersection, we see that 0.1 hy. and 0.1 μfd. intersect at approximately 1,500 cycles and 1,000 ohms. Thus, the reactance of either the coil or condenser taken alone is about 1000 ohms, and the resonant frequency about 1,500 cycles.

To find the reactance of 0.1 hy. at, say, 10,000 cycles, simply follow the inductance line diagonally up towards the upper left till it intersects the horizontal 10,000 kc. line. Following vertically downward from the point of intersection, we see that the reactance at this frequency is about 6000 ohms.

To facilitate use of the chart and to avoid errors, simply keep the following in mind: The vertical lines indicate reactance in ohms, the horizontal lines always indicate the frequency, the diagonal lines sloping to the lower right represent inductance, and the diagonal lines sloping toward the lower left indicate capacitance. Also remember that the scale is *logarithmic*. For instance, the next horizontal line above 1000 cycles is *2000* cycles. Note that there are 9, not 10, divisions between the heavy lines. This also should be kept in mind when interpolating between lines when best possible accuracy is desired; halfway between the line representing 200 cycles and the line representing 300 cycles is *not* 250 cycles, but approximately 230 cycles. The 250 cycle point is approximately 0.7 of the way between the 200 cycle line and the 300 cycle line, rather than halfway between.

Use of the chart need not be limited by the physical boundaries of the chart. For instance, the 10-$\mu\mu$fd. line can be extended to find where it intersects the 100-hy. line, the resonant frequency being determined by projecting the intersection horizontally back on to the chart. To determine the reactance, the logarithmic ohms scale must be extended.

R. F. Tank Circuit Calculations When winding coils for use in radio receivers and transmitters, it is desirable to be able to determine in advance the full coil specifications for a given frequency. Likewise, it often is desired to determine how much capacity is required to resonate a given coil so that a suitable condenser can be used.

Fortunately, extreme accuracy is not required, except where fixed capacitors are used across the tank coil with no provision for trimming the tank to resonance. Thus, even though it may be necessary to estimate the stray circuit capacity present in shunt with the tank capacity, and to take for granted the likelihood of a small error when using a chart instead of the formula upon which the chart was based, the results will be sufficiently accurate in most cases, and in any case give a reasonably close point from which to start "pruning."

The inductance required to resonate with a certain capacitance is given in the chart in figure 41. By means of the r.f. chart, the inductance of the coil can be determined, or the capacitance determined if the inductance is known. When making calculations, be sure to allow for stray circuit capacity, such as tube interelectrode capacity, wiring, sockets, etc. This will normally run from 5 to 25 micromicrofarads, depending upon the components and circuit.

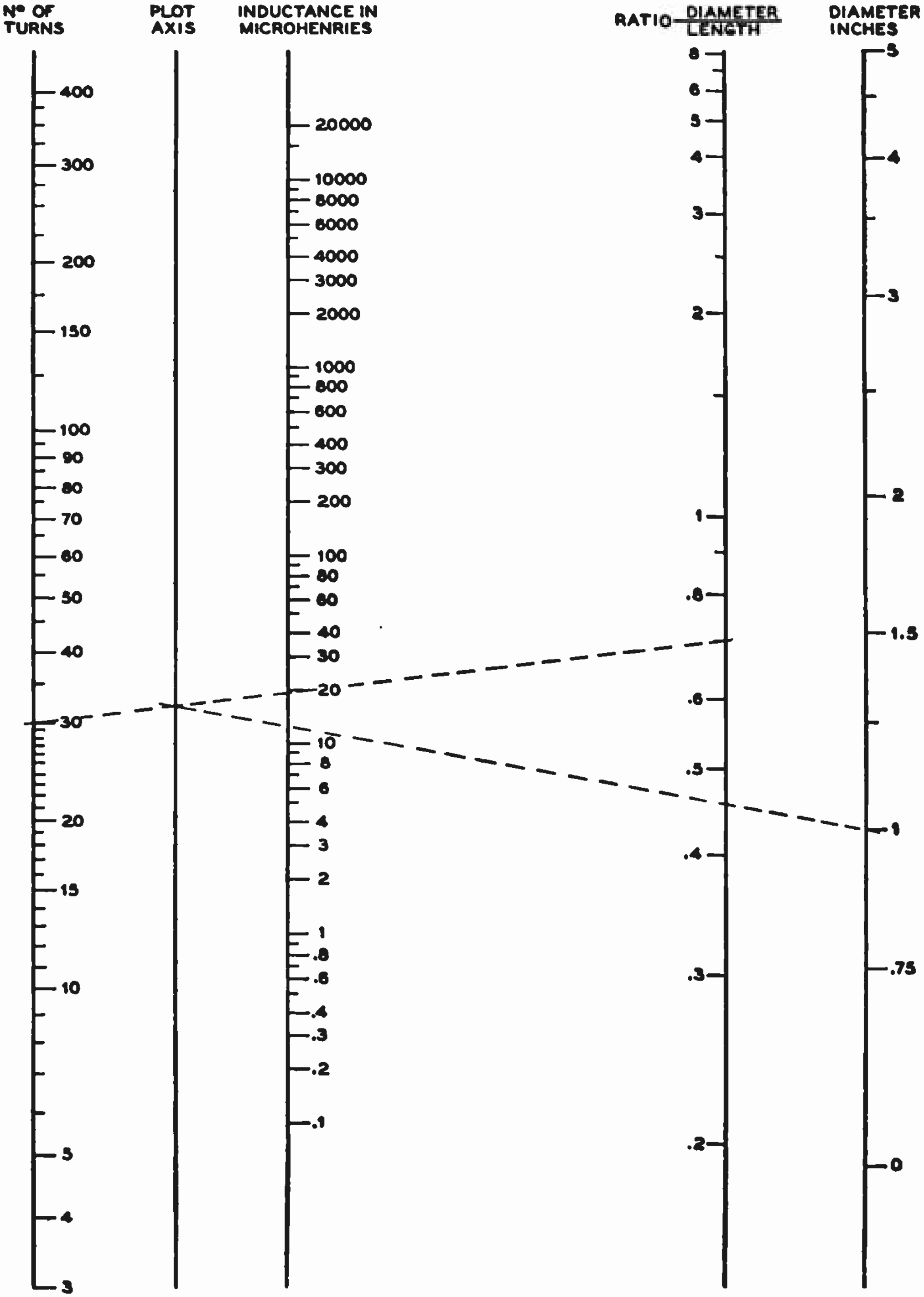
To convert the inductance in microhenries to physical dimensions of the coil, or vice versa, the nomograph chart in figure 45 is used. A pin and a straightedge are required. The inductance of a coil is found as follows:

The straightedge is placed from the correct point on the turns column to the correct point on the diameter-to-length ratio column, the latter simply being the diameter divided by the length. Place the pin at the point on the plot axis column where the straightedge crosses it. From this point lay the straightedge to the correct point on the diameter column. The point where the straightedge intersects the inductance column will give the inductance of the coil.

From the chart, we see that a 30 turn coil having a diameter-to-length ratio of 0.7 and a diameter of 1 inch has an inductance of approximately 12 microhenries. Likewise any one of the four factors may be determined if the other three are known. For instance, to determine the number of turns when the desired in-

Figure 45. COIL CALCULATOR NOMOGRAPH

For single layer solenoid coils, any wire size. See text for instructions.



ductance, the D/L ratio, and the diameter are known, simply work backwards from the example given. In all cases, remember that the straightedge reads either turns and D/L ratio, *or* it reads inductance and diameter. It can read no other combination.

The actual wire size has negligible effect upon the calculations for commonly used wire sizes (no. 10 to no. 30). The number of turns of insulated wire that can be wound per inch (solid) will be found in a copper wire table.

Significant Figures

In most radio calculations, numbers represent quantities which were obtained by measurement. Since no measurement gives absolute accuracy, such quantities are only approximate and their value is given only to a few significant figures. In calculations, these limitations must be kept in mind and one should not finish for instance with a result expressed in more significant figures than the given quantities at the beginning. This would imply a greater accuracy than actually was obtained and is therefore misleading, if not ridiculous.

An example may make this clear. Many ammeters and voltmeters do not give results to closer than ¼ ampere or ¼ volt. Thus if we have 2¼ amperes flowing in a d.c. circuit at 6¾ volts, we can obtain a theoretical answer by multiplying 2.25 by 6.75 to get 15.1875 watts. But it is misleading to express the answer down to a ten-thousandth of a watt when the original measurements were only good to ¼ ampere or volt. The answer should be expressed as 15 watts, not even 15.0 watts. If we assume a possible error of ⅛ volt or ampere (that is, that our original data are only correct to the *nearest* ¼ volt or ampere) the true power lies between 14.078 (product of 2⅛ and 6⅝) and 16.328 (product of 2⅜ and 6⅞). Therefore, any third significant figure would be misleading as implying an accuracy which we do not have.

Conversely, there is also no point to calculating the value of a part down to 5 or 6 significant figures when the actual part to be used cannot be measured to better than 1 part in one hundred. For instance, if we are going to use 1% resistors in some circuit, such as an ohmmeter, there is no need to calculate the value of such a resistor to 5 places, such as 1262.5 ohm. Obviously, 1% of this quantity is over 12 ohms and the value should simply be written as 1260 ohms.

There is a definite technique in handling these approximate figures. When giving values obtained by measurement, no more figures are given than the accuracy of the measurement permits. Thus, if the measurement is good to two places, we would write, for instance, 6.9 which would mean that the true value is somewhere between 6.85 and 6.95. If the measurement is known to three significant figures, we might write 6.90 which means that the true value is somewhere between 6.895 and 6.905. In dealing with approximate quantities, the added cipher at the right of the decimal point has a meaning.

There is unfortunately no standardized system of writing approximate figures with many ciphers to the left of the decimal point. *69000* does not necessarily mean that the quantity is known to 5 significant figures. Some indicate the accuracy by writing 69×10^3 or 690×10^2 etc., but this system is not universally employed. The reader can use his own system, but whatever notation is used, the number of significant figures should be kept in mind.

Working with approximate figures, one may obtain an idea of the influence of the doubtful figures by marking all of them, and products or sums derived from them. In the following example, the doubtful figures have been underlined.

```
 603
  34.6
   0.120
 -------
 637.720    answer: 638
```

Multiplication:

```
   654                        654
   0.342                      0.342
  ------                     ------
    1308                     196|2
   2616                       26|16
  1962                         1|308
 -------                     ------
 223.668   answer: 224       224
```

It is recommended that the system at the right be used and that the figures to the right of the vertical line be omitted or guessed so as to save labor. Here the partial products are written in the reverse order, the most important ones first.

In division, labor can be saved when after each digit of the quotient is obtained, one figure of the divisor be dropped. Example:

```
           1.28
 527 ) 673
       527
 53  ) 146
       106
 5   )  40
        40
```

Appendix

STANDARD COLOR CODE—RESISTORS AND CAPACITORS

AXIAL LEAD RESISTOR

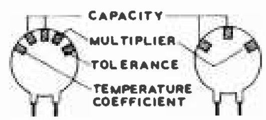
BROWN—INSULATED
BLACK—NON-INSULATED

WIRE-WOUND RESISTORS HAVE 1ST DIGIT BAND DOUBLE WIDTH.

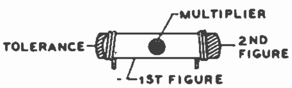
INSULATED UNINSULATED *COLOR*	FIRST RING BODY COLOR *FIRST FIGURE*	SECOND RING END COLOR *SECOND FIGURE*	THIRD RING DOT COLOR *MULTIPLIER*
BLACK	0	0	NONE
BROWN	1	1	0
RED	2	2	00
ORANGE	3	3	,000
YELLOW	4	4	0,000
GREEN	5	5	00,000
BLUE	6	6	000,000
VIOLET	7	7	0,000,000
GRAY	8	8	00,000,000
WHITE	9	9	000,000,000

DISC CERAMIC RMA CODE

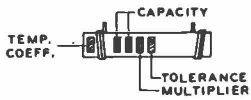
5-DOT 3-DOT



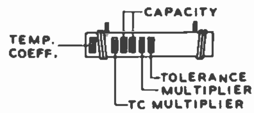
RADIAL LEAD DOT RESISTOR



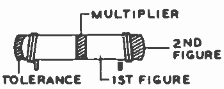
5-DOT RADIAL LEAD CERAMIC CAPACITOR



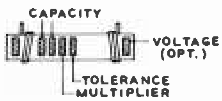
EXTENDED RANGE TC CERAMIC HICAP



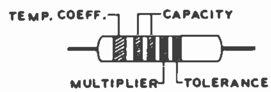
RADIAL LEAD (BAND) RESISTOR



BY-PASS COUPLING CERAMIC CAPACITOR

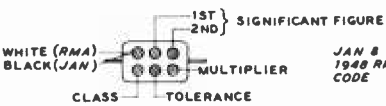


AXIAL LEAD CERAMIC CAPACITOR



MOLDED MICA TYPE CAPACITORS

CURRENT STANDARD CODE

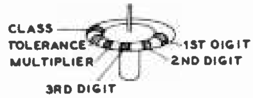


JAN & 1948 RMA CODE

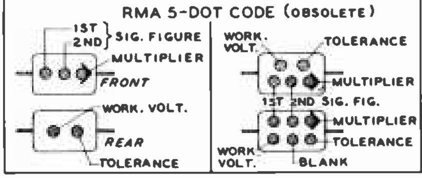
RMA 3-DOT (OBSOLETE)

RATED 500 V.D.C. ± 20% TOL.

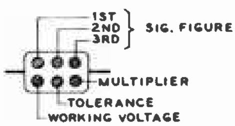
BUTTON SILVER MICA CAPACITOR



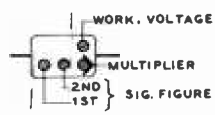
RMA 5-DOT CODE (OBSOLETE)



RMA 6-DOT (OBSOLETE)



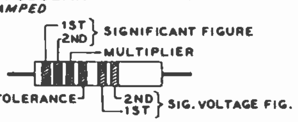
RMA 4-DOT (OBSOLETE)



MOLDED PAPER TYPE CAPACITORS

TUBULAR CAPACITOR

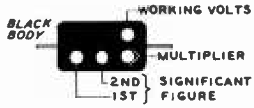
NORMALLY STAMPED FOR VALUE



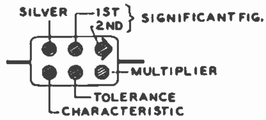
A 2-DIGIT VOLTAGE RATING INDICATES MORE THAN 900 V. ADD 2 ZEROS TO END OF 2 DIGIT NUMBER.

MOLDED FLAT CAPACITOR

COMMERCIAL CODE



JAN CODE CAPACITOR



STANDARD COLOR CODE FOR RESISTORS AND CAPACITORS

The standard code provides the necessary information required to properly identify color coded resistors and capacitors. Refer to the color code for numerical values and the number of zeros (or multiplier) assigned to the colors used. A fourth color band on resistors determines the tolerance rating as follows: Gold = 5%, silver = 10%. Absence of the fourth band indicates a 20% tolerance rating.

Tolerance rating of capacitors is determined by the color code. For example: Red = 2%, green = 5%, etc. The voltage rating of capacitors is obtained by multiplying the color value by 100. For example: Orange = 3 × 100, or 300 volts.

AIRWOUND INDUCTORS

COIL DIA. *INCHES*	TURNS PER INCH	B & W	AIR DUX	INDUCTANCE *µH*
1/2	4	3001	404T	0.18
	6	—	406T	0.40
	8	3002	408T	0.72
	10	—	410T	1.12
	16	3003	416T	2.90
	32	3004	432T	12.0
5/8	4	3005	504T	0.28
	6	—	506T	0.62
	8	3006	508T	1.1
	10	—	510T	1.7
	16	3007	516T	4.4
	32	3008	532T	18.0
3/4	4	3009	604T	0.39
	6	—	606T	0.87
	8	3010	608T	1.57
	10	—	610T	2.45
	16	3011	616T	6.40
	32	3012	632T	26.0
1	4	3013	804T	1.0
	6	—	806T	2.3
	8	3014	808T	4.2
	10	—	810T	6.6
	16	3015	816T	16.8
	32	3016	832T	68.0

COIL DIA. *INCHES*	TURNS PER INCH	B & W	AIR DUX	INDUCTANCE *µH*
1 1/4	4	—	1004	2.75
	6	—	1006	6.30
	8	—	1008	11.2
	10	—	1010	17.5
	16	—	1016	42.5
1 1/2	4	—	1204	3.9
	6	—	1206	8.6
	8	—	1208	15.6
	10	—	1210	24.5
	16	—	1216	63.0
1 3/4	4	—	1404	5.2
	6	—	1406	11.8
	8	—	1408	21.0
	10	—	1410	33.0
	16	—	1416	85.0
2	4	—	1604	6.6
	6	—	1606	15.0
	8	3900	1608	26.5
	10	3907-1	1610	42.0
	16	—	1616	108.0
2 1/2	4	—	2004	10.1
	6	3905-1	2006	23.0
	8	3906-1	2008	41.0
	10	—	2010	108.0
3	4	—	2404	14.0
	6	—	2406	31.5
	8	—	2408	56.0
	10	—	2410	89.0

NOTE:
COIL INDUCTANCE APPROXIMATELY PROPORTIONAL TO LENGTH. I.E., FOR 1/2 INDUCTANCE VALUE, TRIM COIL TO 1/2 LENGTH.

Copper Wire Table

Gauge No. B. & S.	Diam. in Mils[1]	Circular Mil Area	Turns per Linear Inch[2]				Turns per Square Inch[2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Correct Capacity at 1500 C.M. per Amp.[3]	Diam. in mm.
			Enamel	S.S.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel	D.C.C.	Bare	D.C.C.			
1	289.3	82690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189
5	181.9	33100	—	—	—	—	—	-	—	9.980	—	.3195	22.0	4.621
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	6.9	2.588
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	1.3	1.150
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	1.1	1.024
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733
24	20.10	404.0	46.3	45.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546
31	8.928	79.70	101.	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268
32	7.950	63.21	113.	101.	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019
33	7.080	50.13	127.	110.	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798
34	6.305	39.75	143.	120.	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601
35	5.615	31.52	158.	132.	104.	73.5	10900	8700	—	10480	6737	335.0	.021	.1426
36	5.000	25.00	175.	143.	111.	77.0	12200	10700	—	13210	7877	423.0	.017	.1270
37	4.453	19.83	198.	154.	118.	80.3	—	—	—	16660	9309	533.4	.013	.1131
38	3.965	15.72	224.	166.	126.	83.6	—	—	—	21010	10666	672.6	.010	.1007
39	3.531	12.47	248.	181.	133.	86.6	—	—	—	26500	11907	848.1	.008	.0897
40	3.145	9.88	282.	194.	140.	89.7	—	—	—	33410	14222	1069	.006	.0799

[1] A mil is 1/1000 (one thousandth) of an inch.
[2] The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.
[3] The current-carrying capacity at 1000 C.M. per ampere is equal to the circular-mil area (Column 3) divided by 1000.

CONVERSION TABLE — UNITS OF MEASUREMENT

MICRO = (μ) ONE-MILLIONTH — KILO = (K) ONE THOUSAND

MILLI = (m) ONE-THOUSANDTH — MEGA = (M) ONE MILLION

TO CHANGE FROM	TO	OPERATOR
UNITS	MICRO-UNITS	$\times$ 1,000,000 or $\times 10^6$
	MILLI-UNITS	$\times$ 1,000 or $\times 10^3$
	KILO-UNITS	$\div$ 1,000 or $\times 10^{-3}$
	MEGA-UNITS	$\div$ 1,000,000 or $\times 10^{-6}$
MICRO-UNITS	MILLI-UNITS	$\div$ 1,000 or $\times 10^{-3}$
	UNITS	$\div$ 1,000,000 or $\times 10^{-6}$
MILLI-UNITS	MICRO-UNITS	$\times$ 1,000 or $\times 10^3$
	UNITS	$\div$ 1,000 or $\times 10^{-3}$
KILO-UNITS	MEGA-UNITS	$\div$ 1,000 or $\times 10^{-3}$
	UNITS	$\times$ 1,000 or $\times 10^3$
MEGA-UNITS	KILO-UNITS	$\times$ 1,000 or $\times 10^3$
	UNITS	$\times$ 1,000,000 or $\times 10^6$

COMPONENT COLOR CODING

POWER TRANSFORMERS

PRIMARY LEADS	*BLACK*
IF TAPPED:	
COMMON	*BLACK*
TAP	*BLACK/YELLOW*
END	*BLACK/RED*
HIGH VOLTAGE WINDING	*RED*
CENTER-TAP	*RED/YELLOW*
RECTIFIER FILAMENT WINDING	*YELLOW*
CENTER-TAP	*YELLOW/BLUE*
FILAMENT WINDING Nº 1	*GREEN*
CENTER-TAP	*GREEN/YELLOW*
FILAMENT WINDING Nº 2	*BROWN*
CENTER-TAP	*BROWN/YELLOW*
FILAMENT WINDING Nº 3	*SLATE*
CENTER-TAP	*SLATE/YELLOW*

I-F TRANSFORMERS

PLATE LEAD	*BLUE*
B+ LEAD	*RED*
GRID (*OR DIODE*) LEAD	*GREEN*
A-V-C (*OR GROUND*) LEAD	*BLACK*

AUDIO TRANSFORMERS

PLATE LEAD (*PRI.*)	*BLUE OR BROWN*
B+ LEAD (*PRI.*)	*RED*
GRID LEAD (*SEC.*)	*GREEN OR YELLOW*
GRID RETURN (*SEC.*)	*BLACK*

Index

INDEX