The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION

A R R L

1945 EDITION

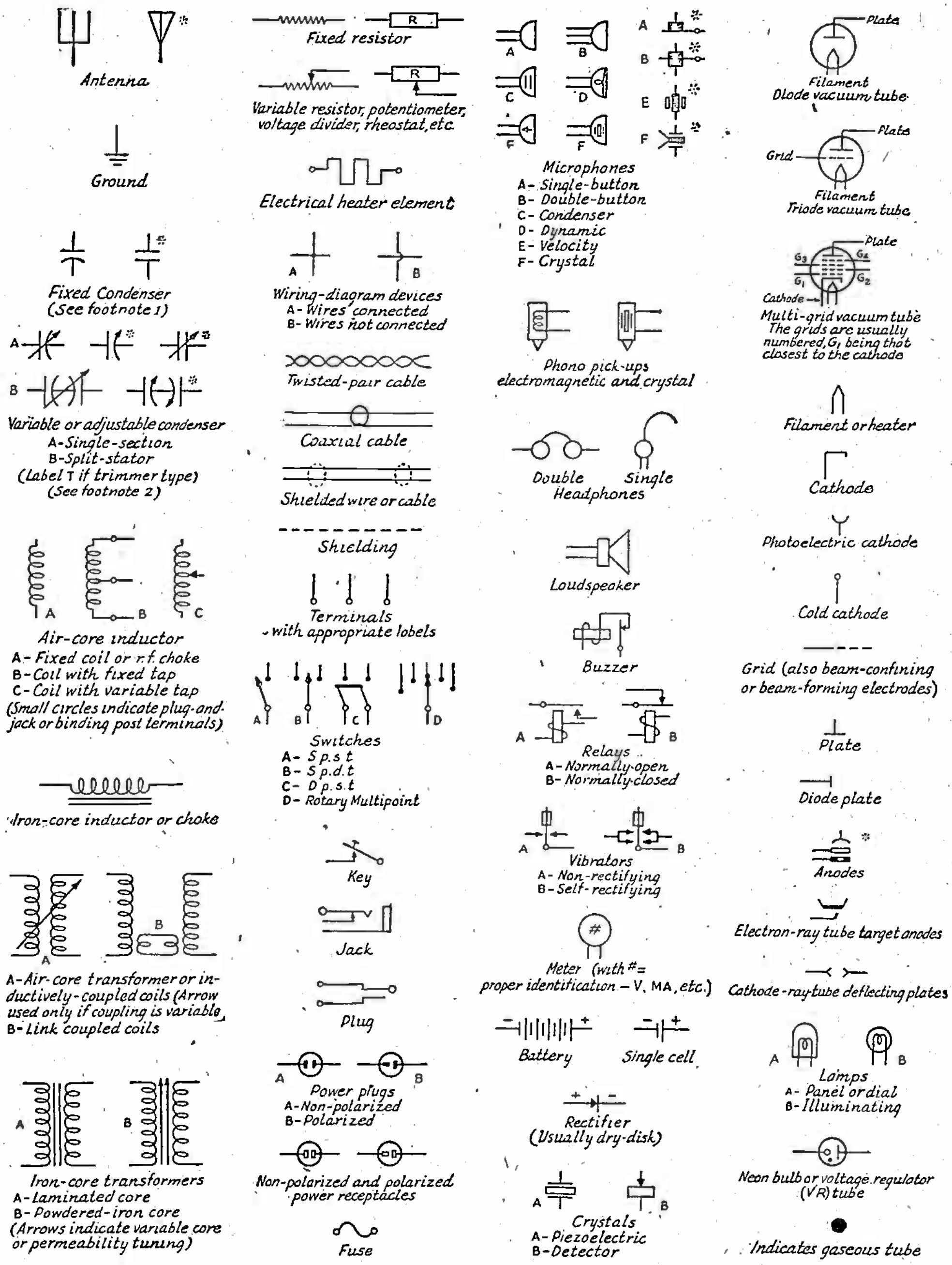
$1 IN CONTINENTAL U.S.A.

PUBLISHED BY

THE AMERICAN RADIO RELAY LEAGUE

THE RADIO AMATEUR'S HANDBOOK

STANDARD SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



For convenience and simplicity, schematic wiring diagrams employing conventionalized symbols to represent various components, as shown above, are utilized to indicate the circuit connections used in radio apparatus. The symbols used in this *Handbook* follow the new standardized forms adopted by the radio industry under the ASA standardization program in 1944.

Alternative symbols marked with an asterisk are the conventional radio forms used prior to mid-1944. These are included for reference information in instances where the original symbol has undergone appreciable change.

[1] Where it is necessary or desirable to identify the electrodes, the curved element represents the *outside* electrode (marked "outside foil," "ground," etc.) in fixed paper- and ceramic-dielectric condensers, and the *negative* electrode in electrolytic condensers.

[2] In the new symbol, the curved line indicates the moving element (rotor plates) in variable and adjustable air- or mica-dielectric condensers. To distinguish trimmers, the letter "T" should appear adjacent to the symbol.

In the case of switches, jacks, relays, etc., only the basic combinations are shown. Any combination of these symbols may be assembled as required, following the elementary forms shown.

TWENTY-SECOND EDITION
NINETEEN FORTY-FIVE

THE RADIO AMATEUR'S HANDBOOK

BY THE HEADQUARTERS STAFF OF
THE AMERICAN RADIO RELAY LEAGUE

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE, INC.
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Foreword

The *Radio Amateur's Handbook* has had a remarkable history. It was back in early 1926 when we of the headquarters staff of the radio amateur's organization, the American Radio Relay League, conceived a need for a small pamphlet of operating instructions intended to improve the performance of amateur stations. The task of writing this operating manual was, logically, assigned the communications manager of the League, now Lieut. Colonel F. E. Handy of the U. S. Army Air Forces. When it was finished we found that, with the necessary additional material on technical topics, CM Handy had written a considerable-size book. Upon its publication it enjoyed an instant success and quickly became an important factor in the literature of the radio amateur. As successive reprintings were undertaken new material was added by the technical members of the ARRL Headquarters staff; and an evolutionary process began which became the policy of the *Handbook:* to present the kind of information required to get *results*, to avoid the academic intricacies of the classroom, and, as time and experience showed us how, to refine and perfect the explanations so that they were thoroughly helpful to practical amateurs.

Thus the book grew through the years — in size, in value, in acceptance. It became the right-hand guide of practical radio amateurs in every country of the globe. Revised annually in recent years, it was always designed to present not necessarily the newest or novel but rather the best. Our editorial problem in these annual revisions always lay largely in so adjusting its scope and perspective that the book would be of maximum usefulness to the reader in his problems and activities of that particular year. The volume became a comprehensive compendium of the whole art of high-frequency radio — corresponding literally to the designation long carried on its cover: "The Standard Manual of Amateur Radio Communication." It became both a text and a reference book, an engineering-design handbook and a practical "how-to-build-it" manual.

When war came to this nation at the end of 1941, it was discovered by the military and other agencies that a book written as the *Handbook* was written was precisely what was needed to provide exactly the kind of information required to make practical radiomen for the Army and Navy and to help those who were training themselves for wartime radio work. Not only was — and is — the *Handbook* used directly in many training programs, either as basic text or reference, but its information has provided source-data for many of the service-written special courses.

With the *Handbook* doing a tremendous wartime job, with its total distribution soon well over a million copies, we have undertaken each recent revision with a heavy sense of our responsibilities as writers and publishers. In the happier prewar days of active amateur operation the task of annual editorial revision had been chiefly one of selecting the ideas and creating the apparatus designs upon which the greatest operating reliance could be placed, of providing the latest and best "construction" information of which we had knowledge. But with amateur stations silenced and all our people geared to the needs of war, the perspective necessarily shifted sharply. Particularly as we approached the present revision — the twenty-second edition, the forty-fourth printing — we have felt the dual responsibility of continuing the *Handbook's* wartime usefulness and of preparing both the prewar amateur body and the newer members of its ranks for the greatly expanded field of postwar amateur radio.

Thus this edition is what it is — a comprehensive, authentic digest of the older, familiar phases of radio theory and practice, interspersed with forecasts of the more advanced techniques of tomorrow. Many of the theory sections of this *Handbook* will appear with no appreciable revision. In such cases the language employed is, or could have been, that used to describe these same topics ten or twenty years ago. They relate to the basic elements of the game — the game as it once was and will continue to be, in part, at least. But after the war we can anticipate that amateur radio will follow its evolutionary trend, ever broadening the base of its activities. Where radio, to the broadcast listener, once was simply sound from 550 to 1500 kilocycles, he now has a.m. radio, f.m. radio, television and facsimile to contemplate. The amateur, to an even greater degree, will have open to him a wholly new and diversified range of opportunities for technical developments. These we cannot as yet even forecast in this edition of the *Handbook*. The techniques, as applicable by the amateur, have been developed only to a basic degree; all the work remains to be done; but we can and, within the limits imposed, have endeavored to present some of the basic principles. For these are as broadly applicable to phases of these new activities as those referred to above have been to the game we have known for the past ten or twenty years.

This edition, we are proud to state, continues the *Handbook's* unbroken (and, incidentally, unparalleled) record of growth in size and content, with a substantial increase in the total number of pages. This, together with the enlarged page format adopted as a conservation measure in 1944, has enabled the inclusion of the new material without curtailing the treatment of any of the older phases. Continuing in the style of the last few editions, it is divided into two main parts. In the first is grouped all the material treating of principles, theory and design considerations — the enduring basis of the art. In the second part, embodying the practical employment of the basic knowledge presented in the first part, are the examples of practical equipment — in general with at least one representative example of each accepted type or combination — together with essential constructional data and instructions for adjustment and use.

The first ten chapters constitute a textbook on the principles of radio. Basically, they are the culmination of several years of work by George Grammer, technical editor of *QST* now on leave, not only in the writing but in the refining of the presentation in the crucible of actual use — by teaching experimental classes and by surveying the progress of typical self-taught students. The aim has been to write an understandable nonmathematical treatment for busy, practical people of average education. Necessarily compact (as is any good text), information is deliberately presented without sugar-coating, but every effort has been made to make it understandable and to avoid saying things in such a way that they are intelligible only to those who already know the subject! The material has been so arranged in topical sections as to make it readily possible to find what is wanted, a multitude of subheadings identifying subjects at a glance. The information is presented concisely but with copious cross-references, to permit the background always to accompany the subject under consideration. Subjects are arranged in a logical order which can serve as the basis for a well-ordered radio study course. Indeed, Mr. Grammer's companion work, *A Course in Radio Fundamentals*, also published by the League, is written around this portion of the *Handbook*, providing for the student a proved and effective series of study assignments, directions for experiments, and examination questions.

The second part of the book is that which has been dearest to the heart of the practicing amateur. That amateur today may be engaged in rebuilding his station to improve its performance after the war, but much more probably he is working for Uncle Sam — in the armed forces, perhaps at the front but more likely at a key post behind the lines, or in a research or development laboratory. Wherever he is, he and his similars need a reliable guide for the construction of various pieces of radio apparatus. The second part of the *Handbook* deals only with practical considerations, but reference to the first part of the book always will lead the reader quickly to any needed information on the whys thereof. The apparatus designs are the best we know for their respective jobs and they will be found reliable. The classified vacuum-tube tables, always a most important feature of the *Handbook*, have been revised to include data on all released new types, and remain the most comprehensive compilation available. As a special convenience to facilitate locating a tube whose classification is not known, a cross-index by type numbers is provided. For those still on the home front, mention should be made of the revised chapter on carrier-current communication, the alternative field which the amateur has found most interesting and fruitful during the war's restrictions. The chapter on the War Emergency Radio Service has been revised and reorganized, to enable it to provide the best possible guidance to those engaged in this essential service.

Most of the technically skilled specialists on the League's headquarters staff at West Hartford have participated in the preparation of this *Handbook*. The present revision is the work largely of Clinton B. DeSoto, the editor of *QST*; Donald H. Mix, *QST*'s acting technical editor; Hollis M. French and J. Venable Fitzhugh, *QST*'s assistant technical editors, and Walter E. Bradley, of ARRL's Technical Information Service. The production of the book has fallen on the production staff of *QST*, including most particularly Louisa B. Dresser, *QST*'s editorial assistant, Harry Hick (for thirty years our staff draftsman), Elsie Hochmeister, and, in the concluding phase, Assistant Editor A. D. Middelton.

A word about the reference system: It will be noted that each chapter is divided into sections and that these are numbered serially within each chapter. The number takes the form of two digits or groups separated by a hyphen. The first figure is the chapter number, the second the section number within the chapter. Cross-references in the text take such a form as (§ 4-7), for example, which means that the subject referred to will be found discussed in Chapter Four, Section 7. Throughout the book, illustrations are serially numbered within each chapter. Thus Fig. 1107 can be readily identified as the seventh illustration in Chapter Eleven. The carefully prepared index at the rear, in itself one of the most valuable and unique features of the book, constitutes practically a glossary of the entire high-frequency radio art.

In the past year we have heard from many sources of the great utility the *Handbook* has had in the prosecution of the war effort. We earnestly hope that this edition will prove as helpful to its wartime readers as earlier editions have been to the amateurs of peacetime.

KENNETH B. WARNER
Managing Secretary, A.R.R.L.

WEST HARTFORD, CONN.
November, 1944

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THE AMATEUR'S CODE

☆ ☆ ☆

1. *The Amateur is Gentlemanly*

He never knowingly uses the air for his own amusement in such a way as to lessen the pleasure of others. He abides by the pledges given by the ARRL in his behalf to the public and the Government.

2. *The Amateur is Loyal*

He owes his amateur radio to the American Radio Relay League, and he offers it his unswerving loyalty.

3. *The Amateur is Progressive*

He keeps his station abreast of science. It is built well and efficiently. His operating practice is clean and regular.

4. *The Amateur is Friendly*

Slow and patient sending when requested, friendly advice and counsel to the beginner, kindly assistance and coöperation for the broadcast listener; these are marks of the amateur spirit.

5. *The Amateur is Balanced*

Radio is his hobby. He never allows it to interfere with any of the duties he owes to his home, his job, his school, or his community.

6. *The Amateur is Patriotic*

His knowledge and his station are always ready for the service of his country and his community.

☆ ☆ ☆ ☆ ☆ ☆ ☆ ☆ ☆ ☆ ☆ ☆

Chapter One

Introduction to Radio

First experiences always remain most clearly fixed in the mind. Everyone who has ever been active in amateur radio can recall the experience of putting together his first simple receiver — how he proceeded, with doubts in his mind, frequently checking back to the diagrams in the book, trying to understand their meaning. Coils, condensers, fixed and variable resistors, grid leaks, sockets, tubes, transformers, batteries, headphones. Finally, the last connection was soldered; the antenna and ground and the power leads connected. Headphones on, he turned the dial — and then came the first sounds of code, music or the human voice. *It worked!*

This chapter is written especially for the individual who has yet to enjoy the satisfaction that comes from building a radio set that works; for the neophyte who finds himself confused by the array of tubes and other parts hidden within the cabinet of his home broadcast receiver or behind the panel of even the simplest transmitter. There is more behind those panels than a switch, a dial, a loudspeaker and a knob to make the sounds strong or weak, but until certain fundamentals are understood the purpose and functioning of these components can have little significance.

Understanding how and why radio receivers and transmitters work is merely a matter of seeing in one's mind what is happening. It requires little more effort than forming mental pictures of other applications of natural phenomena, such as why airplanes fly or how gasoline engines make wheels go around. The only difference is that the latter are mechanical operations, and therefore more readily visualized.

A very simple explanation of the mechanics of how messages are transmitted and received by means of radio waves may, therefore, help prepare the way for the more detailed discussion of fundamentals which will follow.

1-1 Radio Waves

Communication — the process of conveying intelligence from one point to another — by radio is similar to communication by any other means — wire telephone, wire telegraph, visual signaling or messenger. The difference is that in radio communication the physical means for conveying or carrying the intelligence is a progressive chain of high-frequency radio waves traveling in space, rather than electrical impulses on a wire or other means.

At the start, and also at the end of the communicating process, radio technique is identical with that of wire telephony and telegraphy. Where radio differs from other methods of communication is in the means used to connect the transmitting and receiving points — the invisible electromagnetic fields of radio replacing the metallic strands of the wire system.

The science of radio centers around these electromagnetic fields or waves. In fact, the term radio is derived from "radiation" — the process of propagating or sending the waves into space.

When an electric current flows through a conductor — whether that conductor be part of the house wiring or the extension cord to an electric light bulb or a radio antenna — magnetic lines of force are set up surrounding the wire as a result of the current flow. These lines of force constitute what is known as an electromagnetic field. This field is not detectable by any ordinary human sense, but it does exist — much like the force exerted by a magnet which, although also invisible, can be readily demonstrated.

The energy in the field thus created alternately builds up and subsides with each alternation of the current. If the length of the conductor is great enough and the frequency of alternation slow enough, the energy in the field will have time to return to the conductor before the next reversal occurs and the cycle is repeated. If, however, the frequency or rate of reversal is sufficiently rapid, the field set up by current flow in one direction will not be completely collapsed before it is displaced by a new field from the next cycle. Thus not all of the energy in it can return to the wire.

With each successive cycle portions of additional fields are detached, each following its predecessor away from the wire. Soon there is a continuous series of these detached fields traveling out into space at a rate of nearly 200,000 miles (actually, 300,000,000 meters) per second — the speed of light.

This is the process of radiation, and it occurs in some measure whenever alternating current flows through any conductor. Unless the length of the conductor approaches the wavelength of the alternating current very little actual energy is released, however.

Consider, for example, the wavelength corresponding to 60 cycles per second (the frequency commonly used in electric power transmission). A single wave at this frequency covers 5 million meters, or about 3,100 miles per cycle. A practical radiating circuit of that length obviously being inconceivable, the radiation from any wire carrying 60-cycle current is negligible. On the other hand, an alternating

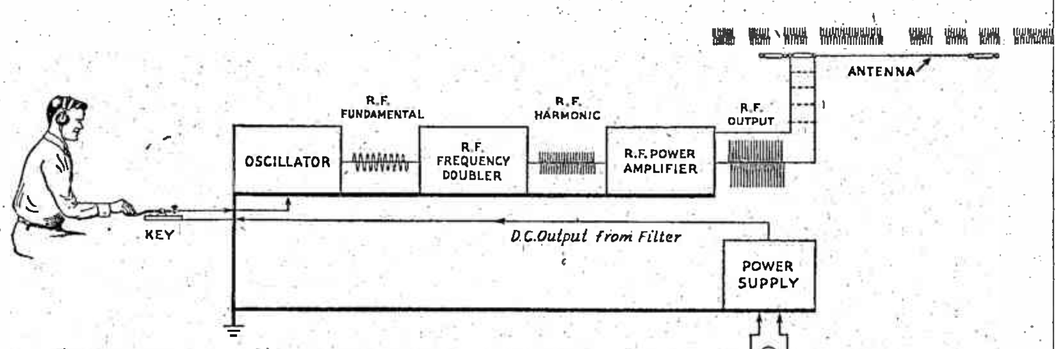


Fig. 101 — Radiotelegraph transmission of continuous-wave (c.w.) signals.

current whose frequency is 60 million cycles per second (60 megacycles) will radiate waves having a length of only 5 meters — and wire lines with dimensions of this order can be readily constructed.

In practice, the frequencies used for radio transmission range from about 15,000 cycles per second up into the billions.

ℂ 1-2 Radio Transmission

The basic elements of any system of radio communication — or for employing radio waves in any other application — are (1) means for generating a radio-frequency carrier (transmitter); (2) means for impressing intelligence upon the carrier (telegraph key; modulator); and (3) means for radiating the modulated carrier wave (transmitting antenna).

At the receiving end there must be (1) a means for intercepting radio waves (receiving antenna); (2) means for selecting the desired wave and amplifying the intercepted energy (receiver); (3) means for converting the carrier modulation into useful form (headphones, loudspeaker, cathode-ray tube, relay, etc.).

Figs. 101 through 104 illustrate the use of these elements in the transmission and reception of voice and code signals by radio.

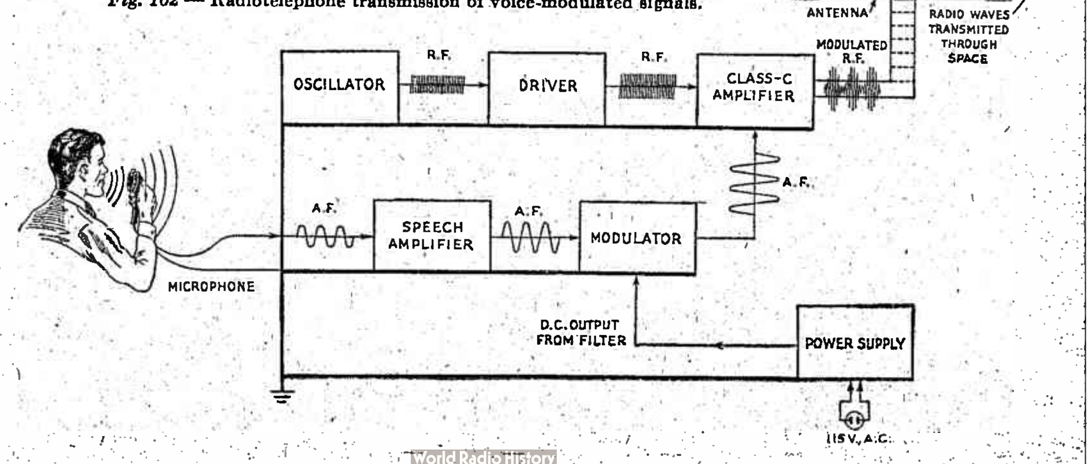
Compared with the range of tones audible to the human ear as sound, the alternations of radiated electromagnetic waves occur so rapidly that, even if converted into vibrations in the air, they are indistinguishable from the non-alternating or direct current from the battery or equivalent source which supplies the "carrier" current for a wire telephone or telegraph system. Thus it is that radio waves, even though they are only invisible fields of energy in space, can be substituted for the telephone or telegraph wires used in those systems.

To make radiated high-frequency alternating current useful for communication, the desired intelligence first must be impressed upon it in a manner that will permit its interpretation or conversion at the receiving end. This process is called modulation.

For code transmission it is necessary merely to interrupt the current at suitable intervals, producing a series of trains of waves whose duration and spacing conform to an artificial code representing the letters and numerals of the alphabet. This can be accomplished by simply turning on and off the power source from which the carrier energy is derived, using a special hand switch, called a "key," as pictured in Fig. 101.

The transmitted carrier is supplied by an oscillator, consisting of a vacuum tube and a resonant circuit connected to serve as a generator of alternating current at the desired radio frequency. This current is then amplified, either at the original frequency of the oscilla-

Fig. 102 — Radiotelephone transmission of voice-modulated signals.



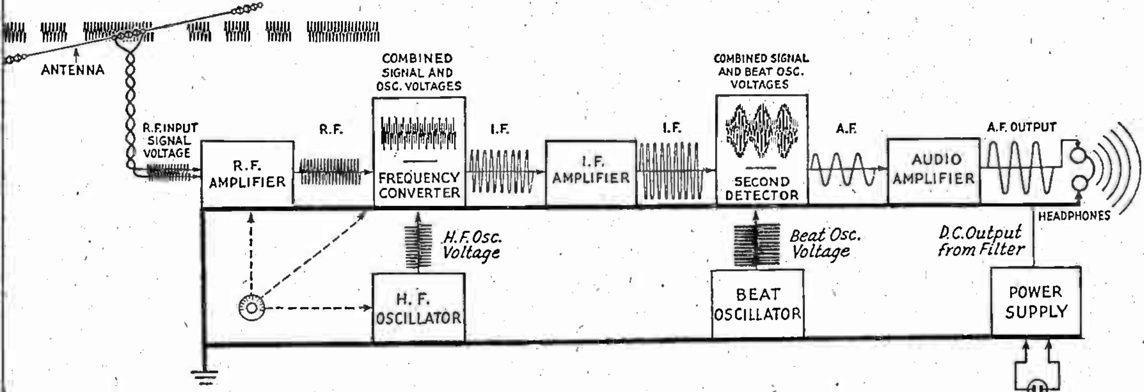


Fig. 103 — Superheterodyne reception of c.w. code signals.

tor or (as in Fig. 101) at a multiple thereof. After passing through as many amplifier stages as are needed to give the required power, the carrier wave trains corresponding to the dots and dashes of the code are conveyed to the transmitting antenna via a transmission line, and radiated.

The method by which the sounds of voice or music are transmitted by radio is somewhat more complex, as shown in Fig. 102. Radiotelephone transmission begins with a microphone placed in the path of the sound waves. Variations in the pressure or velocity of the air produce corresponding motions of a pliant diaphragm or ribbon in the microphone and by an electromechanical process made to generate a feeble alternating current having the same characteristics as the sound waves.

The resulting audio-frequency current, as it is termed, is amplified by vacuum tubes in an intermediate speech amplifier and then delivered to the modulator, where it is applied to the power in the radio-frequency carrier.

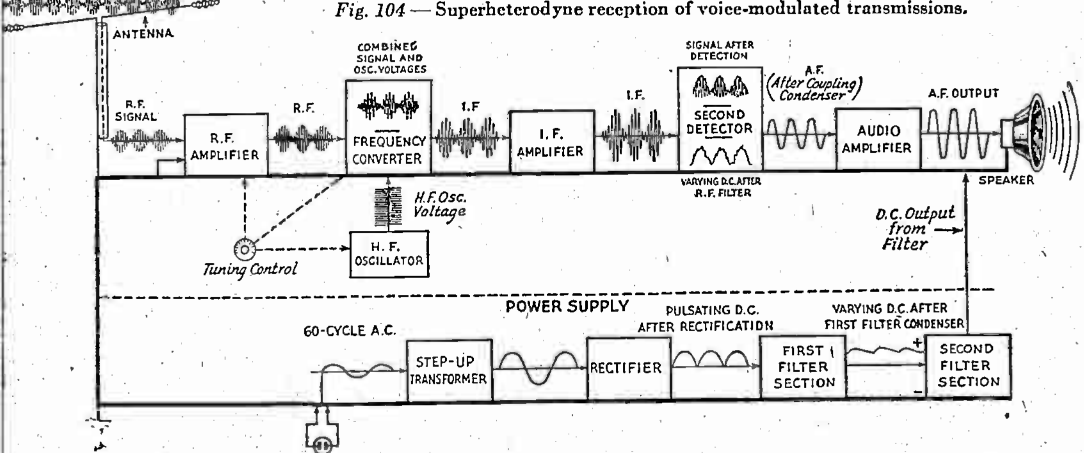
Up to the point of modulation the carrier is generated and amplified in the same manner as in the case of code transmission. Until it reaches the final radio-frequency amplifier stage the carrier has constant amplitude and frequency and waveform. The amplified audio frequency voltage from the modulator, which also employs vacuum tubes, then is made to vary either the amplitude or the frequency of the carrier. These variations are, of course, proportional to the amplitude and other characteristics of the audio-frequency voltage, and therefore represent a replica of the fluctuations in the original sound. The modulated carrier, in turn, is delivered to the transmitting antenna, which radiates the electrical energy in the form of electromagnetic waves.

ℂ 1-3 Radio Reception

When the waves from the transmitting antenna reach the antenna of a radio receiver, they create in it an alternating voltage which is identical with the characteristics of the carrier, whether it be the interrupted pulses of code or the modulated carrier of radiotelephony.

The manner in which this voltage is translated into sound is illustrated in Fig. 103, in the case of code transmission, and in Fig. 104 in the case of voice. The type of receiver depicted in these diagrams is the superheterodyne, which involves an intermediate conversion step. In the simplest possible type of receiver the received signal would be delivered directly to a vacuum-tube detector. If any preliminary amplification is performed in this type of re-

Fig. 104 — Superheterodyne reception of voice-modulated transmissions.



ceiver, an r.f. stage is inserted between the antenna and detector which operates at the carrier frequency.

In the superheterodyne receiver, however, such preliminary amplification is only incidental, and is followed by the conversion of the original carrier frequency to a fixed intermediate frequency. This is accomplished by means of what is known as a "beating" or heterodyne process, in which the intermediate frequency corresponds to the slow "beat" note heard when two musical tones of different pitch are struck simultaneously.

In code reception an auxiliary device is required to make audible sounds out of the steady pulses of the high-frequency carrier, which by themselves are inaudible to the human ear. This device, called a beat oscillator, also operates on the heterodyne principle.

With voice modulation, the modulated carrier is detected, which means that the audio-frequency envelope is separated from the i.f. carrier. The resulting audio current is further amplified to a point where sufficient power is available to operate headphones or a loudspeaker, which in turn sets up trains of sound waves in the surrounding air. These are a more or less exact reproduction of the sound vibrations which originally entered the microphone at the transmitter.

ℂ 1-4 Pulse Transmission

In addition to the conventional modes of communication by voice modulation and by c.w. transmission, other new methods of signaling or transmitting intelligence by radio, as well as new mechanisms for performing these older functions, are being developed.

For the most part these methods are based on the transmission of pulses of various kinds which, upon suitable selection at the receiving end, can be employed for such diversified uses as recreating television pictures, remote control of almost any mechanism, sounding the upper atmosphere, obstacle detection, high-speed automatic telegraphy, automatic landing systems for aircraft, and a multitude of other uses.

The technique of pulse transmission and reception is based on the translation, transmission and interpretation of various kinds of pulses. Ordinary c.w. (code) transmission employs pulse transmission in its most elementary form, wherein two kinds of pulses — one long and one short — are utilized to convey intelligence according to a prescribed code. The original intelligence conveyed by these pulses may be interpreted, aurally, by the ears of a trained operator, or through automatic devices which differentiate between them by means of their relative length. This then becomes one of the forms of pulse selection — identification in terms of length, or duration. Other forms of pulse selection are based on amplitude, on polarity and on shape.

The method used in the transmission of television pictures is similar to that used for sound transmission. The microphone is replaced by a television camera which scans the scene to be viewed and converts the variations of light and shade into an alternating voltage. This "video" voltage is superimposed on a carrier, transmitted, received, and demodulated, and finally is reconverted into light by a cathode-ray tube which is equivalent to the loudspeaker of a sound receiver.

It is logical to anticipate that the application by amateurs of the methods and mechanisms of pulse transmission will increase. Already some of the techniques thereof have been employed in such devices as electronic keys and automatic station control systems. The basic methods themselves have been applied to the remote control of models, and even, in some instances, to automatic coded transmission and ionosphere soundings. Prewar experimental work with television and facsimile has already demonstrated that these fields will have major significance in postwar amateur radio.

ℂ 1-5 Underlying Fundamentals

While all of these applications involve specialized techniques in certain elements of the individual systems, each employs the common medium of radio waves for conveying the energy employed in accomplishing the ultimate desired result. Basically, therefore, the fundamental techniques of radio are common to all, and a knowledge of the concepts and instruments employed in radio communication is prerequisite to acquiring the specialized knowledge utilized in any of these allied fields.

There is another aspect common alike to communication and these allied fields. That is the electronic vacuum tube. The fact that this invaluable tool is employed in so many new processes and devices (the more recent ones currently enshrouded still in military secrecy) has led to their classification under the categorical definition of "electronics" or electronic devices. Some of these specialized devices, while employing vacuum tubes, do not use a radio link; thus, while practically all radio devices are electronic devices, not all electronic devices are also radio devices.

This *Handbook,* therefore, while created specifically for the radio amateur and dealing with the art from the amateur's point of view, covers all the common and, to a great extent, the specialized forms of equipment found in the radio and electronic services.

A glance at the chapter headings will show that each element of the radio system is considered in appropriate sequence for readiest comprehension, first from the theoretical standpoint and then in terms of actual apparatus.

First, however, the component parts utilized in each of these elements must be known and the principles of their operation understood — most notably in the case of the vacuum tube. The two chapters immediately following treat, therefore, of the basic electrical and radio fundamentals and of vacuum tubes.

Chapter Two

Electrical and Radio Fundamentals

2-1 The Nature of Electricity

ALL matter — solids, liquids and gases — is made up of fundamental units, called *molecules*. The smallest subdivision of a substance retaining all its characteristic properties, the molecule in turn is constructed of *atoms* of the various elements comprising the substance.

The atom is made up of a central part, called the *nucleus*, around which minute particles or charges of electricity, called *electrons*, circulate. The atom can be compared roughly to the solar system, with the sun representing the nucleus and the planets the electrons. By far the greater part of the mass or weight of an atom is in the nucleus, but because of its extreme compactness the nucleus occupies only a small part of the space taken up by the atom. In the normal or *neutral* atom the electrical "charge" on each electron is balanced by an equal charge of opposite kind associated with the nucleus. The kind of electricity represented by the electron is called *negative*, while that associated with the nucleus is called *positive*.

The greater mass of the nucleus (the nucleus is more than 1850 times as heavy as the electrons associated with it) is considered to be principally in neutral particles — that is, particles which exhibit no electrical effects — bound together by some means. These neutral particles each may actually be the result of the combination of a positive and a negative particle, so that the charge on each is neutralized. The net positive charge associated with the nucleus can be regarded as an excess of positive particles, or as an absence of enough negative charges to neutralize all the positive charges present.

Ordinary electrical activity is the result of movements of the electrons, or negative charges, so it is customary to consider electrical phenomena as caused by the presence or absence of these negative particles.

2-2 Electrons

Electron flow — In the atoms of many substances, one or two of the outer electrons associated with the nucleus can be detached from the atom, thus leaving the atom as a whole with a net positive charge.

The electrons outside the nucleus move in planetary elliptic orbits about it. The radius of the different orbits varies within a single atom, and as a consequence the strength of the bond existing between the nucleus and the different electrons varies. The outer electrons, in general, are rather loosely bound to the nucleus, and under favorable conditions may be completely dissociated from the remainder of the atom. This process, to be treated subsequently in more detail, is known as ionization. It is the process by which electrons are emitted from the heated filament in a vacuum tube.

In conduction through solids the electrons constituting the current flow are the outer orbital electrons of the atoms. Since these electrons are less tightly bound to the atom, they are spoken of as free electrons. As these electrons move through the solid under the influence of an electric field, they collide with the atoms and continuously lose the energy gained from the field. As a consequence the velocity of electrons in the direction of the field is comparatively small — a ten-millionth or less the velocity of free electrons.

The free electrons which contribute to the electric current have a low drift velocity in the negative direction of the field within the conductor. Moving through the metal in a common general direction they enter into frequent collisions with the molecules of the metal, and in consequence are continually retarded in their forward motion. The maximum velocity they are able to attain depends on the strength of the field and the nature of the substance.

The collisions which tend to reduce the drift velocity of the electrons act as a retarding force. When current flows, this retarding force must exactly equal the accelerating force of the field. The retarding force is proportional to the number of free electrons per unit length of conductor and to their drift velocity. The accelerating force is proportional to the field per unit length of conductor, to the number of electrons per unit length, and to the electronic charge.

The unit of quantity — The amount of electricity represented by a single electron is extremely small — far too small to be used as a unit of quantity in practical electrical work. The practical unit of electrical quantity is the *coulomb*. One coulomb is equal to about 6.2×10^{18} electrons. Because the electron is so minute, the "granular" nature of electricity is not apparent in practical work.

Static and current electricity — An electrical charge may be either at rest (*static*) or moving. *Electrostatics* is that branch of electrical theory which deals with the behavior of electricity at rest. If an electrical charge is moving, its movement constitutes a *current* of electricity. The movement may take place through a vacuum, through a gas or liquid, or through solid materials (usually metals) called *conductors*. When the movement is through a solid, the collection of electrons constituting the original charge does not move as a unit through the entire path; instead, individual electrons all along the path are urged to leave the atoms to which they are attached. Each electron travels only a relatively short distance before finding another atom which is electron-deficient, and to which it tends to attach itself. The motion is, therefore, transmitted along the path from electron to electron, much in the same way that motion in a chain is transmitted from link to link. Naturally, the ease with which the electron motion is transmitted depends upon the ease with which an electron can be detached from an atom of the substance through which the current is moving.

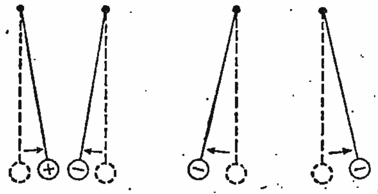


Fig. 201 — Attraction and repulsion of charged objects, as demonstrated by the familiar pith-ball experiment.

The flow of current through a conductor is accompanied by heating of the conductor, which may be explained as resulting from collisions between moving electrons and atoms, setting the latter into vibration. Thus there is a loss of energy, in the form of heat, accompanying the flow of current.

Insulators and conductors — Materials whose atoms will readily give up an electron are called *conductors*, while those in which all the electrons are firmly bound in the atom are called *insulators* or *dielectrics*. Most metals are good conductors, as also are acid or salt solutions. Among the insulators are wood, hard rubber, bakelite, quartz, glass, porcelain, textiles and many other nonmetallic materials.

Resistance — No substance is a perfect conductor, and there is also no such thing as a perfect insulator. The measure of the difficulty in moving an electron by electrical means is called *resistance*. Good conductors have low resistance (high *conductivity*), good insulators very high resistance. Between the two are materials which are neither good conductors nor good insulators, but nonetheless are useful since there often is need for intermediate values of resistance in electrical circuits.

Circuits — A circuit is simply a complete path along which electrons can move. There will normally be a source of energy (a battery, for instance), and a *load*, or portion of the circuit where the current is made to do useful work. There must be an unbroken path through which the electrons can move, with the source of energy acting as an electron pump and sending them around the circuit. The circuit is said to be *open* when no charges can move, because of a break in the path. It is *closed* when no break exists — when switches are closed and all connections are made.

2-3 Static Electricity

The electric charge — Many materials that have a high resistance can be made to acquire a charge (surplus or deficiency of electrons) by mechanical means, such as friction. The familiar crackling when a hard-rubber comb is run through hair on a dry winter day is an example of an electric charge generated by friction. Objects can have either a surplus or a deficiency of electrons — a surplus of electrons is called a *negative* charge; a lack of them is called a *positive* charge. The kind of charge is called its *polarity*. A negatively charged object is frequently called a negative *pole*, while a positively charged object similarly is called a positive pole.

Attraction and repulsion — Unlike charges (one positive, one negative) exert an attraction on each other. This can be demonstrated by giving charges of opposite polarity to two very light, well-insulated conductors, such as bits of metal foil suspended from dry thread (Fig. 201). Pith balls covered with foil frequently are used in this experiment.

When the two charged objects are brought close together, it will be observed that they will be attracted to each other. If the charges are equal and the charged bodies are permitted to touch, the surplus electrons on the negatively charged object will transfer to the positively charged object (i.e., the one deficient in electrons) and the two charges will neutralize, leaving both bodies uncharged. If the charges are not equal, the weaker charge neutralizes an equal amount of the stronger when the two bodies touch, upon which the excess of the stronger charge distributes itself over both. Both bodies then have charges of the same polarity, and a force of repulsion is exercised between them. Consequently, the bits of foil tend to spring away from each other.

Remember this rule: *Unlike charges attract, like charges repel.*

Electrostatic field — From the foregoing it is evident that an electric charge can exert a force through the space surrounding the charged object. The region in which this force is exerted is considered to be pervaded by an *electrostatic field*, this concept of a field being adopted to explain the "action at a distance" of the charge. The field is pictured as consisting of *lines of force* originating on the charge and

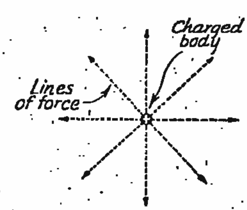


Fig. 202 — Lines of force from a charged object extend outward radially. Although only two dimensions are shown, the field extends in all directions from the charge, and should be visualized in three dimensions.

spreading in all directions, finally terminating on other charges of opposite polarity. These other charges may be a very large distance away. The number of lines of force per unit area is, however, a measure of the intensity of the field.

The general picture of a charged object in isolated space is shown in Fig. 202. This is an idealized situation, since in practice the charged object could not be completely isolated. The presence of other charges, or simply of insulators or conductors, in the vicinity will greatly change the configuration of the field. The direction of the field, as indicated by the arrowheads, is away from a positively charged object; if the charge were negative, the direction would be toward the charge.

It should be understood that the field picture as represented above is merely a convenient method of explaining observed effects, and is not to be taken too literally. The electric force does not consist of separate lines like strings or rods; instead, it completely pervades the medium through which the force is exerted. With this understanding in mind, it is *convenient* to talk of lines of force and to measure the field intensity in terms of number of lines per unit area.

The intensity of the field dies away with distance from the charged object in a manner determined by its shape and the circumstances of its surroundings. In the case of an isolated charge at a point (an infinitesimally small object), the field strength is inversely proportional to the square of the distance. However, this relationship is not true in many other cases; in some important practical applications the field intensity is inversely proportional to the distance involved, and not to its square.

Electrostatic induction — If a piece of conducting material is brought near a charged object, the field will exert a force on the electrons of the metal so that those free to move will do so. If the object is positively charged, as indicated in Fig. 203, the free electrons will move toward the end of the conductor nearest the charged body, leaving a deficiency of electrons at the other end. Hence, one end of the conductor becomes negatively charged while the other en[illegible] has an equal positive charge. The lines of [illegible]rce from the charged body terminate on [illegible] conductor, where sufficient electrons ac[illegible]ulate to provide an electric intensity e[illegible] and opposite to that of the field at that [illegible]. Because of this effect, the electrosta[illegible]nside the conductor is completely n[illegible] by the induced charge; in other wo[illegible]ld does not penetrate the conduct[illegible] work this principle provides, t[illegible] which electrostatic fields may b[illegible]n regions where they are not w[illegible]

Ch[illegible] conductor as shown in Fig. [illegible] existence by the field from [illegible]. On taking the conductor out of the field the electrons will redistribute themselves so that the charges disappear. However, if the conductor is connected to the earth through a wire while under the influence of the field, as shown in Fig. 203-B, the induced positive charge will tend to move as far as possible from the source of the

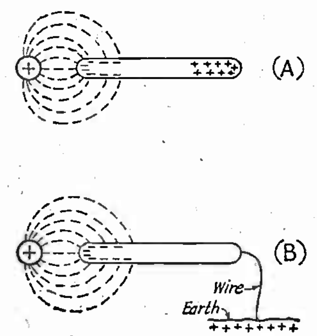


Fig. 203 — Electrostatic induction. The field from the positively charged body attracts electrons, which accumulate to form a negative charge. The opposite end of the conductor consequently acquires a positive charge. This charge may be "drained off" to earth as shown at B.

field (that is, electrons will flow from the earth to the conductor). If the grounding wire is then removed, the conductor will be left with an excess of electrons and will have acquired a "permanent" charge — permanent, that is, so long as the conductor is well enough insulated to prevent the charge from escaping to earth or to other objects. The polarity of the induced charge always is opposite to the polarity of the charge which set up the original field.

Energy in the electrostatic field — The expenditure of energy is necessary to place an electrical charge upon an object and thus establish an electrostatic field. Once the field is established and is constant, no further expenditure of energy is required. The energy supplied to establish the field is stored in the field; thus the field represents *potential* energy (that is, energy available for use). The potential energy is acquired in the same way that potential energy is given any object (a 10-pound weight, for instance) when it is lifted against the gravitational pull of the earth. If the weight is allowed to drop, its potential energy is changed into the energy of motion. Similarly, if the electrostatic field is made to disappear its potential energy is transformed into a movement of electrons or into what is known as an electric current.

The potential energy of the lifted weight is measured by its weight and the distance it is lifted; that is, by the work done in lifting it. Similarly, the potential energy (called simply *potential*) of the electrostatic field at any point is measured by the work done in moving a charge of specified value to that point, against the repulsion of the field. In practice, absolute potential is of less interest than the *difference of potential* between two points in the field.

Potential difference — If two objects are charged differently, a potential difference exists between them. Potential difference is measured by an electrical unit called the *volt*. The greater the potential difference, the higher (numerically) the voltage. This voltage exerts an electrical pressure or *force* as explained above, and is often called *electromotive force* or, simply, *e.m.f.* It is not necessary to have unlike charges in order to have a difference of potential; both, for instance, may be negative, so long as one charge is more intense than the other. From the viewpoint of the stronger charge, the weaker one appears to be positive in such a case, since it has a smaller number of excess electrons; in other words, its *relative polarity* is positive. The greater the potential difference, the more intense is the electrostatic field between the two charged objects.

Capacity — More work must be done in moving a given charge against the repulsion of a strong field than against a weak one; hence, potential is proportional to the strength of the field. In turn, field strength is proportional to the charge or quantity of electricity on the charged object, so that potential also is proportional to charge. By inserting a suitable constant, the proportionality can be changed to an equality:

$$Q = CE$$

where Q is the quantity of charge, E is the potential, and C is a constant depending upon the charged object (usually a conductor) and its surroundings and is called the *capacity* of the object. Capacity is the ratio of quantity of charge to the potential resulting from it, or

$$C = \frac{Q}{E}$$

When Q is in coulombs and E in volts, C is measured in *farads*. A conductor has a capacity of one farad when the addition of one coulomb to its charge raises its potential by one volt.

The farad is much too large a unit for practical purposes. In radio work, the *microfarad* (one millionth of a farad) and the *micro-microfarad* (one millionth of a microfarad) are the units most frequently used. They are abbreviated *μfd.* and *μμfd.*, respectively.

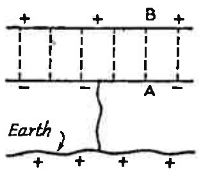


Fig. 204 — The principle of the condenser.

The capacity of a conductor in air depends upon its size and shape. A given charge on a small conductor results in a more intense electrostatic field in its vicinity than the same charge on a larger conductor. This is because the charge distributes itself over the surface, hence its density (the quantity of electricity per unit area) is smaller on the larger conductor. Consequently, the potential of the larger conductor is smaller, for the same amount of charge. In other words, its capacity is greater because a greater charge is required to raise its potential by the same amount.

Condensers — If a grounded conductor, A (Fig. 204), is brought near a second conductor, B, which is charged, the former will acquire a charge by electrostatic induction. Since the charge on A is opposite in polarity to that on B, the field set up by the induced charge on A will oppose the original field set up by the charge on B, hence the potential of B will be lowered. Because of this, more charge must be placed on B to raise its potential to its original value; in other words, its *capacity has been increased* by the presence of the second conductor. The combination of the two conductors separated by a dielectric is called a *condenser*.

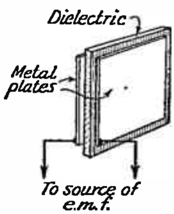


Fig. 205 — A simple condenser, consisting of two metal plates separated by dielectric material.

The capacity of a condenser depends upon the areas of the conductors, as before, and also becomes greater as the distance between the conductors is decreased, since, with a fixed amount of charge, the potential difference between them decreases as they are moved closer together.

If insulating or dielectric material other than air is inserted between the conductors, it is found that the potential difference is lowered still more — that is, there is a further increase in capacity. This lowering of the potential difference is considered to be the result of *polarization* of the dielectric. By this it is meant that the molecules of the substance tend to be distorted under the influence of the electrostatic field in such a way that the negative charges within the molecule are drawn toward the positively charged conductor, leaving the other end of the molecule with a positive charge facin[illegible]e negatively charged conductor. Since [illegible]lectrons are firmly bound in the ato[illegible]he dielectric, there is no flow of curre[illegible] total charge on each atom is st[illegible] there is a *tendency* toward [illegible]h causes a reaction on the[illegible] The dielectric of a c[illegible] is under mechanic[illegible]ial difference bet[illegible]denser is great [illegible]k down mech[illegible]

[illegible]er with a [illegible] plates [illegible]ith air as [illegible]*uctive* *capacity* of the dielectric, or, probably more commonly, the *dielectric constant*. Strictly speaking, the comparison should be made to empty space (i.e., a vacuum) rather than to air, but the dielectric constant of air is so nearly that of a vacuum that the practical difference is negligible. A table of dielectric constants is given in the Appendix.

Condensers have many uses in electrical and radio circuits, all based on their ability to store energy in the electric field when a potential difference or voltage is caused to exist between the plates — energy which later can be released to perform useful functions.

ℂ 2-4 The Electric Current

Conduction in metals — When a difference of potential is maintained between the ends of a metallic conductor, there is a continuous drift of electrons through the conductor toward the end having a positive potential (relative polarity positive). This electron drift constitutes an electric current through the metal (§ 2-2). The speed with which the electron movement is established is very nearly the speed of light (300,000,000 meters, or approximately 186,000 miles, per second), so that the current is said to travel at nearly the speed of light. By this it is meant that the time interval between the application of the electromotive force and the flow of current in all parts of a circuit, even one extending over hundreds of miles, is negligible. However, the individual electrons do not move at anything approaching such a speed. The situation is similar to that existing when a mechanical force is transmitted by means of a rigid rod. A force applied to one end of the rod is transmitted practically instantaneously to the other end, even though the rod itself moves relatively slowly or not at all.

The magnitude of the electric current is the rate at which electricity is moved past a point in the circuit. If the rate is constant, then the current is equal to the quantity of electricity moved past a given point in some selected time interval. That is,

$$I = \frac{Q}{t}$$

where I is the intensity or magnitude of the current, Q is the quantity of electricity, and t is the time. If Q is in coulombs and t in seconds, the unit for I is called the *ampere*. One ampere of current is equal to one coulomb of electricity moving or "flowing" past a given point in a circuit in one second.

The currents used by different electrical devices vary greatly in magnitude. The current which flows in an ordinary 60-watt lamp, for instance, is about one-half ampere, the current in an electric iron is about 5 amperes, and that in a radio tube may be as low as 0.001 ampere.

When a current flows through a metallic conductor there is no visible or chemical effect on the conductor. The only physical effect is the heat developed (§ 2-2) as the result of energy loss in the conductor. Under normal conditions the rate at which heat is generated and that at which it is radiated by the conductor will quickly reach equilibrium. However, if the heat is developed at a more rapid rate than it can be radiated, the temperature will continue to rise until the conductor burns or melts.

Experimental measurements have shown that the current which flows in a given metallic conductor is directly proportional to the applied e.m.f., so long as the temperature of the conductor is held constant. There is no e.m.f. so small but that some current will flow as a result of its application to a metallic conductor.

Gaseous conduction — In any gas or mixture of gases (such as air, for example) there are always some free electrons — that is, electrons not attached to an atom — and also some atoms lacking an electron. Thus there are both positively and negatively charged particles in the gas, as well as many neutral atoms. An atom lacking an electron is called a *positive ion*, while the free electron is called a *negative ion*. The term *ion* is, in fact, applied to any elemental particle which has an electric charge.

If the gas is in an electric field, the free electrons will be attracted toward the source of positive potential and the positive ions will be attracted toward the source of negative potential. If the gas is at atmospheric pressure neither particle can travel very far before meeting an ion of the opposite kind, when the two combine to form a neutral atom. Since a neutral atom is not affected by the electric field, there is no flow of current through the gas.

However, if the gas is enclosed in a glass container in which two separate metal pieces called *electrodes* are sealed, and the gas pressure is then reduced by pumping out most of the gas, a different set of conditions results. At low pressure there is a comparatively large distance between each atom, and when an electric field is established by applying a difference of potential to the electrodes the ions can travel a considerable distance before meeting another ion or atom. The farther the ion travels the greater the velocity it acquires, since the effect of the field is to accelerate its motion. If the field is strong enough the ions will acquire such velocity that when one happens to collide with a neutral atom the force of the collision will knock an electron out of

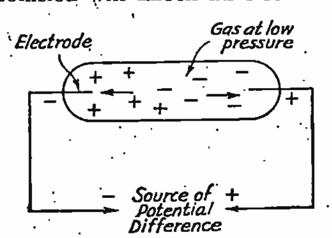


Fig. 206 — Illustrating conduction through a gas at low pressure. Positive ions are attracted to the negative electrode, while electrons are attracted to the positive electrode. This takes place only after the gas is ionized.

the atom, so that this atom also becomes *ionized.* The process is cumulative, and the freed electrons are attracted to the positive electrode while the positive ions are attracted to the negative electrode. This movement of charged particles constitutes an electric current through the gas.

Since an ion must acquire a certain velocity before it can knock an electron out of a neutral atom, a definite field strength is required before conduction can take place in a gas. That is, a certain value of potential difference, called the *ionizing potential,* must be applied to the electrodes. If less voltage is applied, the gas does not ionize and the current is negligible. On the other hand, once the gas is ionized an increase in potential does not have much effect on the current, since the ions already have sufficient velocity to maintain the ionization. The ionizing potential required depends upon the kind of gas and the pressure. Ionization is usually accompanied by a colored glow, different gases having different characteristic colors.

Current flow in liquids — A very large number of chemical compounds have the peculiar characteristic that, when they are put into solution, the component parts become ionized. For example, common table salt (sodium chloride), each molecule of which is made up of one atom of sodium and one of chlorine, will, when put into water, break down into a sodium ion (positive, with one electron deficient) and a chlorine ion (negative, with one excess electron). This can only occur so long as the salt is in solution — take away the water and the ions are recombined into the neutral sodium chloride. This spontaneous *dissociation* in solution is another form of ionization. If two wires with a difference of potential between them are placed in the solution, the negative wire will attract the positive sodium ions while the positive wire will attract the negative chlorine ions and an electric current will flow through the solution. When the ions reach the wires the electron surplus or deficiency will be remedied, and a neutral atom will be formed.

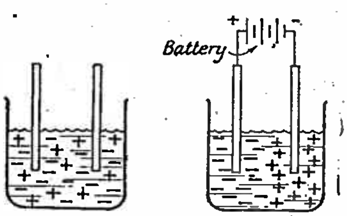


Fig. 207 — Electrolytic conduction. When an e.m.f. is applied to the electrodes, negative ions are attracted to the positively charged plate and positive ions to the negatively charged plate. The battery, which is the source of the e.m.f., is indicated by its customary symbol.

In this process, the water is decomposed into its gaseous constituents, hydrogen and oxygen. The energy used up in decomposing the water and in moving the ions is supplied by the source of potential difference. The energy used in decomposing the water is equivalent to an opposing e.m.f., of the order of a volt or two. If this constant "back voltage" is subtracted from the applied voltage, it is found that the current flowing through a given solution, or *electrolyte,* is proportional to the difference between the two voltages.

Current flow in vacuum — If a suitable metallic conductor is heated to a high temperature in a vacuum, electrons will be emitted from the surface. The electrons are freed from this *filament* or *cathode* because it has been

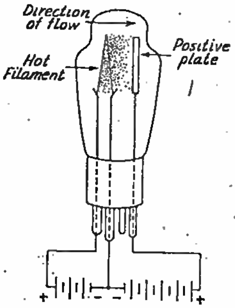


Fig. 208 — Conduction by thermionic emission in a vacuum tube. One battery is used *only* to heat the filament to a temperature where it will emit electrons. The other battery places a potential on the plate which is positive with respect to the filament, and as a result the electrons are attracted to the plate. The electron flow from filament to plate completes the electrical circuit.

heated to a temperature that gives them sufficient energy of motion to allow them to break away from the surface. The process is called *thermionic electron emission.* Now, if a metal plate is placed in the vacuum and given a positive charge with respect to the cathode, this plate or *anode* will attract a number of the electrons that surround the cathode. The passage of the electrons from cathode to anode constitutes an electric current. All thermionic vacuum tubes depend for their operation on the emission of electrons from a hot cathode.

Since the electrons emitted from the hot cathode are negatively charged, it is evident that they will be attracted to the plate only when the latter is at a positive potential with respect to the cathode. If the plate is negatively charged with respect to the cathode the electrons will be repelled back to the cathode, hence no current will flow through the vacuum. Consequently, a thermionic vacuum tube conducts current *in one direction only.* When the plate is positive, it is found that (if the potential is not too large) the current increases with an increase in potential difference between the plate and cathode. However, the relationship between current and applied voltage is not a simple one. If the voltage is made large enough all the electrons emitted by the cathode will be drawn to the plate, and a further increase in voltage therefore cannot cause a further increase in current. The number of electrons emitted by the cathode depends upon the temperature of the cathode and the material of which it is constructed.

Direction of current flow — Use was being made of electricity for a long time before its electronic nature was understood. While it is now clear that current flow is a drift of negative electrical charges or electrons toward a source of positive potential, in the era preceding the electron theory it was *assumed* that the current flowed from the point of higher positive potential to a point of lower (i.e., less positive or more negative) potential. While this assumption turned out to be wholly wrong, it is still customary to speak of current as flowing "from positive to negative" in many applications. The practice often causes confusion, but this distinction between "current" flow and "electron" flow often must be taken into account. If electron flow is specifically mentioned there can be, of course, no doubt as to the meaning; but when the direction of *current* flow is specified, it may be taken, by convention, as being opposite to the direction of electron movement.

Primary cells — If two electrodes of dissimilar metals are immersed in an electrolyte, it is found that a small difference of potential exists between the electrodes. Such a combination is called a *cell*. If the two electrodes are connected together by a conductor external to the cell, an electric current will flow between them. In such a cell, chemical energy is converted into electrical energy. The difference of potential arises as a result of the fact that material from one or both of the electrodes goes into solution in the electrolyte, and in the process ions are formed in the vicinity of the electrodes. The electrodes acquire charges because of the electric field associated with the charged ions. The difference of potential between the electrodes is principally a function of the metals used, and is more or less independent of the kind of electrolyte or the size of the cell.

When current is supplied to an external circuit, two principal effects occur within the cell. The negative electrode (negative as viewed from outside the cell) loses weight as its material is used up in furnishing energy, and hydrogen bubbles form on the positive electrode. Since the gas bubbles are non-conducting, their accumulation tends to reduce the effective area of the positive electrode, and consequently reduces the current. The effect is cumulative, and eventually the electrode will be completely covered and no further current can flow. This effect is called *polarization*. If the bubbles are removed, or prevented from forming by chemical means, polarization is reduced and current can flow as long as there is material in the negative electrode to furnish the energy. A chemical which prevents the formation of hydrogen bubbles in a cell is called a *depolarizer*.

In addition to polarization effects, a cell has a certain amount of *internal resistance* because of the resistance of the electrodes and the electrolyte and the contact resistance between the electrodes and electrolyte. The *internal resistance* depends upon the materials used and the size and electrode spacing of the cell. Large cells with the electrodes close together will have smaller internal resistance than small cells made of the same materials.

A collection of cells connected together is called a *battery*. The term battery also is applied (although incorrectly) to a single cell.

Dry cells — The most familiar form of primary cell is the *dry cell*. Like the elementary type of cell just described, it has a liquid electrolyte, but the liquid is mixed with other materials to form a paste. The cell therefore can be used in any position and handled as though it actually were dry.

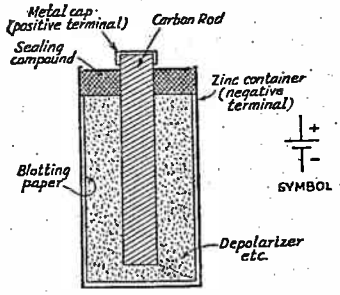


Fig. 209 — Construction of a dry cell.

The construction of an ordinary dry cell is shown in Fig. 209. The container is the negative electrode and is made of zinc. Next to it is a section of blotting material saturated with the electrolyte, a solution of sal ammoniac. The positive electrode is a carbon rod, and the space between it and the blotting paper is filled with a mixture of carbon, manganese dioxide (the depolarizer) and the electrolyte. The top is filled with sealing compound to prevent evaporation, since the cell will not work when the electrolyte drys out. The e.m.f. of a dry cell is about 1.5 volts.

Dry cells are made in various sizes, depending upon the current which they will be called upon to furnish. The construction frequently varies from that shown in Fig. 209, although in general the basic materials are the same in all dry cells. Batteries of small cells are assembled together as a unit for furnishing plate current for the vacuum tubes used in portable receiving sets; such "B" batteries, as they are called, can supply a current of a few hundredths of an ampere continuously. Larger cells, such as the common "No. 6" cell, can deliver currents of a fraction of an ampere continuously, or currents of several amperes for very short periods of time. The total amount of energy delivered by a dry cell is larger when the cell is used only intermittently, as compared with continuous use. The cell will deteriorate even without use, and should be put into service within a year or so from the time it is manufactured. The period during which it is usable (without having been put in service) is known as the "shelf life" of the cell or battery.

Secondary cells — The types of cells just described are known as *primary cells*, because the electrical energy is obtained directly from chemical energy. In some types of cells the chemical actions are reversible; that is, forcing a current through the cell, in the opposite direction to the current flow when the cell is delivering electrical energy, causes just the reverse chemical action. This tends to restore the cell to its original condition, and electrical energy is transformed into chemical energy. The process is called *charging* the cell. A cell which must first be charged before it can deliver electrical energy is called a *secondary cell.*

A simple form of secondary cell can be made by immersing two lead electrodes in a dilute solution of sulphuric acid. If a current is forced through the cell, the surface of the electrode which is connected to the positive terminal of the charging e.m.f. will be changed to lead peroxide and the surface of the electrode connected to the negative terminal will be changed to spongy lead. After a period of charging the charging source can be disconnected, and the cell will be found to have an e.m.f. of about 2.1 volts. It will furnish a small current to an external circuit for a period of time. This *discharge* of electrical energy is accompanied by chemical action which forms lead sulphate on both electrodes. When the lead peroxide and spongy lead are converted to lead sulphate there is no longer a difference of potential, since both electrodes are now the same material, and the cell is completely discharged.

The lead storage battery — The most common form of secondary cell is the lead storage cell. The common storage battery for automobile starting consists of three such cells connected together electrically and assembled in a single container. The principle of operation is similar to that just described, but the construction of the cell is considerably more complicated. To obtain large currents it is necessary to use electrodes having a great deal of surface area and to put them as close together as possible. The electrodes are made in the form of rectangular flat plates, consisting of a latticework or grid of lead or an alloy of lead. The interstices of the latticework are filled with a paste of lead oxide. The electrolyte is a solution of sulphuric acid in water. When the cell is charged, the lead oxide in the positive plate is converted to lead peroxide and that in the negative plate to spongy lead. To obtain high current capacity, a cell consists of a number of positive plates, all connected together, and a number of negative plates likewise connected together. They are arranged as shown in Fig. 210, with alternate negative and positive plates kept from touching by means of thin *separators* of insulating material, generally treated wood or perforated hard rubber. The separators preferably should be porous, so that the electrolyte can pass through them freely; thus they do not impede the passage of current from one plate to the next. There is always one extra negative plate in such an assembly, because the active material in the positive plate expands when the cell is being charged and if all the expansion took place on one side the plate would be distorted out of shape.

The e.m.f. of a fully charged storage cell is about 2.1 volts. When the e.m.f. drops to about 1.75 volts on discharge, the cell is considered to be completely discharged. Discharge beyond this limit may result in the formation of so much lead sulphate on the plates that the cell cannot be recharged, since lead sulphate is an insulator. During the charging process water in the electrolyte is used up, with the result that the sulphuric acid solution becomes more concentrated. The higher concentration increases the specific gravity of the solution, so that the specific gravity may be used to indicate the state of the battery with respect to charge. In the ordinary lead storage cell the solution is such that a specific gravity of 1.285 to 1.300 indicates a fully charged cell, while a discharged cell is indicated by a specific gravity of 1.150 to 1.175. The specific gravity can be measured by means of a *hydrometer*, shown in Fig. 211. For use with portable batteries, the hydrometer usually consists of a glass tube fitted with a syringe so that some of the electrolyte can be drawn from the cell into the tube. The hydrometer float is a smaller glass tube, air-tight and partly filled with shot to make it sink into the solution. The lower the specific gravity of the solution, the farther the float sinks into it. A graduated scale on the float shows the specific gravity directly, being read at the level of the solution.

Storage cells are rated in *ampere-hour capacity*, based on the number of amperes which can be furnished continuously for a stated period of time. For example, the cell may have a rating of 100 ampere-hours at an 8-hour discharge rate. This means that the cell will deliver 100/8 or 12.5 amperes continuously for 8 hours after having been fully charged. The ampere-hour capacity of a cell will vary with the discharge rate, becoming smaller as the rated time of discharge is made shorter. It also depends upon the size of the plates and their number. In automobile-type batteries the dimensions of the plates are fairly well standardized, so that the ampere-hour capacity is chiefly determined by the number of plates in a cell. It is, therefore, common practice to speak of "11-plate,"

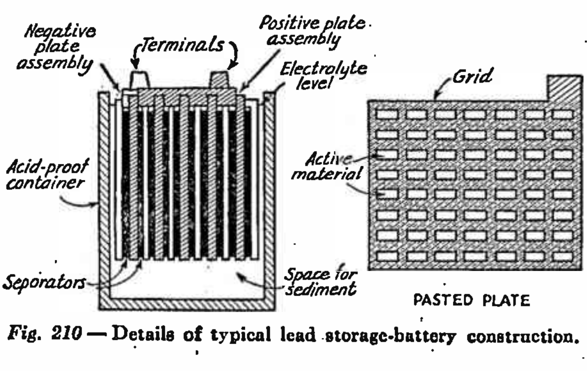


Fig. 210 — **Details of typical lead storage-battery construction.**

"15-plate," etc., batteries as an indication of the battery capacity.

Lead storage batteries must be kept fully charged if they are to stay in good condition. If a discharged battery is left standing idle, lead sulphate will form on the plates and eventually the battery will be useless. When the battery is being charged, hydrogen bubbles are given off by the electrolyte which, in bursting at the surface, throw out fine drops of the electrolyte. This is called "gassing." The sulphuric-acid solution spray from gassing will attack many materials, and consequently care must be used to see that it is not permitted to fall on near-by objects. It should also be wiped off the battery itself.

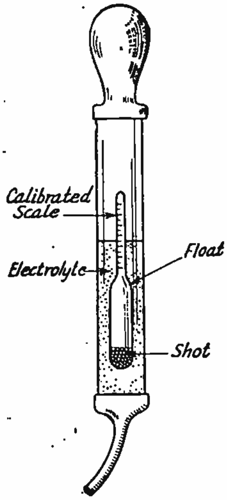


Fig. 211 — The hydrometer, a device with a calibrated scale for measuring the specific gravity of the electrolyte, used to determine the state of charge of a lead storage battery.

A lead battery may be charged at its nominal discharge rate; i.e., a 100-ampere-hour battery, 8-hour rating, can be charged at 100/8, or 12.5 amperes. The charging voltage required is slightly more than the output voltage of the cell. The preferred method is to charge at the full rate until the cells start to "gas" freely, after which the charging rate should be dropped to about half its initial value until the battery is fully charged, as indicated by the hydrometer reading. Alternatively, the battery may be charged from a constant-potential source (about 2.3 volts per cell), when the rise of terminal voltage of the battery as it accumulates a charge will automatically "taper" the charging rate.

The solution in a lead storage battery will freeze at a temperature of about zero degrees Fahrenheit when the battery is discharged, but a fully charged battery will not freeze until the temperature reaches about 90 degrees below zero. Keeping the battery charged therefore will prevent damage by freezing.

Cells in series and parallel — For proper operation, many electrical devices require higher voltage or current than can be obtained from a single cell. If greater voltage is needed, cells may be connected in *series*, as shown in Fig. 212-A. The negative terminal of one cell is connected to the positive terminal of the next, so that the total e.m.f. of the battery is equal to the sum of the e.m.f.s of the individual cells. For radio purposes, batteries of 45 and 90 volts or more are built up in this way from 1.5-volt dry cells. An automobile storage battery consists of three lead storage cells in series, totalling 6.3 volts — or, in round figures, 6 volts. The current which may be taken safely from a battery composed of cells in series is the same as that which may be taken safely from one cell alone; since the same current flows through all cells, the current capacity is unchanged.

When the device or load to which the battery is to be connected requires more current than can be taken safely from a single cell, the cells may be connected in parallel, as shown in Fig. 212-B. In this case the total current is the sum of the currents contributed by the individual cells, each contributing the same amount if the cells are all alike. When cells are connected in parallel it is essential that the e.m.f.s all be the same, since if one cell generated a larger voltage than the others it would force current through the other cells in the reverse direction and thus would take most, if not all, of the load. Also, if one cell has a lower terminal voltage than the others it will take current from the others rather than carrying its fair share.

Cells may be connected in series-parallel, as in Fig. 212-C, to increase both the voltage and the current-carrying capacity of the battery.

2-5 Electromagnetism

The magnetic field — Everyone is familiar with the fact that a bar or horseshoe magnet will attract small pieces of iron. Just as in the case of electrostatic attraction (§ 2-3) the concept of a *field*, in this case a field of *magnetic* force, is adopted to explain the magnetic action. The field is visualized as being made up of *lines* of magnetic force, the number of which per unit area determines the field strength. As in the case of the electrostatic field, the lines of force do not have physical existence but simply represent a convenient way of describing the properties of the force.

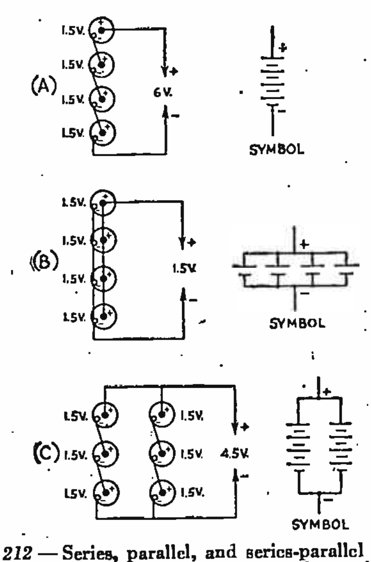


Fig. 212 — Series, parallel, and series-parallel connection of cells. Series connection increases the total voltage without changing current capacity; parallel connection increases current capacity without increasing voltage.

Magnetic attraction and repulsion — The forces exerted by the magnetic field are analogous to electrostatic forces. Corresponding to positive and negative electric charges, it is found that there are two kinds of magnetic poles. Instead of being called "positive" and "negative," however, the magnetic poles are called "north" (*N*) and "south" (*S*) poles. These names arise from the fact that, when a magnetized steel rod is freely suspended, it will turn into such a position that one end points toward the north. The end which points north is called the "north-seeking," or simply the "north," pole.

Unlike electric lines of force, which terminate on charges of opposite polarity (§ 2-3), magnetic lines of force are *closed upon themselves*. This is illustrated by the field about a bar magnet, as shown in Fig. 213-A. The lines extend through the magnet, the direction being taken from *S* to *N* inside the magnet and from *N* to *S* outside the magnet. If similar poles of two magnets are brought near each other, there is a force of repulsion between them, while dissimilar poles are attracted when brought close together. As in the case of electric charges, like poles repel, unlike poles attract.

If a bar magnet is cut in half, as in Fig. 213-B, it is found that the cut ends also are poles, of opposite kind to the original poles on the same piece. Such cutting can be continued indefinitely, and, no matter how small the pieces are made, there are always two opposite poles associated with each piece. In other words, a single magnetic pole cannot exist alone; it must always be associated with a pole of the opposite kind.

To explain this property of a magnet, it is considered that each molecule of a magnetic substance is itself a miniature magnet. If the material is not magnetized, the molecules are in random positions and the total magnetic effect is zero since there are just as many molecules tending to set up a magnetic field in one direction as there are others tending to set up a field in the opposite direction. When the substance becomes magnetized, however, the molecules are aligned so that most or all of the *N* poles of the molecular magnets are turned toward one end of the material while the *S* poles point toward the other end.

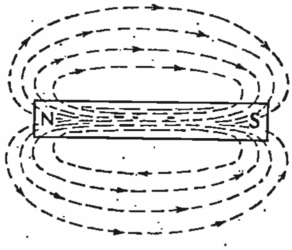
Magnetic induction — When an unmagnetized piece of iron is brought into the field of a magnet, its molecules tend to align themselves as described in the preceding paragraph. If one end of the iron is near the *N* pole of the magnet, the *S* poles of the molecules will turn toward that end and an *S* pole is said to be *induced* in the iron. An *N* pole will appear at the opposite end. Because of the attraction between opposite poles, the iron will be drawn toward the magnet. Since the iron has become a magnet under the influence of the field, it also possesses the property of attracting other pieces of iron.

When the magnetic field is removed, the molecules may or may not resume their random positions. If the material is soft iron the magnetism disappears quite rapidly when the field is removed, but in some types of steel the molecules are slow to resume their random positions and such materials will retain magnetism for a long time. A magnet which loses its magnetism quickly when there is no external magnetizing force is called a *temporary magnet*, while one which retains its magnetism for a long time is called a *permanent magnet*. The tendency to retain magnetism is called *retentivity*. The process of destroying magnetism can be hastened by heating, which increases the motion of the molecules within the substance, as well as by mechanical shock, which also tends to disturb the molecular alignment.

Electric current and the magnetic field — Experiment shows that a moving electron generates a magnetic field of exactly the same nature as that existing about a permanent magnet. Since a moving electron, or group of electrons moving together, constitutes an electric current, it follows that the flow of current is accompanied by the creation of a magnetic field. When the conductor is a wire the magnetic lines of force are in the form of concentric circles around it and lie in planes at right angles to it, as shown in Fig. 214. The direction of this field is controlled by the direction of current flow.

There is an easily remembered method for finding the relative directions of the current and of the magnetic field it sets up. Imagine the fingers of the right hand curled about the wire, with the thumb extended along the wire in the direction of current flow (the conventional direction, from positive to negative, not the direction of electron movement). Then the fingers will be found to point in the direction of the magnetic field; that is, from *N* to *S*.

(A)



(B)

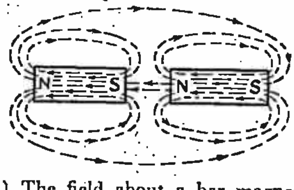


Fig. 213 — (A) The field about a bar magnet. The magnetic lines of force are continuous, part of the path being inside the magnet and part outside. (B) Cutting a magnet produces two magnets, each complete with N and S poles. With the magnets in the positions shown, some of the lines of force are common to both magnets.

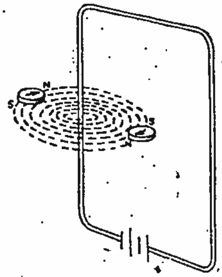


Fig. 214 — Whenever electric current passes through a wire, magnetic lines of force are set up, in the form of concentric circles, at right angles to the wire, and a magnetic field is said to exist around the wire. The direction of this field is controlled by the direction of current flow, and can be traced by means of a small compass.

Magnetomotive force — The force which causes the magnetic field is called *magnetomotive force*, abbreviated *m.m.f.* It corresponds to electromotive force or e.m.f. in the electric circuit. The greater the magnetomotive force, the stronger the magnetic field; that is, the larger the number of magnetic lines per unit area. Magnetomotive force is proportional to the current flowing. When the wire carrying the current is formed into a coil so that the magnetic flux will be concentrated instead of being spread over a large area, the m.m.f. also is proportional to the number of turns in the coil. Consequently magnetomotive force can be expressed in terms of the product of current and turns, and the *ampere-turn*, as this product is called, is in fact the common unit of magnetomotive force. The same magnetizing effect can be secured with a great many turns and a weak current or with a few turns and a strong current. For example, if 10 amperes flow in one turn of wire, the magnetizing effect is 10 ampere-turns. If there is one ampere flowing in 10 turns of wire, the m.m.f. also is 10 ampere-turns.

The magnetic circuit — Since magnetic lines of force are always closed upon themselves, it is possible to draw an analogy between the magnetic *circuit* and the ordinary electrical circuit. The electrical circuit also must be closed so that a complete path is provided around which the electrons or current can flow. However, there is no insulator for the magnetic field, so that the magnetic circuit is always complete even though no magnetic material (such as iron) may be present.

The number of lines of magnetic force, or *flux*, is equivalent in the magnetic circuit to current in the electric circuit. However, it is usual practice to express the strength of the field in terms of the number of lines per unit area, or *flux density*. The unit of flux density is the *gauss*, which is equal to one line per square centimeter, but the terms "lines per square centimeter" or "lines per square inch" are commonly used instead.

Corresponding to resistance in the electric circuit is the tendency to obstruct the passage of magnetic flux, which is called *reluctance*. The reluctance of good magnetic materials, such as iron and steel, is quite low.

The *permeability* of a material is the ratio of the flux which would be set up in a closed magnetic path or circuit of the material to the flux that would exist in a path of the same dimensions in air, the same m.m.f. being used in both cases. The permeability of air is assigned the value 1. The permeability of steels of various types varies from about 50 to several thousand, depending upon the materials alloyed with the steel. Very high permeabilities are attained in certain special magnetic materials, such as "permalloy," which is an alloy of iron and nickel.

The permeability of magnetic materials depends upon the density of magnetic flux in the material. At very high flux densities the permeability is less than its value at low or moderate flux densities. This is because the flux in magnetic materials is proportional to the applied m.m.f. only over a limited range. As the m.m.f. increases more and more of the molecular magnets within the material become aligned, until eventually a point is reached where a very great increase in m.m.f. is required to cause a relatively small increase in flux. This is called *magnetic saturation*. In this region of saturation the permeability decreases, since the *ratio* between the number of lines in the material and the number in air, for the same m.m.f., is smaller than when the flux density is below the saturation point.

Energy in the magnetic field — Like the electrostatic field (§ 2-3), the magnetic field represents potential energy. Consequently the expenditure of energy is necessary to set up a magnetic field, but once the field has been established and remains constant no further energy is consumed in maintaining it. If by some means the field is caused to disappear, the stored-up magnetic energy is converted to energy in some other form. In other words the energy undergoes a transformation when the magnetic field is *changing*, being stored in the field when the field strength is increasing and being released from the field when the field strength is decreasing.

When a magnetic field is set up by a current flowing in a wire or coil, a certain amount of energy is used initially in bringing the field into existence. Thereafter the current must continue to flow, if the field is to be maintained at steady strength, but no expenditure of energy is required for this purpose. (There will be a steady energy loss in the circuit, but only because of the resistance of the wire.) If the current stops the energy of the field is transformed back into electrical energy, tending to keep the current flowing. The amount of energy stored and subsequently released depends upon the strength of the field, which in turn depends upon the intensity of the current and the circuit conditions; i.e., it depends upon the relationship between field strength and current in the circuit.

Induced voltage — Since a magnetic field is set up by an electric current, it is not surprising to find that, in turn, a magnetic field can cause a current to flow in a closed electrical circuit.

That is, an e.m.f. can be *induced* in a wire in a magnetic field. However, since a *change* in the field is required for energy transformation, an e.m.f. will be induced only when there is a change in the field with respect to the wire.

This change may be an actual change in the field strength or may be caused by relative motion of the field and wire; e.g., a moving field and a stationary wire, or a moving wire and a stationary field. It is convenient to consider this induced e.m.f. as resulting from the wire's "cutting through" the lines of force of the field. The strength of the e.m.f. so induced is proportional to the *rate* of cutting of the lines of force.

If the conductor is moving parallel with the lines of force in a field, no voltage is induced since no lines are cut. Maximum cutting results when the conductor moves through the field in such a way that both its longer dimension and direction of motion are perpendicular to the lines of force, as shown in Fig. 215. When the conductor is stationary and the field strength varies, the induced voltage results from the alternate increase and decrease in the number of lines of force cutting the wire as the m.m.f. varies in intensity.

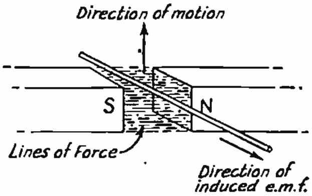


Fig. 215 — Showing how e.m.f. is induced in a conductor moving through a stationary magnetic field, cutting the lines of force. Conversely, a current sent through the conductor in the same direction by means of an external e.m.f. will cause the conductor to move downward.

Lenz's Law — When a voltage is induced and current flows in a conductor moving in a magnetic field, energy of motion is transformed into electrical energy. That is, mechanical *work* is done in moving the conductor when an induced current flows in it. If this were not so the induced voltage would be creating electrical energy, in violation of the fundamental principle of physics that energy can neither be created nor destroyed but only transformed. It is found, therefore, that the flow of current creates an opposing magnetic force tending to stop the movement of the wire. The statement of this principle is known as Lenz's Law: "In all cases of electromagnetic induction, the induced currents have such a direction that their reaction tends to stop the motion which produces them."

Motor principle — The fact that current flowing in a conductor moving through a magnetic field tends to oppose the motion indicates that current sent through a *stationary* conductor in a magnetic field would tend to *set the conductor in motion.* Such is the case. If moving the conductor through the field in the direction indicated in Fig. 215 causes a current to flow as shown, then, if the conductor is stationary and an e.m.f. is applied to send a current through the conductor in the same direction, the conductor will tend to move across the field in the *opposite* direction.

This principle is used in the electric motor. The same rotating machine frequently may be used either as a generator or motor; as a generator it is turned mechanically to cause an induced e.m.f., and as a motor electric current through it causes mechanical motion.

Self-induction — When an e.m.f. is applied to a wire or coil, current begins to flow and a magnetic field is created. Just before closing the circuit there was no field; just after closing it the field exists. Consequently, at the instant of closing the circuit the *rate of change* of the field is very rapid. Since the wire or coil carrying the current is a conductor in a changing field, an e.m.f. will be induced in the wire. This induced voltage is the *e.m.f. of self-induction,* so called because it results from the current flowing in the wire itself.

By the principle of conservation of energy (and Lenz's Law), the polarity of the induced voltage must be such as to oppose the applied voltage; that is, the induced voltage must tend to send current through the circuit in the direction opposite to that of the current caused by the applied voltage. At the instant of closing the circuit the field changes at such a rate that the induced voltage equals the applied voltage (it cannot exceed the applied voltage, because then it would be supplying energy to the source of applied e.m.f.), but after a short interval the rate of change of the field no longer is so rapid and the induced voltage decreases. Thus the current flowing is very small at first when the applied and induced e.m.f.s are about equal, but rises as the induced voltage becomes smaller. The process is cumulative, the current eventually reaching a final value determined only by the resistance in the circuit.

In forcing current through the circuit against the pressure of the induced or "back" voltage, work is done. The total amount of work done during the time that the current is rising to its final value is equal to the amount of energy stored in the magnetic field, neglecting heat losses in the wire itself. As explained before, no further energy is put into the field once the current becomes steady. However, if the circuit is opened and current flow caused by the applied e.m.f. ceases, the field collapses. The rate of change of field strength is very great in this case, and a voltage is again induced in the coil or wire. This voltage causes a current flow in the *same* direction as that of the applied e.m.f., since energy is now being restored to the circuit. The energy usually is dissipated in the spark which occurs when such a circuit is opened. Since the field collapses very rapidly when the switch is opened, the induced e.m.f. at such a time can be extremely high.

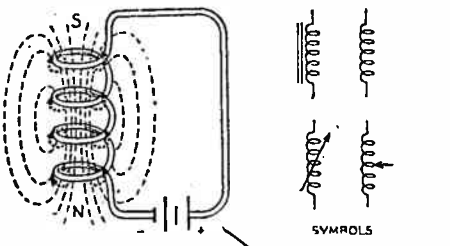


Fig. 216 — When the conducting wire is coiled, the individual magnetic fields of each turn are in such a direction as to produce a field similar to that of a bar magnet. The schematic symbols for inductance are shown at the right. The symbol at the left in the top row indicates an iron-core inductance; at the right, air core. Variable inductances are shown in the bottom row.

Inductance — As explained above, the strength of the self-induced voltage is proportional to the rate of change of the field. However, it is also apparent from the foregoing that the voltage also depends upon the properties of the circuit, since, if a number of similar conductors are in the same varying field, the same voltage will be induced in each. By combining the conductors properly, the total induced voltage in such a case will be the sum of the voltages induced in each wire. Also, the rate of change of field strength depends upon the strength of the field set up by a given amount of current flowing in the wire or coil, and this in turn depends upon the ampere-turns, permeability, length and cross-section of the magnetic path, etc.

For a given circuit, however, the field strength will be determined by the current, and the rate of change of the field consequently will be determined by the rate of change of current. Hence, it is possible to group all of these other factors into one quantity, a property of the circuit. This property is called *inductance.* When this is done, the equation giving the value of the induced voltage becomes:

Induced voltage
$= L \times$ rate of change of current

where L is the value of inductance in the circuit.

Inductance is a property associated with all circuits, although in many cases it may be so small in comparison to other circuit properties (such as resistance) that no error results from neglecting it. The inductance of a straight wire increases with the length of the wire and decreases with increasing wire diameter. The inductance of such a wire is small, however. For a given length of wire, much greater inductance can be secured by winding the wire into a coil so that the total flux from the wire is concentrated into a small space and the flux density correspondingly increased. The unit of inductance is the *henry.* A circuit or coil has an inductance of one henry if an e.m.f. of one volt is induced when the current changes at the rate of one ampere per second. In radio work it is frequently convenient to use smaller units; those commonly used are the *millihenry* (one thousandth of a henry) and the *microhenry* (one millionth of a henry).

It will be recognized that the relationship between inductance and the magnetic field is similar to that between capacity and the electrostatic field. The greater the inductance, the greater the amount of energy stored in the magnetic field for a given amount of current; the greater the capacity, the greater the amount of energy stored in the electrostatic field for a given voltage.

The inductance of a coil of wire depends upon the number of turns, the cross-sectional dimensions of the coil, and the length of the winding. It also depends upon the permeability of the material on which the coil is wound, or *core.* Formulas for computing the inductance of air-core coils of the type commonly used in radio work, are given in Chapter Twenty-One.

Mutual inductance — If two coils are arranged with their axes coinciding, as shown in Fig. 217, a current sent through Coil 1 will cause a magnetic field which cuts Coil 2. Consequently, an e.m.f. will be induced in Coil 2 whenever the field strength is changing. This induced e.m.f. is similar to the e.m.f. of self-induction; that is,

Induced e.m.f.
$= M \times$ rate of change of current

where M is a quantity called the *mutual inductance* of the two coils. The mutual inductance may be large or small, depending upon the self-inductances of the coils and the proportions of the total flux set up by one coil which cuts the turns of the other coil. If all the flux set up by one coil cuts all the turns of the other coil the mutual inductance has its maximum possible value, while if only a small part of the flux set up by one coil cuts the turns of the other the mutual inductance may be relatively small. Two coils having mutual inductance are said to be *coupled.*

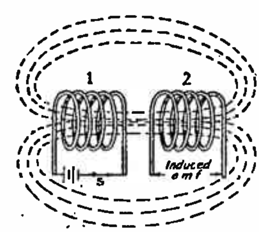


Fig. 217 — Mutual inductance. When the switch, S, is closed current flows through coil No. 1, setting up a magnetic field which induces an e.m.f. in the turns of coil No. 2.

The *degree of coupling* expresses the ratio of actual mutual inductance to the maximum possible value. Coils which have nearly the maximum possible mutual inductance are said to be *closely,* or *tightly,* coupled, while if the mutual inductance is relatively small the coils are said to be *loosely* coupled. The degree of coupling depends upon the physical spacing between the coils and how they are placed with respect to each other. Maximum coupling exists when they have a common axis, as shown in Fig. 217, and are as close together as possible.

If two coils having mutual inductance are connected in the same circuit, the directions of the respective magnetic fields may be such as to add or oppose. In the former case the mutual inductance is said to be "positive"; in the latter case, "negative." Positive mutual inductance in such a circuit means that the total inductance is greater than the sum of the two individual inductances, while negative inductance means that the total inductance is less than the sum of the two individual inductances. The mutual inductance may be made either positive or negative simply by reversing the connections to *one* of the coils.

2-6 Fundamental Relations

Direct current — A current which always flows in the same direction through a circuit is called a *direct current*, frequently abbreviated *d.c.* Current flow caused by batteries, for example, is direct current. One terminal of each cell is always positive and the other always negative, hence electrons are attracted only in the one direction around the circuit. To make the current change direction, the connections to the battery terminals must be reversed.

Work, energy and power — When a quantity of electricity is moved from a point of one potential to a point at a second potential, work is done. The work done is the product of the quantity of electricity and the difference of potential through which it is moved; that is,

$$W = QE$$

In the practical system of units, with Q in coulombs and E in volts, the unit of work is called the *joule*. Energy, which is the capacity for doing work, is measured in the same units.

Since $I = Q/t$ when the current is constant (§ 2-1), $Q = It$. Substituting for Q in the equation above gives

$$W = EIt$$

where E is in volts, I in amperes, and t in seconds. One ampere flowing through a difference of potential of one volt for one second does one joule of work. *Power* is the time rate at which work is done, so that, if the work is done at a uniform rate, dividing the equation by t will give the electrical power:

$$P = EI$$

The unit of electrical power, P, is the *watt*.

In practical work, the term "joule" is seldom used for the unit of work or energy. The more common name is *watt-second* (one joule is equal to one watt applied for one second). The watt-second is a relatively small unit; a larger one, the *watt-hour* (one watt of power applied for one hour) is more frequently used. Again, for some purposes the watt is too small a unit, and the *kilowatt* (1000 watts) is used instead. A still larger energy unit is the *kilowatt-hour*, the meaning of which is easily interpreted.

Fractional and multiple units — As illustrated by the examples in the preceding paragraph, it is frequently convenient to change the value of a unit so that it will not be necessary to use very large or very small numbers. As applied to electrical units, the practice is to add a prefix to the name of the fundamental unit to indicate whether the modified unit is larger or smaller. The common prefixes are *micro* (one millionth), *milli* (one thousandth), *kilo* (one thousand) and *mega* (one million). Thus, a microvolt is one millionth of a volt, a milliampere is one thousandth of an ampere, a kilovolt is one thousand volts, and so on.

Unless there is some indication to the contrary, it should be assumed that, whenever a formula is given in terms of unprefixed letters (E, I, P, R, etc.), the fundamental units are meant. If the quantities to be substituted in the equation are given in fractional or multiple units, conversion to the fundamental units is necessary before the equation can be used.

Ohm's Law — In any metallic conductor, the current which flows is directly proportional to the applied electromotive force. This relationship, known as *Ohm's Law*, can be written

$$E = RI$$

where E is the e.m.f., I is the current, and R is a constant, depending on the conductor, called the *resistance* of the conductor. By definition, a conductor has one unit of resistance when an applied e.m.f. of one volt causes a current of one ampere to flow. The unit of resistance is called the *ohm*.

Ohm's Law does not apply to all types of conduction, particularly to conduction through gases and in a vacuum. The law is of very great importance, however, because practically all electrical circuits use metallic conduction.

By transposing the equation, the following equally useful forms are obtained:

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

The three equations state that, in a circuit to which Ohm's Law applies, the voltage across the circuit is equal to the current multiplied by the resistance; the resistance of the circuit is equal to the voltage divided by the current; and the current in the circuit is equal to the voltage divided by the resistance.

Resistance and resistivity — The resistance of a conductor is determined by the material of which it is made and its temperature, and is directly proportional to the length of the conductor (that is, the length of the path of the current through the conductor) and inversely proportional to the area through which the current flows. If the temperature is constant,

$$R = k\frac{L}{A}$$

where R is the resistance, k is a constant depending upon the material of which the conductor is made, L is the length and A the area. For the purpose of giving a specific value to k, L is taken as one centimeter and A as one square centimeter (a cube of the material measuring one centimeter on a side); k is then the resistance in ohms of such a cube at a specified temperature. It is called the *specific resistance* or *resistivity* of the material. If the resistivity is known, the resistance of any conductor of known length and uniform cross-section readily can be determined by the formula above. The length must be in centimeters and the area in square centimeters.

The relationships given above are true only for unidirectional (direct) currents and low-frequency alternating currents. Modifications must be made when the current reverses its direction many times each second (§ 2-8).

Conductance and conductivity — The reciprocal of resistance is called *conductance*, and has the opposite properties to resistance. The lower the resistance of a circuit, the higher is the conductance, and vice versa. The symbol of conductance is G, and the relationship to resistance is

$$G = \frac{1}{R} \qquad R = \frac{1}{G}$$

The unit of conductance is called the *mho*. A circuit or conductor which has a resistance of one ohm has a conductance of one mho. By substituting $1/G$ for R in Ohm's Law,

$$G = \frac{I}{E} \qquad I = EG \qquad E = \frac{I}{G}$$

The reciprocal of resistivity is called the *specific conductance* or *conductivity* of a material, and is measured in mhos per centimeter cube. It is frequently useful to know the *relative* conductivity of different materials. This is usually expressed in *per cent conductivity*, the conductivity of annealed copper being taken as 100 per cent. A table of per cent conductivities is given in Chapter Twenty-One.

Power used in resistance — If two conductors of different resistances have the same current flowing through them, then by Ohm's Law the conductor with the larger resistance will have a greater difference of potential across its terminals. Consequently, more energy is supplied to the larger resistance, since in a given period of time the same amount of electricity is moved through a greater potential difference. The energy appears in the form of heat in the conductor. With a steady current, the heat will raise the temperature of the conductor until a balance is reached between the heat generated and that radiated to the surrounding air or otherwise carried away.

Since $P = EI$, substituting for E the appropriate form of Ohm's Law ($E = IR$) gives

$$P = I^2R$$

and making a similar substitution for I gives

$$P = \frac{E^2}{R}$$

That is, the power used in heating a resistance (or *dissipated* in the resistance) is proportional to the square of the voltage applied or to the square of the current flowing. In these formulas P is in watts, E in volts and I in amperes.

Further transposition of the equations gives the following forms, useful when the resistance and power are known:

$$E = \sqrt{PR} \qquad I = \sqrt{\frac{P}{R}}$$

Unless the circuit containing the resistance is being used for the specific purpose of generating heat, the power used in heating a resistance is generally considered as a loss. However, there are very many applications in radio circuits where, despite the loss of power, a useful purpose is served by introducing resistance deliberately. Resistances made to specified values and provided with connecting terminals are called *resistors*. They are frequently wound on ceramic or other heat-resisting tubing with wire having high resistivity.

Temperature coefficient of resistance — The resistance of most pure metals increases with an increase in temperature. The resistance of a wire at any temperature is given by

$$R = R_0\,(1 + at)$$

where R is the required resistance, R_0 the resistance at 0°C. (temperature of melting ice), t is the temperature (Centigrade), and a is the *temperature coefficient of resistance*. For copper, a is about 0.004; that is, starting at 0°C., the resistance increases 0.4 per cent per degree above zero.

Temperature coefficient of resistance becomes of importance when conductors operate at high temperatures. In the case of resistors used in electrical and radio circuits, the heat developed by current flow may raise the temperature of the resistance wire to several hundred degrees F. Thus the resistance at operating temperatures can be very much higher than the resistance at room temperature. Consequently such resistors are wound with wire which has a low temperature coefficient of resistance, so that the resistance will be more nearly constant under all conditions.

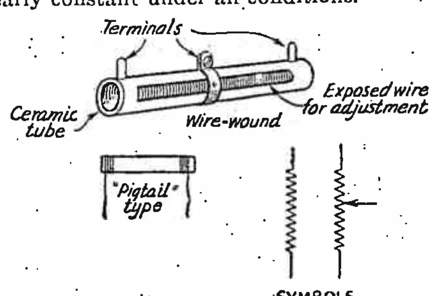


Fig. 218 — Two common types of fixed resistors. The wire-wound type is used for dissipating power of the order of 5 watts or more. "Pigtail" resistors, usually made of carbon or other resistance material in the form of a molded rod or as a thin coating on an insulating tube, rather than being wound with wire, are small in size but do not safely dissipate much power. Schematic symbols for fixed and variable resistors are shown at lower right.

Resistances in series — When two or more resistances are connected so that the same current flows through each in turn, as shown in Fig. 219, they are said to be connected *in series*. Then, by Ohm's Law,

$$E_1 = IR_1$$
$$E_2 = IR_2$$
$$E_3 = IR_3$$

etc., where the subscripts 1, 2, 3 indicate the first, second and third resistor, and the voltages E_1, E_2 and E_3 are the voltages appearing across the terminals of the respective resistors. Adding the three voltages gives the total voltage across the three resistors:

$$E = E_1 + E_2 + E_3 = IR_1 + IR_2 + IR_3 = I\,(R_1 + R_2 + R_3) = IR$$

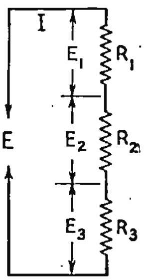


Fig. 219 — Resistances in series.

That is, the voltage across the resistors in series is equal to the current multiplied by the sum of the individual resistances. In the above equation, R, which denotes this sum, may be called the *equivalent* resistance or *total* resistance. The equivalent resistance of a number of resistors connected in series is, therefore, equal to the sum of the values of the individual resistors.

Resistances in parallel — When a number of resistances are connected so that the same voltage is applied to all, as shown in Fig. 220,

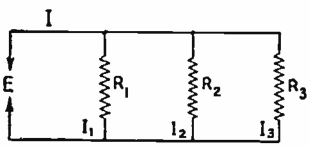


Fig. 220 — Resistances in parallel.

they are said to be connected *in parallel.* By Ohm's Law,

$$I_1 = \frac{E}{R_1} \qquad I_2 = \frac{E}{R_2} \qquad I_3 = \frac{E}{R_3}$$

so that the total current, I, which is the sum of the currents in the individual resistors, is

$$I = I_1 + I_2 + I_3 = \frac{E}{R_1} + \frac{E}{R_2} + \frac{E}{R_3} =$$

$$E\left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right) = E\,\frac{1}{R}$$

where R is the equivalent resistance — i.e., the resistance through which the same total current would flow if such a resistance were substituted for the three shown. Therefore,

$$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$$

That is, the reciprocal of the equivalent resistance of a number of resistances in parallel is equal to the sum of the reciprocals of the individual resistances. Since the reciprocal of resistance is conductance,

$$G = G_1 + G_2 + G_3$$

where G is the total conductance and G_1, G_2, G_3, etc., are the individual conductances in parallel.

To obtain R instead of its reciprocal the equation above may be inverted, so that

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}}$$

The number of terms in the denominator of this equation will, of course, be equal to the actual number of resistors in parallel.

For the special case of only two resistances in parallel, the equation reduces to

$$R = \frac{R_1\,R_2}{R_1 + R_2}$$

Series-parallel connection of resistors is shown in Fig. 221. When circuits of this type are encountered the equivalent or total resistance can be found by first adding the series resistances in each group, then treating each group as a single resistor so that the formula for resistors in parallel can be used.

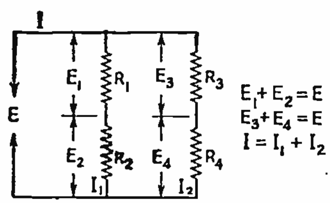


Fig. 221 — Series-parallel connection of resistances. Voltage and current relationships are given at the right.

Voltage dividers and potentiometers — Since the same current flows through resistors connected in series, it follows from Ohm's Law that the voltage (termed *voltage drop*) across each resistor of a series-connected group is proportional to its resistance. Thus, in Fig. 222-A, the voltage E_1 across R_1 is equal to the applied voltage, E, multiplied by the ratio of R_1 to the total resistance, or

$$E_1 = \frac{R_1}{R_1 + R_2 + R_3} \cdot E$$

Similarly, the voltage, E_2, is equa lto

$$\frac{R_1 + R_2}{R_1 + R_2 + R_3} \cdot E$$

Such an arrangement is called a *voltage divider,* since it provides a means for obtaining smaller voltages from a source of fixed voltage. When current is drawn from the divider at the various tap points the above relations are no longer strictly true, for then the same current does not flow in all parts of the divider. Design data for such cases are given in § 8–10.

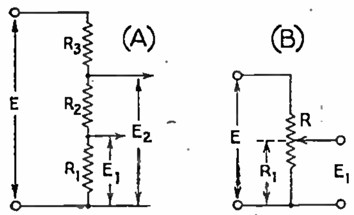


Fig. 222 — Voltage divider (A) and potentiometer (B).

A similar arrangement is shown in Fig. 222-B, where the resistor, R, is equipped with a sliding tap for fine adjustment. Such a variable resistor is frequently called a *potentiometer.*

Inductances in series and parallel — As explained in § 2-5, inductance determines the voltage induced when the current changes at a given rate. That is, $E = L \times$ rate of change of current. This resembles Ohm's Law, if L corresponds to R and the rate of change of current to I. Thus, by reasoning similar to that used in the case of resistors, it can be shown that, for inductances in series,

$$L = L_1 + L_2 + L_3$$

and for inductances in parallel,

$$L = \cfrac{1}{\cfrac{1}{L_1} + \cfrac{1}{L_2} + \cfrac{1}{L_3}}$$

where the number of terms in either equation is determined by the actual number of inductances connected in series or parallel.

These equations do not hold if there is mutual inductance (§ 2-5) between the coils.

Condensers in series and parallel — When a number of condensers are in parallel, as in Fig. 223-A, the same e.m.f. is applied to all. Consequently, the quantity of electricity stored in each is in proportion to its capacity. The total quantity stored is the sum of the quantities in the individual condensers:

$$Q = Q_1 + Q_2 + Q_3 = C_1E + C_2E + C_3E = (C_1 + C_2 + C_3)E = CE$$

where C is the equivalent capacity. The equivalent capacity of condensers in parallel is equal to the sum of the individual capacities.

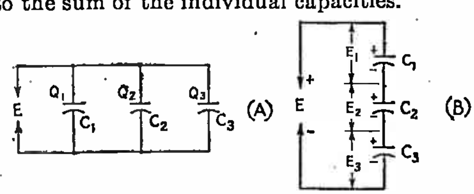


Fig. 223 — Condensers in parallel (A) and in series (B).

When condensers are connected in series, as in Fig. 223-B, the application of an e.m.f. to the circuit causes a certain quantity of electricity to accumulate on the top plate of C_1. By electrostatic induction, an equal charge of opposite polarity (negative in the illustration) appears on the bottom plate of C_1, and, since the lower plate of C_1 and the upper plate of C_2 are connected together, this must leave an equal positive charge on the upper plate of C_2. This, in turn, causes the lower plate of C_2 to assume an equal negative charge, and so on down to the plate connected to the negative terminal of the source of e.m.f. In other words the same quantity of electricity is placed on each condenser, and this is equal to the total quantity stored. The voltage across each condenser will depend upon its capacity, and the sum of these voltages must equal the applied voltage. Thus,

$$E = E_1 + E_2 + E_3 = \frac{Q}{C_1} + \frac{Q}{C_2} + \frac{Q}{C_3} = Q\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right) = \frac{Q}{C}$$

where C is the equivalent capacity. This leads to an expression similar to that for resistances in parallel:

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

where the number of terms in the denominator should be the same as the actual number of condensers in series.

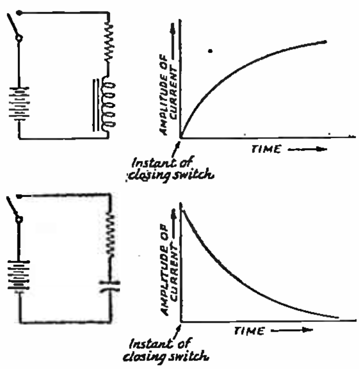
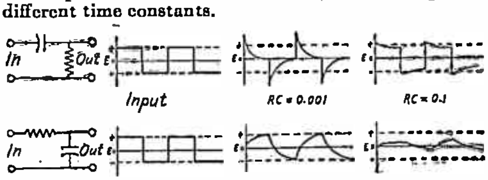


Fig. 224 — The RC and LC circuits at the left, together with the curves of current amplitude vs. time, show how the current in a circuit combining resistance with inductance or capacity takes a finite period of time to reach a steady-state value.

Below — With square wave input, the voltage waveshapes across R and C respectively in an RC circuit have the shapes shown. Note the variation in waveshapes for different time constants.



Time Constant — A charged condenser which had infinite resistance between its plates would hold a charge indefinitely at its full initial value. However, since all practical condensers do have a finite value of resistance (through the dielectric material and between the connecting terminals), the charge gradually leaks off. Good-quality condensers have very high "leakage resistance," however, and will hold a charge for days if left undisturbed.

In a circuit containing only capacity and resistance, the time required for the potential difference between the charged plates of a condenser to fall to a definite percentage of its initial value is determined by the capacity of the condenser and the value of the resistance. When a condenser and resistor are connected in series with a source of e.m.f., such as a battery, the initial flow of current into the condenser is limited by the resistance, so that a longer period of time is required to complete the charging of the condenser than would be the case without the resistor. Likewise, when the condenser is discharged through a resistor a measurable period of time is taken for the current flow to reach a negligible value. In the case of either charge or discharge the time required is proportional to the capacity and resistance, the product of which is called the *time constant* of the circuit. If C is in farads and R in ohms, or C in microfarads and R in megohms, the product gives the time in seconds required for the voltage across a discharging condenser to drop to $1/\epsilon$, or approximately 37 per cent of its original value. (The constant ϵ is the base of the natural series of logarithms.)

A circuit containing inductance and resistance also has a time constant, for similar reasons. The time constant of an inductive circuit is equal to L/R, where L is in henrys and R in ohms. It gives the time in seconds required for the current to reach $1-1/\epsilon$, or approximately 63 per cent of its final steady value when a constant voltage is applied.

The significant point concerning these relationships is that, by proper application to associated circuits, it is possible to create almost any desired wave or pulse shape with extreme accuracy and recurrent uniformity including timing with the precision of millionths of a second. The relation is of practical importance in many circuit applications in amateur transmission and reception, as in electronic keyers, time delay with automatic volume control, resistance-capacity filters and remote control. Apart from these subsidiary applications, on this principle are based the fundamentals of the specialized techniques employed in television and many electronic devices of both a military and civilian character.

Measuring instruments — Instruments for measuring d.c. current and voltage make use of the force acting on a coil carrying current in a magnetic field (§ 2-5), produced by a permanent magnet, to move a pointer along a calibrated scale. The magnetic field may be produced by a permanent magnet acting upon a moving coil, or by a fixed coil acting upon a moving iron vane or plunger.

The first type of instrument, based on what is known as the d'Arsonval moving-coil movement, is shown at the left in Fig. 225. The moving-iron vane instrument shown at the right is less accurate and requires higher energizing current, making it relatively insensitive as compared to the moving-coil type. Only the cheaper

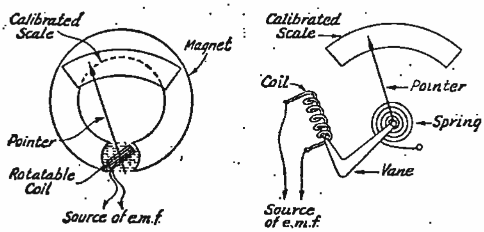


Fig. 225 — *Left* — The d'Arsonval or moving-coil meter for d.c. current measurement. Current flowing through the rotatable coil in the field of the permanent magnet causes a force to act on the coil, tending to turn it. The turning tendency is counteracted by springs (not shown) so that the amount of movement is proportional to the value of the current in the coil. *Right* — In the simpler moving-iron-vane type, a light-weight soft-iron plunger is attracted by current flowing in a fixed coil. As the plunger moves the pointer to which it is linked also moves, until the magnetic force in the coil is balanced by the spiral spring restraining the plunger movement.

measuring instruments available to amateurs are based on this principle.

In such instruments the current required for full-scale deflection of the pointer varies from several milliamperes to a few microamperes, according to the sensitivity required. If the instrument is to read high currents, it is *shunted* (paralleled) by a low resistance through which most of the current flows, leaving only enough flowing through the instrument to give a full-scale deflection corresponding to the total current flowing through both meter and shunt. An instrument which reads microamperes is called a *microammeter* or *galvanometer;* one calibrated in milliamperes is called a *milliammeter;* one calibrated in amperes is an *ammeter*. A *voltmeter* is simply a milliammeter with a high resistance in series so that the current will be limited to a suitable value when the instrument is connected across a voltage source; it is calibrated in terms of the voltage which must appear across the terminals to cause a given value of current to flow. The series resistance is called a *multiplier*. A *wattmeter* is a combination voltmeter and ammeter in which the pointer deflection is proportional to the power in the circuit.

An ammeter or milliammeter is connected in series with the circuit in which current is being measured, so that the current flows through the instrument. A voltmeter is connected in parallel with the circuit.

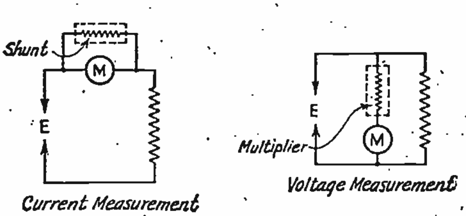


Fig. 226 — Circuit connections for measuring current and voltage. The shunt resistor is used for increasing the value of the current which the instrument can measure, by providing an alternate path through which some of the current can flow. The series multiplier limits the current when the instrument is used to measure voltage.

2-7 Alternating Current

Description — An alternating current is one which periodically reverses its direction of flow. In addition to this alternate change in direction, usually the amount or *amplitude* of the current also varies continually during the period when the current is flowing in one direction. These variations are accompanied by corresponding variations in the magnetic field set up by the current, and it is this feature which makes the alternating current so useful. By means of the varying field, energy may be continually transferred (by induction) from one circuit to another without direct connection, and the voltage may be changed in the process. Neither of these is possible with direct current because, except for brief periods when the circuit is closed or opened, the field accompanying a steady direct current is unchanging, and hence there is no way of inducing an e.m.f. except by moving a conductor through the field (§ 2-5).

Alternating currents may be generated in several ways. Rotating electrical machines (*a.c. generators* or *alternators*) are used for developing large amounts of power when the rate of reversal is relatively slow. However, such machines are not suitable for producing currents which reverse direction thousands or millions of times each second. The thermionic vacuum tube is used for this purpose, as described in Chapter Three.

The simplest form of alternating current (or voltage) is shown graphically in Fig. 227. This chart shows that the current starts at zero value, builds up to a maximum in one direction, comes back down to zero, builds up to a maximum in the opposite direction and comes back to zero. The curve follows the sine law and is known as a *sine wave*, because of the wavelike nature of the curve which results when sine values are plotted on rectangular coördinates as a function of angle or time.

An alternating current or voltage of sine-wave form has angular velocity, period, frequency and phase.

Units of frequency — Alternating currents are identified by their frequency, the basic unit for which is the number of cycles per second. In radio work, where frequencies are extremely large, it is convenient to use two other units, *kilocycles* per second (cycles per second ÷ 1000) and megacycles per second (cycles per second ÷ 1,000,000). These are usually abbreviated kc. and Mc., respectively. Occasionally these abbreviations are written kcs. and Mcs. to indicate "kilocycles per second" and "megacycles per second" rather than simply "kilocycles" and "megacycles," but in both written and spoken usage it is understood that "per second" is an integral part of the term when the shorter forms are used.

Peak, instantaneous effective and average values — The highest value of current or voltage during the time when the current is flowing in one direction is called the *maximum* or *peak* value. For the sine wave, the peak has the same absolute value on both the positive and negative halves of the cycle. This is not necessarily true of waves having shapes other than the true sine form.

The value of current or voltage existing at any particular point of time in the cycle is called the *instantaneous* value. The instant for which a particular value is to be found can be specified in terms of time (fraction of the period) or of angle.

Since both the voltage and current are swinging continuously between their positive maximum and negative maximum values, it might be wondered how one can speak of so many amperes of alternating current when the value is changing continuously. The problem is simplified in practical work by considering that an alternating current has an effective value of one ampere when it produces heat, in flowing through a given resistance, at the same average rate as one ampere of continuous direct current flowing through the same resistance. This effective value is the square root of the mean of all of the instantaneous current values squared. In the case of the sine-wave form,

$$E_{eff} = \sqrt{\tfrac{1}{2}E^2_{max}}$$

For this reason, the effective value of an alternating current or voltage is also known as the *root-mean-square*, or *r.m.s.*, value. Hence, the effective value is the square root of ½, or 0.707, times the maximum value.

In a purely a.c. circuit the average current over a whole cycle must be zero, because if the average current on, say, the positive half of the cycle were greater than the average on the negative half, there would be a net current flow in the positive direction. This would correspond to a direct (although intermittent) current, and hence must be excluded because a purely alternating current was assumed. The "average" value of an alternating current is defined as the average current during the part of the cycle when the current is flowing in one direction only. It is of particular importance when alternating current is changed to direct current by the methods considered in later chapters. For a sine wave, the average value is equal to 0.636 of the peak value.

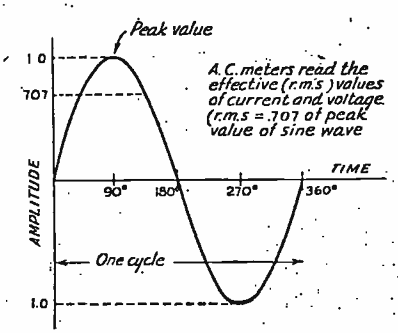


Fig. 227 — Sine wave of alternating current or voltage.

In the sine wave the three voltage values, peak, effective and average, are related to each other as follows:

$$E_{max} = E_{eff} \times 1.414 = E_{ave} \times 1.57$$
$$E_{eff} = E_{max} \times 0.707 = E_{ave} \times 1.11$$
$$E_{ave} = E_{max} \times 0.636 = E_{eff} \times 0.9$$

The relationships for current are equivalent to those given above for voltage.

Current, voltage and power in an inductance — When alternating current flows through an inductance, the continually varying magnetic field causes the continuous generation of an e.m.f. of self-induction (§ 2-5). The induced voltage at any instant is proportional to the rate at which the current is changing at that instant. If the current is a sine wave, it can be shown that the rate of change is greatest when the current is passing through zero and least when the current is maximum. For this reason, the induced voltage is maximum when the current is zero and zero when the current is maximum. The direction or polarity of the induced voltage is such as to tend to sustain the current flow when the current is decreasing and to prevent it from flowing when the current is increasing (§ 2-5). As a result, the induced voltage in an inductance lags 90 degrees behind the current.

By Lenz's Law, the induced voltage must always oppose the applied voltage; that is, the induced and applied voltages must be in phase opposition, or 180 degrees out of phase. Consequently, the *applied* voltage leads the current by 90 degrees. These relationships are shown in Fig. 228. Using the voltage as a reference,

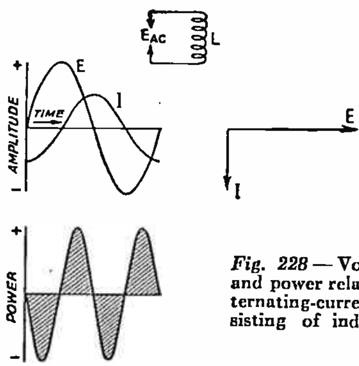


Fig. 228 — Voltage, current and power relations in an alternating-current circuit consisting of inductance only.

the current in an inductance lags 90 degrees, or one-quarter cycle, behind the voltage. (In a vector diagram, the current and voltage cannot be added vectorially because they are different kinds of quantities. In diagrams showing both current and voltage, one is generally used simply as a reference, to establish phase relationships. If several currents are shown in the same circuit, they can be added together; similarly, several voltages in the same circuit can be added. However, where use is made of vector diagrams to show phase relationships in *different* circuits, the vectors representing current and/or voltage in different circuits cannot be added.

When the current is increasing in either direction, energy is being stored in the magnetic field. At such times the voltage has the same polarity as the current, so that the product of the two, which gives the instantaneous power fed to the inductance, is positive. When the current is decreasing energy is being restored to the circuit and the applied voltage has the opposite polarity, so that the product of current and voltage is negative. This is also shown in Fig. 228. Positive power means power taken from the source (i.e., the source of the applied e.m.f.), while negative power means power returned to the source. Power is alternately taken and given back in each quarter cycle, and, since the amount given back is the same as that taken, the *average* power in an inductance is zero when considering a whole cycle. In a practical inductance the wire will have some resistance, so that some of the power supplied will be consumed in heating the wire, but if the resistance of the circuit is small compared to the inductance the power consumption is very small compared to the power which is alternately stored and returned.

Current, voltage and power in a condenser — When an alternating voltage is applied to a condenser, the condenser acquires a charge while the voltage is rising and loses its charge while the voltage is decreasing. The quantity of electricity stored in the condenser at any instant is proportional to the voltage across its terminals at that instant ($Q = CE$). Since current is the *rate* of transfer of quantity of electricity, the current flowing into the condenser (when it is being charged) or out of it (when it is discharging) consequently will be proportional to the rate of change of the applied voltage. If the voltage is a sine wave, its rate of change will be greatest when passing through zero and least when the voltage is maximum. As a result, the current flowing into or out of the condenser is greatest when the voltage is passing through zero and least when the voltage reaches its peak value.

This relationship is shown in Fig. 229. Whenever the voltage is rising (in either direction) the current flow is in the same direction as the applied voltage. When the voltage is decreasing and the condenser is discharging, the current flows in the opposite direction. This is of course the normal condition for charge and discharge of a condenser. The energy stored in the condenser on the charging part of the cycle is restored to the circuit on the discharge part, and the total energy consumed in a whole cycle therefore is zero. A condenser operating on a.c. takes no average power from the source, except for such actual energy losses as may occur as the result of heating of the dielectric (§ 2-3). The energy loss in air condensers used in radio circuits is negligibly small except at extremely high frequencies.

As shown by Fig. 229, the phase relationship between current flow and applied voltage is such that the current leads the voltage by 90 degrees. This is just the opposite to the inductance case.

Current, voltage and power in resistance — In a circuit containing resistance only there are no energy storage effects, and consequently the current and voltage are in phase. The current therefore always flows in the same direction as the applied voltage, and, since the power is always positive, there is continual power *dissipation* in the resistance. The relationships are shown in Fig. 230.

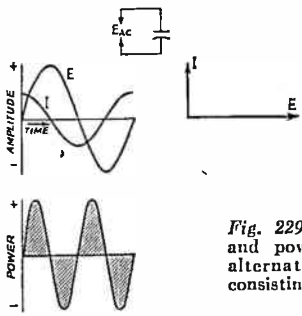


Fig. 229 — Voltage, current and power relations in an alternating-current circuit consisting of capacity only.

Strictly speaking, no circuit can have resistance only, because the flow of current always is accompanied by the creation of a magnetic field and every conductor also has a certain amount of capacity. Whether or not such residual inductance and capacity are large enough to require consideration is determined by the frequency at which the circuit is to operate.

The a.c. spectrum — Alternating currents of different frequencies have different properties and are useful in a variety of ways. For the transmission of power to light homes, run motors and perform familiar everyday tasks by electrical means, low frequencies are most suitable. Frequencies of 25, 50 and 60 cycles are in common use, the latter being most widely used in this country. The range of frequencies between about 15 and 15,000 cycles is known as the *audio-frequency* range, because when frequencies of this order are converted from a.c. into air vibrations, as by a loudspeaker or telephone receiver, they are distinguishable as sounds having a tone pitch proportional to the frequency. Frequencies above 15,000 cycles (15 kilocycles) are used for radio communication, because at frequencies of this order it is possible to convert electrical energy into radio waves which can be radiated over long distances.

For convenience in reference, the following classifications for radio frequencies have been recommended by an international technical conference and are now increasingly in use:

10 to 30 kilocycles	Very-low frequencies
30 to 300 kilocycles	Low frequencies
300 to 3000 kilocycles	Medium frequencies
3 to 30 megacycles	High frequencies
30 to 300 megacycles	Very-high frequencies
300 to 3000 megacycles	Ultrahigh frequencies
3000 to 30,000 megacycles	Superhigh frequencies

Until recently, other terminology was used; for example, the region above 30 megacycles formerly was considered the "ultrahigh" frequencies.

2-8 Ohm's Law for Alternating Currents

Resistance — Since current and voltage are always in phase through a resistance, the instantaneous relations for a.c. are equivalent to those in d.c. circuits. By definition, the effective units of current and voltage for a.c. are made equal to those for d.c. in resistive circuits (§ 2-7). Therefore the various formulas expressing Ohm's Law for d.c. circuits apply without any change to a.c. circuits containing resistance only, or for purely resistive parts of complex a.c. circuits. See § 2-6.

In applying the formulas, it must be remembered that consistent units must be used. For example, if the instantaneous value of current is used in finding voltage or power, the voltage found will be the instantaneous voltage and the power will be the instantaneous power. Likewise, if the effective value is used for one quantity in the formula, the unknown will be expressed in effective value. Unless otherwise indicated, the effective value of current or voltage is always understood to be meant when reference is made to "current" or "voltage."

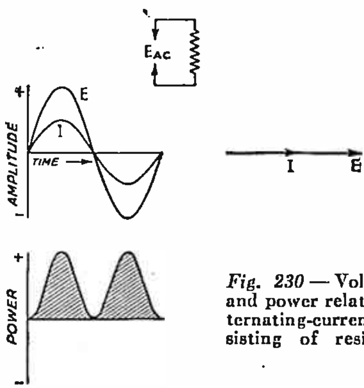


Fig. 230 — Voltage, current and power relations in an alternating-current circuit consisting of resistance only.

Reactance — In an a.c. circuit containing inductance or capacity, but no resistance, there is no consumption of power — simply a continuous back-and-forth transfer of energy between the magnetic or electric field and the circuit (§ 2-7). As shown in section 2-7, neither a perfect condenser nor a perfect inductance will absorb power from the source of an applied voltage since the energy absorbed or stored on one half cycle of alternation is released in the next. In the process of absorbing and giving up this energy, however, opposition is offered to the flow of current through the circuit. This opposition is called *reactance*. Since the average power is zero, the Ohm's Law formulas cannot be applied in terms of peak or effective voltages and currents. Nevertheless, if the frequency is constant the current which flows in an inductive or capacitive circuit is directly proportional to the voltage applied; that is,

$$E = XI$$

where X is a constant depending upon the circuit, and is called the *reactance* of the circuit. By transposition, the formula for reactance can be written

$$X = \frac{E}{I} \qquad I = \frac{E}{X}$$

These expressions are quite similar to those for the resistive circuit, and the quantity X has the same effect upon current flow as does resistance in a resistive circuit. Consequently, the ohm is used as the unit of reactance, just as it is for resistance. Unlike resistance in a circuit, however, reactance does not use up or dissipate power.

Reactance may be compared with resistance by saying that, where the latter represents the inertia or opposition to electron flow in a conductor which can be overcome only by a continuing expenditure of actual energy, reactance represents the inertia or opposition of the circuit to the varying electric charges or fields created by the moving electrons. Such opposition, once initially overcome, requires no actual expenditure of energy.

Inductive reactance — A voltage may be induced in a wire (§ 2-5) either by moving a wire through a magnetic field or by moving a magnetic field through the wire. In an inductance coil the expanding and contracting field is analogous to the latter case. When alternating current flows through an inductance, the induced voltage and applied voltage are equal (§ 2-7). Since the induced voltage is equal to the inductance of the coil multiplied by the rate of change of current, it is evident that the same given value of voltage (to oppose a fixed applied voltage) can be induced either by (a) using a large inductance and a small rate of change of current or (b) by using a large rate of current change and a small inductance, so long as the product of the two is constant. Thus the amount of such voltage is automatically increased by simply increasing the frequency of the applied voltage, inasmuch as the magnetic field changes at the same rate as the applied voltage.

The rate of change of current is determined

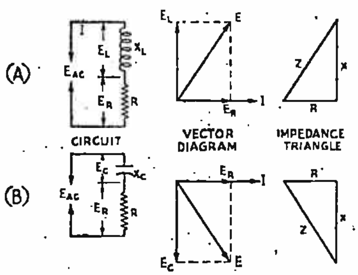


Fig. 231 — Voltage and current relationships in a.c. circuits having resistance and inductance or resistance and capacity. Vector diagrams should be drawn to scale for particular values of voltage or reactance and resistance.

by the amplitude of the current and the angular velocity.

It is the induced voltage acting in opposition to the applied voltage that creates what is termed the reactance of a coil or its ability to impede the flow of alternating current through it. The ohmic value of reactance, however, does not equal the numerical value of induced voltage for the induced voltage actually impedes the flow of current one moment and assists it the next. As a result the voltage, E, necessary to push a given current, I, through a given inductance, L, at a given frequency, f, is a combination of these factors. Therefore, for sine-wave current,

$$E = 2\pi fLI, \text{ or } \frac{E}{I} = 2\pi fL$$

Since $X = E/I$, then

$$X_L = 2\pi fL.$$

where the subscript L indicates that the reactance is inductive.

It is apparent that inductive reactance is proportional both to inductance and to the applied frequency. At low frequencies a large inductance must be used to obtain high reactance, but at very high frequencies the same value of reactance can be obtained readily with quite a small inductance.

The value of inductive reactance equals the rate of current change ($2\pi f$) times the inductance in the coil. Therefore the reactance, X_L, varies directly with either frequency and/or inductance; i.e.: doubling the frequency doubles the reactance; doubling the inductance doubles the reactance, and doubling both at the same time quadruples the reactance.

Observe that the term I in the above formula is a coefficient of E but not of X_L. This means that, although the induced voltage in a coil increases with increased current flow, the increased current flow has no effect upon the reactance of the coil.

The fundamental units (ohms, cycles, henrys) must be used in the above equation, or appropriate factors inserted if other units are employed. If inductance is in millihenrys, the frequency should be stated in kilocycles; if inductance is in microhenrys, the frequency should be given in megacycles, to bring the answer in ohms.

Capacitive reactance — The inherent ability of a condenser to impede current flow (reactance) is mainly a result of the opposing voltage it can offer to oppose the applied voltage but, as in the case of inductance, the value of reactance does not equal the numerical value of the opposing voltage since this voltage, too, alternately opposes and assists current flow. One of the important factors affecting current flow in a condenser is the angular velocity ($2\pi f$) or the rate of voltage change. With a fixed value of applied voltage, the quantity of electricity stored in a condenser of given capacity is always the same ($Q = CE$), and, if losses are negligible, the same quantity of electricity is taken out of the condenser on discharge. When an a.c. voltage is applied to a condenser the alternate charge and discharge, as the applied voltage rises and falls and reverses polarity, constitutes current flow "through" the condenser. The amplitude of the current is proportional to the rate of change of the voltage (§ 2-7) and also to the capacity of the condenser, since both these quantities increase the amount of electricity transferred in the circuit in a given period of time. The larger the capacity the greater is the charge required to establish a given potential difference between its plates. A condenser which requires one coulomb of electricity to bring the potential difference between its plates to a value of one volt is said to have a capacity of one farad. Doubling the capacity of the condenser will double the alternating current flow, hence current flow in a condenser varies directly with a change in its capacity.

Obviously, current flow in a resistor, coil or condenser across which a fixed voltage is applied cannot increase unless the opposition to current flow is correspondingly decreased. If the current doubles, the opposition must be halved. Doubling the capacity has, therefore, halved the opposition (reactance) of the condenser. Hence, the value of reactance, X_C, of a condenser varies inversely with a change in its capacity.

When two equal voltages, regardless of their frequencies, are applied separately to the same condenser, the condenser will take equal charges in both instances ($Q = CE$, § 2-3), but the current flow will double when the frequency alone is doubled. Since rate of change of voltage is proportional to the amplitude of the voltage and the angular velocity, then, for sine-wave voltage,

$$I = 2\pi fCE, \text{ or } \frac{E}{I} = \frac{1}{2\pi fC}$$

Since $X = E/I$, then

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

where the subscript C indicates that the reactance is capacitive. Capacitive reactance is *inversely* proportional to capacity and to the applied frequency. For a given value of capacity, the reactance decreases as the frequency increases.

Fundamental units (farads, cycles per second) must be used in the right-hand side of the equation to obtain the reactance in ohms. Conversion factors must be used if the frequency and capacity are in units other than cycles and farads. If C is in microfarads and f in megacycles, the conversion factors cancel and the reactance will be given in ohms.

Impedance — In any series circuit the same current flows through all parts of the circuit. If a resistance and inductance are connected in series to form an a.c. circuit they both carry the same current, but the voltage across the resistance is in phase with the current while the voltage across the inductance leads the current by 90 degrees. In a d.c. circuit with resistances in series, the applied voltage is equal to the sum of the voltages across the individual resistances (§ 2-6). This is also true of the a.c. circuit with resistance and inductance in series if the *instantaneous* voltages are added algebraically to find the instantaneous value of applied voltage. But, because of the phase difference between the two voltages, the maximum value of the applied voltage will not be the sum of the maximum values of the two voltages (§ 1-9), so that the effective values cannot be added directly.

The relationships are shown by means of vectors in Fig. 231. If the current vector is used as a reference, the voltage across the resistance is in phase with the current and hence lies on the same line. The voltage across the inductance is 90 degrees ahead of the current, and therefore is drawn at right angles upward. The resultant voltage is, consequently, the hypothenuse of a right triangle, and, by geometry,

$$E^2 = E^2_L + E^2_R, \text{ or } E = \sqrt{E^2_R + E^2_L}$$

Since $E_R = IR$ and $E_L = IX_L$, substitution gives

$$E = I\sqrt{R^2 + X^2_L}, \text{ or } \frac{E}{I} = \sqrt{R^2 + X^2_L}$$

E/I is called the *impedance* of the circuit and is designated by the letter Z. The impedance determines the voltage which must be applied to the circuit to cause a given current to flow. The unit of impedance is, therefore, the ohm, just as in the case of resistance and reactance, which also determine the ratio of voltage to current. However, the phase angle between voltage and current must be specified, along with the impedance, for the true nature of the impedance to be known.

Similar consideration of resistance and capacity in series leads to the same expression for the impedance of such a circuit. However, in this case the voltage across the condenser lags the current, so that the "impedance triangle" is drawn with the condenser voltage or reactance extending downward. The general formula is

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

Ohm's Law for alternating current circuits then becomes

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

It should be noted that the equivalent Ohm's Law relationship for power in a d.c. circuit does not apply directly in the case of an a.c. circuit where Z replaces R. As will be explained, the power factor of the circuit must be taken into consideration.

In summary, impedance is a generalized quantity applying to a.c. or d.c. circuits, simple or complex. In a d.c. circuit or in an a.c. circuit containing resistance only, the phase angle is zero (current and voltage are in phase) and the impedance is equal to the resistance.

In an a.c. circuit containing reactance only the phase angle is 90 degrees, with current lagging the voltage if the reactance is inductive and current leading the voltage if the reactance is capacitive. In either case, the impedance is equal to the reactance.

In an a.c. circuit containing both resistance and reactance the phase angle may have any value between zero and 90 degrees, with the current lagging the voltage if the reactance is inductive and leading the voltage if the reactance is capacitive. The value of impedance, in ohms, may be found from the equation on page 35.

Power is consumed in a circuit only when the current flow produced by the applied voltage is in phase or less than 90 degrees out of phase with that voltage. Power consumption decreases from maximum with in-phase conditions to zero at 90 degrees out-of-phase conditions.

Series circuits with L, C and R — When inductance, capacity and resistance all are in series in an a.c. circuit, the voltage relations are a combination of the separate cases just considered. The voltage across each element will be proportional to the resistance or reactance of that element, since the current is the same through all. The voltages across the inductance and capacity are 180 degrees out of phase, since one leads the current by 90 degrees and the other lags the current by 90 degrees. This means that the two voltages tend to cancel; in fact, if the voltage across only the inductance and capacity in series is considered (leaving out the resistance), the total voltage is the *difference* between the two voltages.

This is shown by the vector diagram of Fig. 232. Since the angles of lead and lag are both 90 degrees, the reactance voltage lines are oppositely directed. The effect is exactly the same as though the *difference* between the two voltages (or reactances, in the impedance diagram) had been found first, and this difference then used as though it were a single voltage (or reactance). The net reactance in a

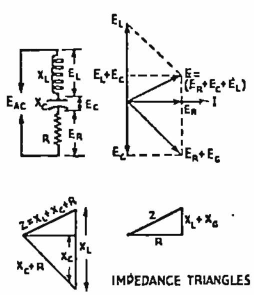


Fig. 232 — Current and voltage relationships in an a.c. circuit containing resistance, inductance and capacity in series. Step-by-step addition of voltage vectors is shown. The + signs are used in the algebraic sense, capacitive voltage or reactance being considered negative.

series circuit is, therefore, the difference between the individual inductive and capacitive reactances; or

$$X = X_L - X_C$$

If more than one inductance element is present in the circuit, the total inductive reactance is the sum of the individual reactances; similarly, the same is true for capacitive reactances. Inductive reactance is conventionally taken as "positive" (+) in sign and capacitive reactance as "negative" (−). With this convention, algebraic addition of all the reactances in a series circuit gives the total or net reactance of the circuit.

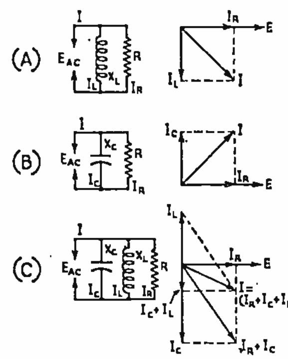


Fig. 233 — Parallel reactance and resistance in a.c. circuits. The various currents add vectorially as indicated.

Parallel circuits with L, C and R — The equivalent resistance of a number of resistances in parallel in an a.c. circuit is found by the same rules as in the case of d.c. circuits (§ 2-6). Parallel reactances *of the same kind* have an equivalent reactance given by a similar rule:

$$X = \frac{1}{\frac{1}{X_1} + \frac{1}{X_2} + \frac{1}{X_3} \cdots\cdots}$$

This formula applies to reactances of the same sign; it cannot be used if both inductive and capacitive reactance are in parallel.

When reactances and resistances are in parallel, the same voltage is applied to the various circuit elements. The current which flows is the *vector* sum of the currents in the various branches; that is, the phase of the currents with respect to the applied voltage must be taken into account in finding the total current. Fig. 233-A shows a resistance and inductance in parallel, with the corresponding vector diagram. The voltage is taken as the reference, since it is common to both branches. The current through the resistance is in phase with the voltage and coincides in direction with the voltage line. The current through the inductance lags the applied voltage by 90 degrees, hence is drawn at right angles downwards. The amplitudes of both currents are found by dividing the voltage by the resistance and reactance, respectively. The total current combines by the right-triangle rule; that is,

$$I = \sqrt{I_R^2 + I_X^2}$$

The impedance of the circuit is equal to E/I, so

$$Z = \frac{E}{\sqrt{I_R^2 + I_X^2}}$$

By assuming some convenient value for the applied voltage and then solving for the currents in the resistance and reactance, the values so found may be substituted in this equation to find the impedance of the circuit.

Resistance and capacity in parallel are shown in Fig. 233-B, with the corresponding vector diagram. Except that the current now leads the voltage, the relationships are the same as before; the total current and impedance can be found by using the same formulas.

When resistance, inductance and capacity are combined in parallel, as in Fig. 233-C, the vector diagram has the typical form shown. The currents in the inductance and capacity are 180 degrees out of phase, so that the total current through these elements (neglecting the resistance) is the *difference* between the two currents. Because of this, the current flowing in the line is always smaller than the largest reactive current, indicating that when inductance and capacity are connected in parallel the resultant impedance is larger than either of the individual reactances. The net reactive current may either lead or lag the applied voltage, depending upon whether the inductive or capacitive reactance is larger. The current taken by the parallel resistance is determined solely by the applied voltage. The total current and impedance of such a circuit can be found by the formulas used above, if for I_X the *difference* between the currents in the inductance and capacity is used.

With series-parallel circuits the solution becomes considerably more complicated, since the phase relationships in any parallel branch may not be either 90 degrees or zero. However, the majority of parallel circuits used in radio work can be solved by the rather simple approximate methods described in § 2-10.

Power factor — The power dissipated in an a.c. circuit containing both resistance and reactance is consumed entirely in the resistance, hence is equal to I^2R. However, the reactance is also effective in determining the current or voltage in the circuit, even though it consumes no energy. Hence the product of volts times amperes (which gives the power consumed in d.c. circuits) for the whole circuit may be several times the actual power used up. The ratio of power dissipated (watts) to the *volt-ampere* product is called the power factor of the circuit, or

$$\textit{Power factor} = \frac{\textit{Watts}}{\textit{Volt-amperes}}$$

The power factor of the dielectric in a condenser having a value of capacity, C, in microfarads (μfd.) at a certain frequency, f, is commonly expressed as a percentage:

$$\cos\theta = \frac{W \times 10^6}{2\pi f C V^2}$$

A table giving the power factor for various dielectrics at different frequencies appears on page 458 of the Appendix.

Distributed capacity and inductance — It should not be thought that the reactance of coils becomes infinitely high as the frequency is increased to a high value and, likewise, that the reactance of condensers becomes infinitely low at high frequencies. All coils have some capacity between turns, and the reactance of this capacity can become low enough at some high frequencies to tend to cancel the high reactance of the coil. Likewise, the leads and plates of condensers will have considerable inductance at very high frequencies, which will tend to offset the capacitive reactance of the condenser itself. For these reasons, coils constructed for high-frequency use must be designed to have low "distributed" capacity. Similarly, condensers must be made with short, heavy leads so that they will have low self-inductance.

Units and instruments — The units used in a.c. circuits may be divided or multiplied to give convenient numerical values to different orders of magnitude, just as in d.c. circuits (§ 2-6). Because the rapidly reversing current is accompanied by similar reversals in the magnetic field, instruments used for measurement of d.c. (§ 2-6) will not operate on a.c.

At low frequencies suitable instruments can be constructed by making the current produce both magnetic fields, one by means of a fixed coil and the other by the moving coil. Instruments having movements of this kind are variously known as *dynamometer*, *electrodynamometer* and *electrodynamic* types.

Another type of instrument suitable for measuring alternating current is less expensive in construction and therefore more widely used. This is the *repulsion-type* moving-iron a.c. ammeter shown in Fig. 234. Fundamentally, the movement is based on the same principle as the inexpensive moving-iron-vane meter for d.c. shown in Fig. 225. In the repulsion-type instrument current flowing through the stationary coil magnetizes two iron vanes, one fixed and the other attached to the movable pointer shaft. Inasmuch as the two vanes are in the same plane and magnetized by the same

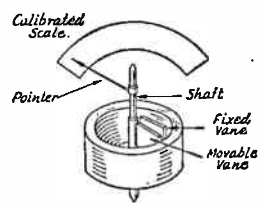


Fig. 234 — Ammeter based on a repulsion-type moving-iron movement used for a.c. measurements.

source, the magnetic effect upon them by the current through the coil will be identical regardless of its polarity. When the two vanes are magnetized they repel each other (§ 2-2) and the movable vane moves away from the fixed vane, causing the pointer to travel along the scale. The degree of travel is controlled by a spring which brings the pointer to rest at a point where the electrical and mechanical forces balance, and returns the pointer to zero on the scale when current flow ceases.

Such instruments are used for measurement of either current or voltage. However, when employed for voltage measurement by the use of high-resistance series multipliers, the minimum current drain required by such instruments because of their inherent insensitivity is so great that excessive load is placed upon the measurement source. For this reason, in radio work it is more common practice to convert the a.c. voltage to d.c. by means of a copper-oxide or vacuum-tube rectifier and then measure the resulting indication on a d.c. instrument, as described in § 2-6.

At radio frequencies instruments of the type described above are inaccurate because of distributed capacity and other effects; and the only reliable type of direct-reading instrument is the *thermocouple* ammeter or milliammeter. This is a power-operated device consisting of a resistance wire heated by the flow of r.f. current through it, to which is attached a thermocouple or pair of wires of dissimilar metals joined together and possessing the property of developing a small d.c. voltage between the terminals when heated. This voltage, which is proportional to the heat applied to the couple, is used to operate a d.c. instrument of ordinary design.

ℂ 2-9 The Transformer

Principles — It has been shown in the preceding sections that, when an alternating voltage is applied to an inductance, the flow of alternating current through the coil causes an induced e.m.f. which is opposed to the applied e.m.f. The induced e.m.f. results from the varying magnetic field accompanying the flow of alternating current. If a second coil is brought into the same field, a similar e.m.f. likewise will be induced in this coil. This induced e.m.f. may be used to force a current through a wire, resistance or other electrical device connected to the terminals of the second coil.

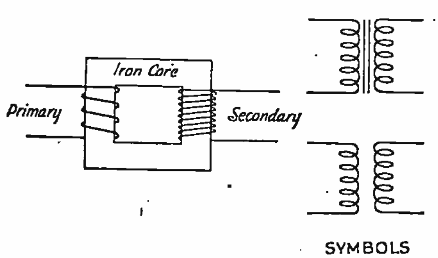


Fig. 235 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

Two coils operating in this way are said to be *coupled*, and the pair of coils constitutes a *transformer*. The coil connected to the source of energy is called the *primary* coil, and the other is called the *secondary* coil. Energy may be taken from the secondary, being transferred from the primary through the medium of the varying magnetic field.

Voltage and turns ratio — For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns on the coil. If the two coils of a transformer are in the same field, it follows that the induced voltages will be proportional to the number of turns on each coil. In the case of the primary, or coil connected to the source of power, the induced voltage is practically equal to, and opposes, the applied voltage. Hence, for all practical purposes,

$$E_s = \frac{n_s}{n_p} E_p$$

where E_s is the secondary voltage, E_p is the primary voltage, and n_s and n_p are the number of turns on the secondary and primary, respectively. The ratio n_s/n_p is called the *turns ratio* of the transformer.

This relationship is true *only* when all the flux set up by the primary current cuts all the turns of the secondary. If some of the magnetic flux follows a path which does not make it cut the secondary turns then the secondary voltage is less than given by this formula, since this reduces the number of lines of force (and thus reduces the effective strength of the magnetic field affecting the secondary) by causing the rate of change of flux to be less in the secondary than in the primary. In general, the equation can be used only when both coils are wound on a closed core of high permeability, so that practically all of the flux can be confined to definite paths.

Types of transformers — The usefulness of the transformer lies in the fact that energy can be transferred from one circuit to another without direct connection, and in the process can be readily changed from one voltage level to another. Thus, if a device to be operated requires, for example, 120 volts and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. The transformer, of course, can be used only on a.c., since no voltage will be induced in the secondary if the magnetic field is not changing. If d.c. is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

Transformers for use at radio frequencies are usually wound on nonmagnetic material ("air core") because the losses in ordinary iron cores are excessive at these frequencies. As a general rule, the equation given in the preceding paragraph does not apply to such transformers, because only a small part of the flux set up by the primary cuts the secondary turns. Even when special iron cores are used this statement is usually true of r.f. transformers, for reasons considered in § 2-11.

Transformers for use at power frequencies and audio frequencies are wound on iron cores, hence nearly all of the primary flux cuts the secondary turns. The turns-ratio equation can be used at these frequencies. The following discussion will be confined to such transformers.

Effect of secondary current — The primary current which has been discussed above is usually called the *magnetizing current* of the transformer. Like the current in any inductance, it lags the applied voltage by 90 degrees, neglecting the small energy losses in the resistance of the primary coil and in the iron core.

When current is drawn from the secondary winding, the secondary current sets up a magnetic field of its own in the core. The phase relationship between this field and that caused by the magnetizing current will depend upon the phase relationship between current and voltage in the secondary circuit. In every case there will be an effect upon the original field. To maintain the induced primary voltage equal to the applied voltage, however, the original field *must* be maintained. Consequently, the primary current must change in such a way that the effect of the field set up by the secondary current is completely canceled. This is accomplished when the primary draws an additional current that sets up a field exactly equal to the field set up by the secondary current, but which always opposes the secondary field. The additional primary current is thus 180 degrees out of phase with the secondary current. (This assumes that all the flux cuts both coils.) The total primary current is then the vector sum of the magnetizing current and this additional *load current*.

In rough calculations on transformers it is convenient to neglect the magnetizing current and to assume that the primary current is caused entirely by the secondary load. This is justifiable, because in any well-designed transformer the magnetizing current is quite small in comparison to the load current when the latter is near the rated value.

For the fields set up by the primary and secondary load currents to be equal, the number of ampere turns in the primary must equal the number of ampere turns in the secondary. That is,

$$n_s I_s = n_p I_p$$

Hence,

$$I_p = \frac{n_s}{n_p} I_s$$

The load current in the primary for a given load current in the secondary is proportional to the turns ratio, secondary to primary. This is the opposite of the voltage relationships.

If the magnetizing current is neglected, the phase relationship between current and voltage in the primary circuit will be identical with that existing between the secondary current and voltage. This is because the applied voltage and induced voltage are 180 degrees out of phase, and the primary current and secondary current likewise are 180 degrees out of phase.

Energy relationships; efficiency — A transformer cannot create energy; it can only transfer and transform it. Hence, the power taken from the secondary cannot exceed that taken by the primary from the source of applied e.m.f. Since there is always some power loss in the resistance of the coils and in the iron core, the power taken from the source always will exceed that taken from the secondary. Thus,

$$P_o = n P_i$$

where P_o is the power taken from the secondary, P_i is the power input to the primary, and n is a factor which always is less than 1. It is called the *efficiency* of the transformer and is usually expressed as a percentage. The efficiency of small power transformers such as are used in radio receivers and transmitters may vary between about 60 per cent and 90 per cent, depending upon the size and design.

Leakage reactance — In a practical transformer not all of the magnetic flux is common to both windings, although in well-designed transformers the amount of flux which cuts one coil and not the other is only a small percentage of the total flux. This *leakage flux* acts in the same way as flux about any coil which is not coupled to another coil; that is, it gives rise to self-induction. Consequently, there is a small amount of *leakage inductance* associated with both windings of the transformer, but not common to them. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of inductance and the frequency. This reactance is called *leakage reactance*.

In the primary the practical effect of leakage reactance is equivalent to a reduction in applied voltage, since the primary current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing primary current, hence it in-

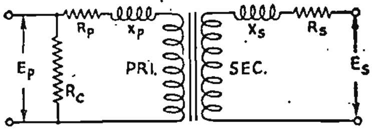


Fig. 236 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance R_c is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected in many approximate calculations.

creases as more current is drawn from the secondary. The induced voltage consequently decreases, since the applied voltage (which the induced voltage must equal in the primary) has been effectively reduced. The secondary induced voltage also decreases proportionately. When current flows in the secondary circuit the secondary leakage reactance causes an additional voltage drop, which results in a further reduction in the voltage available from the secondary terminals. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistance of the primary and secondary windings of the transformer also causes voltage drops when current is flowing, and, although these voltage drops are not in phase with those caused by leakage reactance, together they result in a lower secondary voltage under load than is indicated by the turns ratio of the transformer. At power frequencies (60 cycles) the voltage at the secondary, with a reasonably well-designed transformer, should not drop more than about 10 per cent under load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies, however, since the leakage reactance in a transformer increases directly with the frequency.

Impedance ratio — In an ideal transformer having no losses or leakage reactance, the primary and secondary volt-amperes are equal; that is,

$$E_p I_p = E_s I_s$$

On this assumption, and by making use of the relationships between voltage, current and turns ratio previously given, it can be shown that

$$\frac{E_p}{I_p} = \frac{E_s}{I_s}\left(\frac{n_p}{n_s}\right)^2$$

Since $Z = E/I$, E_s/I_s is the impedance of the load on the secondary circuit, and E_p/I_p is the impedance of the loaded transformer as viewed from the line. The equation states that the impedance presented by the primary of the transformer to the line, or source of power, is equal to the secondary load impedance multiplied by the square of the primary-to-secondary turns ratio. This primary impedance is called the *reflected impedance* or *reflected load*. The reflected impedance will have the same phase angle as the secondary load impedance, as previously explained. If the secondary load is resistive only, then the input terminals of the transformer primary will appear to the source of e.m.f. as a pure resistance.

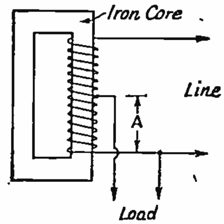


Fig. 238 — The auto-transformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (A) flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

In practice there is always some leakage reactance and power loss in the transformer, so that the relationship above does not hold exactly. However, it gives results which are adequate for many practical cases. The *impedance ratio* of the transformer consequently is considered to be equal to the square of the turns ratio, both ratios being taken from the same winding to the other.

Impedance matching — Many devices require a specific value of load resistance (or impedance) for optimum operation. The resistance of the actual load which is to dissipate the power may differ widely from this value, hence the transformer, with its impedance-transforming properties, is frequently called upon to change the actual load to the desired value. This is called *impedance matching*. From the preceding paragraph,

$$\frac{n_s}{n_p} = \sqrt{\frac{Z_s}{Z_p}}$$

where n_s/n_p is the required secondary-to-primary turns ratio, Z_s is the impedance of the actual load, and Z_p is the impedance required for optimum operation of the device delivering the power.

Transformer construction — Transformers are generally built so that flux leakage is minimized insofar as possible. The magnetic path is laid out so that it is as short as possible, since this reduces its reluctance and hence the number of ampere-turns required for a given flux density, and also tends to minimize flux leakage. Two core shapes are in common use, as shown in Fig. 237. In the shell type both windings are placed on the inner leg, while in the core type the primary and secondary windings may be placed on separate legs, if desired. This is sometimes done when it is necessary to minimize capacity effects between the primary and secondary, or when there is a large difference of potential between primary and secondary.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of thin sheets, called

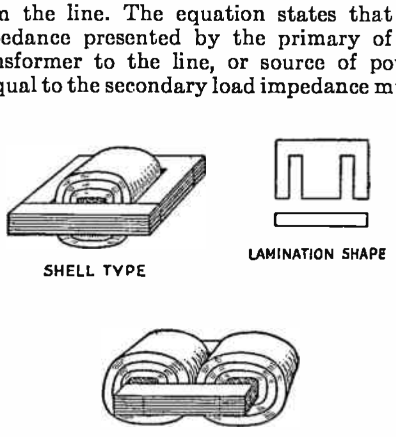


Fig. 237 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of *eddy currents* which are induced in the iron at right angles to the direction of the field. If allowed to flow, these eddy currents would cause considerable loss of energy in overcoming the resistance of the core material. The separate laminations are overlapped, to make the magnetic path as continuous as possible and thus reduce leakage.

The number of turns required on the primary for a given applied e.m.f. is determined by the maximum permissible flux density in the type of core material used, the frequency, and the magnetomotive force required to force the flux through the iron. As a rough indication, windings of small power transformers frequently have about two turns per volt for a core of 1 square inch cross-section and a magnetic path 10 or 12 inches in length. A longer path or smaller cross section would require more turns per volt, and vice versa.

In most transformers the coils are wound in layers, with a thin sheet of paper insulation between each layer. Thicker insulation is used between separate coils and between the coils and the core.

In power transformers distributed capacity in the windings is of little consequence, but in audio-frequency transformers it may cause undesired resonance effects (see § 2-10 for a discussion of resonance). High-grade audio transformers often have special types of windings designed to minimize distributed capacity.

The autotransformer — The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 236; the principles just discussed apply equally well. The *autotransformer* has the advantage that, since the line and load currents are out of phase, the section of the winding common to both circuits carries less current than the remainder of the coil. This advantage is not very marked unless the primary and secondary voltages do not differ very greatly, while it is frequently disadvantageous to have a direct connection between primary and secondary circuits. For these reasons, application of the autotransformer is usually limited to boosting or reducing the line voltage by a relatively small amount for purposes of voltage correction.

2-10 Resonant Circuits

Principle of resonance — It has been shown (§ 2-8) that the inductive reactance of a coil and the capacitive reactance of a condenser are oppositely affected by frequency. In any series combination of inductance and capacitance, therefore, there is one particular frequency for which the inductive and capacitive reactances are equal. Since these two reactances cancel each other, the net reactance in the circuit becomes zero, leaving only the resistance to impede the flow of current. The frequency at which this occurs is known as the *resonant frequency* of the circuit and the circuit is said to be *in resonance* at that frequency, or *tuned* to that frequency.

Series circuits — The frequency at which a series circuit is resonant is that for which $X_L = X_C$. Substituting the formulas for inductive and capacitive reactance (§ 2-8) gives

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

Solving this equation for frequency gives

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

This equation is in the fundamental units — cycles per second, henrys and farads — and so, if fractional or multiple units are used, the appropriate factors must be inserted to change them to the fundamental units. A formula in units commonly used in radio circuits is

$$f = \frac{1}{2\pi\sqrt{LC}} \times 10^6$$

where f is the frequency in kilocycles per second, 2π is 6.28, L is the inductance in microhenrys (μh.), and C is the capacitance in micromicrofarads ($\mu\mu$fd.).

The resistance that may be present does not enter into the formula for resonant frequency.

When a constant a.c. voltage of variable frequency is applied, as shown in Fig. 239-A, the current flowing through such a circuit will be maximum at the resonant frequency. The magnitude of the current at resonance will be determined by the resistance in the circuit. The curves of Fig. 239 illustrate this, curve *a* being for low resistance and curves *b* and *c* being for increasingly greater resistances.

In the circuits used at radio frequencies the reactance of either the coil or condenser at resonance is usually several times as large as the resistance of the circuit, although the *net* reactance is zero. As the applied frequency departs from resonance, say on the low-frequency side, the reactance of the condenser increases and that of the inductance decreases, so that the net reactance (which is the difference between the two) increases rather rapidly. When it becomes several times as high as the resistance, it becomes the chief factor in determining the amount of current flowing. Hence, for cir-

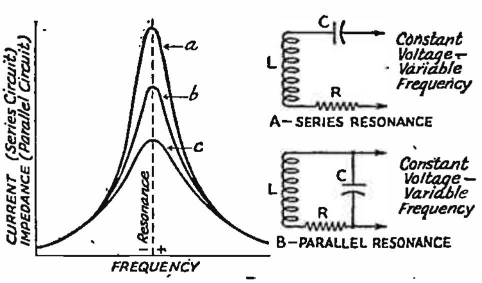


Fig. 239 — Characteristics of series-resonant and parallel-resonant circuits with variations in resistance, *R*.

cuits having the same values of inductance and capacity but varying amounts of resistance, the resonance curves tend to coincide at frequencies somewhat removed from resonance. The three curves in the figure show this tendency.

Parallel circuits — The parallel-resonant circuit is illustrated in Fig. 239-B. This circuit also contains inductance, capacitance and resistance in series, but the voltage is applied in parallel with the combination instead of in series with it as in A. As explained in connection with parallel inductance and capacity (§ 2-8), the total current through such a combination is less than the current flowing in the branch having the smaller reactance. If the currents through the inductive and capacitive branches are equal in amplitude and exactly 180 degrees out of phase, the total current, called the *line* current, will be zero no matter how large the individual branch currents may be. The impedance ($Z = E/I$) of such a circuit, viewed from its parallel terminals, would be infinite. In practice the two currents will not be exactly 180 degrees out of phase, because there is always some resistance in one or both branches. This resistance makes the phase relationship between current and voltage less than 90 degrees in the branch containing it, hence the phase difference between the currents in the two branches is less than 180 degrees and the two currents will not cancel completely. However, the line current may be very small if the resistance is small compared to the reactance, and thus the parallel impedance at resonance may be very high.

As the applied frequency is increased or decreased from the resonant frequency, the reactance of one branch decreases and that of the other branch increases. The branch with the smaller reactance takes a larger current, if the applied voltage is constant, and that with the larger reactance takes a smaller current. As a result, the difference between the two currents becomes larger as the frequency is moved farther from resonance. Since the line current is the difference between the two currents, the current increases when the frequency moves away from resonance; in other words, the parallel impedance of the circuit decreases.

The variation of parallel impedance of a parallel-resonant circuit with frequency is illustrated by the same curves of Fig. 239 that show the variation in current with frequency for the series-resonant circuit. The parallel impedance at resonance increases as the series resistance is made smaller.

In the case of parallel circuits, resonance may be defined in three ways: the condition which gives maximum impedance, that which gives a power factor of 1 (impedance purely resistive), or (as in series circuits) when the inductive and capacitive reactances are equal. If the resistance is low, the resonant frequencies obtained on the three bases are practically identical. This condition usually is satisfied in radio work, so that the resonant frequency of a parallel circuit is generally computed by the series-resonance formula given above.

Resistance at high frequencies — At radio frequencies the resistance of a conductor may be considerably higher than its resistance to direct current or low-frequency a.c. This is because of what is known as the *skin effect*. The magnetic field set up inside the wire tends to force the current to flow in the outer part of the wire to a degree where it will travel a long, round-about surface path around a conductor such as a condenser plate rather than a much shorter direct path through the metallic conductor. In the case of inductance coils the current concentrates not only on the outer surface of the conductor, but also in the portions most closely linked by the mutual magnetic fields. The importance of this effect increases with frequency. At high radio frequencies this skin effect is so pronounced that practically all the current flows very near the surface of the conductor, thereby in effect reducing the cross-sectional area and hence increasing the resistance. For this reason low resistance can be achieved only by using conductors with large surface area, but, since the inner part of the conductor does not carry current at the higher frequencies, thin-walled tubing will serve even more efficiently than solid wire of the same diameter.

A further effect occurs in coils at radio frequencies. The magnetic fields cause a concentration of current in certain parts of the conductors, again causing an effective decrease in the conductor size and raising the resistance. These effects, plus the effects of stray currents caused by distributed capacity (§ 2-8), raise the effective resistance of a coil at radio frequencies to many times the d.c. resistance of the wire.

Magnetic materials at high frequencies — At frequencies above the audio range (upper limit in the vicinity of 15,000 cycles) ordinary iron and steel cores are not useful for increasing the inductance of coils, although they are highly effective for this purpose at low frequencies. This is principally because losses from currents induced in the iron (eddy currents) increase to a prohibitive extent at high frequencies, since the induced current is proportional to frequency. Coils for radio-frequency purposes either are constructed without magnetic material (*air-core*) or have special types of iron cores particularly designed to reduce losses. Cores for radio-frequency use are made from finely divided iron of selected grades, held together with an insulating binding material in such a way that each iron particle is effectively insulated from the others. This prevents, or greatly reduces, the loss from eddy currents. The permeability of such a "powdered-iron" core is high enough, and the core losses are low enough, so that it is possible to construct an iron-core coil having lower effective resistance than an air-core coil of the same inductance. Although coils of this type now are used chiefly at the lower radio frequencies (below about 2 Mc.), practicable iron-core coils have been constructed for frequencies approaching the very-high-frequency range.

Sharpness of resonance — As the internal series resistance is increased the resonance curves become "flatter" for frequencies near the resonance frequency, as shown in Fig. 239. The relative sharpness of the resonance curve near resonance frequency is a measure of the *sharpness of tuning* or *selectivity* (ability to discriminate between voltages of different frequencies) in such circuits. This is an important consideration in tuned circuits for radio work.

Flywheel effect; Q — A resonant circuit may be compared to a flywheel in its behavior. Just as such a wheel will continue to revolve after it is no longer driven, so also will oscillations of electrical energy continue in a resonant circuit after the source of power is removed. The flywheel continues to revolve because of its stored mechanical energy; current flow continues in a resonant circuit by virtue of the energy stored in the magnetic field of the coil and the electric field of the condenser. When the applied power is shut off the energy surges back and forth between the coil and condenser, being first stored in the field of one, then released in the form of current flow, and then restored in the field of the other. Since there is always resistance present some of the energy is lost as heat in the resistance during each of these *oscillations* of energy, and eventually all the energy is so dissipated. The length of time the oscillations will continue is proportional to the ratio of the energy stored to that dissipated in each cycle of the oscillation. This ratio is called the Q (quality factor) of the circuit.

Since energy is stored by either the inductance or capacity and may be dissipated in either the inductive or capacitive branch of the circuit, a Q can be established for either the inductance or capacity alone as well as for the entire circuit. It can be shown that the energy stored is proportional to the reactance and that the energy dissipated is proportional to the resistance, so that, for either inductance or capacity associated with resistance,

$$Q = \frac{X}{R}$$

This relationship is useful in a variety of circuit problems.

In resonant circuits at frequencies below about 28 Mc. the internal resistance is almost wholly in the coil; the condenser resistance may be neglected. Consequently, the Q of the circuit as a whole is determined by the Q of the coil. Coils for use at frequencies below the very-high-frequency region may have Qs ranging from 100 to several hundred, depending upon their size and construction.

The sharpness of resonance of a tuned circuit is directly proportional to the Q of the circuit. As an indication of the effect of Q, the current in a series circuit drops to a little less than half its resonance value when the applied frequency is changed by an amount equal to $1/Q$ times the resonant frequency. The parallel-impedance of a parallel circuit similarly decreases with change in frequency. For example, in a circuit having a Q of 100, changing the applied frequency by 1/100th of the resonant frequency will decrease the parallel impedance to less than half its value at resonance.

Damping, decrement — The rate at which current dies down in amplitude in a resonant circuit after the source of power has been removed is called the *decrement* or *damping* of the circuit. A circuit with high decrement (low Q) is said to be highly damped; one with low decrement (high Q) is lightly damped.

Voltage rise — When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage which appears across either the coil or condenser is considerably higher than the applied voltage. This is because the current in the circuit is limited only by the actual resistance of the coil-condenser combination in the circuit, and hence may have a relatively high value; however, the same current flows through the high reactances of the coil and condenser, and consequently causes large voltage drops (§ 2-8). As explained above, the reactances are of opposite types and hence the voltages are opposite in phase, so that the net voltage around the circuit is only that which is applied. The ratio of the reactive voltage to the applied voltage is proportional to the ratio of reactance to resistance, which is the Q of the circuit. Hence, the voltage across either the coil or condenser is equal to Q times the voltage inserted in series with the circuit.

This can be better understood by referring to Fig. 240 and the discussion of parallel-resonant impedance in the following paragraphs. It should be noted that the curve labelled "resistance" in the figure is not the actual a.c. resistance of the circuit, as referred to above, but rather the parallel impedance of the combination, which becomes a pure resistance at resonance.

If, for example, the inductive reactance of a circuit is 200 ohms, the capacitive reactance is 200 ohms, the resistance 5 ohms, and the applied voltage is 50, there will be but the 5 ohms of pure resistance to oppose the current

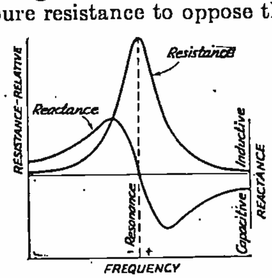


Fig. 240 — The impedance of a parallel-resonant resistance circuit is shown here separated into its reactance and resistance components. The parallel resistance of the circuit is equal to the parallel impedance at resonance.

flow through the network. Thus the actual current in the combination of coil and condenser and resistance will be 50/5 or 10 amperes. The voltage developed across either the coil or the equal condenser will be its reactance (200) times the current (10) or $200 \times 10 = 2000$ volts.

The ratio of reactive voltage to applied voltage is proportional to the ratio of the reactance of the coil or the condenser to the resistance. Since the latter ratio equals the Q of the circuit, the reactive voltage equals the applied voltage times the Q (200/5 or $40 \times 50 = 2000$ volts).

Parallel-resonant circuit impedance — The parallel-resonant circuit offers pure resistance (its *resonant impedance*) between its terminals because the line current is practically in phase with the applied voltage. At frequencies off resonance the current increases through the branch having the lower reactance (and vice versa) so that the circuit becomes reactive, and the resistive component of the impedance decreases as shown in Fig. 240.

If the circuit Q is 10 or more, the parallel impedance at resonance is given by the formula

$$Z_r = X^2/R = XQ$$

where X is the reactance of either the coil or the condenser and R is the internal resistance.

Q of loaded circuits — In many applications, particularly in receiving, the only power dissipated is that lost in the resistance of the resonant circuit itself. Hence the coil should be designed to have as high Q as possible. Since, within limits, increasing the number of turns raises the reactance faster than it raises the resistance, coils for such purposes are made with relatively large inductance for the frequency under consideration.

On the other hand, when the circuit delivers energy to a load, as in the case of the resonant circuits used in transmitters, the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit can be represented as shown in Fig. 241-A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is greater by 10 times or more than the power lost in the coil and condenser, the parallel impedance of the resonant circuit alone will be so high compared to the resistance of the load that the latter may be considered to determine the impedance of the combined circuit. (The parallel impedance of the tuned circuit alone is resistive at resonance, so that the impedance of the combined circuit may be calculated from the formula for resistances in parallel. If one of two resistances in parallel has 10 times the resistance of the other, the resultant resistance is practically equal to the smaller resistance.) The error will be small, therefore, if the losses in the tuned circuit alone are neglected. Then, since $Z = XQ$, the Q of a circuit loaded with a resistive impedance, Z, is

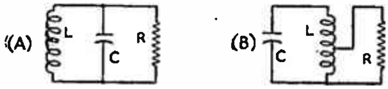


Fig. 241 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At (B) the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

$$Q = \frac{Z}{X}$$

where Z is the load resistance connected across the circuit and X is the reactance of either the coil or condenser. Hence, for a given parallel impedance, the effective Q of the circuit including the load is inversely proportional to the reactance of either the coil or the condenser. A circuit loaded with a relatively low resistance (a few thousand ohms) must therefore have a large capacity and relatively small inductance to have reasonably high Q.

From the above it is evident that connecting a resistance in parallel with a resonant circuit decreases the impedance of the circuit. However, the reactances in the circuit are unchanged, hence the reduction in impedance is equivalent to a reduction in the Q of the circuit. The same reduction in impedance also could be brought about by increasing the *series* resistance of the circuit. The *equivalent series resistance* introduced in a resonant circuit by an actual resistance connected in parallel is that value of resistance which, if added in series with the coil and condenser, would decrease the circuit Q to the same value it has when the parallel resistance is connected. When the resistance of the resonant circuit alone can be neglected, the equivalent resistance is

$$R = \frac{X^2}{Z}$$

the symbols having the same meaning as in the formula above.

The effect of a load of given resistance on the Q of the circuit can be changed by connecting the load across only part of the circuit. The most common method of accomplishing this is by tapping the load across part of the coil, as shown in Fig. 241-B. The smaller the portion of the coil across which the load is tapped, the less the loading on the circuit; in other words, tapping the load "down" is equivalent to connecting a higher value of load resistance across the whole circuit. This is similar in principle to impedance transformation with an iron-core transformer (§ 2-9). However, in the high-frequency resonant circuit the impedance ratio does not vary exactly as the square of the turn ratio, because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

L/C ratio — The formula for resonant frequency of a circuit shows that the same frequency always will be obtained so long as the *product* of L and C is constant. Within this limitation, it is evident that L can be large and C small, L small and C large, etc. The relation between the two for a fixed frequency is called the *L/C ratio*. A *high-C* circuit is one which has more capacity than "normal" for the frequency; a *low-C* circuit one which has less than normal capacity. These terms depend to a considerable extent upon the particular application considered, and have no exact numerical meaning.

LC Constants — The frequency of a resonant circuit varies inversely as the square root of the product of the inductance and capacitance. Therefore the product of the values of L and C required to resonate at any given frequency is a numerical constant. The formula for this constant any frequency is

$$LC = \frac{25330.3}{(f_{Mc.})^2} = \mu\text{h.} \times \mu\mu\text{fd.}$$

Doubling both the capacitance and the inductance (giving a product of 4 ×) halves the frequency; reducing the capacitance by one-half and the inductance by one-half doubles the frequency; leaving the inductance fixed and reducing the capacitance to one-half increases the frequency 40 per cent. To double the frequency, it is necessary to reduce either the capacitance or the inductance to one-fourth (leaving the other value fixed).

Any combination of capacity and inductance will tune to the same frequency provided the product of C and L equals the LC constant for that frequency.

ℂ 2-11 Coupled Circuits

Energy transfer; loading — Two circuits are said to be *coupled* when energy can be transferred from one to the other. The circuit delivering energy is called the primary circuit; that receiving energy is called the secondary circuit. The energy may be practically all dissipated in the secondary circuit itself, as in receiver circuits, or the secondary may simply act as a medium through which the energy is transferred to a load resistance where it does work. In the latter case, the coupled circuits may act as a radio-frequency impedance-matching device (§ 2-9) where the matching can be accomplished by adjusting the loading on the secondary (§ 2-10) and by varying the coupling between the primary and secondary.

Coupling by a common circuit element — One method of coupling between two resonant circuits is to have some type of circuit element common to both circuits. The three variations of this type of coupling (often called *direct coupling*) shown at A, B and C of Fig. 241, utilize a common inductance, capacity and resistance, respectively. Current circulating in one LC branch flows through the common element (L_c, C_c, or R_c) and the voltage developed across this element causes current to flow in the other LC branch. The degree of coupling between the two circuits becomes greater as the reactance (or resistance) of the common element is increased in comparison to the remaining reactances in the two branches.

If both circuits are resonant to the same frequency, as is usually the case, the common impedance — reactance or resistance — required for maximum energy transfer is generally quite small compared to the other reactances in the circuits.

Capacity coupling — The circuit at D shows electrostatic coupling between two resonant circuits. The coupling increases as the capacity of C_c is made greater (reactance of C_c is decreased). When two resonant circuits are coupled by this means, the capacity required for maximum energy transfer is quite small if the Q of the secondary circuit is at all high. For example, if the parallel impedance of the secondary circuit is 100,000 ohms, the reactance of the coupling condenser need not be lower than 10,000 ohms or so for ample coupling. The corresponding capacity required is only a few micromicrofarads at high frequencies.

Inductive coupling — Fig. 241-E illustrates inductive coupling, or coupling by means of the magnetic field. A circuit of this type resembles the iron-core transformer (§ 2-9) but, because only a small percentage of the flux lines set up by one coil cut the turns of the other coil, the simple relationships between turns ratio, voltage ratio and impedance ratio in the iron-core transformer do not hold. To determine the operation of such circuits, it is necessary to take account of the mutual inductance (§ 2-5) between the coils.

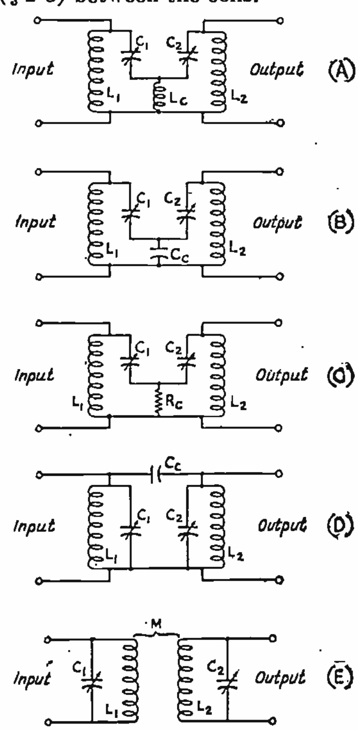


Fig. 242 — **Basic methods of circuit coupling,**

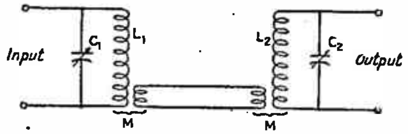


Fig. 243 — Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

Link coupling — A variation of inductive coupling, called *link coupling,* is shown in Fig. 243. This gives the effect of inductive coupling between two coils which may be so separated that they have no mutual inductance; the link may be considered simply as a means of providing the mutual inductance. Because mutual inductance between coil and link is involved at each end of the link, the total mutual inductance between two link-coupled circuits cannot be made as great as when normal inductive coupling is used. In practice, however, this ordinarily is not disadvantageous. Link coupling frequently is convenient in the design of equipment where inductive coupling would be impracticable for constructional reasons.

The link coils generally have few turns compared to the resonant-circuit coils, since the coefficient of coupling is relatively independent of the number of turns on either coil.

Coefficient of coupling — The degree of coupling between two coils is a function of their mutual inductance and self-inductances:

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

where k is called the *coefficient of coupling.* It is often expressed as a percentage. The coefficient of coupling cannot be greater than 1, and generally is much smaller in resonant circuits.

Inductively coupled circuits — Three types of circuits with inductive coupling are in general use. As shown in Fig. 244, one type has a tuned-secondary curcuit with an untuned-primary coil, the second a tuned-primary circuit and untuned-secondary coil, and the third uses tuned circuits in both the primary and secondary. The circuit at A is frequently used in receivers for coupling between amplifier tubes when the tuning of the circuit must be varied to respond to signals of different frequencies. Circuit B is used principally in transmitters, for coupling a radio-frequency amplifier to a resistive load. Circuit C is used for fixed-frequency amplification in receivers. The same circuit also is used in transmitters for transferring power to a load which has both reactance and resistance.

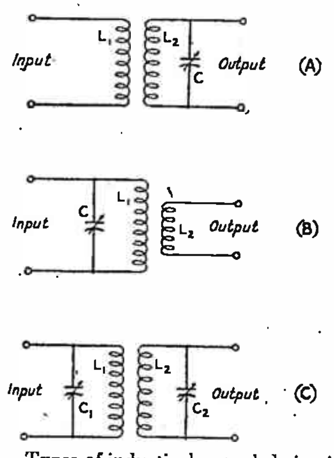


Fig. 244 — Types of inductively coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.

If the coupling between the primary and secondary is "tight" (coefficient of coupling large), the effect of inductive coupling in circuits A and B, Fig. 244, is much the same as though the circuit having the untuned coil were tapped on the tuned circuit (§ 2-10). Thus any resistance in the circuit to which the untuned coil is connected is coupled into the tuned circuit in proportion to the mutual inductance. This is equivalent to an increase in the series resistance of the tuned circuit, and its Q and selectivity are reduced (§ 2-10). The higher the coefficient of coupling, the lower the Q for a given value of resistance in the coupled circuit. These circuits may be used for impedance matching by adjustment of the coupling and of the number of turns in the untuned coil.

If the circuit to which the untuned coil is connected has reactance, a certain amount of reactance will be "coupled in" to the tuned circuit depending upon the amount of reactance present and the degree of coupling. The chief effect of this coupled reactance is to require readjustment of the tuning when the coupling is increased, if the tuned circuit has first been adjusted to resonance under conditions of very loose coupling.

Coupled resonant circuits — The effect of a tuned-secondary circuit on a tuned primary is somewhat more complicated than in the simpler circuits just described. When the secondary is tuned to resonance with the applied frequency, its impedance is resistive only. If the primary also is tuned to resonance, the current flowing in the secondary circuit (caused by the induced voltage) will, in turn, induce a voltage in the primary which is opposite in phase to the voltage acting in series in the primary circuit. This opposing voltage reduces the effective primary voltage, and thus causes a reduction in primary current. Since the actual voltage *applied* in the primary circuit has not changed, the reduction in current can be looked upon as being caused by an increase in the resistance of the primary circuit. That is, the effect of coupling a resonant secondary to the primary is to increase the primary resistance. The resistance under consideration is the *series* resistance of the primary circuit, not the parallel impedance or resistance. The parallel resistance decreases, since the increase in series resistance reduces the Q of the primary circuit.

If the secondary circuit is not tuned to resonance, the voltage induced back in the primary by the secondary current will not be exactly out of phase with the voltage acting in the primary; in effect, reactance is coupled into the primary circuit. If the applied frequency is fixed and the secondary circuit tuning is being varied, this means that the primary circuit will have to be retuned to resonance each time the secondary tuning is changed.

If the two circuits are initially tuned to resonance at a given frequency and then the applied frequency is varied, both circuits become reactive at all frequencies off resonance. Under these conditions, the reactance coupled into the primary by the secondary retunes the primary circuit to a new resonant frequency. Thus, at some frequency off resonance, the primary current will be maximum, while at the actual resonant frequency the current will be smaller because of the resistance coupled in from the secondary at resonance. There is a point of maximum primary current both above and below the true resonant frequency.

These effects are almost negligible with very "loose" coupling (coefficient of coupling very small), but increase rapidly as the coupling increases. Because of them, the selectivity of a pair of coupled resonant circuits can be varied over a considerable range simply by changing the coupling between them. Typical curves showing the variation of selectivity are shown in Fig. 245, lettered in order of increasing coefficient of coupling. At loose coupling, A, the voltage across the secondary circuit (induced voltage multiplied by the Q of the secondary circuit) is less than the maximum possible because the induced voltage is small with loose coupling. As the coupling increases the secondary voltage also increases, until critical coupling, B, is reached. At still closer coupling the effect of the primary current "humps" causes the secondary voltage to show somewhat similar humps, while when the coupling is further increased the frequency separation of the humps becomes greater. Resonance curves such as those at C and D are called "flat-topped," because the output voltage is substantially constant over an appreciable frequency range. Such a characteristic is desirable in many receiver applications.

Critical coupling — It will be observed that maximum secondary voltage is obtained in the curve at B in Fig. 245. With tighter coupling the resonance curve tends to be double-peaked, but in no case is such a peak higher than that shown for curve B. The coupling at which the secondary voltage is maximum is known as *critical coupling*. With this coupling the resistance coupled into the primary circuit is equal to the resistance of the primary itself, corresponding to the condition of matched impedances. Hence, the energy transfer is maximum at critical coupling. The over-all selectivity of the coupled circuits at critical coupling is intermediate between that obtainable with loose coupling and tight coupling. At very loose coupling, the selectivity of the system is very nearly equal to the product of the selectivities of the two circuits taken separately; that is, the effective Q of the circuit is equal to the product of the Qs of the primary and secondary.

Effect of circuit Q — Critical coupling is a function of the Qs of the two circuits taken independently. A higher coefficient of coupling is required to reach critical coupling when the Qs are low; if the Qs are high, as in receiving applications, a coupling coefficient of a few per cent may give critical coupling.

With loaded circuits it is not impossible for the Q to reach such low values that critical coupling cannot be obtained even with the highest practicable coefficient of coupling (coils as close physically as possible). In such case the only way to secure sufficient coupling is to increase the Q of one or both of the coupled circuits. This can be done either by decreasing the L/C ratio or by tapping the load down on the secondary coil (§ 2-10). One or the other of these methods often must be used with link coupling, because the maximum coefficient of coupling between two coils seldom runs higher than 50 or 60 per cent and the net coefficient is approximately equal to the products of the coefficients at each end of the link. If the load resistance is known beforehand, the circuits may be designed for a Q in the vicinity of 10 or so with assurance that sufficient coupling will be available; if unknown, the proper Qs can be determined by experiment.

Shielding — Frequently it is necessary to *prevent* coupling between two circuits which, for constructional reasons, must be physically near each other. Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called *shields*. The electrostatic field from the circuit components does not penetrate the shield, because the lines of force are short-circuited (§ 2-3). A metallic plate called a *baffle shield*, inserted between two components, may suffice to prevent electrostatic coupling between them, since very little of the field tends to bend around such a shield if it is large enough to make the components electrostatically invisible to each other.

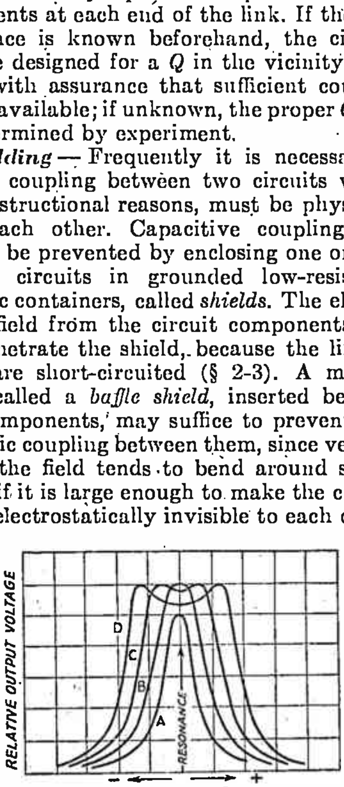


Fig. 245 — Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The input voltage is held constant in amplitude while the frequency is varied.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. In this case the magnetic field induces a current (*eddy current*) in the shield, which in turn sets up its own magnetic field opposing the original field (§ 2-5). The induced current is proportional to the frequency and also to the conductivity of the shield, hence the shielding effect increases with frequency and with the conductivity and thickness of the shielding material. A closed shield is required for good magnetic shielding; in some cases separate shields, one about each coil, may be required. The baffle shield is rather ineffective for magnetic shielding, although it will give partial shielding if placed at right angles to the axes of, as well as between, the two coils to be shielded from each other.

Cancellation of part of the field of the coil reduces its inductance, and, since some energy is dissipated in the shield, the effective resistance of the coil is raised as well. Hence the Q of the coil is reduced. The effect of shielding on coil Q and inductance becomes less as the distance between the coil and shield is increased. The losses also decrease with an increase in the conductivity of the shield material. Copper and aluminum are satisfactory materials. The Q and inductance will not be greatly reduced if the spacing between the sides of the coil and the shield is at least half the coil diameter, and is not less than the coil diameter at the ends of the coil.

At audio frequencies the shielding container should be made of magnetic material, preferably of high permeability (§ 2-5), so as to effectively short-circuit the external flux about the coil to be shielded. A nonmagnetic shield is quite ineffectual at these low frequencies since the induced current is small because of eddy currents that oppose flux penetration.

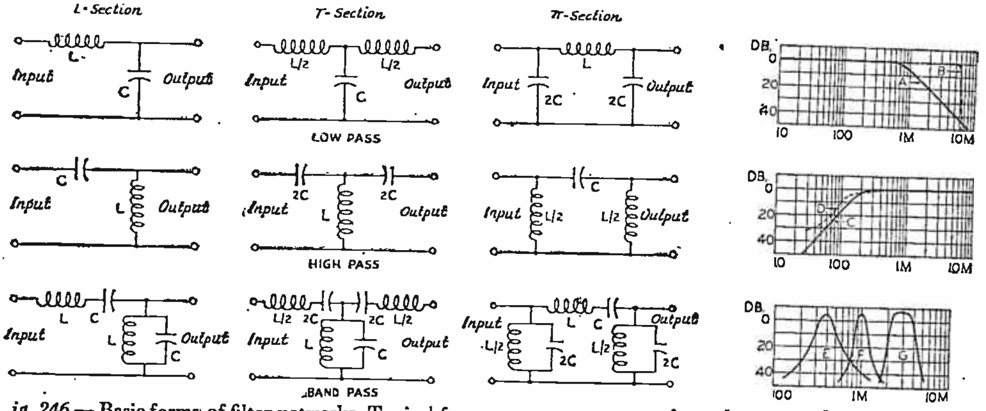
Filters — By suitable choice of circuit elements a coupling system may be designed to pass, without undue attenuation, all frequencies below and reject all frequencies above a certain value, called the *cut-off frequency*. Such a coupling system is called a *filter*, and in the above case is known as a *low-pass filter*.

If frequencies above the cut-off frequency are passed and those below attenuated, the filter is a *high-pass filter*. Simple filter circuits of both types are shown in Fig. 246, along with typical frequency-response curves. The fundamental circuit, from which more complex filters are constructed, is the *L-section*. Fig. 246 also shows *π-section* and *T-section* filters, both constructed from the basic L-section.

A *band-pass* filter, also shown in Fig. 246, is a combination of high- and low-pass filter elements designed to pass without attenuation all frequencies between two selected cut-off frequencies, and to attenuate all frequencies outside these limits. The group of frequencies which is passed by the filter is called the *pass-band*. Two resonant circuits with greater than critical coupling represent a common form of band-pass filter.

In curves of Fig. 246, A shows the attenuation at high frequencies of a single-section low-pass filter with high-Q components; B illustrates the extremely sharp cut-off obtainable with a more elaborate three-section filter. Curve C is that of a high-pass section having high Q, comparable to A. D shows the attenuation by a less-efficient section having some resistance in the inductance branch. Curves E, F and G illustrate various band-pass characteristics, E being a low-Q filter designed for narrow-band response, F a high-Q narrow-band filter, and G a wide-band high-Q two-section filter.

Filter circuits are frequently encountered both in low-frequency and r.f. applications. The proportions of L and C for proper operation depend upon the load resistance connected across the output terminals, L being larger and C smaller as the load resistance is increased. The type of section does not affect the attenuation curve, provided the input and output resistances are correct. In a symmetrical filter the input and output impedances must be equal to the characteristic impedance for which the filter is designed. Assuming these relationships, the design equations on the following page apply to the sections illustrated in Fig. 246.



ig. 246 — Basic forms of filter networks. Typical frequency response curves for each type are shown at the right.

Low-pass filter:

$$L = \frac{R}{\pi f_c} \qquad C = \frac{1}{\pi f_c R}$$

$$R = \frac{\sqrt{L_1}}{C_2} \qquad fc = \frac{1}{\pi\sqrt{L_1 C_2}}$$

High-pass filter:

$$L = \frac{R}{4\pi f_c} \qquad C = \frac{1}{4\pi f_c R}$$

$$R = \frac{\sqrt{L_2}}{C_1} \qquad f_c = \frac{1}{4\pi\sqrt{L_2 C_1}}$$

Band-pass filter:

$$L_1 = \frac{R}{\pi(f_2-f_1)} \qquad C_1 = \frac{f_2-f_1}{4\pi f_1 f_2 R}$$

$$L_2 = \frac{(f_2-f_1)R}{4\pi f_1 f_2} \qquad C_2 = \frac{1}{\pi(f_2-f_1)R}$$

$$R = \frac{\sqrt{L_1}}{C_2} = \frac{\sqrt{L_2}}{C_1} \qquad f_M = \sqrt{f_1 f_2}$$

$$f_M = \frac{1}{2\pi\sqrt{L_1 C_1}} = \frac{1}{2\pi\sqrt{L_2 C_2}}$$

In these formulas, R is the terminal impedance and f_c the design cut-off frequency for low-pass and high-pass filters. For band-pass filters, f_1 and f_2 are the pass-band limits and f_M the middle frequency.

The *resistance-capacity filter*, shown in Fig. 247, is used where both d.c. and a.c. are flowing through the circuit and it is desired to provide greater attenuation for the alternating current than the direct current. It is usually employed where the direct current has a low value so that the d.c. voltage drop is not excessive, or when a d.c. voltage drop actually is required. The time constant, RC, (§ 2-6) must be large compared to the time of one cycle of the lowest frequency to be attenuated. In determining the time constant, the resistance of the load must be included as well as that in the filter itself.

Bridge circuits — A *bridge circuit* is a device primarily used in making measurements of resistance, reactance or impedance (§ 2-8), and frequency, although bridges also have other applications in radio circuits.

The fundamental form is shown in Fig. 248-A. It consists of four resistances (called *arms*) connected in series-parallel to a source of voltage, E, with a sensitive galvanometer, M, connected between the junctions of the series-connected pairs. When the equation

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

is satisfied there is no potential difference between points A and B, since the drop across R_2

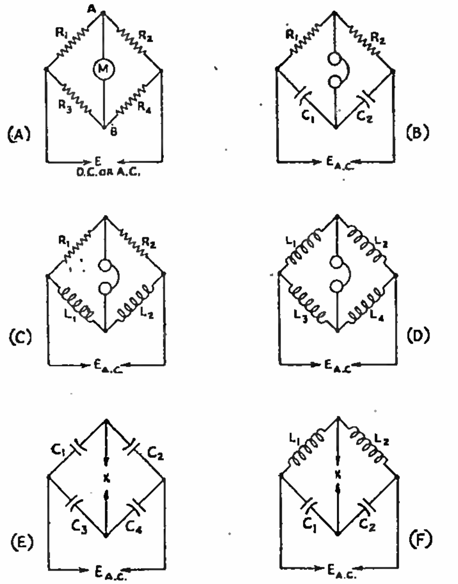


Fig. 248 — Bridge circuits utilizing resistance, inductance and capacity arms, both alone and in combination.

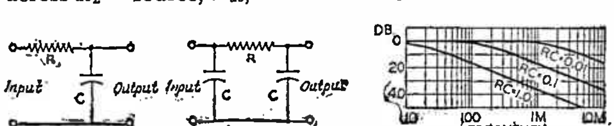
equals that across R_4 and the drop across R_1 equals that across R_3. Under these conditions the bridge is said to be *balanced*, and no current flows through M. If R_3 is an unknown resistance and R_4 is a variable known resistance, R_3 can be found from the following equation after R_4 has been adjusted to balance the bridge (*null* indication on M):

$$R_3 = \frac{R_1}{R_2} R_4$$

R_1 and R_2 are known as the *ratio arms* of the bridge; the ratio of their resistances is usually adjustable (frequently in steps of 1, 10, 100, etc.), so that a single variable resistor, R_4, can serve as a standard for measuring widely different values of unknown resistance.

Bridges similarly can be formed with arms containing capacity or inductance, and with combinations of either with resistance. Typical simple arrangements are shown in Fig. 248. For measurements involving alternating current the bridge must not introduce phase shifts which will destroy the balance, hence similar impedances should be used in each branch, as shown in Fig. 248, and the Qs of the coils and condensers should be the same. When bridges are used at audio frequencies, a telephone headset is a suitable null indicator. The bridges at E and F are commonly used in r.f. neutralizing circuits (§ 4-7); the voltage from the source, E_{ac}, is balanced out at X.

Fig. 247 — L-section and π-section resistance-capacity filter circuits (left) and curves showing the attenuation in db. for three different RC products at various frequencies in the audio-frequency range.



2-12-A Linear Circuits

Standing waves — If an electrical impulse is started along a wire, it will travel at approximately the speed of light until it reaches the end. If the end of the wire is open circuited, the impulse will be *reflected* at this point and will travel back again. When a high-frequency alternating voltage is applied to the wire a current will flow toward the open end, and reflection will occur continuously. If the wire is long enough so that time comparable to a half cycle or more is required for current to travel to the open end, the phase relations between the reflected current and outgoing current will vary along the wire. At one point the two currents will be 180° out of phase and at another in phase, with intermediate values between. Assuming negligible losses, the resultant current along the wire will vary in amplitude from zero to a maximum value. Such a variation is called a *standing wave*. The voltage along the wire also goes through standing waves, reaching its maximum value where the current is minimum and vice versa.

These phenomena are useful in various ways. For example, if a single wire is cut to such a length that standing waves can exist upon it at a certain frequency and energy of that frequency is applied, that energy will be radiated into space as though there were an infinite extension of the wire. It is by this means that radio signals are transmitted and received. If, however, the wire is arranged so that immediately alongside it lies another wire carrying standing waves of similar frequency and amplitude but precisely 180° out of phase, no radiation will occur because the one set of standing waves will cancel the other. Again, if this pair of wires, which is termed a transmission line, is provided with a suitable load circuit at the far end, the energy carried by the wires will be delivered to this load without loss other than that resulting from the ohmic resistance of the wires themselves.

Frequency and wavelength — It is possible to describe the constants of such line circuits in terms of inductance and capacitance, or inductance and capacitance per unit length, but it is more convenient to give them simply in terms of fundamental resonant frequency or of length. Since the velocity at which the current

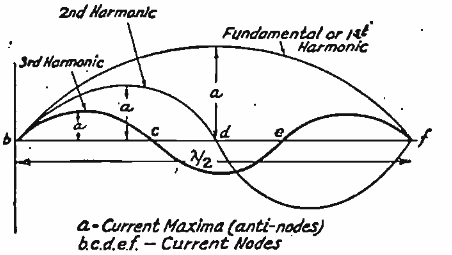


Fig. 248 — Standing-wave current distribution on a wire operating as an oscillatory circuit, at the fundamental, second harmonic and third harmonic frequencies.

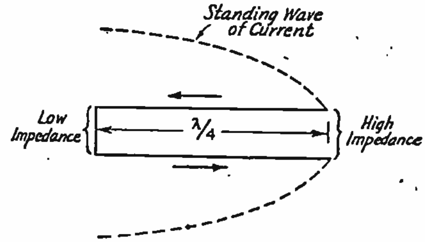


Fig. 249 — Standing wave and instantaneous current (shown by the arrows) in a folded resonant-line circuit.

travels is 300,000 kilometers (186,000 miles) per second, the *wavelength*, or distance the current will travel in the time of one cycle, is

$$\lambda = \frac{300,000}{f_{kc.}}$$

where λ is the wavelength in meters and $f_{kc.}$ is the frequency in kilocycles.

Wavelength is also used interchangeably with frequency in describing not only antennas but also tuned circuits, complete transmitters, receivers, etc. Thus, the terms "high-frequency receiver" and "short-wave receiver," or "75-meter antenna" and "4000-kilocycle antenna," are synonymous.

Harmonic resonance — Although a coil-condenser combination having lumped constants (capacitance and inductance) resonates only at one frequency, circuits such as antennas which contain distributed constants resonate readily at frequencies which are very nearly integral multiples of the fundamental frequency. These frequencies are, therefore, in *harmonic* relationship to the fundamental frequency, and hence are referred to as *harmonics* (§ 2-7). In radio practice the fundamental itself is called the *first harmonic*, the frequency twice the fundamental is called the *second harmonic*, and so on.

Fig. 248 illustrates the distribution of current on a wire for fundamental, second and third harmonic excitation. There is one point of maximum current with fundamental operation, two when operation is at the second harmonic, and three at the third harmonic; the number of current maxima corresponds to the order of the harmonic and the number of standing waves on the wire. As noted in the figure, the points of maximum current are called *anti-nodes* (also known as "loops") and the points of zero current are called *nodes*.

Radiation resistance — Since a line circuit has distributed inductance and capacity, current flow causes storage of energy in magnetic and electrostatic fields (§ 2-3, 2-5). At low frequencies practically all the energy so stored is returned to the wire during another part of the cycle (§ 2-8), but above 15,000 cycles or so (radio frequency) some escapes — is *radiated* — in the form of electromagnetic waves. Since energy radiated by a line or antenna is dissipated, insofar as the line is concerned, the loss can be considered as taking place in an *equivalent resistance*. This equivalent resistance is known as *radiation resistance*.

Resonant-line circuits — The effective resistance of a resonant straight wire, such as an antenna, is considerable, because of the power radiated. The resonance curve of such a straight-line circuit is quite broad; in other words, its Q is relatively low. However, by folding the line, as suggested by Fig. 249, the fields about the adjacent sections largely cancel each other and very little radiation takes place. The radiation resistance is greatly reduced, and the line-type circuit can be made to have a very sharp resonance curve, or high Q.

A circuit of this type will have a standing wave on it, as shown by the dashed-line of Fig. 249, with the instantaneous current flow in each wire opposite in direction to the flow in the other, as indicated by the arrows on the diagram. This opposite current flow accounts for the cancellation of radiation, since the fields about the two wires oppose each other. Furthermore, the impedance across the open ends of the line will be very high (thousands of ohms) while the impedance across the line near the closed end will be very low. This is because the current is low and the voltage is high at the open end of the line, but the current is high and the voltage low at the closed end.

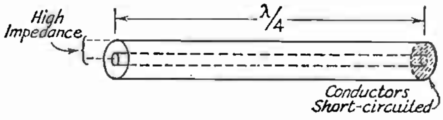


Fig. 250 — A quarter-wave coaxial-line resonant circuit.

A folded line may be made in the form of two coaxial or concentric conductors, as shown in Fig. 250. In effect, this line is directly comparable with the parallel conductor line, except that one conductor may be said to have been rotated around the other in a complete circle. The *coaxial line* has even lower radiation resistance than the folded-wire line, since the outer conductor acts as a shield. Standing waves exist but are confined to the outside of the inner conductor and the inside of the outer conductor, since skin effect prevents the currents from penetrating to the other sides. Thus such a line will have no radio-frequency potentials on its exposed surfaces, and no radiation can occur. Because of the low radiation resistance and the relatively large conducting surfaces, such self-enclosed lines can be made to have much higher Qs than are attainable with coils and condensers. They are most applicable at very high frequencies (very short wavelengths) (§ 2-7), where the dimensions are small.

A modified form of construction for coaxial lines is the "trough" line in which a tubular inner conductor is enclosed within a rectangular sheet-metal box or trough, usually left open on one side to facilitate tapping or other adjustments. The absence of shielding on one side does not affect the performance materially, and the simplicity of construction is an advantage.

Nonresonant lines — The foregoing has been concerned primarily with "short" lines of a length resonant at a given frequency. With a transmission line of infinite length, the power would travel along the line until eventually it would be entirely dissipated; consequently, none would be reflected and no standing waves would exist on the line. Theoretically, such a line would present a constant impedance in the form of a pure resistance to an input at any frequency. In practice the characteristics of such a line can be simulated by terminating a line of finite length with a load resistance or impedance equal to the characteristic impedance of the line.

Whether lines are classified as *resonant* or *nonresonant* depends upon the standing-wave ratio. If the ratio is near 1, the line is said to be nonresonant. Reactive effects will be small, and consequently no special tuning provisions need be made for canceling them (§ 2-10) even when the line length is not an exact multiple of a quarter wavelength. If the standing-wave ratio is large, the input reactance must be canceled or "tuned out" unless the line is resonant — i.e., a multiple of a quarter wavelength.

Characteristic impedance — The *characteristic impedance* of a transmission line, also known as the *surge impedance*, is defined as that impedance which a long line would present to an electrical impulse induced in the line. Mathematically, it is the square root of the ratio of inductance to capacity per unit length of the line.

The characteristic impedance of air-insulated transmission lines may be calculated from the following formulas:

Parallel-conductor line:

$$Z = 276 \log \frac{b}{a} \qquad (5)$$

where Z is the surge impedance, b the spacing, center to center, and a the radius of the conductor. The quantities b and a must be measured in the same units (inches, cm., etc.).

Coaxial or concentric line:

$$Z = 138 \log \frac{b}{a} \qquad (6)$$

where Z again is the surge impedance. In this case, b is the *inside diameter* (not radius) of the outer conductor and a is the *outside diameter* of the inner conductor. The formula is true for lines having air as the dielectric, and approximately so with ceramic insulators so spaced that the major part of the insulation is air.

The surge impedance for both parallel and coaxial lines using various sizes of conductors is given in chart form on page 52.

When a solid insulating material is used b- tween the conductors, because of the i- in line capacity the impedance d- the factor $1/\sqrt{K}$, where K i- constant of the insulatin-

The impedance of a single-wire transmission line varies with conductor size, height above ground, and orientation with respect to ground. An average figure is about 500 ohms.

Standing-wave ratio — The lengths of transmission lines used at radio frequencies are of the same order as the operating wavelengths, and therefore standing waves of current and voltage may appear on the line. The ratio of current (or voltage) at a loop to the value at a node (*standing-wave ratio*) depends upon the ratio of the resistance of the load connected to the output end of the line (its *termination*) to the characteristic impedance of the line itself. That is,

$$\text{Standing-wave ratio} = \frac{Z_s}{Z_t} \text{ or } \frac{Z_t}{Z_s} \qquad (7)$$

where Z_s is the characteristic impedance of the line and Z_t is the terminating resistance. Z_t is generally called an impedance, although it must be non-reactive and therefore must correspond to a pure resistance for the line to operate as described. This means that, when the load or termination is an antenna, it must be resonant at the operating frequency.

The formula is given in two ways because it is customary to put the larger number in the numerator, so that the ratio will not be fractional. As an example, a 600-ohm line terminated in a resistance of 70 ohms will have a standing wave ratio of 600/70, or 8.57. The ratio on a 70-ohm line terminated in a resistance of 600 ohms would be the same. Thus, if the current as measured at a node is 0.1 ampere, the current at a loop will be 0.857 ampere.

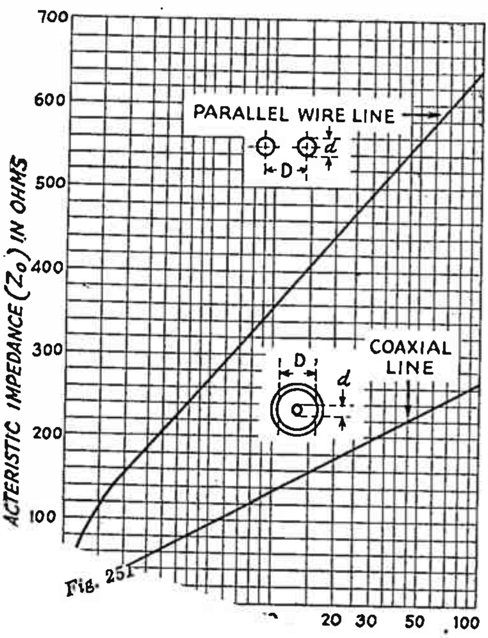


Fig. 25

~ of uniform lines.

A line terminated in a resistance equal to its characteristic impedance is equivalent to an infinitely long line; consequently there is no reflection, and no standing waves will appear. The standing wave ratio therefore is 1. The input end of such a line appears as a pure resistance of a value equal to the characteristic impedance of the line.

Input and output ends — The input end of a line is that connected to the source of power; the output end is that connected to the power-absorbing device. When a line connects a transmitter to an antenna, the input end is at the transmitter and the output end at the antenna; with the same line and antenna connected to a receiver, however, the energy flows from the antenna to the receiver, hence the input end of the line is at the antenna and the output end at the receiver.

Reactance, resistance, impedance — The input end of a line may show reactance as well as resistance, and the values of these quantities will depend upon the nature of the load at the output end, the electrical length of the line, and the line characteristic impedance. The reactance and resistance are important in determining the method of coupling to the source of power. Assuming that the load at the output end of the line is purely resistive, which is essentially the case since the load circuit is usually tuned to resonance, a line less than a quarter wavelength long electrically will show inductive reactance at its input terminals when the output termination is *less* than the characteristic impedance, and capacitive reactance when the termination is *higher* than the characteristic impedance. If the line is more than a quarter wave but less than a half wave long, the reverse conditions exist. With still longer lengths, the reactance characteristics reverse in each succeeding quarter wavelength. The input impedance is purely resistive if the line is an exact multiple of a quarter wave in length. The reactance at intermediate lengths is higher the greater the standing-wave ratio, being zero for a ratio of 1.

Transmission lines as circuit elements — Sections of transmission lines, together with combinations of such sections, can be used to simulate practically any electrical circuit element. Transmission lines can be used as resistance, inductance and capacity, as well as for resonant circuits, impedance-matching transformers, filters, and even as insulators.

When a quarter-wavelength line is connected between a "hot" circuit and ground, the input offers a high resistive impedance which is infinite. In other words, the transmission line is virtually an insulator. Insulating lines of this sort are commonly employed in ultrahigh frequency work. Such insulators can be used to provide a d.c. path between the r.f. conductor and chassis, and at the same time effectively block the flow of r.f. current.

A transmission line terminated in its characteristic impedance affords a pure resistance at high frequencies, and so may be used as a non-reactive resistor. Unterminated lines afford a variety of reactive properties. Lengths of short-circuited line less than a quarter wavelength represent pure inductive reactance, while open-circuited lines have pure capacitive reactance. Thus the former can be used in lieu of r.f. chokes, while the latter serve as by-pass condensers. Longer lengths of these elements have the opposite kind of reactance.

These characteristics are summarized in Fig. 252.

Resonant lines as tuned circuits — In resonant circuits as employed at the lower frequencies it is possible to consider each of the reactance components as a separate entity. A coil is used to provide the required inductance and a condenser is connected across it to provide the necessary capacity. The fact that the coil has a certain amount of self-capacity of its own, as well as some resistance, while the condenser also possesses a small self-inductance, can usually be disregarded.

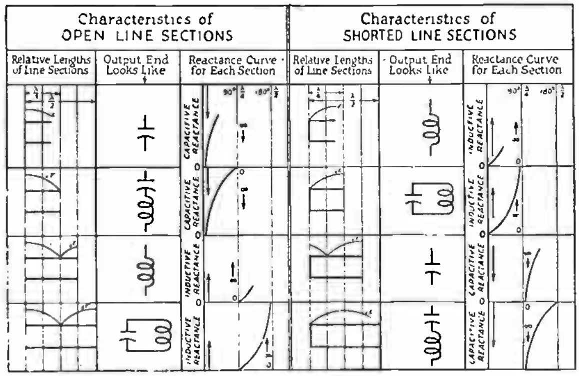


Fig. 253 — Open and closed transmission lines as circuit elements.

At the very-high and ultrahigh frequencies, however, it is no longer possible to separate these components. The connecting leads which, at lower frequencies, would serve merely to join the condenser to the coil now may have more inductance than the coil itself. The required inductance coil may be no more than a single turn of wire, yet even this single turn may have dimensions comparable to a wavelength at the operating frequency. Thus the energy in the field surrounding the "coil," instead of being confined to the wire by adjacent turns, may in part be radiated. The part which is radiated is, of course, lost. At a sufficiently high frequency the loss by radiation may represent a major portion of the total energy in the circuit. Since energy which cannot be utilized as intended is wasted, regardless of whether it is consumed as heat by the resistance of the wire or simply radiated into space, the effect is as though the resistance of the tuned circuit were greatly increased and its Q reduced to a negligible quantity.

For this reason, it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 100 Mc. A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other, exhibits large standing waves. When a voltage of the frequency at which such a line is resonant is applied to the open end, the response is very similar to that of a parallel resonant circuit; it will have very high input impedance at resonance and a large current flowing at the short-circuited end, multiplied many times over that flowing in at the open end.

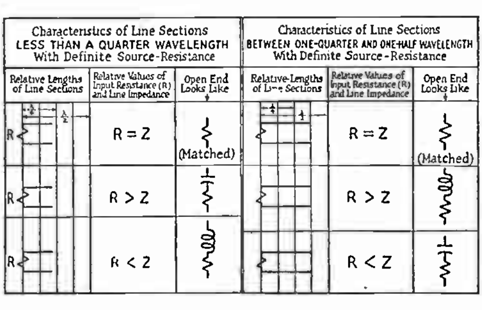


Fig. 252 — **Terminated lines as reactive circuit elements.**

The action of a resonant quarter-wavelength line can be compared with that of a coil-and-condenser combination whose constants have been adjusted to resonance at a corresponding frequency. Around the point of resonance, in fact, the line will display very nearly the same characteristics as those of the tuned circuit. The equivalent relationships are shown in Fig. 253. At frequencies off resonance the line displays qualities comparable to the inductive and capacitive reactances of the coil and condenser circuit, although the exact relationships involved are somewhat different. For all practical purposes, however, sections of resonant wire or transmission line can be used in much the same manner as coils or condensers.

In very-high-frequency circuits operating at frequencies in the neighborhood of 300 Mc., the spacing between conductors becomes an appreciable fraction of a wavelength. To keep the radiation loss as small as possible the parallel conductors should not be spaced farther apart than 10 per cent of the wavelength, center to center. On the other hand, the spacing of large-diameter conductors should not be reduced to much less twice the diameter because of what is known as the *proximity effect*, whereby another form of loss is introduced through eddy currents set up by the adjacent fields. Because the cancellation is no longer complete, radiation from an open line becomes so great that the Q is greatly reduced. Consequently, at these frequencies coaxial lines must be used. While the coaxial line is advantageous at the lower frequencies, as well, because it is more complicated to construct and adjustments are more difficult the open type of line is generally favored at these frequencies.

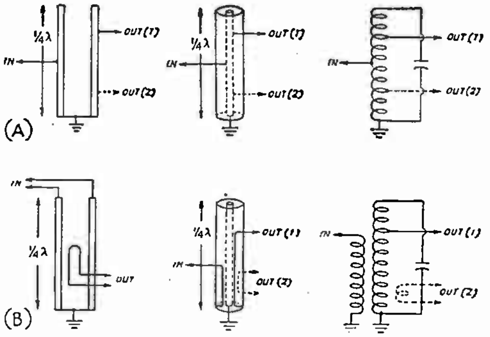


Fig. 254 — Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

Impedance transformation — Regardless of the standing-wave ratio, the input impedance of a line a half-wave long electrically will be equal to the impedance connected at its output end; the same thing is true of a line any integral multiple of a half-wave in length. Such a line can be considered to be a one-to-one transformer. However, if the line is a quarter-wave (or an odd multiple of a quarter-wave) long, the input impedance will be equal to

$$Z_i = \frac{Z_0{}^2}{Z_t} \qquad (8)$$

where Z_0 is the characteristic impedance of the line and Z_t the impedance connected to the output end. That is, a quarter-wave section of line will match two impedances, Z_i and Z_t, provided its characteristic impedance, Z_s, is equal to the geometric mean of the two impedances. A quarter-wave line may, therefore, be used as an *impedance transformer*. By suitable selection of constants, a wide range of impedance-matching values can be obtained.

Since the impedance measured between the two conductors anywhere along the line will vary between the two end values, a short-circuited quarter-wave line (§ 2-12) can be used as a *linear transformer* with an adjustable impedance ratio. For best operation, the two terminating impedances must be of the same order of magnitude. However, a series of quarter-wave sections can be used to obtain a step-by-step match of two terminal impedances efficiently if they are widely different.

Impedance-matching or transformation with transmission-line sections may also be affected by taps on quarter-wave resonant lines employed as coupling circuits in the same manner as conventional coil-condenser circuits. The equivalent relationships between parallel-line, coaxial-line and coil-and-condenser circuits for this purpose are shown in Fig. 254.

Other impedance-matching arrangements employ the use of matching stubs or equivalent sections so arranged so as to balance out the reactive component introduced by the coupled circuit. These are employed primarily in connection with antenna feed systems and are described in detail in § 10-8.

Transmission-line filter networks — The same general equations can be applied to any type of electrical network whether it be an actual section of transmission line, a combination of lumped-circuit elements, or a combination of transmission-line elements. Ordinary electric filters as used at lower frequencies use combinations of coils and condensers to fulfill this purpose, but coils cannot be used at extremely high frequencies. Combinations of transmission-line sections or combinations of transmission lines and parallel-plate condensers may, however, be used for the elements of very-high-frequency filter networks, instead.

Construction — Practical information concerning the construction of transmission lines for such specific uses as feeding antennas and as resonant circuits in radio transmitters will be found in the constructional chapters of this *Handbook*. Certain basic considerations applicable in general to resonant lines used as circuit elements may be considered here, however.

While either parallel-line or coaxial sections may be used, the latter are preferred for higher-frequency operation. Representative methods for adjusting the length of such lines to resonance are shown in Fig. 255. At the left, a sliding shorting disc is used to reduce the effective length of the line by altering the position of the short circuit. In the center, the same effect is accomplished by using a telescoping tube in the end of the inner conductor to vary its length and therefore the effective length of the line. At the right, two possible methods of mounting parallel plate condensers, used to tune a foreshortened line to resonance, are illustrated. The arrangement with the loading capacity at the open end of the line has the greatest tuning effect per unit of capacity; the alternative method, which is equivalent to "tapping" the condenser down on the line, has less effect on the Q of the circuit.

The short-circuiting disc at the end of the line must be designed to make perfect electrical contact. The voltage is a minimum at this end of the line; therefore, it will not break down some of the thinnest films. Usually a soldered connection or a tight clamp is used to secure good contact. When the length of line must be readily adjustable, the shorting plug is provided with spring collars which make contact on the inner and outer conductors at some distance away from the shorting plug at a point where the voltage is sufficient to break down the film between the collar and the conductor.

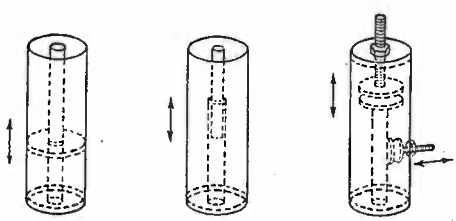


Fig. 255 — Methods of tuning coaxial resonant lines.

2-12-B Wave Guides and Cavity Resonators

Hollow wave guides — In Fig. 256-A, several closed quarter-wave stubs are shown connected in parallel across a two-wire transmission line. Since the open end of each stub is equivalent to an open circuit, the line impedance is not affected by their presence. Enough stubs may be added to form a U-shaped rectangular tube with solid walls, as at B, and another identical U-shaped tube may be added edge-to-edge to form the rectangular pipe shown in Fig. 256-C. As before, the line impedance still will not be affected. But now, instead of a two-wire transmission line, the energy is being conducted within a hollow rectangular tube. This is the fundamental operating mechanism of what is known as the hollow wave guide.

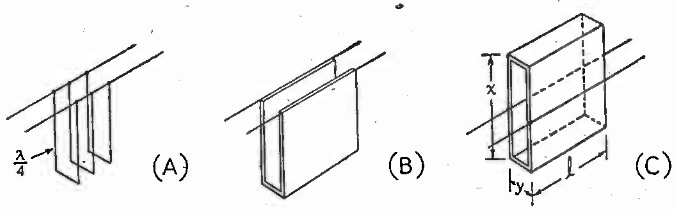


Fig. 256 — Evolution of a wave guide from a two-wire transmission line.

The frequency-determining dimension of such a wave guide is x, its height. This must be a half wavelength or multiple thereof. The width of the tube is important only in that it determines the breakdown voltage.

Operating principles of wave guides — Around any conductor carrying current there exist magnetic and electric fields, each in a separate plane, and each at right angles to the other. Use is made of this fact to direct energy lengthwise through hollow metallic tubes. The v.h.f. energy is injected at one end, either through capacitive or inductive coupling or by radiation, and is received at the other end. The wave guide then merely confines the energy of the fields, which are propagated through it to the receiving end by means of reflections against its inner walls.

Typical illustrations of reflected magnetic waves inside of a hollow wave guide are shown in Fig. 257. At A, the points of incidence of the wave are relatively far apart. Since energy in the wave moves down the guide faster when the reflections are few and far between, a relatively high frequency is being used in this case. At B, there are more reflections within a given length of the tube, which means that the propagated frequency must be lower. A still lower frequency is used at C, since the wave as a whole is slower reaching the end of the tube.

Along the line of propagation, a right-angle relationship exists between the electric and magnetic fields. Therefore, as shown in Fig. 257-D, by selecting the proper frequency an angle of incidence, α, can be obtained whereby the two fields do not overlap each other at any point because at this frequency there will be an equal number of reflections of both fields, all at the same angle with respect to the inside of the tube. Under these conditions, maximum propagation of energy through the tube occurs. The length of wave guide at which maximum energy transfer occurs, for a given frequency, is known as the *critical length*.

When the angle of incidence becomes less and less, as indicated in Fig. 257-E, there will be increasingly fewer reflections of the magnetic field against the walls of the guide. On the other hand, there will be more reflections of the electric field, as can be seen by following the paths of the arrows which are shown crossing the line of projection of the magnetic field. As the angle of incidence, α, approaches zero, the path of the magnetic field becomes more nearly parallel to the lengthwise axis of the tube, but the electric field is reflected more frequently and it strikes the walls of the tube at points which are progressively closer together. Finally, when the magnetic field reaches a plane entirely parallel to the axis, the electric field merely "bounces" up and down at right angles to the axis and no energy is transmitted down the tube.

Modes of propagation — The term *mode* refers to the arrangement of the magnetic and electric fields and the extent to which each is propagated through a hollow wave guide. All modes of transmission may be placed in two general groups. That which has a component of the electric field in the direction of propagation but no component of the magnetic field in that direction is known as a *TM* (*transverse magnetic*) mode, and the waves are called *E*

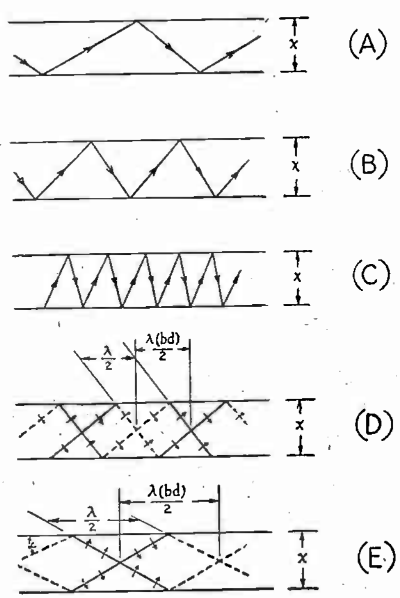


Fig. 257 — Reflected magnetic waves in a hollow guide.

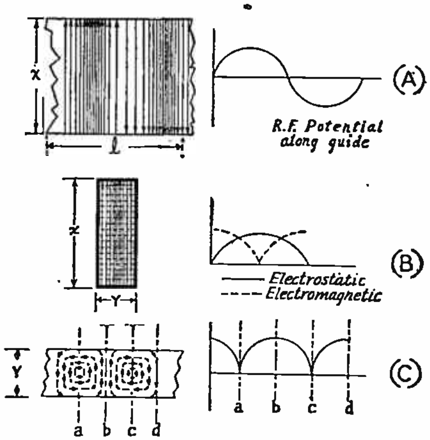


Fig. 258 — Field distribution in a hollow wave guide.

waves. A mode is designated as a *TE* (*transverse electric*) mode with *H* waves when only a component of the magnetic field and no components of the electric field are in the direction of propagation. If the field has some mode of transmission other than these two, it will be classified under one or the other general classification and designated with an appropriate subscript, such as $TM_{1,1}$ or $TM_{0,2}$.

Illustrations of a typical mode of simple form are given in Fig. 258. The side view at A shows the grouping of electrostatic lines inside the tube for a propagation path similar to that in Fig. 257-C. The same subject viewed from the front end of the tube is shown at Fig. 258-B. A top view showing the magnetic field is given in Fig. 258-C. From these diagrams it may be seen that the maximum and minimum intensities of the electric and magnetic fields occur simultaneously at the same positions inside the tube. Different modes result where the propagation is such as to produce more complex wave patterns inside the wave guide than those discussed here.

Hollow wave guides of circular form may be used following the same general principles which apply to the rectangular tubes discussed above. Different modes of transmission occur in circular tubes, however, because of the different configuration of the propagated waves within the tube.

Cavity resonators — The resonant circuits heretofore discussed (§ 2-12-A) are composed of what are termed lumped constants of *L* and *C*. As the frequency is increased, however, these constants, if used in their accustomed form, must be reduced to impracticably small physical dimensions. As explained previously (§ 2-12-B), this difficulty may be overcome in part by using transmission lines as linear resonant circuits. Another kind of linear circuit even more generally applicable at extremely high frequencies is the cavity resonator, the derivation of which is shown in Fig. 259. The cavity resonator may be compared with a coaxial line from which the central conductor has been removed, waves being propagated in the space within the cavity rather than along an inner conductor.

Considering that even a straight piece of wire has appreciable inductance at the very-high frequencies, it may be seen in Fig. 259-A and -B that a direct short across a two-plate condenser with air dielectric is the equivalent of a tuned circuit with a typical coiled inductance. With two wires between the plates, as shown in Fig. 259-C, the circuit may be thought of as a resonant-line section. For d.c. or even low frequency r.f., this line would appear as a short across the two condenser plates. At the ultra-high frequencies, however, as shown in Fig. 252, such a section of line a quarter-wavelength long would appear as an open circuit when viewed from one of the plates with respect to the other end of the section.

Increasing the number of parallel wires between the plates of the condenser would have no effect on the equivalent circuit, as shown at D. Eventually, the closed figure at E will be developed. Since each wire which is added in D is like connecting inductances in parallel, the total inductance across the condenser becomes increasingly smaller as the solid form is approached, and the resonant frequency of the figure therefore becomes higher.

If energy from some v.h.f. source now is introduced into the cavity in a manner such as that shown at F, the circuit will respond like any equivalent coil-condenser tank circuit at its resonant frequency. A cavity resonator may therefore be used as a u.h.f. tuning element, along with a tube of suitable design, to form the main components of an oscillator circuit capable of functioning at frequencies considerably beyond the maximum limits of conventional tubes, coils and condensers.

The frequency of a cavity resonator may be varied by the plunger arrangement shown at F. Other methods also may be used for this purpose, such as placing a small sphere at various positions inside the cavity in order to alter the field pattern of standing waves in both the electromagnetic and electrostatic fields.

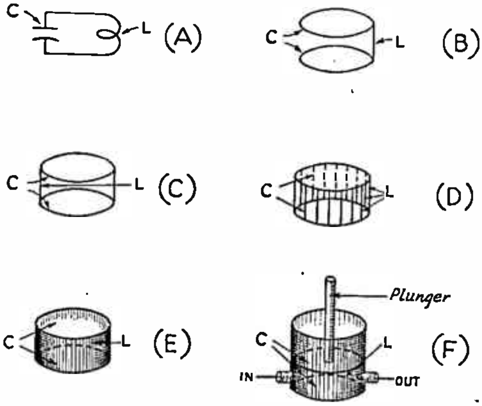


Fig. 259 — Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.

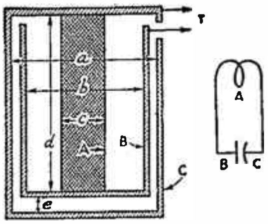


Fig. 260 — Concentric-cylinder or "pot"-type tank for v.h.f. The equivalent circuit diagram is also shown. Connections are made to the terminals marked T. For maximum Q the ratio of b to c should be between 3 and 5.

ℂ 2-12-C Lumped-Constant Circuits

V.h.f. resonator circuits — At the very-high frequencies the low values of L and C required make ordinary coils and condensers impracticable, while linear circuits offer mechanical difficulties in making tuning adjustments over a wide frequency range.

To overcome these difficulties, special high-Q lumped-constant circuits have been developed in which connections from the "condenser" to the "coil" are an inherent part of the structure. Integral design minimizes both resistance and inductance and increases the C/L ratio.

The simplest of these circuits is based on the use of discs combining half-turn inductance loops with semi-circular condenser plates. By connecting several of these half-turn coils in parallel, the effective inductance is reduced to a value appreciably below that for a single turn. Tuning is accomplished by interleaving grounded rotor plates between the turns. Both by shielding action and short-circuited-turn effect, these further reduce the inductance.

Another type of high-C circuit is a single-turn toroid in a form commonly termed the "hat" resonator. It consists of two copper shells with wide, flat "brims" mounted facing each other on an axially aligned copper rod. The capacity in the circuit is chiefly that between the wide shells, while the large-diameter central rod comprises the inductance.

"Pot"-type tank circuits — The lumped-constant concentric-element tank in Fig. 260, commonly referred to as the "pot" circuit, equivalent to a very short coaxial line (no linear dimension should exceed 1/20th wavelength), loaded by a large integral capacity.

The inductance is supplied by the field surrounding the central copper rod, A. The capacity is provided by the concentric cylinders, B and C, plus the capacity between the plates at the bottoms of the cylinders.

Approximate values of capacity and inductance for tank circuits of the "pot" type can be determined by the following:

$$L = 0.0117 \log \frac{b}{c} \ \mu\text{h.}$$

$$C = \left(\frac{0.6225\, d}{\log \dfrac{a}{b}}\right) + \left(\frac{0.1775\, b^2}{c}\right) \mu\mu\text{fd.}$$

where the symbols are as indicated in Fig. 260, and all dimensions are in inches. The left-hand term for capacity applies to the concentric cylinders, B and C, while the second term gives the capacity between the bottom plates.

"Butterfly" circuits — The tank circuits described in the preceding section are primarily fixed-frequency devices. The "butterfly" circuits shown in Fig. 261 are capable of being tuned over an exceptionally wide range, while still having high Q and reasonable physical dimensions. The circuit at A is derived from a conventional balanced-type variable condenser. The inductance is in the wide circular band connecting the stator plates. At its minimum setting the rotor plate fills the opening of the loop, reducing the inductance to a minimum. Connections are made to points 1 and 2. This basic structure eliminates all connecting leads and avoids all sliding or wiping electrical contacts to a rotating member. A disadvantage is that the electrical midpoint shifts from point 3 to point 3′ as the rotor is turned. Constant output coupling may be obtained electromagnetically by means of a coupling loop located at point 4, however.

In the modification shown at C, two sectoral stators are spaced 180°, thereby achieving the electrical symmetry required to permit tapping for balanced operation. Connections to the circuit should be made at points 1 and 2 and it may be tapped at points 3 and 3′, which are the electrical midpoints. Where magnetic coupling is employed, points 4 and 4′ are suitable locations for coupling links.

The capacity of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in the Appendix. Similarly, the maximum inductance can be obtained approximately by finding the inductance of a full ring of the same diameter and multiplying the result by a factor of 0.17. The ratio of minimum to maximum inductance varies between 1.5 and 4 with usual construction.

Any number of butterfly sections may be connected in parallel. In practice, units of four to eight plates prove most satisfactory. The ring and stator may either be made in one piece or with separate sectoral stator plates and spacing rings assembled with machine screws.

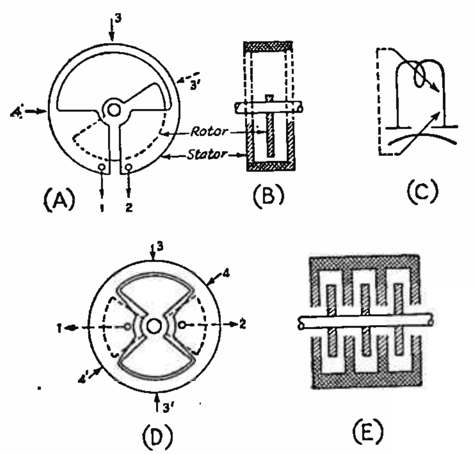


Fig. 261 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

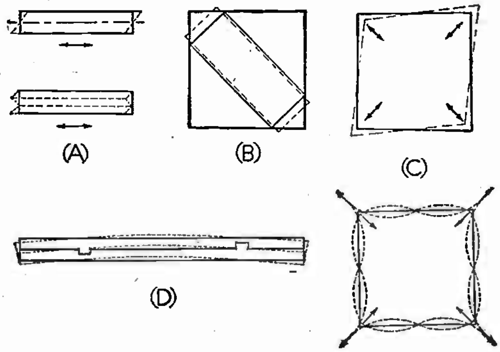


Fig. 262 — Modes of vibration for various crystal cuts. A — Fundamental (above) and harmonic (below) of the AT and BT cuts. B — The GT cut. C — CT and DT cuts (above) and ET and FT cuts (below). D — NT cut.

ℂ 2-12-D Piezoelectric Crystals

Piezoelectricity — Properly ground plates or bars of quartz and certain other crystalline materials, such as Rochelle salts, show a mechanical strain when subjected to an electric charge and, conversely, a difference in potential between two faces when subjected to mechanical stress. The relationship between mechanical force and electrical stress is known as the *piezoelectric effect*. The charges appearing on the crystal as a result of mechanical force applied to the crystal or mechanical vibration of the crystal itself are termed *piezoelectricity*.

Piezoelectric crystals may be employed as devices for changing either mechanical energy to electrical energy or electrical energy to mechanical energy. In the former category are crystal microphones and phonograph pickups; in the latter, crystal headphones, crystal loudspeakers and crystal recording heads.

A properly cut crystal is a mechanical vibrator electrically equivalent to a series-resonant circuit of very high Q, and so can be also used for many of the purposes for which ordinary resonant circuits are used. The resonant frequency, as in the case of other highly elastic solids such as metal reeds and violin strings, depends upon shape, thickness, length and cut.

Natural quartz crystals are usually in the form of a hexagonal prism terminated at one or both ends by a six-sided pyramid. Joining the vertices of these pyramidal ends, and perpendicular to the plane of the hexagonal cross section, is the optical or Z axis. The three electrical or X axes lie in a plane perpendicular to the optical axis and passing through opposite corners of the hexagon. The three mechanical or Y axes lie in the same plane but perpendicularly to the sides of the hexagon.

Active plates cut from a raw crystal at various angles to its optical, electrical and mechanical axes have differing characteristics as to thickness, frequency-temperature coefficient, power-handling capabilities, etc. The basic cuts are designated X and Y after their respective axes, but a variety of specialized cuts, such as the AT, are in more common use.

Frequency-thickness ratio — The linear oscillating dimension of the crystal is the principal frequency-determining factor, the other dimensions being of relatively minor importance. For most crystal cuts, the thickness and frequency are related by a constant, K, when

$$f = \frac{K}{t}$$

where f is the frequency in megacycles and t the thickness of the crystal in mils. For the X-cut, $K = 112.6$; for the Y-cut, $K = 77.0$; for the AT-cut, $K = 66.2$.

At frequencies above about 10 Mc. the crystal becomes very thin and correspondingly fragile, so that crystals seldom are manufactured for fundamental operation above this frequency. Direct crystal control on 14 and 28 Mc. is secured by use of "harmonic" crystals, which are ground to be active oscillators when excited at a harmonic (usually the third).

Temperature coefficient of frequency — The resonant frequency of a crystal varies with temperature, the variation depending upon the type of cut. The frequency change is usually expressed as a coefficient relating the number of cycles of frequency change per megacycle per °C. It may be either positive (increasing frequency with increasing temperature) or negative (decreasing frequency with increasing temperature). X-cut crystals have a negative coefficient of 15 to 25 cycles/Mc./°C. The coefficient of Y-cut crystals may vary from −20 cycles/Mc./°C. to +100 cycles/Mc./°C.

Variations in frequency caused by temperature changes can be minimized by proper cutting of the plate. By orienting the plate through various angles in relation to its optical, electrical and mechanical axes, a compensatory relationship can be derived between the dimensions of the plate, its density, and its elastic constants — the components responsible for the temperature coefficient.

The AT cut is the type perhaps most extensively used for transmitter frequency control. This plate can be ground to almost any frequency between 300 to 5000 kc. Its opposite, the BT cut, is used for frequencies within the range 4500 to 10,000 kc.

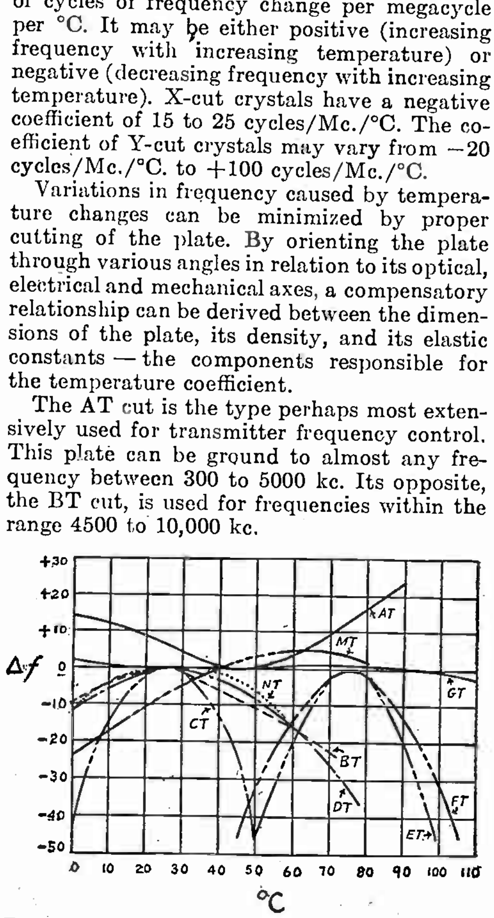


Fig. 263 — Frequency change in parts per million vs. variation in temperature in °C. for various crystal cuts.

For frequencies below 500 kc., CT and DT shear-type cuts have been developed which depend not upon thickness but on length and width for determining frequency. Plates of the CT and DT type vibrating at a harmonic mode are designated ET or FT cuts.

The low-drift types described above show a zero temperature coefficient through only a few degrees of change. Another type of cut, the GT, will drift less than 1 cycle/Mc./°C. over a temperature change of 100° C. In this plate a face shear vibration is changed into two longitudinal vibrations coupled together. At a certain ratio of length to width one mode has a zero temperature coefficient, making it especially useful as a frequency standard. The MT cut, which also vibrates longitudinally, can be used from 50 to 100 kc. The NT crystal is a flexurally vibrating cut having a low temperature coefficient in the range from 4 to 50 kc. MT and NT cuts are useful for phase-modulated f.m. transmitters.

¶ 2-13 Miscellaneous Circuit Details

Combined a.c. and d.c. — There are many practical instances of simultaneous flow of alternating and direct currents in a circuit. When this occurs there is a *pulsating* current, and it is said that an alternating current is *superimposed* on a direct current. As shown in Fig. 264, the maximum value is equal to the d.c. value plus the a.c. maximum, while the minimum value (on the negative a.c. peak) is the difference between the d.c. and the maximum a.c. values. The average value (§ 2-7) of the current is simply equal to the direct-current component alone. The effective value (§ 2-7) of the combination is equal to the square root of the sum of the effective a.c. squared and the d.c. squared:

$$I = \sqrt{I_{ac}^2 + I_{dc}^2}$$

where I_{ac} is the effective value of the a.c. component, I is the effective value of the combination, and I_{dc} is the average (d.c.) value of the combination.

Beats — When two dissimilar frequencies are superimposed electrically or mechanically, at recurrent intervals each original frequency will have its amplitude alternately increased and decreased at a rate corresponding exactly to the difference and sum of the original frequencies. If two or more alternating currents of different frequencies are present in a normal circuit they will have no particular effect upon one another and can be separated again by the proper selective circuits. However, if two (or more) alternating currents of different frequencies are present in an element having unilateral or one-way current flow properties, not only will the two original frequencies be present in the output but also currents having frequencies equal to the sum, and difference, of the original frequencies. These sum and difference frequencies are called the *beat* frequencies. For example, if frequencies of 2000 and 3000 kc. are present in a normal circuit only those two frequencies exist, but if they are passed through a unilateral-element there will be present in the output not only the two original frequencies of 2000 and 3000 kc. but also currents of 1000 (3000 − 2000) and 5000 (3000 + 2000) kc. Suitable circuits can be used to select the desired beat frequency. The human ear has unilateral characteristics and is, therefore, capable of hearing audible beat frequencies. Electronic devices of this nature are called mixers, converters, and detectors.

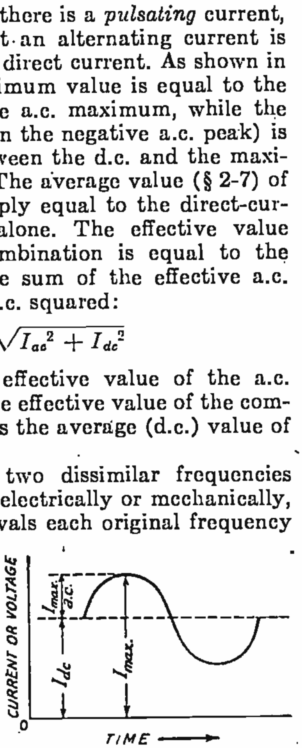


Fig. 264 — Pulsating current, composed of an alternating current or voltage superimposed on a direct current or voltage.

By-passing — In combined circuits, it is frequently necessary to provide a low-impedance path for a.c. around, for instance, a source of d.c. voltage. This can be done by using a *by-pass condenser*, which will not pass direct current but will readily permit the flow of alternating current. The capacity of the condenser should be of such value that its reactance is low (of the order of 1/10th or less) compared to the a.c. impedance of the device being by-passed. The lower the reactance, the more effectively will the a.c. be confined as desired.

Similarly, alternating current can be prevented from flowing through a direct-current circuit to which it may be connected by inserting an inductance of high reactance (called a *choke coil*) between the two circuits. This will permit the d.c. to flow without hindrance, since the resistance of the choke coil may be made quite low, but will effectively prevent the a.c. from flowing where it is not wanted.

If both r.f. and low-frequency (audio or power) currents are present in a circuit, they may be confined to desired paths by similar means, since an inductance of high reactance for radio frequencies will have negligible reactance at low frequencies, while a condenser of low reactance at radio frequencies will have high reactance at low frequencies.

Grounds — The term "ground" is frequently met with in discussions of circuits. Normally it means the voltage reference point in the circuit. There may or may not be an actual connection to earth, but it is understood that a point in the circuit said to be at *ground potential* could be connected to earth without disturbing the operation of the circuit in any way. In direct-current circuits, the negative side generally is grounded. The ground symbol in circuit diagrams is used for convenience in indicating common connections between various parts of the circuit, as through a metal chassis, and, with respect to actual ground, usually has the meaning indicated above.

. Chapter Three

Vacuum Tubes

3-1 Diodes

Rectification — Practically all of the vacuum tubes used in radio work depend upon thermionic conduction (§ 2-4) for their operation. The simplest type of vacuum tube is that shown in Fig. 301. It has two elements, a cathode and a plate, and is called a *diode*. When heated by the "A" battery the cathode emits electrons, which are attracted to the plate if the plate is at a positive potential with respect to the cathode.

Because of the nature of thermionic conduction, the tube is a conductor in one direction only. If a source of alternating voltage is connected between the cathode and plate, then electrons will flow only on the positive half-cycles of alternating voltage; there will be no electron flow during the half cycle when the plate is negative with respect to the cathode. Thus the tube can be used as a *rectifier*, to change alternating current to pulsating direct current. This alternating current can be anything from the 60-cycle kind to the highest radio frequencies.

Rectification finds its chief applications in detecting radio signals and in power supplies. These are treated in Chapters Seven and Eight, respectively.

Characteristic curves — The performance of the tube can be reduced to easily understood terms by making use of tube *characteristic curves*. A typical characteristic curve for a diode is shown at the right, in Fig. 301. It shows the current flowing between plate and cathode with different d.c. voltages applied between the elements. The curve of Fig. 301 shows that, with fixed cathode temperature, the plate current increases as the voltage between cathode and plate is raised. For an actual tube the values of plate current and plate voltage would be plotted along their respective axes.

The power consumed in the tube is the product of the plate voltage multiplied by the plate current, just as in any d.c. circuit. In a vacuum tube this power is dissipated in heat developed in the plate and radiated to the bulb.

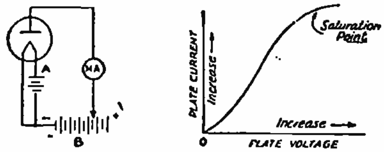


Fig. 301 — The diode or two-element tube and a typical characteristic curve showing plate current vs. voltage.

Space charge — With the cathode temperature fixed the total number of electrons emitted is always the same, regardless of the plate voltage. Fig. 301 shows, however, that less plate current will flow at low plate voltages than when the plate voltage is large. With low plate voltage, only those electrons nearest the plate are attracted to the plate. The electrons in the space near the cathode, being themselves negatively charged, tend to repel the similarly charged electrons leaving the cathode surface and cause them to fall back on the cathode. This is called the *space-charge* effect. As the plate voltage is raised more and more electrons are attracted to the plate, until finally the space charge effect is completely overcome. When this occurs all the electrons emitted by the cathode are attracted to the plate, and a further increase in plate voltage can cause no further increase in plate current. This condition is called *saturation*.

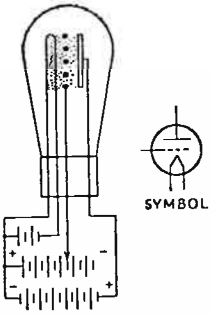
3-2 Triodes

Grid control — If a third element, called the *control grid*, or simply the *grid*, is inserted between the cathode and plate of the diode, the space-charge effect can be controlled. The tube then becomes a *triode* (three-element tube) and is useful for more things than rectification. The grid is usually in the form of an open spiral or mesh of fine wire. If the grid is connected externally to the cathode so that it is at the same potential as the cathode, and a steady voltage from a d.c. supply is then applied between the cathode and plate (the positive of the "B" supply is always connected to the plate), there will be a constant flow of electrons from cathode to plate through the openings of the grid, much as in the diode. However, if the grid is given a positive potential with respect to the cathode, the space charge will be partially neutralized and there will be an increase in plate current. If the grid is made negative with respect to the cathode, the space charge will be reinforced and the current will decrease.

This effect of grid voltage can be shown by curves in which plate current is plotted against grid voltage. At any given value of grid voltage the plate current will still depend upon the plate voltage, so if complete information about the tube is to be secured it is necessary to plot a *series* of curves taken with various values of plate voltage. Such a set of grid voltage vs. plate current curves, typical of a small receiving triode, is shown in Fig. 303.

So long as the grid has a negative potential with respect to the cathode, electrons emitted

Fig. 302 — Illustrating the construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density. Battery symbols follow those of the usual schematic diagrams, while the schematic tube symbol is shown at the right.



by the cathode are repelled (§ 2-3) from the grid, with the result that no current flows to the grid. Hence, under these conditions, the grid consumes no power. However, when the grid becomes positive with respect to the cathode, electrons are attracted to it, and a current flows to the grid; when this *grid current* flows, power is dissipated in the grid circuit.

In addition to the set of curves showing the relationship between grid voltage and plate current at various fixed values of plate voltage, two other sets of curves may be plotted to show the characteristics of a triode. These are the plate voltage vs. plate current characteristic, which shows the relationship between plate voltage and plate current for various fixed values of grid voltage, and the constant-current characteristic, which shows the relationship between plate voltage and grid voltage for various fixed values of plate current.

Amplification — The grid evidently acts as a valve to control the flow of plate current, and it is found that it has a much greater effect on plate current flow than does the plate voltage; that is, a small change in grid voltage is just as effective in bringing about a given change in plate current as is a large change in plate voltage.

The fact that a small voltage acting on the grid is equivalent to a large voltage acting on the plate indicates the possibility of *amplification* with the triode tube; that is, the generation of a large voltage by a small one, or the generation of a relatively large amount of power from a small amount. The many uses of the electronic tube nearly all are based upon this amplifying feature. The amplified power or voltage output from the tube is obtained, not from the tube itself, but from the source of e.m.f. connected between its plate and cathode. The tube simply *controls* the power from this source, changing it to the desired form.

To utilize the controlled power, a device for consuming it, or for transferring it to another circuit, must be connected in the plate circuit, since no particularly useful purpose would be served in having the current merely flow through the tube and the source of e.m.f. Such a device is called the *load*, and may be either a resistance or an impedance. The term "impedance" is frequently used even though the load may be purely resistive.

Amplification factor — The [illegible] of the grid and plate voltages on the plate cur[illegible]rent is measured by the *amplification factor* of the tube, usually represented by the Greek letter μ. Amplification factor is defined as the ratio of the change in plate voltage required to produce a given change in plate current to the change in grid voltage required to produce the same plate-current change. Strictly speaking, very small changes in both grid and plate voltage must be used in determining the amplification factor, because the curves showing the relationship between plate voltage and plate current, and between grid voltage and plate current, are not perfectly straight, especially if the plate current is nearly zero. Hence the slope (§ 1-10) varies at different points along the curves, and different values will be obtained for the amplification factor as larger or smaller voltage differences are taken for the purpose of calculating it. The expression for amplification factor can be written:

$$\mu = \frac{\Delta E_p}{\Delta E_g}$$

where ΔE_p indicates a very small change in plate voltage and ΔE_g is the change in grid voltage producing the same plate current change. The symbol Δ (the Greek letter *delta*) indicates a small increment, or small change.

The amplification factor is simply a ratio, and has no unit.

Plate resistance — Since only a limited amount of plate current flows when a given voltage is applied between plate and cathode, it is evident that the plate-cathode circuit of the tube has resistance. However, there is no simple relationship between plate voltage and plate current, so that in general the plate circuit of the tube does not follow Ohm's Law. Under a given set of conditions the application of a given plate voltage will cause a certain plate current to flow, and if the plate voltage is divided by the plate current a "resistance" value will be obtained which frequently is called the "d.c. resistance" of the tube. This "d.c. resistance" will be different for every value of plate voltage and also for different values of grid voltage, since the plate current also depends upon the grid voltage when the plate voltage is fixed.

In applications of the vacuum tube, it is more

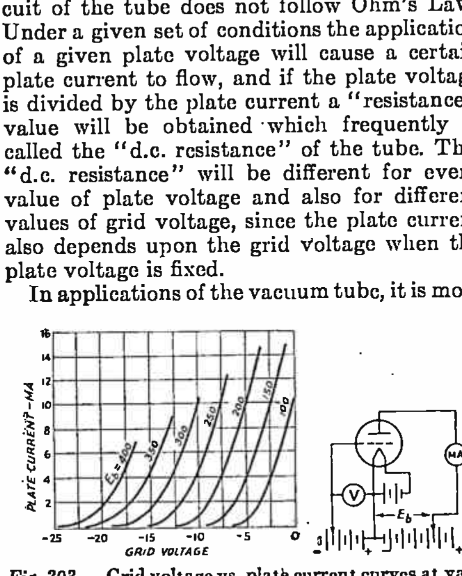


Fig. 303 — Grid voltage vs. plate current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

... The ...change and ...es the *a.c. plate* ...is resistance usually is ...an be found from the plate ...ate current characteristic curves ...ng the slope of the curve at the point ...onsidered. Hence

$$r_p = \frac{\Delta E_p}{\Delta I_p}$$

where ΔE_p is a small change in plate voltage and ΔI_p the corresponding small change in plate current, the grid voltage being fixed.

Plate resistance is expressed in ohms, since it is the ratio of voltage to current. The value of plate resistance will, in general, change with the particular voltages applied to the plate and grid. It depends as well upon the structure of the tube; low-μ tubes have relatively low plate resistance and high-μ tubes have high plate resistance.

Transconductance — The effect of grid voltage upon plate current is expressed by the *grid-plate transconductance* of the tube. Transconductance is a general term giving the relationship between the voltage applied to one electrode and the current which flows, as a result, in a second electrode. As in the previous two cases, it is defined as the *change* in current through the second electrode caused by a change in voltage on the first. Thus the grid-plate transconductance, commonly called the *mutual conductance*, is

$$g_m = \frac{\Delta I_p}{\Delta E_g}$$

where g_m is the mutual conductance, ΔI_p the change in plate current, and ΔE_g the change in grid voltage, the plate voltage being fixed. As before, the sign Δ indicates that the changes must be small. Transconductance is measured in mhos, since it is the ratio of current to voltage. The unit usually employed in connection with vacuum tubes is the *micromho* (one millionth of a mho), because the conductances are small. By combining with the two preceding formulas, it can be shown that $g_m = \mu / r_p$.

The mutual conductance of a tube is a rough indication of its merit as an amplifier, since it

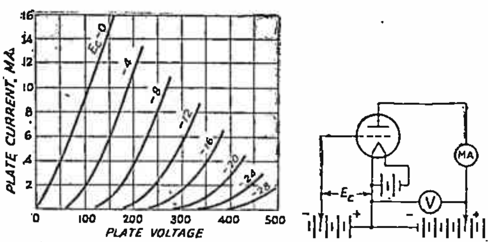


Fig. 304 — **Plate voltage vs. plate current curves at various fixed values of negative grid voltage for the same triode as that used to obtain the curves in Fig. 303.**

...e effects of both amplification factor ... plate resistance. Its value varies with the voltages applied to the plate and grid. With the plate voltage fixed, the mutual conductance decreases when the grid is made increasingly negative with respect to the cathode. This characteristic frequently can be used to advantage in the control of amplification, since the amount of amplification can be varied over wide limits simply by adjusting the value of a steady voltage applied to the grid.

Static and dynamic curves — Curves of the type shown in Figs. 301 and 303 are called *static* curves. They show the current which flows when various voltages are applied directly to the tube electrodes. Another useful set of static curves is the "plate family," or plate voltage vs. plate current characteristic. A typical set of curves of this type is shown in Fig. 304.

A curve showing the relationship between grid voltage and plate current when a load resistance is connected in the plate circuit is called a *dynamic* characteristic curve. Such a curve includes the effect of the load resistance, and hence is more indicative of the performance of the tube as an amplifier. With a fixed value of plate-supply voltage the actual value of voltage between the plate and cathode of the tube will depend upon the amount of plate current flowing, since the plate current also flows through the load resistance and therefore results in a voltage drop which must be subtracted from the plate-supply voltage. The dynamic curve includes the effect of this voltage drop. Consequently, the plate current always is lower, for a given value of grid bias and plate-supply voltage, with the load resistance in the circuit than it is without it.

Representative dynamic characteristics are shown in Fig. 305. These were taken with the same type of tube whose static curves are shown in Fig. 303. Different curves would be obtained with different values of plate-supply voltage, E_b; this set is for a plate-supply voltage of 300 volts. Note that increasing the value of the load resistance reduces the plate current at a given bias voltage, and also that the curves are straighter with the higher values of load resistance. Zero plate current always occurs at the same value of negative grid bias, since at zero plate current there is no voltage drop in the load resistance and the full supply voltage is applied to the plate.

Fig. 306 shows how the plate current responds to an alternating voltage (*signal*) applied to the grid. If the plate current is to have the same waveshape as that of the signal, it is necessary to confine the operation to the straight section of the curve. To do this, it is necessary to select an *operating point* near the middle of the straight portion; this operating point is determined by the fixed voltage (*bias*) applied to the grid. The alternating signal voltage then adds to or subtracts from the grid bias, depending upon whether the instantaneous signal voltage is negative or positive with respect to the cathode, and causes a corresponding variation in plate current. The maximum departure of instantaneous grid voltage or plate current from the operating point is called the *swing*. The varying plate current flows through the load resistance, causing a varying voltage drop which constitutes the useful output voltage of the tube.

The point at which the plate current is reduced to zero is called the *cut-off point*. The value of negative grid voltage at which cut-off occurs depends upon the amplification factor of the tube and the plate voltage. It is approximately equal to the plate-supply voltage divided by the amplification factor.

Interelectrode capacities — Any pair of elements in a tube forms a miniature condenser (§ 2-3), and, although the capacities of these condensers may be only a few micromicrofarads or less, they must frequently be taken into account in vacuum-tube circuits. The capacity from grid to plate (*grid-plate capacity*) has an important effect in many applications. In triodes, the other capacities are the *grid-cathode* and *plate-cathode*. In multi-element tubes (§ 3-5), similar capacities exist between these and other electrodes. With screen-grid tubes, the terms "input" and "output" capacity mean, respectively, the capacity measured from grid to all other elements connected together and from plate to all other elements connected together. The same terms are used with triodes but are not so easily defined, since the effective capacities existing depend upon the operating conditions (§ 3-3).

Tube ratings — Specifications of suitable operating voltages and currents are called *tube ratings*. Ratings include proper values for filament or heater voltage and current, plate voltage and current, and similar operating specifications for other elements. An important rating in power tubes is the *maximum safe plate dissipation*, or the maximum power that can be dissipated continuously in heat on the plate (§ 3-1).

ℂ 3-3 Amplification

Principles — The operation of a simple amplifier, which was described briefly in the preceding section, is shown in more detail in Fig. 307. The load in the plate circuit is the resistor, R_p. For the sake of example, it is assumed that the plate-supply voltage is 300 volts, the negative grid bias is 5 volts, and the plate current at this bias when R_p is 50,000 ohms is 2 milliamperes (0.002 ampere). If no signal is applied to the grid circuit, the voltage drop in the load resistor is 50,000 × 0.002, or 100 volts, leaving 200 volts between the plate and cathode.

If a sine-wave signal having a peak value of 2 volts is applied in series with the bias voltage in the grid circuit, the instantaneous voltage at the grid will swing to − 3 volts at the instant the signal reaches its positive peak and to − 7 volts at the instant the signal reaches its negative peak. The maximum plate current

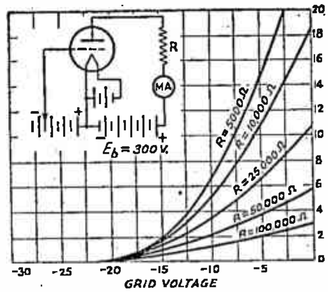


Fig. 305 — Dynamic characteristics of a small triode with various load resistances from 5,000 to 100,000 ohms.

will occur at the instant the grid voltage is − 3 volts and, as shown by the graph, will have a value of 2.65 milliamperes. The minimum plate current occurs at the instant the grid voltage is − 7 volts, and has a value of 1.35 ma. At intermediate values of grid voltage, intermediate plate-current values will occur. The instantaneous voltage between the plate and cathode of the tube also is shown on the graph. When the plate current is maximum the instantaneous voltage drop in R_p is 50,000 × 0.00265 or 132.5 volts, and when the plate current is minimum the instantaneous voltage drop in R_p is 50,000 × 0.00135 or 67.5 volts. The actual voltage between plate and cathode is therefore the difference between the plate-supply voltage, 300 volts, and these voltage drops in the load resistance, or 167.5 and 232.5 volts, respectively.

The varying plate voltage is an a.c. voltage superimposed (§ 2-13) on the steady plate-cathode voltage of 200 volts, which was previously determined for no-signal conditions. The peak value of this a.c. output voltage is the difference between either the maximum or minimum plate-cathode voltage and the no-signal value of 200 volts. In the illustration this difference is 232.5 − 200 or 200 − 167.5, or 32.5 volts. Since the grid signal voltage has a peak value of 2 volts, the voltage amplification ratio of the amplifier is 32.5/2 or 16.25. That is, approximately 16 times as much volt-

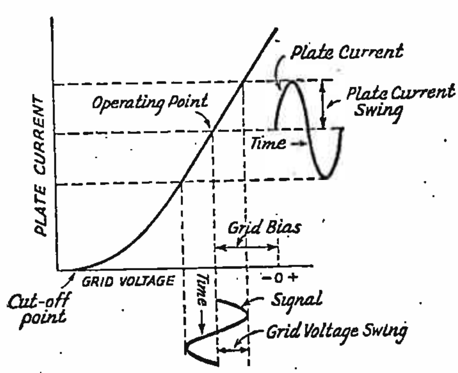


Fig. 306 — Behavior of the plate current of a vacuum tube in response to an alternating signal voltage superimposed on a steady negative grid voltage or bias.

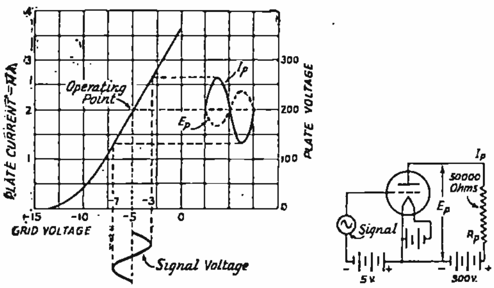


Fig. 307 — Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.

age will be obtained from the plate circuit as is applied to the grid circuit.

It will be observed that only the alternating plate and grid voltages enter into the calculation of the amplification ratio. The d.c. plate and grid voltages are of course essential to the operation of the tube, since they set the operating point, but otherwise their presence may be ignored. This being the case, it is possible to show that the tube can be replaced by an *equivalent generator* which has an internal resistance equal to the a.c. plate resistance of the tube (§ 3-2) at the operating point chosen and which generates a voltage equal to the amplification factor of the tube multiplied by the signal voltage applied to the grid. The equivalent generator, together with the load resistance, R_p, is shown in Fig. 308. This simplification enables ready calculation of the amplification. If the generated voltage is μE_g, then the same current flows through r_p and R_p, and hence the voltage drop across R_p, which is the useful output voltage, is

$$E_o = \mu E_g \frac{R_p}{r_p + R_p}$$

since R_p and r_p together constitute a voltage divider (§ 2-6). The voltage-amplification ratio is given by the output voltage divided by the input voltage, hence dividing the above expression by E_g gives the following formula for the amplification of the tube:

$$\text{Amplification} = \frac{\mu R_p}{r_p + R_p}$$

This expression shows that, to obtain a large voltage-amplification ratio, it is necessary to make the plate load resistance, R_p, large compared to the plate resistance, r_p, of the tube. The maximum possible amplification, obtained when R_p is infinitely larger than r_p, is equal to the μ of the tube. A tube with a large value of μ will, in general, give more voltage amplification than one with a medium or low value. However, the advantage of the high μ is less than might be expected, because a high-μ tube usually also has a correspondingly high value of r_p, so that a high value of load resistance must be used to realize an appreciable part of the possible amplification. This in turn not only requires the use of high values of plate-supply voltage, but has some further disadvantages to be described later.

Amplifiers in which the voltage output, rather than the power output, is the primary consideration are called *voltage amplifiers*.

Power in grid circuit — In the operation depicted in Fig. 306, the grid is always negative with respect to the cathode. If the peak signal voltage is larger than the bias voltage, the grid will be positive with respect to the cathode during part of the signal cycle. Grid current will flow during this time, and the signal source will be called upon to furnish power during the period while grid current is flowing. In many cases the signal source is not capable of furnishing appreciable power, so that care must be taken to avoid "driving the grid positive."

When dealing with small signals the source of signal voltage frequently has high internal resistance, so that a considerable voltage drop occurs in the source itself whenever it is called upon to furnish grid current. Since this voltage drop occurs only during part of the cycle, the voltage applied to the grid undergoes a change in waveshape because of the current flow. This is shown in Fig. 309, where a sine-wave signal is generated but, because of the internal resistance of the source, is *distorted* at the grid of the tube during the time when grid current flows.

If the internal resistance of the signal source is low, so that the internal voltage drop is negligible when current flows, this distortion does not occur. With such a source, it is possible to operate over a greater portion of the amplifier characteristic.

Harmonic distortion — If the operation of the tube is not confined to a straight or linear portion of the dynamic characteristic, the waveshape of the output voltage will not be exactly the same as that of the signal voltage. This is shown in Fig. 310, where the operating point is selected so that the signal voltage swings into the curved part of the characteristic. While the upper half-cycle of plate current reproduces the sine-wave shape of the positive half-cycle of signal voltage, the lower half-cycle of plate current is considerably distorted and bears little resemblance to the upper half-cycle of plate current.

As explained in § 2-7, a non-sinusoidal waveshape can be resolved into a number of sine-wave components or harmonics which are integral multiples of the lowest frequency present. Consequently, this type of distortion is known as *harmonic distortion*. Distortion re-

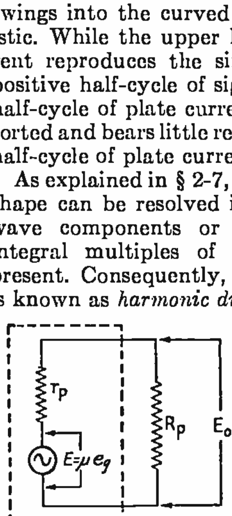


Fig. 308 — Equivalent circuit of the vacuum-tube amplifier. The tube is replaced by an equivalent generator having an internal resistance equal to the a.c. plate resistance of the vacuum tube.

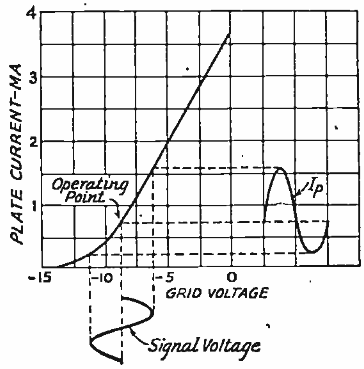


Fig. 310 — Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the grid voltage causing it.

sulting from grid-current flow, described in the preceding paragraph, also is harmonic distortion. Harmonic distortion from either or both causes may arise in the same amplifier.

Harmonic distortion may or may not be tolerable in an amplifier. At audio frequencies it is desirable to keep harmonic distortion to a minimum, but radio-frequency amplifiers are frequently operated in such a way that the r. f. wave is greatly distorted.

Frequency distortion — Another type of distortion, known as *frequency distortion*, occurs when the amplification varies with the frequency of the a.c. voltage applied to the grid circuit of the amplifier. It is not necessarily accompanied by harmonic distortion. It can be shown by a *frequency-response curve* or graph in which the relative amplification is plotted against frequency over the frequency range of interest.

Resistance-coupled amplifiers — An amplifier with a resistance load is known as a "resistance-coupled" amplifier. This type of amplifier is widely used for amplification at audio frequencies. A simplified circuit is shown in Fig. 311, where the amplifier is coupled to a following tube. Since all the power output of a resistance-coupled amplifier is consumed in the load resistor such amplifiers are used almost wholly for voltage amplification, usually working into still another amplifier.

A single amplifier is called a *stage* of amplification, and a number of amplifier stages in succession are said to be in *cascade*.

The purpose of the coupling condenser, C_c, is to transfer to the grid of the following tube the a.c. voltage developed across R_p, and to prevent the d.c. plate voltage on tube A from being applied to the grid of tube B. The grid resistor, R_g, transfers the bias voltage to the grid of tube B and prevents short-circuiting the a.c. voltage through the bias battery. Since no grid current flows, there is no d.c. voltage drop in R_g; consequently the full bias voltage is applied to the grid. In order to obtain the maximum a.c. voltage at the grid of tube B the reactance of the coupling condenser must be small compared to the resistance of R_g, so that most of the voltage will appear across R_g rather than across C. Also, the resistance of R_g must be large compared to R_p because, so far as the a.c. voltage developed in R_p is concerned, R_g is in parallel with R_p and therefore is just as much a part of R_p as though it were connected directly in parallel with it. (The impedance of the plate-supply battery is assumed to be negligible, so that there is no a.c. voltage drop between the lower end of R_p and the common connection between the two tubes.) In practice the maximum usable value of R_g is limited to from 0.5 to about 2 megohms, depending upon the characteristics of the tube with which it is associated. If the value is made too high, stray electrons collecting on the grid may not "leak off" back to the cathode rapidly enough to prevent the accumulation of a negative charge on the grid. This is equivalent to an increase in the negative grid bias, and hence to a shift in the operating point.

The equivalent circuit of the amplifier now includes C_c, R_g, and a shunt capacity, C_s, which represents the input capacity of tube B and the plate-cathode capacity of tube A, together with such stray capacity as exists in the circuit. The reactance of C_s will depend upon the frequency of the voltage being amplified, and, since C_s is in parallel with R_p and R_g, it also becomes part of the load impedance for the amplifier. At low frequencies — below 1000 cycles or so — the reactance of C_s usually is so high that it has practically no effect on the amplification, but, since the reactance decreases at higher frequencies, it is found that the amplification drops off rapidly when the reactance of C_s becomes comparable to the resistance of R_p and R_g in parallel. To maintain the amplification at high frequencies, it is necessary that R_p be relatively small if C_s is large, or that C_s be small if R_p is large.

Under the best conditions, in practice C_s will be of the order of 15 $\mu\mu$fd. or more, while it is

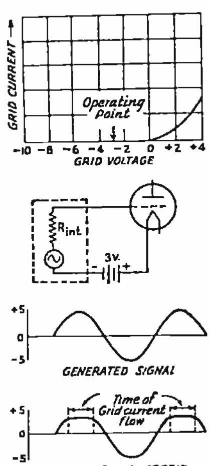


Fig. 309 — Distortion of applied signal because of grid-current flow. With the operating point at 3 volts negative bias, grid current will flow as shown by the curve whenever the applied signal voltage is more than 3 volts positive. If there is appreciable internal resistance, as indicated in the second drawing, there will be a voltage drop in the resistance whenever current is flowing but not during the period when no current flows. The signal will reach the grid unchanged so long as the instantaneous voltage is less than 3 volts positive, but the voltage at the grid will be less than the instantaneous voltage when the latter is above this figure. The shape of the negative half-cycle is unaltered.

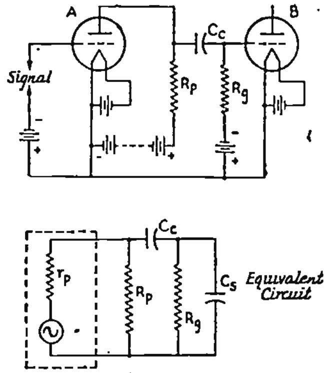


Fig. 311 — Typical resistance-coupled amplifier circuits.

possible for it to reach values as high as a few hundred $\mu\mu$fd. The larger values are encountered when tube *B* is a high-μ triode, as described in a later paragraph. Even with a low value of shunt capacity, the shunt reactance will decrease to a comparatively low value at the upper limit of the audio-frequency range; a shunting capacity of 20 $\mu\mu$fd., for example, represents a reactance of about 0.5 megohm at 15,000 cycles, and hence is of the same order as R_p for the type of tubes with which such a low value of capacity would be associated. In order to secure the same amplification at high as at low frequencies, therefore, it is necessary to sacrifice low-frequency amplification by reducing the value of R_p to the point where the reactance of C_s at the highest frequency of interest is considerably larger than R_p.

At radio frequencies the reactance of C_s becomes so low that the amount of amplification it is possible to realize is negligible compared to that which can be obtained in the audio-frequency range. The resistance-coupled amplifier, therefore, is used principally for audio-frequency work.

Impedance-coupled amplifiers — If either the plate resistor or grid resistor (or both) in the amplifier described in the preceding paragraph is replaced by an inductance, the amplifier is said to be *impedance-coupled.* The inductance or impedance is commonly substituted for the plate load resistor, so that the usual circuit for such an amplifier is as given in Fig. 312.

Considering the operation of the tube from the standpoint of the equivalent circuit of Fig. 308, it is evident that a voltage drop would exist across a reactance of suitable value substituted for the indicated load resistance, R_p, so long as the output of the generator is alternating current. From the physical standpoint, any change in the current flowing through the inductance in Fig. 312 would cause a self-induced e.m.f. having a value proportional to the rate of change of current and to the inductance of the coil. Consequently, if an a.c. signal voltage is applied to the grid of the tube, the resultant variations in plate current cause a corresponding a.c. voltage to appear across the coil terminals. This induced voltage is the useful output voltage of the tube.

The amplitude of the output voltage can be calculated, knowing the μ and plate resistance of the tube and the impedance of the load, in much the same way as in the case of resistance coupling, except that the equation must be modified to take account of the fact that the phase relationship between current and voltage is not the same in an impedance as it is in a resistance. In practice, the plate load inductance is shunted by the tube and stray capacities of the circuit as well as by its own distributed capacity. Since the greatest amplification will be secured when the load impedance is as high as possible, the coil usually is made to have sufficient inductance so that, in combination with these shunting capacities, the circuit as a whole will be parallel-resonant at some frequency near the middle of the audio-frequency range. Under these conditions the load impedance has its highest possible value, and is approximately resistive rather than reactive.

The equation for amplification with resistance coupling shows that, when R_p is several times the plate resistance, r_p, a further increase in R_p results in comparatively little increase in amplification. The load circuit of an impedance-coupled amplifier usually has an impedance value quite high in comparison to the plate resistance of the tube with which it is used, so that the load impedance can vary over a considerable range without much effect on the amplification. This gives the impedance-coupled amplifier an amplification vs. frequency characteristic which is fairly "flat" — that is, the amplification is practically constant with changes in frequency — over a considerable portion of the audio-frequency range. However, the performance of the impedance-coupled amplifier is not as good in this respect as that of a well-designed resistance-coupled amplifier.

If the impedance of the load circuit is high compared to the plate resistance of the tube, which will be the case if the tube is a low-μ triode and normal inductance values (a few hundred henrys) are used in the plate circuit,

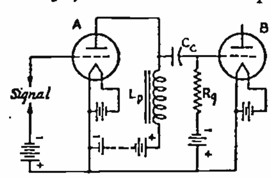


Fig. 312 — Impedance-coupled amplifier.

the amplification in the optimum frequency range will be practically equal to the μ of the tube. At lower frequencies the impedance decreases because of the decreasing reactance of the coil, while at higher frequencies the impedance again decreases because of the decreasing reactance of the shunt capacities. Thus the amplification drops off at both ends of the range, usually more rapidly than with resistance coupling.

The frequency-response characteristic of the impedance-coupled amplifier depends considerably upon the plate resistance of the tube. If impedance coupling is used with tubes of very high plate resistance, the response will be markedly greater at the resonant frequency than at frequencies either higher or lower.

Impedance coupling can be used at radio frequencies, since the inductance can be adjusted to resonate with the shunt capacities at practically any desired frequency.

Transformer-coupled amplifiers — The coupling impedance in Fig. 312 may be replaced by a transformer, connected as shown in Fig. 313. A.c. voltage is developed across the primary of the transformer in the same way as in the case of impedance coupling. The secondary of the transformer serves as a means for transferring the voltage to the grid of the following tube, and if the secondary has more turns than the primary the voltage across the secondary terminals will, in general, be larger than the voltage across the primary terminals.

As in the case of impedance coupling, the effective capacity shunting the primary of an audio-frequency transformer usually causes the primary circuit to be parallel-resonant at some frequency in the middle of the audio-frequency range. At the medium audio frequencies, therefore, the voltage across the primary is practically equal to the applied grid voltage multiplied by the μ of the tube. The voltage across the secondary will be the primary voltage multiplied by the secondary-to-primary turns ratio of the transformer, so that the total voltage amplification is μ times the turns ratio. The amplification at low frequencies depends upon the ratio of the primary reactance to the plate resistance of the tube, as in the case of impedance-coupled amplifiers.

At some high frequency, usually in the range 5000–10,000 cycles with ordinary transformers, the leakage inductance (§ 2-9) of the secondary becomes series resonant with the effective capacity shunting the secondary. At and near this resonant frequency the resonant rise in voltage may increase the amplification considerably, giving rise to a "peak" in the frequency-response curve of the amplifier. At frequencies above this resonance point amplification decreases rapidly, because as the reactance of the shunting capacity decreases it tends to act more and more as a short circuit across the secondary of the transformer. The relative height of the high-frequency peak depends principally upon the effective resistance of the secondary circuit. This effective resistance includes the actual resistance of the secondary coil and the "reflected" (§ 2-9) plate resistance of the tube, this resistance being in parallel with the primary of the transformer. Consequently, the height of the peak is affected by the tube with which the transformer is used. The peak can be reduced by connecting a 0.25 to 1 megohm resistor across the transformer secondary. While this helps to flatten the frequency response curve, it also reduces the amplification at medium and low frequencies.

Transformer coupling is most suitable for triodes of low or medium μ and having medium values of plate resistance. This is because the primary inductance required for good amplification at low frequencies is proportional to the plate resistance of the tube with which the transformer is to be used, and in practice it is difficult to obtain high primary inductance, a large secondary-to-primary turns ratio ("step-up ratio"), and low distributed capacity in the windings all at the same time. Increasing the primary inductance usually means that the turns ratio must be reduced, because the increase in distributed capacity as the coils are made larger tends to bring the resonant peak down to a relatively low frequency unless the secondary inductance is decreased to compensate for the increase in capacity. The step-up ratio seldom is more than 3 to 1 in transformers designed for good frequency response.

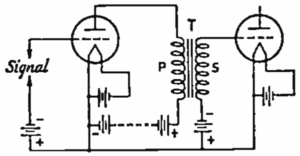


Fig. 313 — Transformer-coupled amplifier.

Transformer coupling can be used at radio frequencies if the transformers are properly designed for the purpose. In such transformers either the primary or secondary (or both) is made resonant at the frequency to be used, so that maximum amplification will be secured.

Phase relations in plate and grid circuits — When the exciting voltage on the grid has its maximum positive instantaneous value, the plate current also is maximum (§ 3-2), so that the voltage drop across the resistance connected in the plate circuit of a resistance-coupled amplifier likewise has its greatest value. The actual instantaneous voltage between plate and cathode is therefore minimum at the same instant, because it is equal to the d.c. supply voltage (which is unvarying) minus the voltage drop across the load resistance. When the signal voltage is at its negative peak the plate current has its least value, with the result that the voltage drop in the load resistance is less than at any other part of the cycle. At this instant, therefore, the voltage between plate and cathode is maximum.

These variations in plate-cathode voltage constitute the a.c. output of the tube, superimposed on the mean or no-signal plate-cathode voltage. Since the alternating plate-cathode voltage is decreasing when the instantaneous grid voltage is increasing (becoming more positive with respect to the cathode), the output voltage is less than the mean value, or negative, when the signal voltage is positive. Likewise, when the signal voltage is negative the output voltage is positive, or greater than the mean value. In other words, the alternating plate voltage is 180 degrees out of phase with the alternating grid voltage. Thus there is a *phase reversal* through the amplifier. The relationships should become clear from the behavior of the signal voltage and E_p in Fig. 307.

The same phase relationship between signal and output voltages holds when the amplifier is impedance- or transformer-coupled, in the frequency region where the load acts like a parallel-resonant circuit. However, if the load is reactive the phase relationship is not exactly 180 degrees but depends upon the kind of reactance present and the relative amounts of reactance and resistance. (This is true also of the resistance-coupled amplifier at low frequencies where the reactance of the coupling condenser affects the amplification, or at high frequencies where the reactance of the shunting capacities becomes important.) Since the reactance varies with the applied signal frequency, the phase relationship between signal voltage and output voltage depends upon the frequency in such cases.

Input capacity and resistance — When an alternating voltage is applied between the grid and cathode of an amplifier tube, an alternating current flows through the small condenser formed by these elements (§ 3-2) just as it would in any other condenser. Similarly, an alternating current also flows in the condenser formed by the grid and plate, since there is an alternating difference of potential between these elements. When the tube is amplifying, the alternating plate voltage and signal voltage are effectively applied in series across the grid-plate condenser, as indicated in Fig. 314. As described in the preceding paragraph, in the resistance-coupled amplifier the two voltages are out of phase with respect to the cathode, but inspection of the circuit shows that they are in phase so far as the grid-plate condenser is concerned. Consequently, the voltage applied to the grid-plate capacity is the sum of the alternating grid and plate voltages, or $E_g + E_p$. Since E_p is equal to $A \times E_g$, where A is the voltage amplification of the tube and circuit, the a.c. voltage between the grid and plate is E_g $(1 + A)$. The current, I, flowing in the grid-plate capacity is E_g $(1 + A)$ divided by the reactance of the grid-plate condenser, and thus is proportional to the grid-plate capacity.

The signal voltage must help in causing this relatively large current to flow, and, since the reactance as viewed from the input circuit is $X_g = E_g/I$, the input reactance becomes smaller as the current becomes larger. That is, the *effective* input capacity of the amplifier is increased when the tube is amplifying. From the above, the increase in input capacity is approximately proportional to the voltage amplification of the circuit and to the grid-plate capacity of the tube. The total input capacity is the sum of the grid-cathode capacity and this additional effective capacity. The total input capacity of an amplifier may reach values ranging from 50 to a few hundred micromicrofarads, if the voltage amplification is high and the grid-plate capacity relatively large. Both usually are true in a high-μ triode.

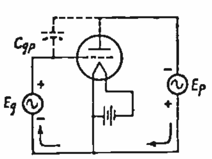


Fig. 314 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

When the load is reactive the a.c. grid and plate voltages still act in series across the grid-plate condenser, but since they are not exactly 180 degrees out of phase with respect to the cathode they are not exactly in phase with respect to the grid-plate capacity. The lack of exact phase relationship indicates that resistance as well as capacity is introduced into the input circuit. Analysis shows that, when the reactance of the load circuit is capacitive, the resistance component is positive — that is, it represents a loss of power in the input circuit — and that when the load circuit has inductive reactance the resistance component is negative. Negative resistance indicates that power is being supplied to the grid circuit from the plate.

Feed-back — If some of the amplified energy in the plate circuit of an amplifier is coupled back into the grid circuit, the amplifier is said to have *feed-back*. If the voltage fed from the plate circuit to the grid circuit is in such phase that, when it is added to the signal voltage already existing, the sum of the two voltages is larger than the original signal voltage, the feed-back is said to be *positive*. Positive feed-back usually is called *regeneration*. If regeneration exists in a circuit the total amplification is increased because the feed-back increases the amplitude of the signal at the grid and this larger signal is amplified in the same ratio, giving a greater output voltage than would exist if the signal voltage alone were present in the grid circuit. Many types of circuits can be used to secure positive feed-back. A simple one is shown in Fig. 315. The feed-back coil, L, a third winding on the grid-circuit transformer, is connected in series with the primary of the transformer in the plate circuit, so that some of the amplified voltage appears across its terminals. This induces a voltage in the secondary, S, of the grid-circuit transformer which, if the winding directions of the two coils are correct, will increase the value of signal voltage applied to the grid.

Positive feed-back is accompanied by a tendency to give maximum amplification at only one frequency, since the feed-back voltage will tend to be highest at the frequency at which the original amplification is greatest. It therefore increases the selectivity of the amplifier, and hence is used chiefly where high gain and sharpness of resonance both are wanted.

If the phase of the voltage fed back to the grid circuit is such that the sum of the feed-back voltage and the original signal voltage is less than the latter alone, the feed-back is said to be *negative*. Negative feed-back frequently is called *degeneration*. In this case the total amplification is decreased, since the grid signal has been made smaller, and hence the amplified output voltage is smaller for a given original signal than it would be without feed-back.

The amount of voltage fed back will depend upon the actual amplification of the tube and circuit, and if the amplification ratio tends to change, as it may at the extreme high or low frequencies in the audio-frequency range, the feed-back voltage will be reduced when the amplification decreases. For example, suppose that an amplifier has a voltage gain of 20 and that it is delivering an output voltage of 50 volts. Without feed-back, the grid signal voltage required to produce 50 volts output is 50/20 or 2.5 volts. But suppose that 10 per cent of the output voltage (5 volts) is fed back to the grid circuit in opposite phase to the applied grid voltage. Then, since it is still necessary to have a 2.5-volt signal to produce 50 volts output, the applied voltage must be 2.5 + 5 or 7.5 volts. Now suppose that at some other frequency the voltage gain drops to 10. Then for the same 50-volt output a 5-volt signal is required, but since the feed-back voltage is still 5 volts the total required signal is now 10 volts. With feed-back the gain in the first case was 50/7.5 volts or 6.66 and in the second case 50/10 or 5, the gain in the second case being 75 per cent as high as in the first. Without feed-back the gain in the second case was 50 per cent as high as in the first. The effect of feed-back therefore is to make the resultant gain more uniform, despite the tendency of the amplifier itself to discriminate against certain frequencies.

Negative feed-back also tends to decrease harmonic distortion arising in the plate circuit of the amplifier. This distortion is present in the amplified output voltage, but not in the original signal voltage applied to the grid. The voltage fed back to the grid circuit contains the distortion but in opposite phase to the distortion components in the plate circuit, hence the two tend to cancel each other. For similar reasons, the over-all amplification is less dependent upon the value of load impedance used in the plate circuit; in fact, if a large amount of negative feed-back is used in an amplifier it is even possible to substitute tubes of rather widely different characteristics without much effect on the over-all performance.

Both positive and negative feed-back may be applied over several stages of an amplifier, rather than being applied directly from the plate circuit to the grid circuit of a single stage.

Power amplification — In the types of amplifiers previously described, the chief consideration was that of securing as much voltage gain as possible within the permissible limits of harmonic distortion and frequency response characteristic. Such amplifiers are principally used to furnish an amplified signal voltage, which in turn can be supplied to a succeeding amplifier. If the succeeding amplifier is operated in such a way that its grid is never driven positive with respect to its cathode, grid current does not flow, and hence the power requirements are negligibly small. However, if an amplifier is used to actuate some power-consuming device, such as a loudspeaker or a succeeding amplifier in which it is permissible to drive the grid into the positive region, the primary consideration is that of obtaining the maximum power output consistent with the permissible distortion. In such a case the voltage at which the power is secured is of little consequence, since a transformer may be used to change the voltage to any desired value, within reasonable limits. Hence, the voltage gain of a power amplifier is of little importance.

In power-amplifier operation the grid may or may not be driven into the positive region, depending upon the particular application. The present discussion will be confined to the triode amplifier operating without grid current; other types are considered in § 3-4. The principles upon which such a power amplifier operates are practically identical with those already described. The chief differences between a voltage amplifier and a power amplifier lie in the selection of tubes and in the choice of the value of load resistance. As previously described, if voltage gain is the primary consideration the load resistance should be as large as possible in comparison to the plate resistance of the tube. It can be shown that, in any electrical circuit, maximum *power* output is secured when the resistance of the load is made equal to the internal resistance of the source of power. This is true whether the power source is a battery, a generator or a vacuum tube. In the case of the vacuum tube the internal resistance is the plate resistance of the tube, so that for maximum power output the load resistance should be made equal to the plate resistance. However, when the tube is operated with so low a value of load resistance there is considerable harmonic distortion, and *optimum* power output, representing an acceptable compromise between distortion and the power obtainable, is secured when the load resistance is approximately twice the plate resistance.

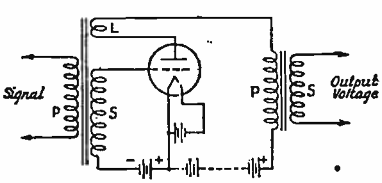


Fig. 315—An elementary form of feed-back circuit. The feed-back may be either positive or negative, depending upon how the coil *L* is connected in the circuit. This type of circuit illustrates the principle of feed-back, but it is not practical for use in an actual audio-frequency amplifier.

Power-amplifier circuits — The plate or output circuit of a power amplifier almost invariably is transformer-coupled to the power-consuming device or load with which it is associated. This is because the impedance of the desired load seldom is the proper value for obtaining optimum power output from the amplifier. Consequently, the load impedance must be changed to a value suitable for the plate circuit of the amplifier tube. This can be done by the use of transformers, as described in § 2-9.

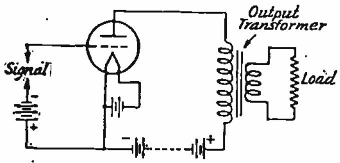


Fig. 316 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

A basic power-amplifier circuit is shown in Fig. 316. So long as the amplifier is operated entirely in the negative-grid region and no grid current flows, any of the previously described types of coupling may be used between the grid of the power amplifier and the preceding amplifier. If there is no preceding amplifier, the method of coupling will depend principally on the characteristics of the source of the signal.

In Fig. 316 the load is represented as a resistance. An actual load may have a reactance as well as a resistance component, but only the resistance will consume power (§ 2-8).

Power amplification ratio — The ratio of a.c. output power to the a.c. power consumed in the grid circuit (*driving power*) is called the *power amplification ratio* or simply *power amplification* of the amplifier. If the amplifier operates without grid current the a.c. power consumed in the grid circuit is negligibly small, so that the power amplification ratio of such an amplifier is extremely large. With other types of operation the power amplification ratio may be relatively small, as described in § 3-4.

Plate efficiency — The ratio of a.c. output power to the d.c. power supplied to the plate circuit is called the *plate efficiency* of the amplifier. It is expressed as a percentage:

$$\% \text{ plate efficiency} = \frac{P_o}{EI} \times 100$$

where P_o is the a.c. output power, E the plate voltage and I the plate current, the latter two being d.c. values.

The plate efficiency of amplifiers designed for minimum distortion and a high power amplification ratio (operation without grid current) is relatively low — of the order of 15 to 30 per cent. For minimum distortion the operation must be confined to the region where the waveshape of the alternating plate current is substantially identical with that of the signal on the grid, and, as previously explained, this requirement can be met only by limiting the plate-current variations (that is, the alternating component of plate current) to the straight portion of the dynamic grid voltage vs. plate current characteristic. Since with a given load resistance the power output is proportional to the square of the alternating component of plate current, it follows that limiting the plate-current variation also limits the power output in comparison to the d.c. plate power input.

Higher plate efficiency can be secured by increasing the alternating component of plate current, but this is accompanied by increased distortion. Special types of amplifiers have been devised to compensate for this distortion, as described in the next section. In some applications, as in r.f. power amplification, the fact that the signal applied to the grid is greatly distorted is of no consequence, so that such amplifiers can have high plate efficiency.

Power sensitivity — The ratio of a.c. power output to alternating grid voltage is called the *power sensitivity* of an amplifier. It provides a convenient measure for comparing power tubes, especially those designed for audio-frequency amplification where the operation is to be without grid current, since it expresses the relationship between power output and the amount of signal voltage required to produce the power.

The term power sensitivity also is used in connection with radio-frequency power amplifiers, in which case it has the same meaning as power amplification ratio. A tube which delivers its rated output power with a relatively small amount of power consumed in the grid circuit is said to have high power sensitivity.

Parallel operation — When it is necessary to obtain more power output than one tube is capable of giving, two or more tubes may be connected in *parallel*. In this case the similar elements in all tubes are connected together. This method is shown in Fig. 317 for a transformer-coupled amplifier. The power output of a parallel stage will be in proportion to the number of tubes used; the exciting voltage required, however, is the same as for one tube.

If the amplifier operates in such a way as to consume power in the grid circuit, the grid power required also is in proportion to the number of tubes used.

Push-pull operation — An increase in power output can be secured by connecting two tubes in *push-pull*, the grids and plates of the two tubes being connected to opposite ends of the circuit as shown in Fig. 317. A "balanced" circuit, in which the cathode returns are made to the midpoint of the input and output devices, is necessary with push-pull operation. At any instant the ends of the secondary winding of the input transformer, T_1, will be at opposite potentials with respect to the cathode connection, so that the grid of one tube is swung positive at the same instant that the grid of the other is swung negative. Hence, in any push-pull-connected stage the voltages and currents of one tube are out of phase with those of the other tube. The plate current of one tube is rising while the plate current of the other is falling, hence the name "push-pull." In push-pull operation the even-harmonic (second, fourth, etc.) distortion is cancelled in the symmetrical plate circuit, so that for the same power output the distortion will be less than with parallel operation.

The exciting voltage measured between the two grids must be twice that required for one tube. If the grids consume power, the driving power for the push-pull stage is twice that taken by either tube alone.

The decibel — The ratio of the power levels at two points in a circuit such as an amplifier can be expressed in terms of a unit called the *decibel*, abbreviated *db.* The number of decibels is 10 times the logarithm of the power ratio, or

$$\text{db.} = 10 \log \frac{P_1}{P_2}$$

The decibel is a particularly useful unit because it is logarithmic, and thus corresponds to the response of the human ear to sounds of varying loudness. One decibel is approximately the power ratio required to make a just noticeable difference in sound intensity. Within wide limits, changing the power by a given ratio produces the same apparent change in loudness regardless of the power level; thus if the power is doubled the increase is 3 db., or three steps of intensity; if it is doubled again the increase is again 3 db., or three further distinguishable steps. Successive amplifications expressed in decibels can be added to obtain the over-all amplification.

A power loss also can be expressed in decibels. A decrease in power is indicated by a minus sign (e.g., − 7 db.), and an increase in power by a plus sign (e.g., + 4 db.). Negative and positive quantities can be added numerically. Zero db. indicates the reference power level, or a power ratio of 1.

Applications of amplification — The major uses of vacuum-tube amplifiers in radio work are for amplifying at audio and radio frequencies (§ 2-7). The audio-frequency amplifier generally is used to amplify without discrimination at all frequencies in a wide range (say from 100 to 3000 cycles for voice communication), and therefore is associated with nonresonant or untuned circuits which offer a uniform load over the desired range. The radio-frequency amplifier, on the other hand, generally is used to amplify selectively at a single radio frequency, or over a small band of frequencies at most, and therefore is associated with resonant circuits tunable to the desired frequency.

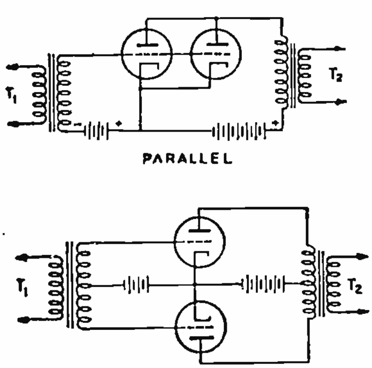


Fig. 317 — Parallel and push-pull a.f. amplifier circuits.

An audio-frequency amplifier may be considered a *broad-band amplifier;* most radio-frequency amplifiers are designed to have relatively narrow bandwidths.

In audio circuits the power tube or output tube in the last stage usually is designed to deliver a considerable amount of audio power, while requiring but negligible power from the input or exciting signal. To get the alternating voltage (*grid swing*) required for the grid of such a tube, voltage amplifiers are used employing high-μ tubes which greatly increase the voltage amplitude of the signal. Voltage amplifiers are used in the radio-frequency stages of receivers as well as in audio amplifiers; power amplifiers are used in the radio-frequency stages of transmitters.

ℂ 3-4 Classes of Amplifiers

Reason for classification — It is convenient to divide amplifiers into groups according to the work they are intended to perform, as related to the operating conditions necessary to accomplish the purpose. This makes identification easy and obviates the necessity for giving a detailed description of the operation when *specific* operating data are not required.

Class A — An amplifier operated as shown in Fig. 306 or 307, in which the output waveshape is a faithful reproduction of the input waveshape, is known as a *Class-A* amplifier.

As generally used, the grid of a Class-A amplifier never is driven positive with respect to the cathode by the exciting signal, and never is driven so far negative that plate-current cut-off is reached. The plate current is constant both with and without grid excitation. The chief characteristics of the Class-A amplifier are low distortion, relatively low power output for a given size of tube, and a high power-amplification ratio. The plate efficiency is relatively low (§ 3-3).

Class-A power amplifiers find application as output amplifiers in audio systems and as drivers for Class-B power amplifiers. Class-A voltage amplifiers are found in the stages preceding the power stage or stages in such applications, and as r.f. amplifiers in receivers.

Class B — The *Class-B* amplifier is primarily one in which the output current, or alternating component of the plate current, is proportional to the amplitude of the exciting grid voltage. Since power is proportional to the square of the current, the power output of a Class-B amplifier is proportional to the square of the exciting grid voltage.

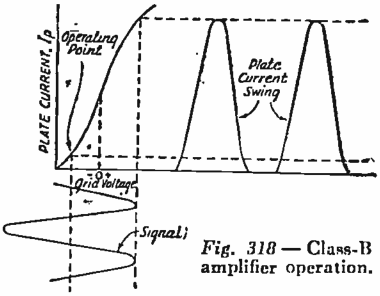


Fig. 318 — Class-B amplifier operation.

In Class-B service the grid bias is set so that the plate current is relatively low without grid excitation; the exciting signal amplitude is made such that the entire linear portion of the characteristic is used. Fig. 318 illustrates operation with the tube biased practically to cut-off. In this condition plate current flows only during the positive half-cycle of excitation. No plate current flows during the negative half-cycle. The shape of the plate current pulse is essentially the same as that of the positive swing of the signal voltage. Since the plate current is driven up toward the saturation point, it is usually necessary for the grid to be driven positive with respect to the cathode during part of the grid swing. Grid current flows, therefore, and the driving source must furnish power to supply the grid losses.

Class-B amplifiers are characterized by medium power output, medium plate efficiency (50 to 60 per cent at maximum signal), and a moderate ratio of power amplification. At radio frequencies they are used as *linear amplifiers* to raise the output power level in radiotelephone transmitters after modulation.

For Class-B audio-frequency amplification two tubes must be used, the second tube working alternately with the first so that both halves of the cycle will be present in the output. A typical method of achieving this is shown in Fig. 319. The signal is fed to a transformer, T_1, whose secondary is divided into two equal parts, with the tube grids connected to the outer terminals and the grid bias fed in at the center. A transformer, T_2, with a similarly divided primary, is connected to the plates of the tubes. When the signal voltage in the upper half of T_1 is positive with respect to the center connection (*center tap*), the upper tube draws plate current while the lower tube is idle; when the lower half of T_1 becomes positive, the lower tube draws plate current while the upper tube is idle. The voltages induced in the primary of T_2 combine in the secondary to produce an amplified reproduction of the signal.

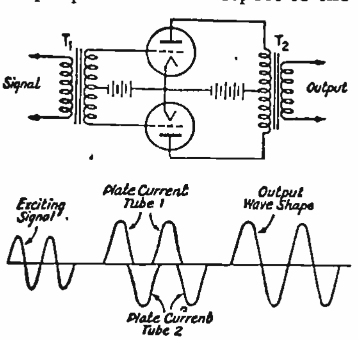


Fig. 319 — Showing how the outputs of the two tubes in push-pull are combined in the Class-B audio amplifier.

Class AB — The similarity between the *Class-AB* amplifier, Fig. 319, and the ordinary push-pull circuit (Fig. 317) will be noted. Actually, the only difference lies in the method of operation. If the bias is adjusted so that the tubes draw a moderate value of plate current with no signal, the amplifier will operate Class A at low signal voltages and more nearly Class B at high signal voltages. This method gives low distortion at moderate signal levels and high plate efficiency at high signal levels, making possible the use of relatively small tubes in audio power amplifiers.

A further distinction can be made between amplifiers which draw grid current and those which do not. The *Class-AB*$_1$ amplifier draws no grid current and thus consumes no power from the driving source. The *Class-AB*$_2$ amplifier draws grid current at higher signal levels, and power must be supplied to its grid circuit.

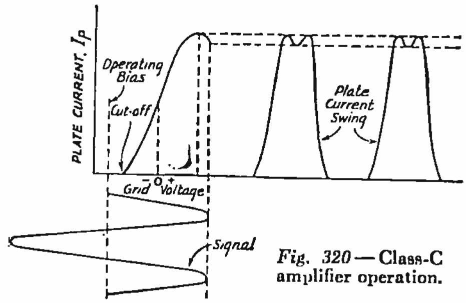


Fig. 320 — Class-C amplifier operation.

Class C — The *Class-C* amplifier is one operated so that the alternating component of the plate current is directly proportional to the plate voltage. The output power is therefore proportional to the square of the plate voltage. Other characteristics inherent to Class-C operation are high plate efficiency, high power output, and relatively low power amplification.

The grid bias is set at a value at least twice that required for plate-current cut-off without excitation. Thus plate current flows during only a fraction of the positive excitation cycle. The exciting signal should be of sufficient amplitude to drive the plate current to the saturation point, as shown in Fig. 320. Since the grid must be driven far into the positive region to cause saturation, considerable numbers of electrons are attracted to the grid at the peak of the cycle, robbing the plate of some that it would normally attract. This causes the droop at the upper bend of the characteristic, and also may cause the plate-current pulse to be indented at the top. The output wave-form is badly distorted, but at radio frequencies the distortion is largely eliminated by the flywheel effect of the tuned output circuit.

3-5-A Multi-Grid Tubes

Radio-frequency amplification — As described in the preceding section, the reactances of the grid-to-cathode and plate-to-cathode capacities (together with unavoidable stray capacities) in a vacuum tube become very low at frequencies higher than the audio-frequency range. As a result, ordinary resistance, impedance or transformer coupling cannot be used at radio frequencies, because these capacities act as low-reactance by-passes across the input and output circuits; hence the total impedance in either the plate or grid circuit is too low for appreciable voltage to be developed.

When an amplifier is to be operated at radio frequencies it is necessary to use resonant circuits as loads, the circuits being tuned to the frequency to be amplified. Since such circuits consist of coils and condensers, the tube and stray capacities become part of the total tuning capacity and are thus made to serve a useful purpose. As described in § 2-10, the parallel impedance of a resonant circuit can reach quite high values when the Q is high. Values of parallel-resonant impedance suitable for effective amplification are readily obtainable with reasonably well-designed tuned circuits.

Since maximum parallel impedance, and consequently maximum amplification when resonant circuits are associated with an amplifier tube, is obtained when the circuit is exactly resonant at the applied frequency, it is necessary that the resonant circuit associated with the grid and that connected to the plate be tuned to the same frequency. In practice, it is difficult to maintain exact tuning over a period of time. If the amplifier tube is a triode, its input circuit will have a negative-resistance characteristic (§ 3-3) when the plate-circuit load has inductive reactance. If the resonant circuit associated with the plate is tuned slightly to the high-frequency side of exact resonance, the circuit will have inductive reactance, and energy will be transferred from the plate circuit to the grid circuit. Such a circuit has positive feed-back, or is regenerative. If enough energy is so transferred (very little is required) the tube will generate a self-sustaining r.f. current, in which case it is said to be *oscillating*. When oscillation commences the circuit ceases to amplify incoming signals, since it is generating a signal of its own.

Oscillation in triode amplifiers can be prevented by using special circuits but in practice these are unsatisfactory in receiving applications. Since the feed-back arises because of the grid-plate capacity of the tube, it is best eliminated by reducing the grid-plate capacity.

Screen-grid tubes — The grid-plate capacity can be eliminated, or at least reduced to a negligible value, by inserting a second grid between the control grid and the plate as indicated in Fig. 321. The second grid, called the *screen grid* or *shield grid*, acts as an electrostatic shield (§ 2-11) between the control grid and plate. It is made in the form of a grid or coarse screen rather than as a solid metal sheet, so that electrons can pass through it to the plate; a solid shield would entirely prevent the flow of plate current. The screen grid is connected to the cathode through a by-pass condenser, which has low impedance at the radio frequency being amplified. The electric lines of force from the plate terminate on the screen grid, very little of the field getting through to the control grid; similarly, the field set up by the control grid does not penetrate past the screen grid. Thus there is no common field between the control grid and plate; hence no capacity between these two tube elements.

Since the electric field from the plate does not penetrate into the region occupied by the control grid, which is the region in which most of the space charge is concentrated, the plate is unable to exert an attraction upon the electrons in this region. Consequently, the plate voltage cannot control the flow of plate current as it does in a triode. In order to get electrons to the plate, it is necessary to apply a positive potential (with respect to the cathode) to the screen. The screen then attracts electrons much as does the plate in a triode tube. In traveling toward the screen the electrons acquire velocity, so that most of them shoot between the screen wires into the field from the plate. Those that pass through and are attracted to the plate constitute the plate current of the tube. A certain proportion do strike the screen, however, with the result that some current also flows to the screen grid. The screen current will be low compared to the plate current in a *tetrode*, or four-element tube, however.

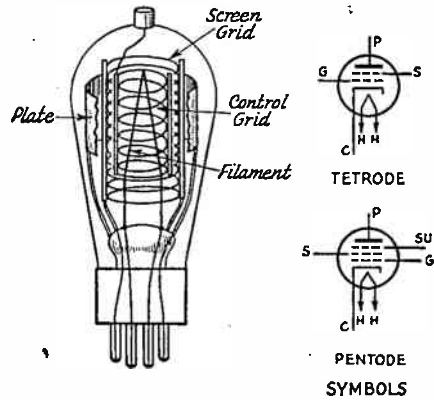


Fig. 321 — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen-grid cut away. The screen grid usually is made longer than either the control grid or plate, so that the shielding will be as effective as possible. In this drawing the control grid connection is made through a cap on the top of the tube, thus eliminating the capacity which would exist between the plate and grid lead wires if both passed through the base. Some modern tubes which have both leads going through the base use special shielding and construction to eliminate capacity. Symbols for pentode and tetrode tubes: H, heater; C, cathode; G, control grid; P, plate; S, screen grid; Sup., suppressor grid.

Secondary emission — When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons. These "splash" from the plate into the interelement space, a phenomenon called *secondary emission*. In a triode ordinarily operated with the grid negative with respect to cathode, secondary electrons are repelled back into the plate and cause no disturbance. In the screen-grid tube, however, the positively charged screen attracts the secondary electrons, causing a reverse current to flow between screen and plate. The effect is particularly marked when the plate and screen potentials are nearly equal, which may be the case during the part of the a.c. cycle when the instantaneous plate current is large and the plate voltage low (§ 3-3).

Pentode tubes — To overcome the effects of secondary emission, a third grid, called the *suppressor grid*, may be inserted between the screen and plate. This grid, which is connected directly to the cathode, repels the relatively low-velocity secondary electrons. They are driven back to the plate without appreciably obstructing the regular plate-current flow.

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, it is quite obvious that the control grid still can control the plate current in essentially the same way that it does in a triode, since the control grid is still in the space-charge region. Consequently, the grid-plate transconductance (or mutual conductance) of a tetrode or pentode will be of the same order of value as in a triode of corresponding structure. On the other hand, since the plate voltage has very little effect on the plate-current flow, both the amplification factor and plate resistance of a pentode or tetrode are very high, as is apparent from the definitions of these constants (§ 3-2). In small receiving pentodes the amplification factor is of the order of 1000 or higher, while the plate resistance may be from 0.5 to 1 or more megohms. Because of the high plate resistance, the actual voltage amplification possible with a pentode is very much less than the large amplification factor might indicate. In resistance-coupled audio-frequency amplifiers, voltage amplification or gain of 100 to 200 is typical.

A typical set of characteristic curves for a small pentode is shown in Fig. 322. That the plate voltage has little effect on the plate current is indicated by the fact that the curves are practically horizontal once the plate voltage is high enough to prevent the electrons in the space between the screen grid and the plate from being attracted back to the screen. The plate potential at which this occurs is less than the screen potential, because the electrons entering the space have considerable velocity and hence tend to move away from the screen despite the fact that it has a positive charge.

In addition to their applications as radio-frequency amplifiers, pentode or tetrode screen grid tubes also can be constructed for audio-frequency power amplification. In tubes designed for this purpose the shielding effect of the screen grid is not so important; the chief function of the screen is to serve as an accelerator of the electrons, so that large values of plate current can be drawn at relatively low plate voltages. Such tubes have quite high power sensitivity (§ 3-4) compared to triodes of the same power output, because the amplification factor of an equivalent triode has to be made quite low in order to secure the same plate current at the same plate voltage. Because of the low μ, the triode requires a relatively large signal voltage for full output, hence has low power sensitivity. The harmonic distortion is somewhat greater with pentodes and tetrodes than with triodes, however.

Variable-mu and sharp cut-off tubes — Receiving screen-grid tetrodes and pentodes for radio-frequency voltage amplification are made in two types, known as *sharp cut-off* and *variable-μ* or "super-control" types. In the sharp cut-off type the amplification factor is practically constant regardless of grid bias, while in the variable-μ type the amplification factor decreases as the negative bias is increased. The purpose of this design is to permit the tube to handle large signal voltages without distortion in circuits in which grid-bias control is used to vary the mutual conductance, and hence the amplification.

The way in which mutual conductance varies with grid bias in two typical small receiving pentodes, similar except in that one is a sharp cut-off type and the other a variable-μ type, is shown in Fig. 323. Obviously, the variable-μ type can handle a much larger signal voltage without swinging beyond either the point of zero grid bias or of plate-current cut-off (zero mutual conductance), if the bias is properly chosen.

Beam tubes — A "beam"-type tube is a tetrode with grids so constructed as to form the electrons traveling to the plate into concentrated beams, resulting in higher plate efficiency and power sensitivity. Suitable design also overcomes the effects of secondary emission without the necessity for a suppressor grid. Tubes constructed on the beam principle are used in receivers as both r.f. and audio amplifiers, and are built in larger sizes for transmitting circuits.

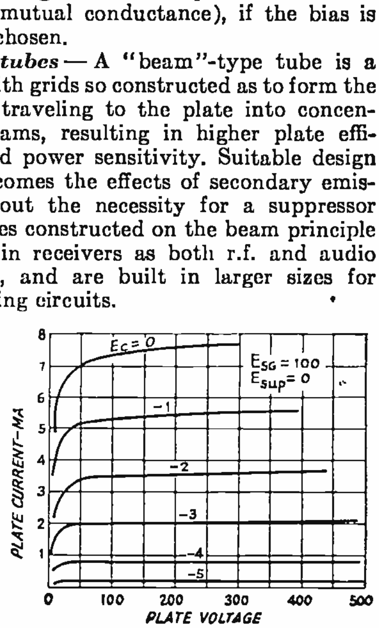


Fig. 322 — Plate voltage vs. plate current curves for a small receiving pentode. Screen-grid voltage, E_{sg}, is 100 volts and suppressor-grid voltage, E_{sup}, is zero.

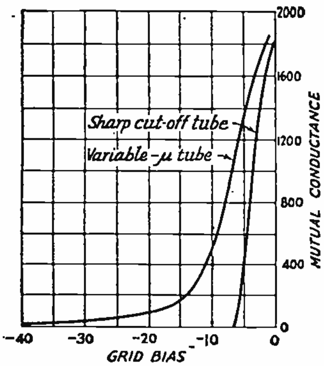


Fig. 323 — Curves showing the relationship between mutual conductance vs. negative grid bias for two small receiving pentodes, one being a sharp cut-off type and the other a variable-μ type.

3-5-B Special-Purpose Tubes

Multi-purpose types — A number of combination types of tubes have been constructed to perform multiple functions, particularly in receiver circuits. For the most part these are multi-unit tubes made up of individual tube element structures, combined in a single bulb for compactness and economy. Among the simplest are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb for Class-B audio amplification. More complex types include duplex-diode triodes, duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on. In many cases the nature can be identified by the name.

Mercury-vapor rectifiers — For a given value of plate current, the power lost in a diode rectifier (§ 3-1) will be lessened if it is possible to decrease the plate-cathode voltage at which the current is obtained. If a small amount of mercury is put in the tube, the mercury will vaporize when the cathode is heated, and, further, will ionize (§ 2-4) when plate voltage is applied. The positive ions neutralize the space charge and reduce the plate-cathode voltage drop to a practically constant value of about 15 volts, regardless of the value of plate current. Since this voltage drop is smaller than can be attained with purely thermionic conduction, there is less power loss in the rectifier. Voltage drop is constant despite variations. Mercury-vapor tubes are widely used in power rectifiers.

Grid-control rectifiers — If a grid is inserted in a mercury-vapor rectifier it is found that with sufficient negative grid bias it is possible to prevent plate current from flowing, but only if the bias is present before plate voltage is applied. If the bias is lowered to the point where plate current can flow, the mercury vapor will ionize and the grid will lose control of plate current, since the space charge disappears when ionization occurs. It can assume control again only after the plate voltage is reduced below the ionizing potential. The same phenomenon also occurs in triodes filled with other gases which ionize at low pressure. Grid-control rectifiers find considerable application in "electronic switching" circuits.

3-6 Vacuum-Tube Cathodes

Types of cathodes — Cathodes are of two general types, directly and indirectly heated. Directly-heated cathodes or filaments are of the *oxide-coated* type, consisting of a wire or ribbon of tungsten coated with an oxide capable of emitting large numbers of electrons with comparatively little cathode-heating power.

When directly-heated cathodes are operated on alternating current, the cyclic variation of current causes the plate current of the tube to vary at the supply-frequency rate, producing hum in the output. Hum from this source is eliminated in the *indirectly heated* cathode, consisting of a thin metal sleeve or thimble, coated with electron-emitting oxides, enclosing a tungsten wire which acts as a heater. The heater brings the cathode to the proper temperature to cause electron emission. This type is also known as the *equi-potential* cathode since all of it is at the same potential, in contrast to the directly heated filament where a voltage drop occurs along the wire.

The source of filament power for a directly heated cathode — battery or transformer — necessarily is directly connected to the tube circuit. With an indirectly heated cathode the source of heating power can be entirely independent of the tube circuit, since the electron-emitting cathode need not be electrically connected to the heating element.

While the oxide-coated cathode emits the largest number of electrons per watt of heating power, it is suitable only for tubes operating at plate voltages of 1000 volts or less. In manufacture there is a tendency for the electron-emitting material to be deposited on the control grid which, while of little consequence in receiving tubes, leads to difficulties in power-tube operation. A receiving tube is usually operated with little or no grid current, so that only very small amounts of power must be dissipated by the grid. On the other hand, in a power tube operated as a Class-C amplifier or oscillator the power the grid must dissipate is relatively high and its temperature rises cor-

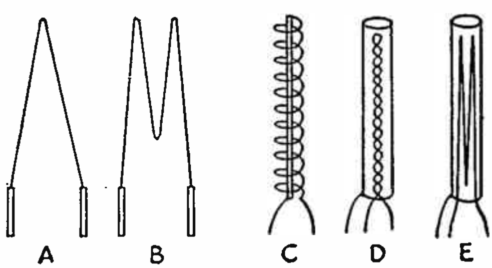


Fig. 324 — Types of cathode construction. Directly heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly heated cathodes at D and E show two types of heater construction, one a twisted loop and the other bunched heater wires. Both types tend to cancel the magnetic fields set up by the current through the heater.

respondingly. In tubes handling large amounts of power, further raising the internal tube temperature, the control-grid temperature may become high enough to cause electron emission from the deposited oxide material. When this occurs, the grid is in effect a second cathode. The electrons so emitted increase the total plate current without increasing the power output, with the result that the plate efficiency is lowered and the plate temperature rises. This in turn tends to increase the grid temperature, causing more electron emission. In a short time the tube will "run away" — its plate current increasing to unsafe values while the power output decreases.

A second factor which makes oxide-coated cathodes unsuitable for power tubes is bombardment of the cathode by positive ions. Although by far the greater part of the air is pumped out of the tube, there is always some residual gas. The gas molecules become ionized (§ 2-4), separate into an electron and a positively charged ion. The heavy positive ion is repelled by the positive charge on the plate and driven at high velocity into the cathode, stripping whole sections of electron-emitting material from the heating wire or sleeve.

Tubes for operation at less than three or four thousand volts usually have *thoriated-tungsten* cathodes. This type is directly heated, consisting of a filament of tungsten containing dissolved thoria which makes the filament much more efficient as an electron emitter. While the thoriated filament is in turn less efficient than the oxide-coated type, it is free from the grid-emission effects described above and is less susceptible to positive ion bombardment than the oxide-coated type. Less gas is trapped or "occluded" in the cathode itself, but the effect of the thorium can be destroyed (if only temporarily) by bombardment.

Tubes built to operate at several thousand volts are provided with cathodes of pure tungsten. This material must be operated at high temperature (white heat) for reasonable electron-emission efficiency, but even then is very much less efficient than either the thoriated or oxide-coated cathode. However, it is the only type satisfactory in high-voltage operation.

The operating temperature of a thoriated tungsten filament is fairly critical, and the specified filament voltage should be maintained within a few per cent. These filaments, as well as oxide-coated cathodes, eventually "lose emission"; that is, the emission efficiency of the cathode decreases sufficient electron emission for satisfactory tube operation cannot be obtained without raising the cathode temperature to an unsafe value. Pure tungsten cathodes do not lose emission, but the high operating temperature makes them more susceptible to "burn-out."

Methods of obtaining grid bias — Grid bias may be obtained from a source of voltage especially provided for that purpose, such as a battery or other type of d.c. power supply. This is indicated in Fig. 325-A. A second method, utilizing a *cathode resistor*, is shown at B; d.c. plate current flowing through the resistor causes a voltage drop which, with the connections shown, has the right polarity to bias the grid negatively with respect to the cathode. The value of the resistor is determined by the bias required and the plate current which flows at that value of bias, as found from the tube characteristic curves; with the voltage and current known, the resistance can be determined by Ohm's Law (§ 2-6):

$$R_c = \frac{E \times 1000}{I_c}$$

where R_c = cathode bias resistor in ohms
E = desired bias voltage
I_c = total d.c. cathode current in milliamperes.

If the tube is a multi-element type, the screen- and suppressor-grid currents should be added to the plate current to obtain the total cathode current. The control-grid current also should be included if the control grid is driven positive.

The a.c. component of plate current flowing through the cathode resistor will cause an a.c. voltage drop which gives negative feed-back (§ 3-3) into the grid circuit, and thus reduces the amplification. To prevent this, the resistor usually is by-passed (§ 2-13), C_c being the *cathode by-pass condenser*. To be effective, the reactance of the by-pass condenser must be small compared to R_c at the frequency being amplified. This condition generally is satisfied if the reactance is 10 percent or less of the cathode resistance. In audio-frequency amplifiers, the *lowest* frequency at which full amplification must be secured should be used in calculating the required capacity.

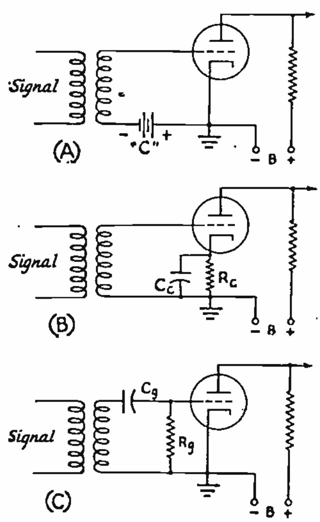


Fig. 325 — The three basic methods of obtaining grid bias. A, fixed bias; B, cathode bias; C, grid-leak bias.

A third biasing method is by use of a *grid leak*, R_g in Fig. 325-C. This requires that the exciting voltage be positive with respect to the cathode during part of the cycle, so that grid current will flow. The flow of grid current through the grid leak causes a voltage drop across the resistor, which gives the grid a negative bias. The time constant (§ 2-6) of the grid leak and grid condenser should be large in comparison to the time of one cycle of the exciting voltage, so that the grid bias will be substantially constant and will not follow the variations in a.c. grid voltage. For grid-leak bias,

$$R_g = \frac{E \times 1000}{I_g}$$

where R_g is the grid-leak resistance in ohms, E the desired bias voltage and I_g the d.c. grid current in ma.

For two tubes operated in push-pull or parallel with a common cathode- or grid-leak resistor, the required resistance becomes one-half that for a single tube. In push-pull Class-A circuits operating at audio frequencies, it is unnecessary to by-pass the cathode resistor. In this case the a.c. component of cathode current in one tube is out of phase with the a.c. component in the other, so that the two cancel each other.

The choice of a biasing method depends upon the type of operation. Fixed bias usually is required where the d.c. plate current of the amplifier varies in operation, as in Class-B audio-frequency amplifiers; if cathode bias is used the bias voltage would vary with the plate current. Since the plate current of a Class-A amplifier is constant with or without signal, such amplifiers almost invariably have cathode bias. Grid-leak bias cannot be used with amplifiers operated so that the grid is always negative with respect to the cathode, since in such a case there is no grid current and hence no voltage drop in the grid leak. Grid-leak bias is chiefly used for r.f. power amplifiers and for certain types of detectors. In power amplifiers, a combination of two or even all three types of bias may be used on one tube.

Cathode circuits; filament center tap — When a filament-type cathode is heated by a.c., hum can be minimized by making the two ends of the filament have equal and opposite potentials with respect to a center point, usually grounded (§ 2-13), to which the grid and plate return circuits are connected. The filament transformer winding may be *center-tapped* for this purpose, as shown in Fig. 326-A. With an untapped winding, a center-tapped resistor of 10 to 50 ohms is used, as at B. The by-pass condensers, C_1 and C_2, are used in r.f. circuits to avoid having the r.f. current flow through the transformer or resistor.

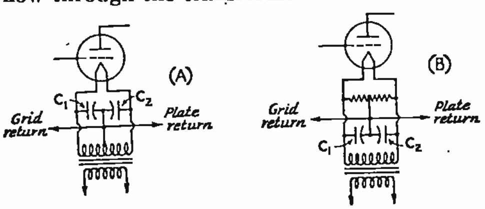


Fig. 326 — Filament transformer center-tap connections.

The heater supply for tubes with indirectly heated cathodes sometimes is center-tapped for the same purpose; more frequently, however, one side of the heater is grounded.

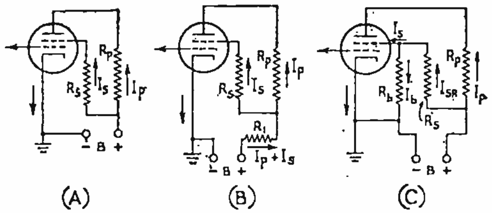


Fig. 327 — Calculation of plate and screen voltages.

Plate and screen voltage — Since the d.c. plate current flows through any resistance placed in the plate circuit of a tube as a load or coupling medium (§ 3-3), the actual voltage at the plate is less than the supply voltage by the voltage drop across the total resistance.

With transformer coupling this effect is not ordinarily of great importance, because the inductance of the transformer primary provides a high-impedance load at audio frequencies, while the d.c. resistance of the winding causes only a small drop in d.c. plate voltage.

In a resistance-coupled or parallel-fed stage the operating voltage is less than the supply voltage by the drop through the load resistor, R_p. Thus, in Fig. 327-A, $E_p = E_b - (I_p \times R_p)$.

Screen voltage is determined in the same way, using the screen current, I_s, to calculate the drop across the screen dropping resistor, R_s.

In Fig. 327-B both plate and screen current flows through a common filter resistor, so that both currents must be added in calculating the voltage drop across R_f. Thus

$$E_p = E_b - (I_p + I_s)(R_1) - I_pR_p$$
$$E_s = E_b - (I_p + I_s)(R_1) - I_sR_s.$$

In Fig. 327-C, the screen voltage, E_s, is obtained from a tap on a voltage divider consisting of R_s and R_b. The screen voltage, E_s, is equal to the voltage drop across R_b. First assigning a value of bleeder current, I_b (§ 8-4), this value is added to I_s to obtain I_{sr}. Then $R_s = E_s/I_{sr}$. The voltage across R_b is the difference between the screen voltage and the supply voltage, or $R_b = E_b - E_s/I_{sr}$. R_p is determined as above.

The resistance-capacity filter (§ 2-11) in Fig. 328, C_fR_f, is a *decoupling circuit* which isolates the stage from the power supply, to eliminate unwanted coupling between stages through the common impedance of the power supply.

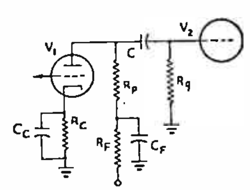


Fig. 328 — Decoupling in a resistance-coupled amplifier.

Frequency response — The frequency response of a resistance-coupled amplifier is determined by the coupling condenser, the following-stage grid resistor, the cathode bias resistor and its by-pass condenser. Amplification at low frequencies depends on the time constant (§ 2-6) of these CR combinations.

The value of the grid resistor, R_g, should be at least four times that of the plate load resistor, R_p, but should not exceed the maximum value for safe operation of the tube. The higher values give greatest low-frequency response with a given size of coupling condenser, C; 0.1 μfd. will provide ample coupling at low frequencies for most tubes with the usual values of load resistance and next-stage grid resistance.

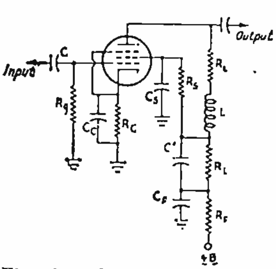


Fig. 329 — Wide-band pentode amplifier with both low- and high-frequency compensation.

The reactance (§ 2-8) of C_c must be small compared to the resistance of R_c. With values of R_l in the vicinity of 10,000 ohms, a condenser of 1 μfd. will suffice, more common practice is to use 5- or 10-μfd. low-voltage electrolytic condensers.

For maximum voltage gain, R_p should have as high resistance as possible without causing too great a voltage drop. Values range from 50,000 ohms to 0.5 megohm, the smaller figure being used with triodes having comparatively low plate resistance. The value of R_c depends upon R_p, which principally determines the plate current. Values for the screen resistor, R_s, may vary from 0.25 to 2 megohms. A screen by-pass condenser (C_s) of 0.1 μfd. will be adequate in most cases.

Table I in Chapter Fourteen shows suitable values for the more popular types of amplifier tubes. The calculated stage gain and peak undistorted output voltage also are given.

Low-frequency compensation — While the amplitude response of a resistance-coupled amplifier usually is satisfactory at low frequencies, the phase angle introduced by the coupling condenser, C_c, and the next-stage grid resistor, R_g, is sufficient to prevent proper reproduction of low-frequency square waves unless very large values are employed. Yet such large values increase the shunt capacity to ground, introduce grid-current difficulties in the following stage, and may even induce relaxation oscillations (motorboating).

The effect of the time constant of C_cR_g may be compensated for by proper design of the decoupling filter, R_FC_F. The design equation is

$$\frac{C_FR_LR_F}{R_L+R_F} = C_cR_g$$

Values of R_FC_F from 0.15 to 0.5 should be used for 0° phase angle above 30 cycles.

High-frequency compensation — The most widely used method for extending high-frequency response is the shunt-peaking circuit, with a resonating (peaking) inductance, L_l, in shunt across the circuit capacity, C_t. The design values of L and R_l are based on the shunt capacity, C_t, and the maximum required frequency, f_{max}. C_t can be estimated by adding 3 to 5 $\mu\mu$fd. (for socket and wiring) to the tube input and output capacities.

The impedance of L is made one-half the impedance of C_t at f_{max}. This is equivalent to making the resonant frequency between L and C_t equal to 1.41 times f_{max}.

Simplified design equations for shunt peaking compensation are as follows:

$$R_L = \frac{1}{2\pi f_{max}C_t}$$

$$L = 0.5C_tR_L^2$$

Typical values of R_L are from 2000 to 10,000 ohms; of L, from 25 to 100 μh.

Cathode follower — The *cathode-coupler* or *cathode follower*, also known as the *grounded-plate* or *inverted amplifier*, differs from a conventional amplifier in that output is taken from the cathode circuit rather than from the plate. Since the cathode impedance is always low compared with that of the plate, the circuit is applicable wherever matching to a low value of load impedance (fifty to several hundred ohms) is required. Because the cathode follower is inherently degenerative, it is particularly useful wherever equalized frequency response and minimum phase shift are important. Power amplification comparable to that of an equivalent plate-coupled stage may be secured, but the voltage gain is always less than unity.

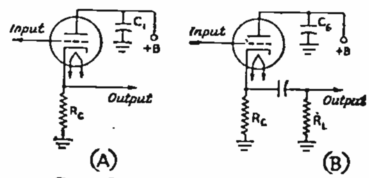


Fig. 330 — Cathode follower or inverted amplifier circuits. A, direct-coupled output; C, resistance-capacity coupling to load. R_c is the usual cathode-bias resistor.

Grounded-grid amplifier — In the circuit of Fig. 311 the grid of a triode amplifier is grounded. Used primarily at very high frequencies above the effective limits of amplification for pentode tubes, in this circuit the grid serves both as a control electrode and as a screen between cathode and plate, greatly reducing the internal feed-back capacity. To enhance the screening effect, tubes designed especially for grounded-grid u.h.f. amplifier service have additional internal shielding associated with the grid. Multiple grid leads connected in parallel minimize lead reactance between tube elements and ground. R.f. chokes in the heater leads maintain the heater at cathode potential.

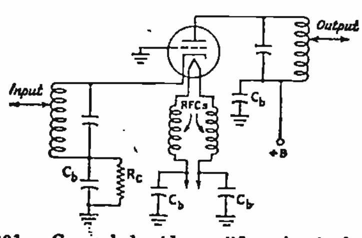


Fig. 331 — Grounded-grid amplifier circuit for v.h.f.

3-7-A Oscillators

Self-oscillation — That an amplifier tube can be made to generate a sustained radio-frequency current (§ 3-5) by virtue of self-oscillation is possible only because of the amplifying action of the tube. In an amplifier stage having positive feed-back, the total amplification is larger than it would be without feed-back (§ 3-3). In general, the greater the feed-back the greater the total amplification. The process of increasing feed-back to obtain greater amplification cannot be carried on indefinitely, however. Because of the amplifying properties of the tube, more energy is developed in the plate circuit than is required in the grid circuit. If enough energy is fed back to the grid, the feed-back process becomes independent of any applied signal voltage. The tube supplies its own grid excitation and continuous oscillations are generated. The actual energy required to overcome the grid losses is, in the end, taken from the d.c. plate supply.

The process of oscillation may also be considered from the standpoint of *negative resistance*. As previously described (§ 3-3), positive feed-back is equivalent to shunting a negative resistance across the input circuit of the tube. When the value of negative resistance becomes lower than the positive resistance of the circuit (if the circuit is parallel resonant the positive resistance will be the resonant impedance of the circuit) the net resistance is negative, indicating that the circuit can be looked upon as a *source* of energy. Such a source is capable of maintaining a constant voltage which can be amplified by the tube. The actual energy, of course, comes from the plate circuit of the tube, so that the two viewpoints are equivalent.

A circuit having the property of generating continuous oscillations is called an *oscillator*. It is not necessary to apply external excitation to such a circuit, since any random variation in current will be amplified to cause oscillation. The frequency of oscillation will be that at which the feed-back voltage has the proper phase and amplitude. Where resonant circuits are associated with oscillators, the oscillation frequency is very nearly that of the tuned circuit.

Excitation and bias — The excitation voltage required depends upon the characteristics of the tube and the losses in the circuit, including the power consumed in the load. In practically all oscillators the grid is driven positive during part of the cycle, so that power is consumed in the grid circuit (§ 3-2). This power must be supplied from the plate circuit. With insufficient excitation, the tube will not oscillate; with over-excitation, the *grid losses* (power consumed in the grid circuit) will be excessive.

Oscillators customarily are grid-leak biased (§ 3-6). This takes advantage of the grid-current flow and gives better operation, the bias adjusting itself to the excitation voltage.

Tank circuit — The resonant circuit associated with the oscillator is commonly called the *tank circuit*, a name derived from the storage of energy associated with a resonant circuit (§ 2-10). The term is applied to any resonant circuit in transmitting applications, whether in an oscillator or in an amplifier.

Plate efficiency — The *plate efficiency* (§ 3-3) of an oscillator depends upon the load resistance, excitation and other operating factors. Usually it is around 50 per cent. It is not as high as in an amplifier, since the oscillator must supply its own grid losses. These may represent 10 to 20 per cent of the output power.

Power output — The *power output* of an oscillator is the useful a.c. power consumed in any load connected to the oscillator. The load may be coupled as described in § 2-11.

Frequency stability — The frequency stability of an oscillator is its ability to maintain constant frequency. The more important factors which may cause a change in frequency are (1) temperature, (2) plate voltage, (3) loading, (4) mechanical variations of circuit elements. Temperature changes will cause vacuum-tube elements to expand or contract slightly, thus causing variations in the interelectrode capacities (§ 3-2). Since these are unavoidably part of the tuned circuit, the frequency will change correspondingly. Temperature changes in the coil or condenser will alter their inductance or capacity slightly, again causing a shift in the resonant frequency. These effects are relatively slow in operation, and the frequency change caused by them is called *drift*.

Load variations act in much the same way as plate voltage variations. A temperature change in the load may also result in drift.

Plate-voltage variations will cause a corresponding instantaneous shift in frequency; this type of frequency shift is called *dynamic* instability. Dynamic instability can be reduced by using a tuned circuit of high effective Q. Since the tube and load represent a relatively low resistance in parallel with the circuit, this means that a low L/C ratio ("high-C") must be used (§ 2-10) and that the circuit should be lightly loaded. Dynamic stability also can be improved by using a high value of grid leak, which gives high grid bias and raises the effective resistance of the tube as seen by the tank circuit, and by using relatively high plate voltage and low plate current. Drift can be minimized by keeping the d.c. input low for the size of tube, by using coils of large wire to prevent undue temperature rise, and by providing good ventilation to carry off heat rapidly. A low L/C ratio in the tank circuit is desirable, because the interelectrode capacity variations have proportionately less effect on the frequency when shunted by a large condenser.

Mechanical variations, usually caused by vibration, cause changes in inductance and/or capacity which in turn cause the frequency to "wobble" in step with the vibration.

Mechanical instability can be minimized by using well-designed components and insulating the oscillator from mechanical vibration.

3-7-B Feed-Back Oscillators

Magnetic feed-back — One form of feed-back is by electromagnetic coupling between plate (output) and grid (input) circuits. Two representative circuits of this type are shown in Fig. 332. That at A is called the *tickler* circuit. The amplified current flowing in the "tickler," L_2, induces a voltage in L_1 in the proper phase when both coils are wound in the same direction and connected as shown in the diagram. The feed-back can be adjusted by adjusting the coupling between L_1 and L_2.

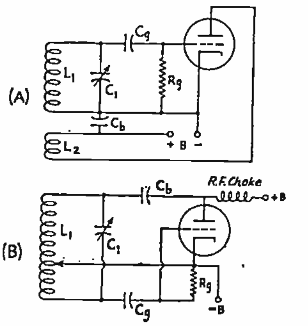


Fig. 332 — Two types of oscillator circuits with magnetic feed-back. A, grid tickler; B, Hartley.

The *Hartley* circuit, B, is similar in principle. There is only one coil, but it is divided so that part of it is in the plate circuit and part in the grid circuit. The magnetic coupling between the two sections provides the feed-back, which can be adjusted by moving the tap on the coil.

Capacity feed-back — The feed-back can also be obtained through capacity coupling, as shown in Fig. 333. In A, the *Colpitts* circuit, the voltage across the resonant circuit is divided, by means of the series condensers, into two parts. The instantaneous voltages at the ends of the circuit are opposite in polarity with respect to the cathode, hence in the right phase to sustain oscillation. The tuned-grid tuned-plate circuit at B utilizes the grid-plate capacity of the tube to provide feed-back coupling. There should be no magnetic coupling between the two tuned-circuit coils. Feed-back can be adjusted by varying the tuning of either the grid or plate circuit. The circuit with the higher Q (§ 2-10) determines the frequency of oscillation. The plate circuit must be tuned to a slightly higher frequency than the grid circuit, so that it will have inductive reactance and hence give positive feed-back (§ 3-3). The amount of detuning is so small it is customary to assume that the circuits are tuned to the same frequency.

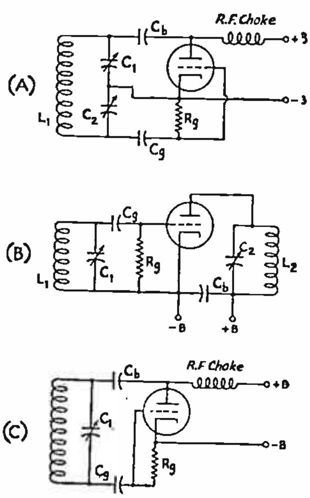


Fig. 333 — Capacity feed-back oscillators. A, Colpitts; B, tuned-plate tuned-grid; C, ultraudion.

The *ultraudion* circuit at C is equivalent to the Colpitts, with the voltage division for oscillation brought about through the grid-to-filament and plate-to-filament capacities of the tube. In this and in the Colpitts circuit, the feed-back can be controlled by varying the ratio of the two capacities. In the ultraudion circuit, this can be done by connecting a small variable condenser between grid and cathode.

The electron-coupled oscillator — The effects of loading and coupling to the next stage can be greatly reduced by use of the *electron-coupled* circuit, in which a screen-grid tube (§ 3-5) is so connected that its screen grid is used as a plate, in conjunction with the control grid and cathode, in an ordinary triode oscillator circuit. The screen is operated at ground r.f. potential (§ 2-13) to act as a shield between the actual plate and the cathode and control grid; the latter two elements therefore must be above ground potential. The output is taken from the plate circuit. Under these conditions the capacity coupling (§ 2-11) between the plate and other ungrounded tube elements is quite small, hence the output power is secured almost entirely by variations in the plate current caused by the varying potentials on the grid and cathode. Since in a screen-grid tube the plate voltage has a relatively small effect on the plate current, the reaction on the oscillator frequency for different conditions of loading is small.

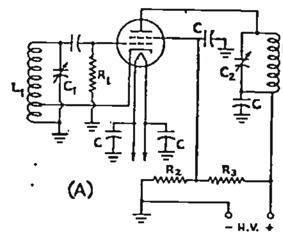


Fig. 334 — Electron-coupled oscillator circuit.

A Hartley (§ 3-7) circuit is used in the frequency-determining portion of the oscillator shown in Fig. 334, where $L_1 C_1$ is the oscillator tank circuit. The screen is grounded for r.f. through a by-pass condenser (§ 2-13), but has the usual d.c. potential. The cathode connection is made to a tap on the tank coil to provide feed-back. The resonant plate circuit, L_2C_2, is tuned either to the oscillation frequency or to a harmonic. Untuned output coupling also may be used; the output voltage and power are considerably lower, but better isolation between oscillator and amplifier is secured.

If the oscillator tube is a pentode having an external suppressor connection the suppressor grid should be grounded. This provides additional internal shielding and further isolates the plate from the frequency-determining circuit.

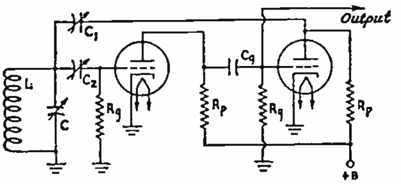


Fig. 335 — Franklin master-oscillator circuit.
C_1, C_2 — Approximately 1 to 2-μμfd. (adjustable).
C_g — 0.001-μfd. R_g, R_p — 50,000 ohms.

Franklin oscillator — The Franklin oscillator circuit of Fig. 335, popular abroad, has characteristics similar to the e.c.o. A high-gain feed-back amplifier is very loosely coupled to a tank circuit, *LC*, via two condensers, C_1 and C_2, of extremely small capacity. So weak is the coupling that the tube circuit has negligible effect upon the frequency-controlling tank.

Crystal oscillators — Since a properly cut quartz crystal is equivalent to a high-*Q* tuned circuit (§ 2-10), it may be substituted for a conventional tuned circuit in an oscillator to control the frequency of oscillation. A simple crystal oscillator circuit is shown in Fig. 336. It is similar to the tuned-plate tuned-grid circuit except that a crystal is substituted for the resonant grid circuit. Detailed information on crystal oscillators is given in Chapter Four.

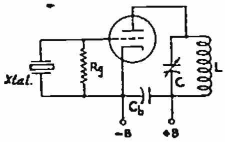


Fig. 336 — Simple crystal oscillator circuit. Many variations of this basic circuit are used in practice.

Series and parallel feed — A circuit such as the grid-tickler circuit of Fig. 332-A is said to be *series fed* because the source of plate voltage and the r.f. plate circuit (the tickler coil) are connected in series; hence the d.c. plate current flows through the coil to the plate. A by-pass (§ 2-13) condenser, C_b, is connected across the plate supply to shunt the r.f. current around the power source. Other examples of series plate feed are shown in Figs. 333-B and 334.

In some cases the source of plate power must be connected in parallel with the tuned circuit to provide a direct-current path to the plate. This is illustrated in Fig. 332-B, where it would be impossible to feed the plate current through the coil because there is a direct connection between the coil and cathode. Hence the voltage is applied to the plate through a radio-frequency choke, which prevents the r.f. current from flowing to the plate supply and thus short-circuiting the oscillator. The blocking condenser, C_b, provides a low-impedance path for radio-frequency current flow but is an open circuit for direct current (§ 2-13). Other examples of parallel feed are shown in Figs. 333-A and 333-C.

Values for the r.f. chokes, by-pass and blocking condensers shown will be determined by the considerations outlined in § 2-13.

3-7-C Negative Resistance Oscillators.

Negative-resistance oscillations — If a resonant circuit were completely free from losses, a current once started would continue indefinitely; that is, sustained oscillations would occur. As previously explained, this condition can be simulated in practice by canceling the actual resistance in the circuit by inserting an equal or greater amount of *negative resistance.* Negative resistance is exhibited by any device showing an increase of current when the applied voltage is decreased, or vice versa. Furthermore, any negative resistance device capable of neutralizing the resistance of a resonant circuit will maintain sustained oscillations in that circuit.

In addition to its ability to create negative resistance by feed-back in a tuned circuit at its resonant frequency, a vacuum tube can in itself be made to show negative resistance by a number of arrangements of electrode potentials.

When a tube so arranged is connected across a two-terminal parallel-resonant circuit which will neutralize its conductance, oscillation will be established. Typical two-terminal oscillator circuits are shown in Fig. 337.

The circuit of Fig. 337-A is that of the *dynatron* oscillator, which functions because of the secondary emission from the plate occurring in certain types of screen-grid tetrodes. The simplest but also the least stable of the negative-resistance or two-terminal oscillators, it makes use of the fact that the plate-current curve of a screen-grid tetrode has a downward slope at certain values of screen voltage, giving a negative plate-resistance characteristic which, by neutralizing the conductance of the resonant circuit, will serve to sustain oscillations when the tuned circuit has a high *L/C* ratio.

In the retarding-field negative-transconductance or *transitron* circuit shown in Fig. 337-B, negative resistance is produced by virtue of the fact that, if the negatively biased suppressor grid of a pentode is given still more negative bias, electrons which normally would pass through to the plate

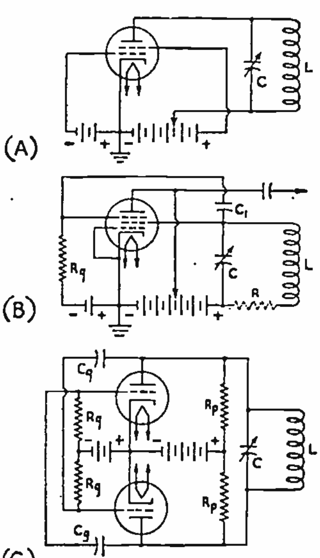


Fig. 337 — Negative-resistance oscillator circuits. A, dynatron; B, transitron; C, push-pull circuit.

are turned back to the screen, thus increasing the screen current and reversing normal tube action (§ 3-2). The negative resistance produced between the screen and suppressor grids is sufficiently low so that ordinary tuned circuits will oscillate readily at frequencies up to 15 Mc. or so.

The operation of the push-pull negative-resistance oscillator at C can best be understood by visualizing it as a push-pull resistance-coupled amplifier in which the output of each tube is coupled to the input of the other, resulting in an equivalent negative resistance between the two plates. The amount of feed-back, and therefore the waveform of the output, can be controlled by the degree of coupling — i.e., by making the reactance of the coupling condensers, C_1, small in comparison to the grid resistors, R_1, and by regulating the grid bias. Low harmonic content and good frequency stability characterize this circuit.

Resistance-capacity tuning — The oscillators described above differ from the feed-back type primarily in that the negative resistance required for oscillation is supplied internally within the vacuum-tube circuit rather than by an associated feed-back circuit. The frequency-controlling element, however, remains a resonant circuit of the inductance-capacity type, for which $f = 1/2\pi\sqrt{LC}$.

It is possible to replace the LC circuit with a resistance-capacity combination having an appropriate time constant, in which case $f = 1/2\pi LC$. Moreover, by varying either R or C the circuit can be tuned over a wide range in the same manner as an LC circuit.

The two more common circuits of this type are shown in Fig. 338. The single-stage RC-tuned oscillator at A has a three-mesh phase-shifting network connected between output and input, so arranged that just enough signal is fed back 180° out of phase at the desired output frequency to sustain oscillation. By careful feed-back adjustment, excellent sine-wave form with exceptionally good frequency stability may be obtained.

The two-tube RC-tuned circuit at B is derived from a two-stage cascade resistance-coupled amplifier with pentode tubes, the second tube constituting the phase-shifting element supplying a regenerative signal to the adjustable C, C' and R_1 combination at the desired frequency, while at all other frequencies the circuit is degenerative.

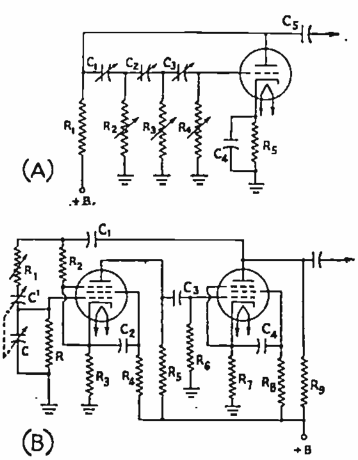


Fig. 338 — Resistance-capacity oscillators. A, phase-shift. B, negative feed-back.

Relaxation oscillators — There is another basic category of oscillators, the *relaxation* type, in which the oscillation frequency is controlled not by a resonant circuit but by the reciprocating change of a current or voltage through the charging or discharging of a condenser when a certain critical value is reached. Relaxation oscillation requires, first, a means for charging a condenser (or other reactive element) at a uniform rate and, second, means for rapidly discharging this condenser once a predetermined voltage has been built up across it. The action is characterized by a period of rapid change or instability followed by a period of relative quiescence or stability during which the stored-up energy is dissipated or transferred.

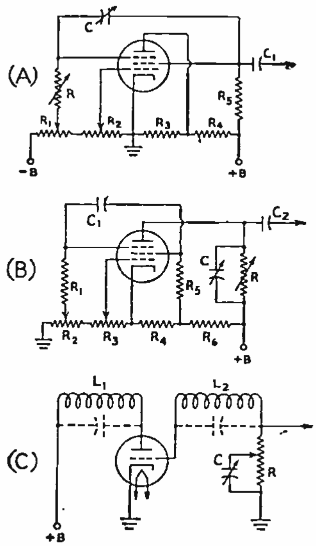


Fig. 339 — Typical relaxation oscillators. A, "dynatron"-type pentode circuit. B, high-frequency pentode circuit. C, squegging oscillator.

Relaxation oscillators have high harmonic content (nonsinusoidal output) and are inherently unstable, permitting ready synchronization with an external controlling voltage.

In the circuit of Fig. 339-A, the operation is based on the reversed screen-current or dynatron characteristic of a pentode tube, the frequency being determined by the rate at which the feed-back condenser, C, discharges through the tube. Apart from the frequency-controlling mechanism, this circuit resembles that of the transitron oscillator (Fig. 337-B).

The alternative pentode circuit at B has the frequency-controlling elements, C and R, in the plate circuit. It is capable of operation at frequencies up to several hundred kilocycles, and affords greater control of wave form.

Operation of the squegging oscillator at C is based on the tendency of any oscillator with high-Q tuned circuits to produce relatively low-frequency intermittent oscillations, controlled by the rate of charge and discharge of L, C and R through the tube plate resistance, if the time constant of the grid leak-condenser, RL, is large compared with the period of oscillation.

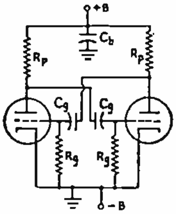


Fig. 340 — The multivibrator, or relaxation oscillator.

The most versatile relaxation oscillator circuit of all, shown in Fig. 340, is known as the *multivibrator*. Two tubes are used with resistance coupling, the output of one tube being fed to the input circuit of the other as in the push-pull negative-resistance oscillator of Fig. 337-C. The frequency of the resulting oscillation is determined by the time constants (§ 2-6) of the resistance-capacity combinations. The principle of oscillation is that of alternately switching conduction from one tube to the other, with one grid at cut-off and the other at zero bias, so that continuous oscillation is maintained, the second tube being necessary to obtain the proper phase relationship (§ 3-3) for oscillation when the energy is fed back.

The multivibrator is a very unstable oscillator, and for this reason its frequency can be readily controlled by a small signal of steady frequency introduced into the circuit. This phenomenon is called *locking* or *synchronization*. The output waveshape of the multivibrator is highly distorted, hence has high harmonic content (§ 2-7). A useful feature is that the multivibrator can be locked at its fundamental frequency by a frequency corresponding to one of its higher harmonics (the tenth harmonic is frequently used), and thus can be used as a *frequency divider*.

ℂ 3-8 Cathode-Ray Tubes

Principles — The cathode-ray tube is a vacuum tube in which the electrons emitted from a hot cathode are first accelerated to give them considerable velocity, then formed into a beam, and finally allowed to strike a special translucent screen which *fluoresces*, or gives off light at the point where the beam strikes. A narrow beam of moving electrons is analogous to a wire carrying current (§ 2-4) and, as in the wire, is accompanied by electrostatic and electromagnetic fields. Hence the beam can be moved laterally, or deflected, by electric or magnetic fields. Such fields exert a force on the beam in the same way as on charged bodies or on wires carrying current (§ 2-3, 2-5).

Since the cathode-ray beam consists only of moving electrons, its weight and inertia are negligibly small. For this reason, it can be made to follow instantly the variations in periodically changing fields even at radio frequencies.

Electron gun — The electrode arrangement which forms the electrons into a beam is called the *electron gun*. In the simple tube structure shown in Fig. 341, the gun consists of the cathode, grid, and anodes Nos. 1 and 2. The intensity of the electron beam is regulated by the grid in the same way as in an ordinary tube (§ 3-2). Anode No. 1 is operated at a positive potential with respect to the cathode, thus accelerating the electrons which pass through the grid, and is provided with small apertures through which the electron stream passes. On emerging from the apertures the electrons are traveling in practically parallel straight-line paths. The electrostatic fields set up by the potentials on anode No. 1 and anode No. 2 form an *electron lens* system, comparable to an optical lens, which makes the electron paths converge to a point at the fluorescent screen in much the same way that a glass lens takes parallel rays of light and brings them to a point focus. Focusing of the electron beam is accomplished by varying the potentials on the anodes, the potential in turn determining the strength of the field. The potential on anode No. 2 is usually fixed, while that on anode No. 1 is varied to bring the beam into focus. Anode No. 1 is, therefore, called the *focusing electrode*.

Sharpest focus is obtained when the electrons of the beam have high velocity, so that relatively high d.c. potentials are common with cathode-ray tubes. A second grid may be placed between the control grid and anode No. 1, for additional acceleration of the electrons.

Methods of deflection — When focused, the beam from the gun produces only a small spot on the screen, as described above. However, if after leaving the gun the beam is deflected by either magnetic or electrostatic fields, the spot will move across the screen in accordance with the force exerted on the beam. If the motion is sufficiently rapid, retentivity of vision makes the path of the moving spot (*trace*) appear as a continuous line.

Electrostatic deflection, the type generally used in the smaller tubes, is produced by *de-*

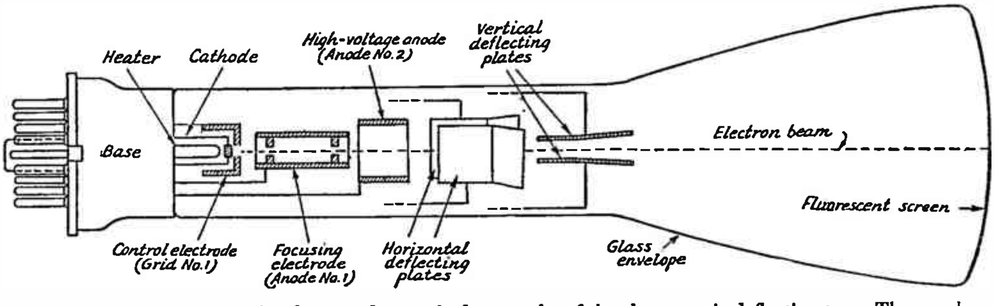


Fig. 341 — Typical construction for a modern cathode-ray tube of the electrostatic-deflection type. The envelope is made of glass, with the fluorescent screen at one end. Leads for the high-voltage anode, the deflection plates, and other electrodes are insulated low-capacity conductors carried inside the envelope to the base.

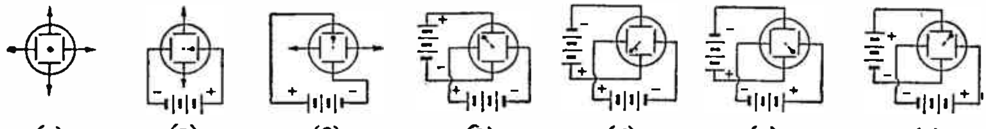


Fig. 342 — Spot diagrams showing the position of the cathode-ray beam on the fluorescent screen for different deflector potentials. A — Both deflectors at zero potential. B — Positive potential on right horizontal deflector. C — Positive potential on upper vertical deflector. D, E, F, G — Equal positive potentials on adjacent plates.

flecting plates. Two sets of plates are placed at right angles to each other, as indicated in Fig. 341. The fields are created by applying suitable voltages between the two plates of each pair. Usually one plate of each pair is connected to anode No. 2, to establish the polarities (§ 2-3) of the vertical and horizontal fields with respect to the beam and to each other.

Tubes for magnetic deflection use the same type of electron gun, but have no deflection plates. Instead, the deflecting fields are set up by means of coils corresponding to the plates used in tubes having electrostatic deflection. The coils are external to the tube, as shown in Fig. 333, but are mounted close to the glass envelope in the relative positions occupied by electrostatic deflection plates. Coils A_1 and A_2 are connected so their fields aid and their axes are on the same line through the tube. Coils B_1 and B_2 likewise are connected with fields aiding and are aligned along the same axis through the tube, but perpendicularly to the A_1A_2 axis.

The beam deflection caused by a given change in the field intensity is called the *deflection sensitivity*. With electrostatic-deflection tubes it is usually expressed in millimeters per volt, which gives the linear movement of the spot on the screen as a function of the voltage applied to a set of deflecting plates. Values range from about 0.1 to 0.6 mm/volt, depending upon the tube construction and gun electrode voltages. The sensitivity is decreased by an increase in anode No. 2 voltage because a higher voltage gives the electrons in the beam higher velocity, and hence they are less easily deflected by a field of given strength.

Fluorescent screens — The fluorescent screen materials used have varying characteristics, according to the type of work for which the tube is intended. The spot color is green, white, yellow or blue, depending upon the screen material. The *persistence* of the screen is the time duration of the after-glow which exists when the excitation of the electron beam is removed. Screens are classified as long-, medium- and short-persistence types. Small tubes for oscilloscope use usually have medium-persistence screens of greenish fluorescence.

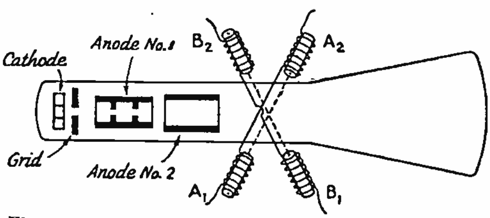


Fig. 343 — A cathode-ray tube with magnetic deflection. The gun is the same as in the electrostatic-deflection tube shown in Fig. 341, but the beam is deflected by magnetic instead of electric fields. Actual deflection coils fit closely to the neck of the tube, so that the field will be as strong as possible for a given coil current.

Tube circuits — A representative cathode-ray tube circuit with electrostatic deflection is shown in Fig. 344. One plate of each pair of deflecting plates is connected to anode No. 2. Since the voltages required normally are rather high, the positive terminal of the supply is usually grounded (§ 2-13) so that the common deflection plates will be at ground potential. This places the cathode and other elements at high potentials above ground, hence these elements must be well insulated. The various electrode voltages are obtained from a voltage divider (§ 2-6) across the high-voltage d.c. supply. R_3 is a variable divider or "potentiometer" for adjusting the negative bias on the control grid and thereby varying the beam current; it is called the *intensity* or *brightness* control. The *focus*, or sharpness of the luminous spot formed on the screen by the beam, is controlled by R_4, which changes the ratio of the anode No. 2 and anode No. 1 voltages. The focusing and intensity controls interlock to some extent, and the sharpest focus is obtained by keeping the beam current low.

Deflecting voltages for the plates are applied to the terminals marked "input voltage," R_1 and R_2 having high values of resistance (1 megohm or more) to drain off any accumulation of charge on the deflecting plates. Usually some provision is made to place an adjustable d.c. voltage on each set of plates, so that the spot can be "centered" when stray electrostatic or magnetic fields are present; the adjustable d.c. voltage neutralizes the effect of such fields.

The tube is mounted so that one set of plates produces a horizontal line when a varying voltage is applied to it, while the other set of plates produces a vertical line under similar conditions. They are called, respectively, the "horizontal" and "vertical" plates, but which set of actual plates produces which line is simply a matter of how the tube is mounted. It is usually necessary to provide a mounting which can be rotated to some extent, so that the lines will actually be horizontal and vertical.

Power supply — The d.c. voltage required for operation of the tube may vary from 500 volts for the miniature type (1-inch diameter screen) to several thousand volts for the larger tubes. The current, however, is very small, so that the power required likewise is small. Because of the low current drain, a power supply with half-wave rectification (§ 8-3) and a single 0.5- to 2-μfd. filter condenser is satisfactory.

3-9 The Oscilloscope

Description — An oscilloscope is essentially a cathode-ray tube in the basic circuit of Fig. 344, but with provision for supplying a suitable deflection voltage on one set of plates (ordinarily those giving horizontal deflection). The deflection voltage is the *time base* or *sweep*. Oscilloscopes frequently are also equipped with vacuum-tube amplifiers for increasing the amplitude of small a.c. voltages to values suitable for application to the deflecting plates. These amplifiers ordinarily are limited to operation in the audio- or video-frequency range.

Formation of patterns — When periodically varying voltages are applied to the two sets of deflecting plates, the path traced by the fluorescent spot forms a *pattern* which is stationary so long as the amplitude and phase relationships of the voltages remain unchanged. Fig. 345 shows how such patterns are formed. The horizontal sweep voltage is assumed to have the "sawtooth" waveshape indicated; with no voltage applied to the vertical plates the trace simply sweeps from left to right across the screen along the horizontal axis X–X' until the instant H is reached, when it reverses direction and returns to the starting point. The sine-wave voltage applied to the vertical plates similarly would trace a line along the axis Y–Y' in the absence of any deflecting voltage on the horizontal plates. However, when both voltages are present the position of the spot at any instant depends upon the voltages on both sets of plates at that instant. Thus at time B the horizontal voltage has moved the spot a short distance to the right and the vertical voltage has similarly moved it upward, so that it reaches the actual position B' on the screen. The resulting trace is easily followed from the other indicated positions, which are taken at equal time intervals.

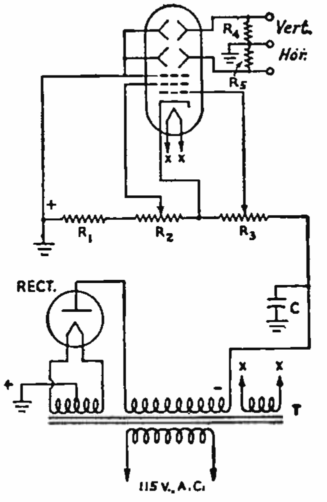


Fig. 344 — Cathode-ray tube circuit. Typical values for a 3-inch (screen-diameter) tube such as the 3AP1/906: R_1, R_2 — 1 to 10 megohms. R_4 — 0.2 megohm. R_3 — 20,000 ohms. R_5 — 0.5 megohm.

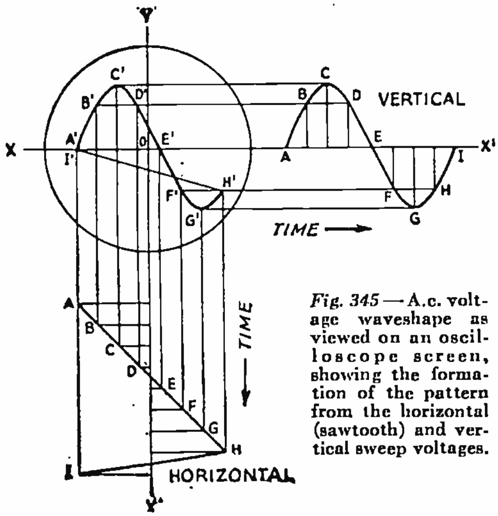


Fig. 345 — A.c. voltage waveshape as viewed on an oscilloscope screen, showing the formation of the pattern from the horizontal (sawtooth) and vertical sweep voltages.

Types of sweeps — A sawtooth sweep-voltage waveshape, such as is shown in Figs. 345 and 347 is called a *linear sweep*, because the deflection in the horizontal direction is directly proportional to time. If the sweep were perfect the "fly-back" time, or time taken for the spot to return from the end (H) to the beginning (I or A) of the horizontal trace, would be zero, so that the line HI would be perpendicular to the axis Y–Y'. Although the fly-back time cannot be made zero in practicable sweep-voltage generators it can be made quite small in comparison to the time of the desired trace AH, at least at most frequencies within the audio range. The fly-back time is somewhat exaggerated in Fig. 345, to show its effect on the pattern. The line $H'I'$ is called the *return trace;* with a linear sweep it is less brilliant than the pattern, because the spot is moving much more rapidly during the fly-back time than during the time of the main trace. If the fly-back time is short enough, the return trace will be invisible.

The linear sweep has the advantage that it shows the shape of the wave applied to the vertical plates in the same way in which it is usually represented graphically (§ 2-7). If the time of one cycle of the a.c. voltage applied to the vertical plates is a fraction of the time taken to sweep horizontally across the screen, several cycles of the vertical or signal voltage will appear in the pattern. The shape of only the last cycle (or the last few cycles, depending upon the number in the pattern and the characteristics of the sweep) to appear will be affected by the fly-back in such a case.

Although the linear sweep generally is most useful, other sweep waveshapes may be desirable for certain purposes. The shape of the pattern obtained, with a given signal waveshape on the vertical plates, obviously will depend upon the shape of the horizontal sweep voltage. If the horizontal sweep is sinusoidal, the main and return sweeps each occupy the same time and the spot moves faster horizontally in the center of the pattern than it does at the ends.

If two sinusoidal voltages of the same frequency are applied simultaneously to both sets of plates, the resulting pattern may be a straight line, an ellipse or a circle, depending upon the amplitude and phase relationships. If the frequencies are harmonically related (§ 2-7) a stationary pattern will result, but if one frequency is not an exact harmonic of the other the pattern will show continuous motion. This is also the case when a linear sweep circuit is used; the sweep frequency and the frequency under observation must be harmonically related or the pattern will not be stationary.

The sweep generator does not ordinarily function as a self-controlled oscillator but rather as an externally controlled or synchronized oscillator which supplies voltage of the required waveform at the periodicity of the signal under study or a multiple thereof.

Sweep circuits — A sinusoidal sweep is easiest to obtain, since it is possible to apply a.c. voltage from the power line, either directly or through a suitable transformer, to the horizontal plates. A variable voltage divider or potentiometer may be used to regulate the width of the horizontal trace.

Practically all sweep circuits employ some version of the relaxation oscillator (§ 3-6). Because of its lack of inherent stability, this type of oscillator is readily controlled or "locked-in" by voltage from an external source even when that voltage is of a different frequency.

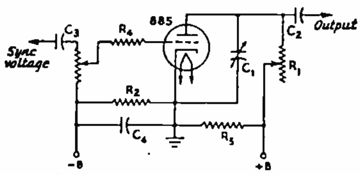


Fig. 346 — A linear-sweep oscillator using a gas triode.

C_1 — 0.001 to 0.25 μfd.
C_2 — 0.5 μfd.
C_3 — 0.1 μfd.
C_4 — 25 μfd. 25-volt electrolytic.
R_1 — 0.3 to 1.5 megohms.
R_2 — 2000 ohms.
R_3 — 0.25 megohm.
R_4 — 25,000 ohms.
R_5 — 0.1 megohm.

The "B" supply should deliver 300 volts. C_1 and R_1 are proportioned to give a suitable sweep frequency; the higher the time constant (§ 2-6), the lower the frequency. R_4 limits grid-current flow during the deionizing period, when positive ions are attracted to the negative grid.

A typical circuit for a linear sweep generator is shown in Fig. 346. The tube is a *gas triode* or grid-control rectifier (§ 3-5), also known as a *thyratron*. The *striking* or breakdown voltage, which is the plate voltage at which the tube ionizes or *fires* and starts conducting, is determined by the grid bias. When plate voltage, E_b, is applied, the condenser, C_1, acquires a charge through R_1. As shown in Fig. 347, the charging voltage rises relatively slowly, as shown by the solid line, until the breakdown or flashing point, V_f, is reached. Then the condenser discharges rapidly through the comparatively low plate-cathode resistance of the tube. When the voltage drops to a value too low to maintain plate-current flow, E_a, the ionization is extinguished and C_1 once more charges through

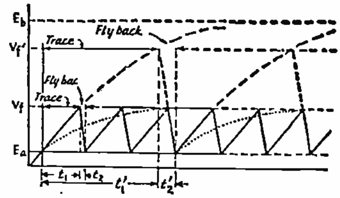


Fig. 347 — Condenser charging curves showing how a sawtooth wave is produced by a gaseous-tube linear sweep oscillator.

R_1. If R_1 is large enough, the voltage across C_1 rises linearly with time, t_1, up to the breakdown point. This linear voltage change is used for the sweep, being applied to the cathode-ray tube plates through C_2. The fly-back time, t_2, is the time required for discharge through the tube; to keep this time small, the resistance during discharge must be low.

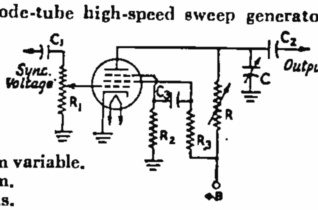
To obtain a stationary pattern, the "sawtooth" rate is controlled by varying C_1 and R_1 and synchronized by introducing some of the voltage being observed on the vertical plates into the grid circuit of the 884 tube. This voltage "triggers" the tube into operation in synchronism with the signal frequency. Synchronization will occur so long as the signal frequency is nearly the same as, or a multiple of, the sweep frequency, provided the circuit constants and the amplitude of the synchronizing voltage are properly adjusted.

The upper frequency limit of gaseous-tube sweep oscillators is in the vicinity of 50,000 cycles, even with the most careful design, because of the fly-back time limitations imposed by the gaseous content of the tube.

To attain a higher-frequency sweep, a "hard"-tube oscillator such as that shown in Fig. 348 must be used. This circuit may be recognized as being similar to that of the pentode relaxation oscillator of Fig. 338-B. With suitable constants it is capable of an upper frequency limit of 100 to 200 kc. or more. If a tube is used which has a high ratio of plate current to screen current, the screen voltage will rise to a very high value during the plate discharge and thus aid in reducing the fly-back time.

A variety of waveshapes may be obtained from this circuit, ranging from the sawtooth or triangular waves which occur at the plate to the rectangular waveform of the screen-grid voltage. The plate-circuit waveforms are those most often employed for oscilloscope work.

Fig. 348 — Pentode-tube high-speed sweep generator.



C — 0.001 to 0.1 μfd.
C_1, C_2 — 0.1 μfd.
C_3 — 1.0 μfd.
R — 25,000 ohms to 5 megohms.
R_1 — 0.5 megohm variable.
R_2 — 0.1 megohm.
R_3 — 25,000 ohms.

The sweep rate is controlled by R and C, but it is influenced also by the value of R_2. R_3 determines the output waveshape by regulating the ratio of charge to discharge time, thus determining the part of the cycle occupied by the rectangular-shaped screen-voltage wave.

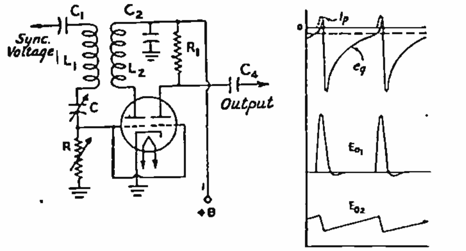


Fig. 349 — Dual-triode blocking-tube oscillator and discharge tube, with characteristic waveforms at the right.

C — 0.001–0.01-μfd. mica. R — 0.25 megohm variable.
C_1, C_3 — 0.005–0.5 μfd. R_1 — 0.1–2 megohm.
C_2 — 0.1 μfd. L_1, L_2 — See text.

The *blocking-tube oscillator* in Fig. 349 is also capable of high-frequency operation, chiefly because the oscillator portion generates a very short, sharp pulse which charges C almost instantaneously. Because of its superiority in this respect, this circuit has received considerable application in television work. Its operation is distinguished from that of the squegging oscillator (Fig. 338-C) in that the intermittent high-frequency oscillations are almost instantly blocked as the bias built up by the grid-leak and condenser, C and R, goes far beyond cut-off. With suitable constants, the build-up time for this blocking bias can be limited to a single high-frequency cycle, resulting in a very short, abrupt pulse of plate current (I_p). Because of the large time constant of C and R, the discharge time is very much slower. Until the charge again leaks off through R, the circuit is paralyzed. When C is discharged, the cycle repeats.

L_1 and L_2 are tightly coupled and designed to be self-resonant at perhaps ten times the maximum sweep frequency.

In the practical form, shown in Fig. 351, the blocking oscillator itself is the left-hand section of the dual triode. The second triode section is used as a discharge tube, the rate of discharge being controlled by the C_2R_1 combination. By giving this combination the proper time constant, the output wave can be made to have almost any desired form. R exercises limited control over the frequency range, while the value of R_1 determines the output amplitude.

Vacuum-tube switching circuits — In contrast to time-base circuits which deliver recurrent output impulses, certain applications in oscilloscope and other electronic work call for what are termed *vacuum-tube* or *electronic* switching circuits.

A *keying* circuit is a non-locking electronic switch which closes (or opens) a circuit when a control voltage is applied and returns the circuit to normal when the control voltage is removed. The keying voltage is usually applied as control-grid bias, although screen- and suppressor-grid voltage also are employed.

A *trigger circuit* may also be operated in this manner, but more strictly it is a type of locking or holding electronic switch, wherein a second impulse is required to restore the circuit. After the initiating control pulse the circuit remains closed, despite removal of the control voltage, until a second releasing impulse is received. Circuits in which values of current or voltage change abruptly from one stable condition to another at some critical value of voltage or resistance, and then change back abruptly at a different critical value of the controlling voltage or resistance, are used for this purpose.

Fig. 352-A shows the basic pentode form of trigger circuit. In this circuit d.c. coupling between the screen and suppressor grids causes the suppressor voltage to change with screen voltage. With a high value of resistance in series with the screen, abrupt changes in these currents occur when the supply voltages or the screen-circuit resistance are varied. For example, by proper choice of voltage and circuit constants the plate current corresponding to a given value of screen current may be made zero. Triggering impulses may be introduced in series with any of the electrodes, but the control grid is the most sensitive. The values of the supply voltages are not critical, but the proper relation must be maintained between them.

In the two-tube trigger circuit of Fig. 352-B, the flow of plate current through the load resistance, r_b, of one tube makes the grid of the other tube negative. If the voltage across r_b is large enough, the other tube is biased to cut-off. As a result, current flows in only one tube at a time. Conduction may be caused to transfer abruptly from one tube to the other at critical values of grid or plate voltage, however, merely by changing the value of any voltage on any electrode in a direction such as to decrease the plate current of the conducting tube or cause current to flow in the other tube. Triggering action can be initiated by a voltage pulse of short duration, or even by merely touching the grid terminal of one tube.

Circuits such as these are the basis of many electronic measuring instruments and controls.

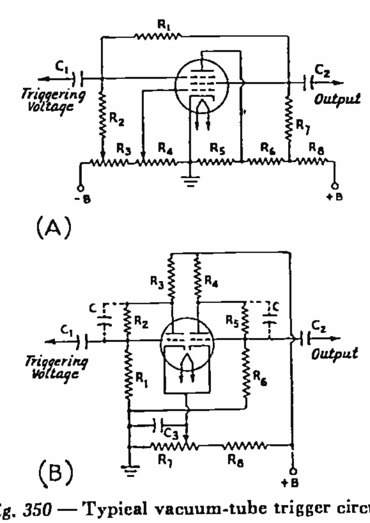


Fig. 350 — Typical vacuum-tube trigger circuits.

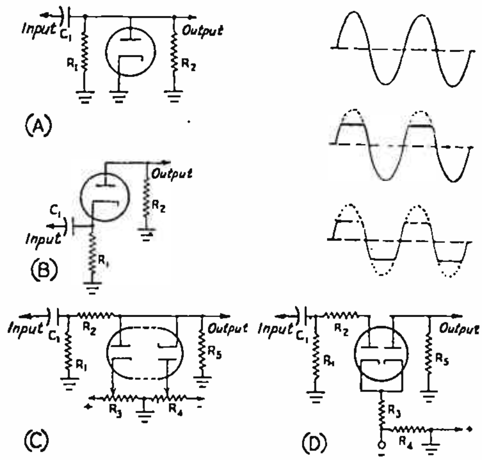


Fig. 351 — Shaping of sine wave to square wave by diode clipping action. The waveforms at the upper right illustrate, progressively, the sinusoidal input wave, the positive peak clipped by the diode parallel limiter (A), and the negative peak clipped by the diode series limiter (B). These are performed jointly in double-diode parallel limiter (C) and double-diode series limiter (D).

3-10 Pulse Technique

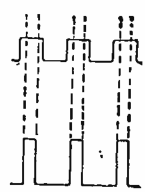
In pulse transmission and reception (§ 1-4), specialized means are employed to generate and shape characteristic pulses on the transmitting end and to recreate and interpret these pulses on the receiving end. One is a process of waveshaping and injection; the other of separation and selection. Certain basic circuit elements are common to both; elementary examples of such circuits will be discussed in this section.

Waveshaping — The primary waveforms employed in pulse transmission, apart from the basic sine wave, are the rectangular wave (from narrow pulse to square wave), trapezoidal wave, triangular wave (from isosceles to right-angle sawtooth), exponential and sawtooth waves.

The nonsinusoidal waveforms obtainable from certain oscillators, particularly those of the relaxation type, approximate the general shapes required. To trim such waves to the ideal form required, auxiliary waveshaping circuits are employed. The basic categories are (1) limiter circuits, which utilize the voltage-limiting action of vacuum tubes, and (2) peaking circuits, which employ *RC* (or *LC*) time-constant circuits.

Fig. 351 shows the use of biased-diode limiters in clipping a sine wave to create a square or trapezoidal waveshape by limiting action.

The diode parallel limiter at A does not limit the output until the input voltage attains a value more positive than that of a negative biasing voltage applied in series with R_2. In the diode series limiter at B, conduction can occur only when the input is more positive than the biasing voltage, inserted in series with R_1. Thus there can be no increase in output during the most negative period of the cycle. The series limiter produces a more squarely clipped wave than the parallel type.



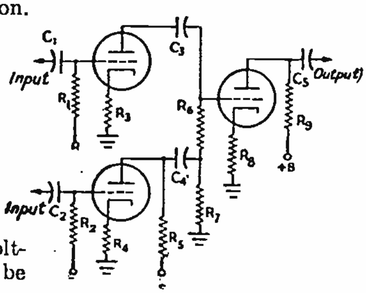


Fig. 353 — Pulse mixer or injector circuit, illustrating how two rectangular pulses of different bases and amplitudes are combined into one complex pulse before transmission.

The operation of either type can be reversed by reversing the polarity of the biasing voltage. In the double diode parallel limiter at C the left-hand diode removes positive peaks while that at the right clips the negative. The degree of limiting is adjusted by varying the fixed bias by means of R_3 and R_4. The double series limiter at D functions in a similar manner but is more critical of adjustment, since the bias source is common to both tubes. Triode limiters may be operated at cut-off or at saturation.

In Fig. 352, the tube is biased near the center of its characteristic. When the signal voltage goes negative, at cut-off, plate current ceases to flow and the bottom of the sine wave is clipped. On the positive peak the plate current is limited by saturation and the top of the sine curve is squared off. The input signal should be 20 or 30 times the grid base for the sine wave to be squared off reasonably sharply.

Limiter circuits may also be employed for generating other types of pulses. If the tube in Fig. 256 is biased beyond cut-off and a con-

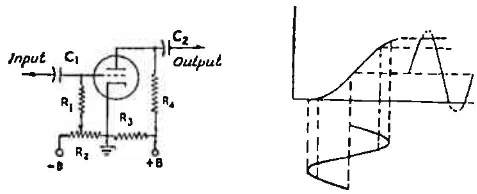


Fig. 352 — Triode limiter action in generating square or trapezoidal wave by clipping peaks of a sinusoidal wave.

denser is connected between plate and ground, a positive rectangular pulse applied to the grid will produce a sawtooth wave. During the interval between pulses the condenser is charged in a relatively slow linear rate through R_4. The sharp front of the positive pulse on the grid causes plate current to flow, and the condenser discharges rapidly through the tube. A triangular waveshape can be obtained by reducing the bias to zero and applying negative pulses to the grid. Between pulses plate current will flow, but each negative pulse biases the tube beyond cut-off, making it nonconducting. The condenser charges through R_4 for the duration of the pulse, then discharges through R_4. The result is a symmetrical triangular pulse.

Pulse selection — Pulse selectivity is based on the following characteristics: (1) polarity; (2) amplitude; (3) shape; and (4) duration (including both "mark" and "space" intervals).

The diode separator functions much like the diode limiters of Fig. 351, except that the action is reversed. Selection by polarity is based on the unilateral conductivity of the diode rectifier, and requires only that the diode be so connected as to pass positive or negative pulses, as desired. For amplitude separation the diode is so biased that only pulses having a large amplitude exceeding the bias voltage will be passed.

The same resemblance applies in the case of triode amplitude separators. In the cut-off or zero-bias separator of Fig. 354, the grid normally is biased to cut-off. When a positive voltage of sufficient amplitude is applied, plate current flows. There will be no response to voltages of lesser amplitude, or to negative pulses.

The positive-grid or blocked-grid separator operates at saturation and is characterized by a series resistor in the grid circuit. Positive pulses drive the tube into the positive-grid region, where grid-current flow increases bias and limits plate-current to a steady value regardless of signal level. Since this circuit passes only negative pulses, it is selective as to polarity.

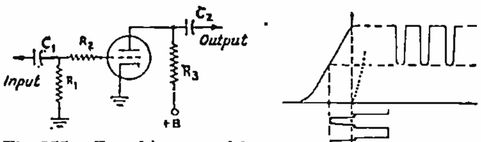


Fig. 355 — Zero-bias or positive-grid limiter-separator.
C_1 — 0.1 μfd. R_1, R_2 — 1 megohm.
C_2 — 0.5 μfd. R_3 — 0.1 megohm

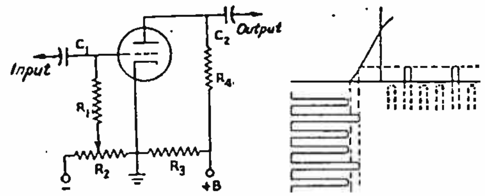


Fig. 354 — Cut-off biased triode amplitude separator.
C_1 — 0.1 μfd. R_1 — 1 megohm. R_3 — 50,000 ohms.
C_2 — 0.5 μfd. R_2 — 2000 ohms. R_4 — 25,000 ohms.

Differentiation — If the front of a rectangular wave is applied to a short time-constant *RC* circuit with series capacity and shunt resistance, the condenser will charge rapidly. Thus the voltage across the load resistor will achieve an instantaneous amplitude equal to the applied voltage and, if the time constant is very short, will have added to it practically the entire charge of the condenser. The instant the condenser is fully charged (which, in the case above, occurs practically instantaneously) it begins to discharge through the load resistor. The condenser voltage during discharge follows an exponential curve (§ 2-6, Fig. 224). Thus the voltage across *R* has the shape of a short pulse with a very sharp peak and an exponential front of relatively short base.

Following this initial positive pulse, a condition of equilibrium in which the voltage across *R* remains constant at the level of the applied voltage continues for the duration of the input wave. Then the process is repeated for the trailing edge of the wave, except that it is a negative pulse with the same characteristics.

By altering the steepness of either the ascending or descending slopes of the input wave the amplitude of the output pulse can be controlled. This is the principle upon which pulse selection by waveshape is based, as illustrated in Fig. 356. A steep front produces a sharp pulse having an amplitude greater than the applied voltage, while a sloping front produces a pulse of correspondingly greater length and lesser amplitude. For sharp pulses the time constant must be considerably shorter than one-half cycle of the input wave. With a longer time constant the charging period becomes correspondingly longer, so that the discharge time, while retaining a logarithmic shape, approaches the duration and form of the wave. Such a network is called a differentiating circuit.

In a reversed time-constant circuit, with the resistor in series and the condenser in shunt, the action is such that with a very short time constant the output wave resembles that of the input except for a slight curvature at the beginning because of the exponential charging characteristic of the condenser. The amplitude is, however, greatly reduced because of the voltage divider effect of the reactance-resistance combination. Increasing the time constant to a value comparable to the duration of the constant-amplitude portion of the input wave increases the amplitude but accentuates also both the ascending and descending slopes of the wave.

Increasing the time constant to more than unity, so that it becomes very long compared with the base of the input wave, results in what is called an integrating circuit. In this circuit discrimination or selection is based on the duration or frequency of the input wave. For example, if a series of short pulses is applied to the input, long-time constant integrator circuit, the energy stored in the condenser by each individual pulse will be discharged before the next pulse arrives. If, however, a series of pulses with longer bases and shorter intervals is applied, only a portion of the energy from each pulse will be discharged before the next begins charging. Energy is therefore accumulated on the condenser until a predetermined amplitude is established. Thus long-base pulses can be separated from shorter pulses.

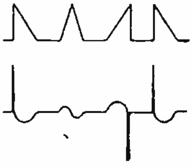


Fig. 356 — Pulse selection based on the discriminating action of a differentiating circuit with inputs of different wavefront shapes. Typical input waves are shown above and the resulting output pulses below.

ℂ 3-11 V.H.F. and U.H.F. Tubes

Negative-grid tubes — In low-frequency operation, the actual time of flight of electrons between the cathode and the anode is negligible in relation to the duration of the cycle. At 1000 kc., for example, transit time of 0.001 microsecond, which is typical of conventional tubes, is only 1/1000 cycle. At 100 Mc., this same transit time represents 1/10 of a cycle, and a full cycle at 1000 Mc. At very high frequencies, therefore, the grid-cathode transit time, together with interelectrode capacities and the inductance of internal leads, determine the highest possible frequency to which a vacuum tube can be tuned. The tube usually will not oscillate up to this limit, however, because of dielectric losses, grid emission, and transit-time effects. These limiting factors establish about 3000 Mc. as the upper frequency limit for negative-grid tubes. Even then the electrodes must be so minute that only a few milliwatts of r.f. power can be obtained.

With tubes of ordinary construction, the upper limit of oscillation is about 150 Mc. For higher frequencies, v.h.f. tubes of special construction are used. The "acorn" and "doorknob" types and the special v.h.f. "miniature" tubes, in which the grid-cathode spacing is made as little as 0.005 inch, are capable of operation up to about 700–800 Mc. The normal frequency limit is around 600 Mc., although output may be obtained up to 800 Mc.

Very low interelectrode capacities and lead inductance have been achieved in the newer tubes of modified construction. In multiple-lead types the electrodes are provided with up to three separate leads which, when connected in parallel, have considerably reduced effective inductance. In double-lead types the plate and grid elements are supported by heavy single wires which run entirely through the envelope, providing terminals at either end of the bulb. When a resonant circuit is connected to each pair of leads, the shunting capacity divides between the two circuits. With linear circuits the leads become a part of the line and have distributed rather than lumped constants. Radiation loss is minimized and the effect of the transit time is reduced. In "lighthouse" tubes or *megatrons* the plate, grid and cathode are assembled in parallel planes instead of coaxially. The uniform coplanar electrode design and disc-seal terminals permit very low interelectrode capacities.

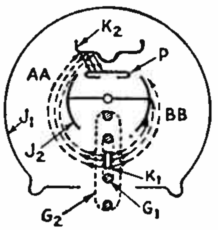


Fig. 357 — Schematic cross-section of the orbital-beam secondary-electron multiplier tube.

In the orbital-beam tube, Fig. 357, a small electrode structure with relatively low initial transconductance is used with a secondary-electron emitter to raise the effective transconductance. Electrons emitted from the cathode, K_1, are accelerated through the control grid, G_1, by a positive grid, G_2, and enter a radial electrostatic field established by the cylindrical electrodes, J_1 and J_2, causing the electrons to move in a circular path and driving them against the secondary-emitter electrode, K_2. About ten secondary electrons are emitted for each primary electron; thus the ultimate electron flow to the plate, P, is considerably greater than the original current emitted. Thus high over-all transconductance (15,000 at 500 Mc. in an experimental tube) is obtained without increasing transit-time losses or internal capacities.

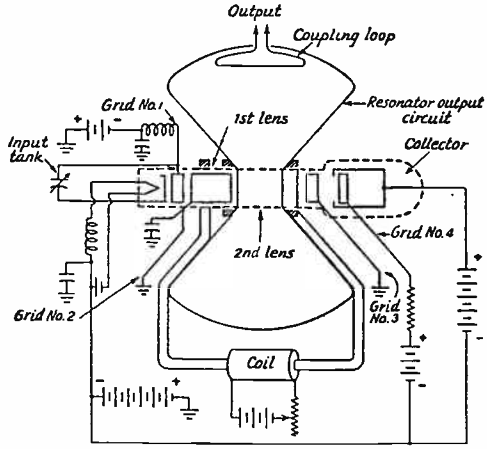


Fig. 358 — Schematic of the inductive output amplifier.

Inductive output amplifier — In the inductive-output tube shown in Fig. 358 the high-velocity electron beam is intensity-modulated by the control grid (grid No. 1). After being accelerated and focused by the combined action of the first and second lenses in the magnetic circuit and the sleeve electrodes (grids No. 2 and 3), the beam moves past a small ¼-inch aperture in the "dimpled sphere" cavity resonator. The potential difference in this gap creates a retarding field which absorbes power from the beam. Electrons passing through the structure are decelerated by a suppressor electrode (grid No. 4) before reaching the final anode or collector. The control-grid structure gives sharp cut-off and large transconductance, while the high accelerating potentials and small apertures result in very short transit time and consequently low input conductance. The inductive-output tube is useful for wide-band operation above 500 Mc., giving efficiencies of 25 per cent or better.

Velocity modulation — In negative-grid operation the negative potential on the grid tends to slow down the electron velocity during the more negative half of the oscillation cycle. while on the other half cycle the positive potential on the grid serves to accelerate their velocity. Thus the electrons tend to separate into groups, those leaving the cathode during the negative half cycle being collectively slowed down, while those leaving on the positive half are accelerated. After passing into the grid-plate space only a part of the electron stream follows the original form of the oscillation cycle, the remainder traveling to the plate at differing velocities. Since these contribute nothing to the power output at the operating frequency, the efficiency is reduced in proportion to the variation in velocity, the output becoming zero when the transit time approaches a half cycle.

This effect, such a disadvantage in conventional tubes, is an advantage in velocity-modulated tubes in that the input signal voltage on

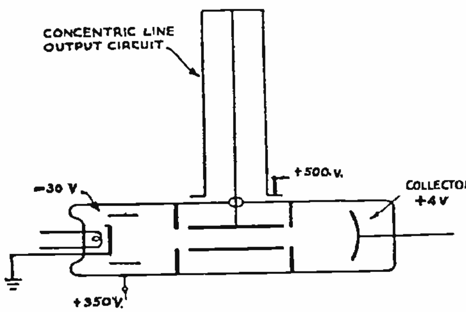


Fig. 359 — Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a super-regenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50 cm. region.

the grid is used to charge the velocity of the electrons in a constant current electron beam, rather than to vary the intensity of a constant velocity current flow as in ordinary tubes.

A simple form of velocity-modulation oscillator tube is shown in Fig. 359. Electrons emitted from the cathode are accelerated through a negatively biased cylindrical grid by a constant positive voltage applied to a sleeve electrode, shown in heavy lines. This electrode, which is the velocity-modulation control grid, consists of two hollow tubes, with a small aperture on each end between the inner tube, through which the electron beam passes, and the discs at the ends of the larger tube portion. With r.f. voltage applied across these gaps, the transit angle of which is made small, electrons entering the tube will be accelerated on positive half cycles and decelerated on the negative half cycles. The length of the tube is made equal to the transit angle, so that the electrons will be further accelerated or decelerated as they leave the tube.

As the beam approaches the collector electrode, which is at nearly zero potential, the electrons are retarded, brought to rest, and ultimately turned back by the attraction of the positive sleeve electrode. The collector electrode is, therefore, also termed a reflector. The point at which electrons are returned depends on their velocity. Thus the velocity modulation is again translated into current modulation.

Velocity-modulated tubes operate satisfactorily up to 6000 Mc. (5 cm.) and higher, with outputs of 100 watts or more.

The Klystron — In the *Klystron* velocity-modulated tube, the electrons emitted by the cathode are accelerated or retarded during their passage through an electric field established by two grids in a cavity resonator, or *rhumbatron*, called the "buncher." The high-frequency electrostatic field between the grids is parallel to the electron stream. This field accelerates the electrons at one moment and retards them at another, in accordance with the variations of the r.f. voltage applied. The resulting velocity-modulated beam travels through a field-free drift space, where the slowly moving electrons are gradually overtaken by the faster ones leaving a half-cycle later. At section points in the disc space the emerging electrons therefore are separated into groups or bunched along the direction of motion. The velocity modulated electron stream is passed to a "catcher" rhumbatron. Again the beam passes through two parallel grids, the higher- and lower-speed electrons inducing different amounts of potential between the grids. The catcher cavity is made resonant at the frequency of the velocity-modulated electron beam, so that an oscillating field is set up within it by the passage of the electron bunches through the grid aperture.

If a feed-back loop is provided between the two rhumbatrons, as shown in Fig. 360, oscillations will occur. The resonant frequency depends on the electrode voltages and on the shape of the cavities, and may be adjusted by varying the supply voltage and altering the dimensions of the rhumbatrons. The output waveform is rich in harmonics.

The maximum theoretical efficiency of the Klystron oscillator is 58 per cent; commercially available types give 5 to 10 watts output at about 10 per cent efficiency.

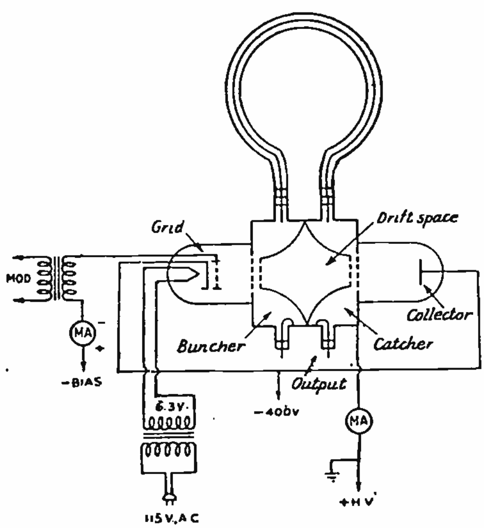


Fig. 360 — Circuit diagram of the Klystron oscillator showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.

Positive-grid electron oscillators — A triode in which the grid is positive with respect to the cathode rather than the plate will oscillate at frequencies higher than those at which transit-time effects cause the tube to be inoperative as a normal negative-grid oscillator.

Oscillators of the positive-grid type, also known as brakefield of electron transit-time oscillators, employ conventional tubes of almost any type. Successful performance is most readily achieved with tube structures having cylindrical grids and plates.

Under static conditions, electrons emitted from the cathode on being accelerated towards the positive grid hit its wires while others pass through the grid into the retarding field between it and the negative plate. No electrons reach the plate, all being driven back toward the grid. Again, some electrons are collected while others pass through into the space between the grid and cathode. Those electrons which have not been collected by the grid are returned to the cathode.

If an a.c. voltage is superimposed on the steady positive voltage of the grid, some electrons leaving the cathode when the a.c. component of the grid voltage is zero and about to increase in a negative direction are driven through the grid and are returned toward it by the grid-to-plate field. If, however, the frequency is such that the time of a half-cycle is equal to the transit time of the electrons from cathode to grid, the voltage reverses direction as the electrons pass through the grid, and the grid-plate retarding field is greater than the cathode-grid accelerating field. The electrons therefore oscillate about the grid, losing energy each time, until they finally return to the grid.

Electrons leaving the cathode when the a.c. component of the grid voltage is zero and about to increase in a positive direction are acted on by the field in the opposite manner. In this case the electron energy is increased, and so these electrons are collected by either the grid, plate or cathode after having completed not more than one cycle.

Since those electrons which lose energy to the field remain free in the tube longer than those which gain energy from the field, there is an average transfer of energy which can be used to maintain high-frequency oscillations in a resonant line connected between the grid and plate of the tube.

In this type of oscillator, shown in Fig. 361, the frequency is controlled primarily by the grid voltage and the tube element spacing. The tuning of the resonant circuit affects the amplitude of the oscillations and thus must be tuned to approximately the oscillation frequency, however.

Positive-grid oscillators can be operated at frequencies up to 10,000 Mc. (3 cm.), but the efficiency is usually only 2 or 3 per cent. Since most of the power is dissipated in the grid, the tube is not capable of delivering much power.

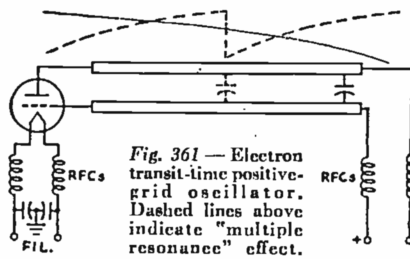


Fig. 361 — Electron transit-time positive-grid oscillator. Dashed lines above indicate "multiple resonance" effect.

Magnetrons — A magnetron is fundamentally a diode with cylindrical electrodes placed in a uniform magnetic field with the lines of electromagnetic force parallel to the elements.

The simple cylindrical magnetron consists of a filamentary cathode surrounded by a concentric cylindrical anode. In the more efficient split-anode magnetron the anode is divided longitudinally.

Magnetron oscillators are operated in two essentially different ways, one corresponding to the negative-grid oscillator. Electrically the two types of operation are similar, the difference being in the relation between electron transit time and the frequency of oscillation.

In the negative-resistance or dynatron oscillator the element dimensions and anode voltage are such that the transit time is short in comparison with the period of the oscillation frequency. Electrons emitted from the cathode are driven towards both halves of the anode. If the potentials of the two halves are unequal, the majority of the electrons travel to that half of the anode which is at the lower potential. In other words, a *decrease* in the potential of either half of the anode results in an *increase* in the electron current flowing to that half, so that the magnetron consequently exhibits negative-resistance characteristics (§ 3-7).

Negative-resistance magnetron oscillators are useful between 100 and 1000 Mc. Under the best operating conditions efficiencies of 20 to 25 per cent may be obtained. Since the power loss in the tube appears as heat in the anode, where it is readily dissipated, relatively large power-handling capacity can be obtained.

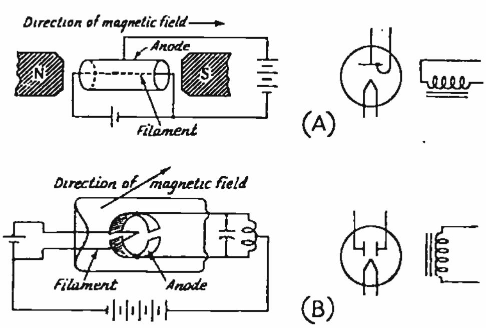


Fig. 362 — Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.

The frequency of a transit-time magnetron is determined primarily by its dimensions and by the intensities of the electric and magnetic fields, rather than by the tuning of the tank circuits. The efficiency is much better than that of a positive-grid oscillator and good power output can be obtained even on the superhighs.

In a nonoscillating magnetron with a weak magnetic field electrons traveling from the cathode to the anode move almost radially, their trajectories being bent only slightly by the magnetic field. Under critical conditions of magnetic field strength outside the immediate neighborhood of the filament a cloud of electrons rotates about the filament. It extends up to the anode but does not actually reach it. With increased magnetic field the electrons tend to spiral around the filament, their radial component of velocity being much smaller than the angular component.

The nature of these electron trajectories is shown in Fig. 363. Cases A, B, and C, correspond to the non-oscillating condition. For a small magnetic field (A) the trajectory is bent slightly near the anode. This bending increases for a higher magnetic field (B) and the electron moves through quite a large angle near the anode before reaching it, signifying a large increase of space charge near the anode. For a strong magnetic field (C) electrons start radially from the cathode but are soon bent and curl about the filament in the form of a long spiral before reaching the anode. This means a very long transit time and a very large space charge in the whole region where the spiraling takes place. Under critical conditions (D), no current flows to the anode and no electron is able to move from cathode to anode, but a large space charge still exists between the cathode and anode. The spiraling becomes a set of concentric circles, and the entire space-charge distribution rotates about the filament.

Figs. 363-E, -F and -G depict higher order (harmonic-type) modes of operation in which the space charge oscillates not only symmetrically but in transverse directions contrasting to the vibrations of the fundamental.

In a transit-time magnetron oscillator the intensity of the magnetic field is adjusted so that, under static conditions, electrons leaving the cathode move in curved paths which just

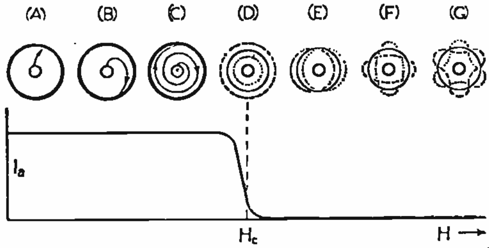


Fig. 363 — Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, *I*. Oscillations commence when *H* reaches a critical value, H_c; progressively higher order modes of oscillation occur beyond this point.

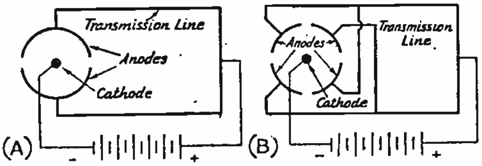


Fig. 364 — U.h.f. magnetron circuits. A, split-anode type. B, four-anode type with opposite electrodes paralleled.

fail to reach the anode. All electrons are therefore deflected back to the cathode, and the anode current is zero. When an alternating voltage is applied between the two halves of the anode, causing the potentials of these halves to vary about their average positive values, the conditions in the tube become analogous to

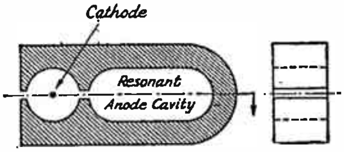


Fig. 365 — Split-anode magnetron with integral resonant anode cavity for u.h.f. use.

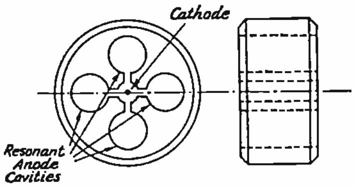


Fig. 366 — Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.

those in a positive-grid oscillator. If the period of the alternating voltage is made equal to the time required for an electron to make one complete rotation in the magnetic field, the a.c. component of the anode voltage reverses direction twice with each electron rotation. Some electrons will lose energy to the electric field, with the result that they are unable to reach the cathode and continue to rotate about it. Meanwhile other electrons gain energy from the field and are returned to the cathode. Since those electrons which lose energy remain in the interelectrode space longer than those which gain energy, the net effect is a transfer of energy from the electrons to the electric field. This energy can be applied to sustain oscillations in a resonant transmission line between the two halves of the anode.

Split anode magnetrons for u.h.f. are constructed with a cavity resonator built into the tube structure, as illustrated in Fig. 366. The assembly is a solid block of copper which assists in heat dissipation. At extremely high frequencies operation is improved by subdividing the anode structure into from 4 to 16 or more segments, the resonant cavities for each anode coupled by slots of critical dimensions to the common cathode region, as in Fig. 366.

The efficiency of multi-segment magnetrons reaches 65 or 70 per cent. Slotted-anode magnetrons with four segments function up to 30,000 Mc. (1 cm.) delivering up to 100 watts at efficiencies greater than 50 per cent. Using larger multiples of anodes and higher-order modes, performance can be attained at 0.2 cm.

Chapter Four

Radio-Frequency Power Generation

4-1 Transmitter Requirements

General Requirements — To minimize interference when a large number of stations must work in one frequency band, the power output of a transmitter must be as stable in frequency and as free from spurious radiations as the state of the art permits. The steady r.f. output, called the *carrier* (§ 5-1), must be free from amplitude variations attributable to ripple from the plate power supply (§ 8-4) or other causes, its frequency should be unaffected by variations in supply voltages or inadvertent changes in circuit constants, and there should be no radiation on other than the intended frequency. The degree to which these requirements can be met depends upon the operating frequency.

Design principles — The design of the transmitter depends on the output frequency, the required power output and the type of operation (c.w. telegraphy or 'phone). For c.w. operation at low power on medium-high frequencies (up to 7 Mc. or so), a simple crystal oscillator circuit can meet the requirements satisfactorily. However, the stable power output which can be taken from an oscillator is limited, so that for higher power the oscillator is used simply as a frequency-controlling element, the power being raised to the desired level by means of amplifiers. The requisite frequency stability can be obtained only when the oscillator is operated on relatively low frequencies, so that for output frequencies up to about 60 Mc. it is necessary to increase the oscillator frequency by multiplication (harmonic generation — § 3-3), which usually is done at fairly low power levels and before the final amplification. An amplifier which delivers power on the frequency applied to its grid circuit is known as a *straight amplifier;* one which gives harmonic output is known as a *frequency multiplier*. An amplifier used principally to isolate the frequency-controlling oscillator from the effects of changes in load or other variations in following amplifier stages is called a *buffer amplifier*. A complete transmitter therefore may consist of an oscillator followed by one or more buffer amplifiers, frequency multipliers and straight amplifiers, the number being determined by the output frequency and power in relation to the oscillator frequency and power. The last amplifier is called the *final amplifier*, and the stages up to the last comprise the *exciter*. Transmitters usually are designed to work in a number of frequency bands so that means for changing frequency harmonic steps usually is provided, generally by means of plug-in inductances.

The general method of designing a transmitter is to decide upon the power output and the highest output frequency required, and also the number of bands in which the transmitter is to operate. The latter usually will determine the oscillator frequency, since it is general practice to set the oscillator on the lowest frequency band to be used. The oscillator frequency seldom is higher than 7 Mc. except in some portable installations where tubes and power must be conserved. A suitable tube (or pair of tubes) should be selected for the final amplifier, and the required grid driving power determined from the tube manufacturer's data. This sets the power required from the preceding stage. From this point the same process is followed back to the oscillator, including frequency multiplication wherever necessary. The selection of a suitable tube complement requires a knowledge of the operating characteristics of the various types of amplifiers and oscillators. These are discussed in the following sections.

At 112 Mc. and higher frequencies these methods of transmitter design tend to become rather cumbersome, because of the necessity for a large number of frequency multiplier stages. However, in this frequency region less severe stability requirements are imposed because the transmission range is limited (§ 9-5) and the possibility of interference to other communication is reduced. Simple oscillator transmitters, without frequency multiplication or buffer amplifiers, are widely used at 112 Mc. and above.

Vacuum tubes — The type of tube used in the transmitter has an important effect on the circuit design. Tubes of high power sensitivity (§ 3-3) such as pentodes and beam tetrodes give larger power amplification ratios per stage than do triodes, hence fewer tubes and stages may be used to obtain the same output power. On the other hand triodes have certain operating advantages, such as simpler power supply circuits and relatively simpler adjustment for modulation (§ 5-3), and in addition are considerably less expensive for the same power output rating. Consequently it is usually more economical to use triodes as output amplifiers, even though an extra low-power amplifier stage may be necessary.

At frequencies in the region of 56 Mc. and above it is necessary to select tubes designed particularly for operation at very-high frequencies, since tubes built primarily for lower frequencies may work poorly or not at all.

4-2 Self-Controlled Oscillators

Advantages and disadvantages — The chief advantage of a self-controlled oscillator is that the frequency of oscillation is determined by the constants of the tuned circuit, and hence readily can be set to any desired value. However, extreme care in design and adjustment are essential to secure satisfactory frequency stability (§ 3-7). Since frequency stability is generally poorer as the load on the oscillator is increased, the self-controlled oscillator should be used purely to control frequency and not for the purpose of obtaining appreciable power output in transmitters intended for working below 60 Mc.

Oscillator circuits — The inherent stability of all of the oscillator circuits described in § 3-7 is about the same, since stability is more a function of choice of proper circuit values and of adjustment than of the method by which feed-back is obtained. However, some circuits are more convenient to use than others, particularly from the standpoint of feed-back adjustment, mechanical considerations (whether the tuning condenser rotor plates can be grounded or not, etc.), and uniform output over a considerable frequency range. In all simple circuits the power output must be taken from the frequency-determining tank circuit, which means that, aside from the effect of loading on frequency stability, the following amplifier stage can react on the oscillator and cause a change in the frequency.

Factors influencing stability — The causes of frequency instability and the necessary remedial steps have been discussed in § 3-7. These apply to all oscillators. In the case of the electron-coupled oscillator the ratio of plate to screen voltage has marked effect on the stability with changes in supply voltage; the optimum ratio is generally of the order of 3:1, but should be determined experimentally for each case. Since the cathode is above ground potential, means should be taken to reduce the effects of heater-to-cathode capacitance or leakage which, by allowing a small a.c. voltage from the heater supply to develop between cathode and ground, may cause modulation (§ 5-1) at the supply frequency.

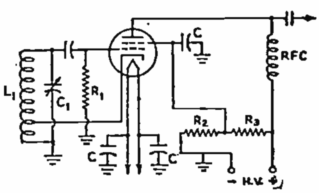


Fig. 401 — Electron-coupled oscillator circuit. R_1 should be 100,000 ohms or more, the grid condenser 100 $\mu\mu$fd. and the other fixed condensers 0.002 to 0.1 μfd.

This effect, which is usually appreciable only at 14 Mc. and higher, may be reduced by by-passing the heater as in Fig. 401 or by operating the heater at the same r.f. potential as the cathode. The latter may be accomplished by the wiring arrangement shown in Fig. 402.

Tank-circuit Q — The most important single factor in determining frequency stability is the Q of the oscillator tank circuit. The effective Q must be as high as possible for best stability. Since oscillation is accompanied by grid-current flow the grid-cathode circuit

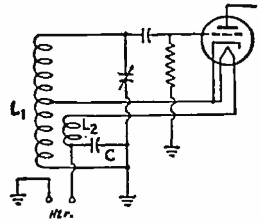


Fig. 402 — Method of operating the heater at cathode r.f. potential in an electron-coupled oscillator. L_2 should have the same number of turns as the cathode section of L_1 and should be closely coupled (preferably interwound). C may be 0.01 to 0.1 μfd.

constitutes a resistance load of appreciable proportions, the effective resistance being low enough to be the determining factor in establishing the effective parallel impedance of the tank circuit. Consequently, if the ends of the tank are connected to plate and grid, as is usual, a high effective Q can be obtained only by decreasing the L/C ratio and making the inherent resistance in the tank as low as possible. The tank resistance can be decreased by using low-loss insulation and by winding the coil with large wire. With ordinary construction, the optimum tank capacity is of the order of 500 to 1000 $\mu\mu$fd. at a frequency of 3.5 Mc.

The effective circuit Q can be raised by increasing the resistance of the grid circuit and thus decreasing the loading. This can be accomplished through reducing the oscillator grid current, using minimum feed-back for stable oscillation, and a high value of grid-leak resistance.

A high-Q tank circuit can also be obtained with a higher L/C ratio by "tapping down" the tube connections on the tank (§ 2-10). This is advantageous in that a coil with higher inherent Q can be used; also, the circulating r.f. current in the tank circuit is reduced so that drift from coil heating is decreased. However, under some conditions parasitic oscillations may be set up (§ 4-10).

Plate supply — Since the oscillator frequency will be affected to some extent by changes in plate-supply voltage, it is necessary that the latter be free from ripple (§ 8-4) which would cause frequency variations at the ripple-frequency rate (*frequency modulation*). It is advantageous to use a voltage-stabilized power supply (§ 8-8). Since the oscillator usually is operated at low voltage and current, VR-type gaseous regulator tubes are quite suitable.

Power level — The self-controlled oscillator should be designed purely for frequency control and not to give appreciable power output, hence small tubes of the receiving type may be used. The power input ordinarily is not more than a watt or two, subsequent buffer amplifiers being used to increase the power to the desired level. The use of receiving tubes is advantageous mechanically, since the small elements are less susceptible to vibration and usually are securely braced to the envelope of the tube.

Oscillator adjustment — The adjustment of an oscillator consists principally in observing the design principles outlined in the preceding paragraphs. Frequency stability should be checked with the aid of a stable receiver. An auxiliary crystal oscillator may be used as a standard for checking dynamic stability and drift, the self-controlled oscillator being adjusted to approximately the same frequency so that an audio-frequency beat (§ 2-13) can be obtained. If it is possible to vary the oscillator plate voltage (an adjustable resistor of 50,000 or 100,000 ohms in series with the plate supply lead will give considerable variation), the change in frequency with change in plate voltage may be observed and the operating conditions varied until minimum frequency shift results. The principal factors affecting dynamic stability will be the tank circuit L/C ratio, the grid-leak resistance, and the amount of feed-back. In the electron-coupled circuit the latter may be adjusted by changing the cathode tap on the tank coil; critical adjustment is required for optimum stability.

Drift may be checked by allowing the oscillator to operate continuously from a cold start, the frequency change being observed at regular intervals. Drift may be minimized by using less than the rated power input to the plate of the tube, by construction which prevents tube heat from reaching the tank circuit elements, and by use of large wire in the tank coil to reduce temperature rise from internal heating.

In the electron-coupled oscillator having a tuned plate circuit (Fig. 401-A), resonance at the fundamental and harmonic frequencies of the oscillator portion of the tube will be indicated by a decrease in plate current as the plate tank condenser is varied. This "dip" is less marked at the fundamental than on harmonics.

ℂ 4-3 Crystal Control

Characteristics — Piezoelectric crystals (§ 2-10) are universally used for controlling the frequency of transmitting oscillators, because the extremely high Q of the crystal and the necessarily loose coupling between it and the oscillator tube make the frequency stability of a crystal-controlled oscillator very high.

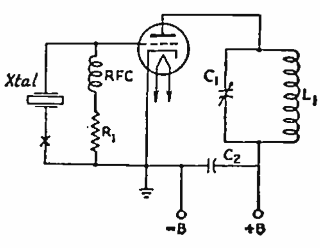


Fig. 403 — Triode crystal oscillator. The tank condenser, C_1, may be a 100-$\mu\mu$fd. variable, with L_1 proportioned so that the tank will tune to the crystal frequency. C_2 should be 0.001 μfd. or larger. The grid leak, R_1, will vary with the type of tube; high-μ tubes take values of 2500 to 10,000 ohms, while medium and low-μ types take values of 10,000 to 25,000 ohms. A small flashlight bulb or r.f. milliammeter (§ 4-3) may be inserted at X.

The ability to adhere closely to a known frequency is the outstanding characteristic of a crystal oscillator. This also is a disadvantage, in that a different crystal is required for each frequency on which the transmitter is to operate.

Power Limitations — The temperature of a crystal depends not only on the temperature of its surroundings but also on the power it must dissipate while oscillating, since power dissipation causes heating (§ 2-6, 2-8). Consequently, the crystal temperature in operation may be considerably above that of the surrounding air. To minimize heating and frequency drift (§ 3-7), the power dissipated must be kept to a minimum.

If the crystal is made to oscillate too strongly, as when it is used in an oscillator circuit with high plate voltage and excessive feed-back, the amplitude of the mechanical vibration will become great enough to crack or puncture the quartz. An indication of the vibration amplitude can be obtained by connecting an r.f. current-indicating device of suitable range in series with the crystal. Safe r.f. crystal currents range from 50 to 200 milliamperes, depending upon the type of cut. A flashlight bulb or dial light of equivalent current rating makes a good current indicator. By choosing a bulb of lower rating than the current specified by the manufacturer as safe for the particular type of crystal used, the bulb will serve as a fuse, burning out before a current dangerous to the crystal is reached. The 60-ma. and 100-ma. bulbs may be used for this purpose. High crystal current means increased power dissipation and heating, so that the frequency change also is greater.

Crystal mountings — To make use of the crystal, it must be mounted between two metal electrodes. There are two types of mountings, one having a small air-gap between the top plate and the crystal and the other maintaining both plates in contact with the crystal. It is essential that the surfaces of the metal plates in contact with the crystal be perfectly flat. In the air-gap type of holder, the frequency of oscillation depends to some extent upon the size of the gap. By using a holder having a top plate with closely adjustable spacing, a controllable frequency variation can be obtained. A suitable 3.5-Mc. crystal will oscillate without great variation in power output over a range of about 5 kc. X- and Y-cut crystals are not generally suitable; they have a tendency to "jump" in frequency with different air gaps.

A holder having a heavy metal bottom plate with a large surface exposed to the air is advantageous in that it radiates quickly the heat generated in the crystal, thereby reducing temperature effects. Different plate sizes, pressures, etc., will cause slight changes in frequency, so that if a crystal is being ground to an exact frequency it should be tested in the same holder and in the same oscillator circuit with which it will be used in the transmitter.

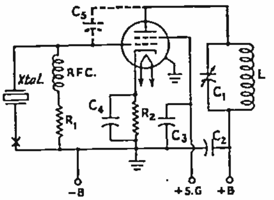


Fig. 404 — Tetrode or pentode crystal oscillator. Typical values: C_1, 100 $\mu\mu$fd., with L wound to suit frequency; C_2, C_3, 0.001 μfd. or larger; C_4, 0.01 μfd.; R_1, 10,000 to 50,000 ohms (value determined by trial); R_2, 250 to 400 ohms.

ℂ 4-4 Crystal Oscillators

Triode oscillators — The triode crystal oscillator circuit (§ 3-7) is shown in Fig. 403. The limit of plate voltage that can be used without endangering the crystal is about 250 volts. With the r.f. crystal current limited to a safe value of about 100 ma., the power output obtainable is about 5 watts. The oscillation frequency is dependent to some extent on the plate tank tuning, because of the change in input capacity with changes in effective amplification (§ 3-3).

Tetrode and pentode oscillators — Since the power output of a crystal oscillator is limited by the permissible r.f. crystal current (§ 4-3), it is advantageous to use an oscillator tube of high power sensitivity (§ 3-3) such as a pentode or beam tetrode (§ 3-5). Thus for a given crystal voltage or current more power output may be obtained than with the triode oscillator, or for a given output the crystal voltage will be lower, thereby reducing crystal heating. In addition, tank-circuit tuning and loading react less on the crystal frequency because of the lower grid-plate capacity (§ 3-3).

Fig. 404 shows a typical pentode or tetrode oscillator circuit. Pentode and tetrode tubes originally designed for audio power work are excellent crystal-oscillator tubes. The screen voltage is generally of the order of half the plate voltage for optimum operation. Small tubes rated at 250 volts for audio work may be operated with 300 volts on the plate and 100–125 on the screen as crystal oscillators. The screen is at ground potential for r.f. and has no part in the operation of the circuit other than to set the operating characteristics of the tube. The larger beam tubes may be operated at 400 to 500 volts on the plate and 250 on the screen for maximum output.

Pentode oscillators operating at 250 to 300 volts will give 4 or 5 watts output under normal conditions. Beam-type tubes such as the 6L6 and 807 will give 15 watts or more at maximum plate voltage.

The grid-plate capacity may be too low to give sufficient feed-back, particularly at the lower frequencies, in which case a feed-back condenser, C_5, may be required. Its capacity should be the lowest value which will give stable oscillation; 1 or 2 $\mu\mu$fd. is generally sufficient. R_2 and C_4 may be omitted, connecting the cathode directly to ground, if plate voltage is limited to 250 volts. C_5 (if needed) may be formed by two metal plates ½-inch square spaced ¼ inch. If the tube has a suppressor grid, it should be grounded. X indicates where a flashlight bulb may be inserted (§ 4-3).

Circuit constants — Typical values for grid-leak resistances and by-pass condensers are given in Figs. 403 and 404. Since the crystal is the frequency-determining element, the Q of the plate tank circuit has a relatively minor effect on the oscillator frequency. A Q of 12 (§ 4-8) is satisfactory for average conditions, but some departure from this figure will not greatly affect the performance of the oscillator.

Adjustment of crystal oscillators — The tuning characteristics and procedure to be followed in tuning are essentially the same for triode, tetrode or pentode crystal oscillators. Using a plate milliammeter as an indicator of oscillation (a 0–100 ma. d.c. meter will have ample range for all low-power oscillators), the plate current will be found to be steady when the circuit is in the non-oscillating state, but will dip when the plate condenser is tuned through resonance at the crystal frequency. Fig. 405 is typical of the behavior of plate current as the tank condenser capacity is varied. An r.f. indicator, such as a small neon bulb touched to the plate end of the tank coil, will show a maximum indication at point A. However, when the oscillator is delivering power to a load it is best to operate in the region B-C since the oscillator will be more stable and there is less likelihood that a slight change in loading will throw the circuit out of oscillation, which is likely to happen when operation is too near the critical point, A. The crystal current also is lower in the B-C region.

When power is taken from the oscillator the dip in plate current is less pronounced, as indicated by the dotted curve. The greater the power output, the smaller the dip in plate current. If the load is made too great, oscillations will not start. Loading is adjusted by varying the coupling to the load circuit (§ 2-11).

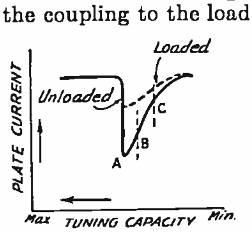


Fig. 405 — Curves showing d.c. plate current vs. plate-circuit tuning in a crystal oscillator, both with and without load. These curves apply equally to the triode, tetrode or pentode crystal oscillator.

The greater the loading, the smaller the voltage fed back to the grid circuit for excitation purposes. This means that the r.f. voltage across the crystal also will be reduced under load, hence there is less crystal heating when the oscillator is delivering power than when it is unloaded.

Failure of a crystal circuit to oscillate may be caused by any of the following:

1) Dirty, chipped or fractured crystal.
2) Imperfect or unclean holder surfaces.
3) Too tight coupling to load.
4) Plate tank circuit not tuning correctly.
5) Insufficient feed-back capacity.

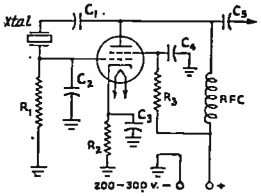


Fig. 406 — Pierce oscillator circuit. R_1 is 25,000 to 50,000 ohms. R_2 is 1000 ohms; R_3, 75,000 ohms for a 6F6; C_1, 0.001 to 0.01 μfd.; C_3 and C_4, 0.01 μfd. For values of C_2 and C_5, see text.

Pierce oscillator — This circuit, Fig. 406, is equivalent to the ultraudion circuit (§3-7), with the crystal replacing the tuned circuit. Although the output is small, it has the advantage that no tuning controls are required. The circuit requires capacitive coupling to a following stage. The amount of feed-back is determined by the condenser, C_2; its capacity must be determined by experiment, usual values being between 50 and 150 μμfd. To sustain oscillation, the net reactance (§ 2-8) of the plate-cathode circuit must be capacitive; this condition is met so long as the inductance of the r.f. choke, together with the inductance of any coils associated with the input circuit of the following stage and the tube and stray capacities, forms a circuit tuned to a lower frequency than that of the crystal.

Tubes such as the triode 6C5 and pentode 6F6 are suitable for use in this circuit. (When a triode is used the screen-voltage dropping resistor, R_3, and by-pass condenser, C_4, in Fig. 406 should, of course, be omitted.) The applied plate voltage should not exceed 300, to prevent crystal fracture. The capacity of the output-coupling condenser, C_5, should be adjusted by experiment so that the oscillator is not overloaded; usually 100 μμfd. is a satisfactory value.

ℂ 4-5 Harmonic-Generating Crystal Oscillators

Tri-tet oscillator — The Tri-tet oscillator circuit is shown in Fig. 407. In this circuit the screen grid is operated at ground potential and the cathode at an r.f. potential above ground. The screen-grid acts as the anode of a triode crystal oscillator, while the plate or output circuit is tuned to the oscillator frequency or, for harmonic output, to a multiple of it.

Besides giving harmonic output, the Tri-tet circuit has the "buffering" feature of electron-coupling between crystal and output circuits (§ 4-2). This makes the crystal frequency less susceptible to changes in loading or tuning, and hence improves the stability.

If the output circuit is to be tuned to the same frequency as the crystal, a tube having low grid-plate capacity (§ 3-2, 3-5) must be used. Otherwise there may be excessive feed-back with consequent danger of fracturing the crystal. The cathode tank circuit, L_1 C_1, is not tuned to the frequency of the crystal, but to a considerably higher frequency. Recommended values for L_1 are given under the diagram. C_1 should be set to as near minimum capacity as is consistent with good output. This reduces the crystal voltage.

With pentode-type tubes having separate suppressor connections, the suppressor may be either connected directly to ground or operated at about 50 volts positive. The latter method will give somewhat higher output.

With transmitting pentodes or beam tubes operated at 500 volts on the plate an output of 15 watts can be obtained on the fundamental and nearly as much on the second harmonic.

Grid-plate oscillator — In the grid-plate oscillator, Fig. 408, the crystal is connected between grid and ground and the cathode tuned circuit, C_2 and *RFC*, is tuned to a frequency lower than that of the crystal. This circuit gives high output on the fundamental crystal frequency with low crystal current. The output on even harmonics (2nd, 4th, etc.) is not so great as that obtainable with the Tri-tet, but on odd harmonics (3rd, 5th, etc.) the output is appreciably better.

If harmonic output is not needed, C_2 may be a fixed capacity of 100 μμfd. The cathode coil, *RFC*, may be a 2.5-mh. choke, since the inductance is not critical.

Output power of 15 to 20 watts at the crystal fundamental may be obtained with a tube such as the 6L6G at plate and screen voltages of 400 and 250, respectively.

Tuning and adjustment — The tuning procedure for the Tri-tet oscillator is as follows: With the cathode tank condenser at about three-quarters scale turn the plate tank condenser until there is a sharp dip in plate cur-

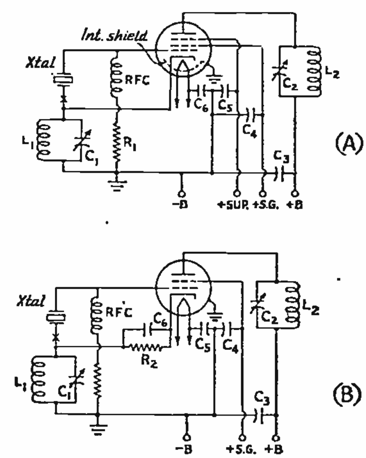


Fig. 407 — Tri-tet oscillator circuit, using pentodes (A) or beam tetrodes (B). C_1 and C_2 are 200-μμfd. variable condensers. C_3, C_4, C_5, C_6, may be 0.001 to 0.01 μfd.; their values are not critical. R_1, 20,000 to 100,000 ohms. R_2 should be 400 ohms for 400- or 500-volt operation. The following specifications for the cathode coils, L_1, are based on a diameter of 1½ inches and a length of 1 inch; turns should be spaced evenly to fill the required length: for 1.75-Mc. crystal, 32 turns; 3.5 Mc., 10 turns; 7 Mc., 6 turns. The screen should be operated at 250 volts or less. Audio beam tetrodes such as the 6L6 and 6L6G should be used only for second-harmonic output. A flash-light bulb may be inserted at the point marked *X* (§ 4-3). The L/C ratio in the plate tank, L_2C_2, should be such that the capacity in use is 75 to 100 μμfd. for fundamental output and about 25 μμfd. for second-harmonic output.

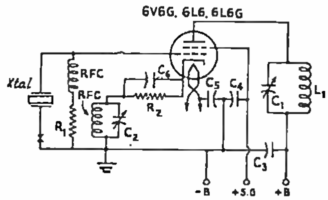


Fig. 408 — Grid-plate crystal oscillator circuit. In the cathode circuit, *RFC* is a 2.5-mh. r.f. choke. Other constants are the same as in Fig. 407. A crystal-current indicator may be inserted at the point marked *X* (§ 4-3).

rent, indicating that the plate circuit is in resonance. The crystal should be oscillating continuously, regardless of the setting of the plate condenser. Set the plate condenser so that plate current is minimum. The load circuit may then be coupled and adjusted so that the oscillator delivers power. The minimum plate current will rise; it may be necessary to retune the plate condenser when the load is coupled to bring the plate current to a new minimum. Fig. 409 shows the typical behavior of plate current with plate-condenser tuning.

After the plate circuit is adjusted and the oscillator is delivering power, the cathode condenser should be readjusted to obtain optimum power output. The setting should be as far toward the low-capacity end of the scale as is consistent with good output; it may, in fact, be desirable to sacrifice a little output if so doing lowers the current through the crystal and thus reduces heating.

For harmonic output the plate tank circuit is tuned to the harmonic instead of the fundamental of the crystal frequency. A plate-current dip will occur at the harmonic. If the cathode condenser is adjusted for maximum output at the harmonic, this adjustment will usually serve for the fundamental as well. The crystal should be checked for excessive heating, the most effective remedy being to lower plate and/or screen voltage or to reduce the loading. Maximum r.f. voltage across the crystal is developed at maximum load, so heating should be checked with the load coupled.

When a fixed cathode condenser is used in the grid-plate oscillator the plate tank circuit is simply resonated, as indicated by the plate-current dip, to the fundamental or a harmonic of the output frequency, loading being adjusted to give optimum power output. If the variable cathode condenser is used, it should be set to give, by observation, the maximum power output consistent with safe crystal current. The variable condenser is useful chiefly in increasing the output on the third and higher harmonics; for fundamental operation, the cathode capacity is not critical and the fixed condenser may be used.

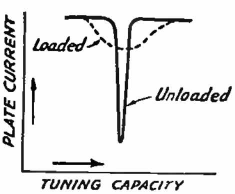


Fig. 409 — Curves showing d.c. plate current vs. plate-condenser tuning, both with and without load, for the Tri-tet oscillator. The setting for minimum plate current may shift with loading.

4-6 Interstage Coupling

Requirements — The purpose of the interstage coupling system is to transfer, with as little energy loss as possible, the power developed in the plate circuit of one tube (the *driver*) to the grid circuit of the following amplifier tube or frequency multiplier. The circuits in practical use are based on the fundamental coupling arrangements described in § 2-11. In the process of power transfer, impedance transformation (§ 2-9) frequently is necessary so that the proper exciting voltage and current will be available at the grid of the driven tube.

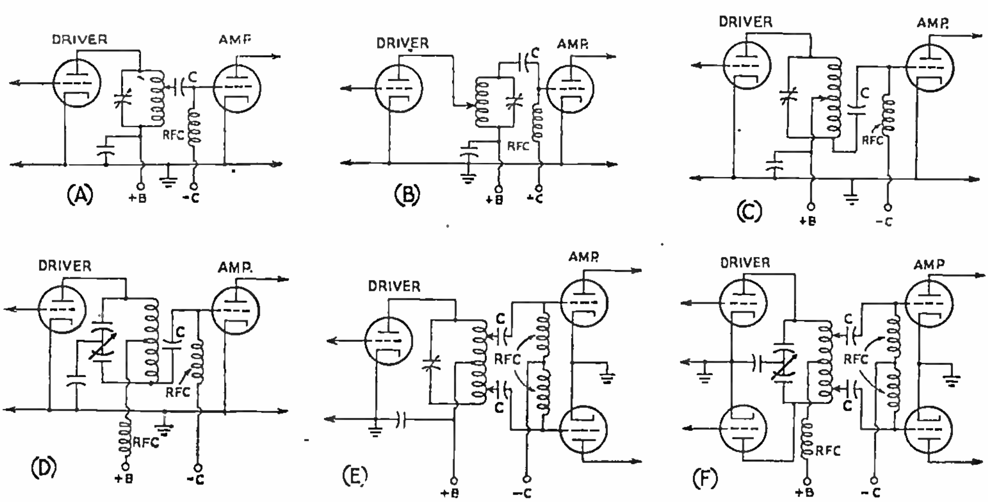


Fig. 410 — Direct- or capacity-coupled driver and amplifier stages. The coupling capacity may be from 50 $\mu\mu$fd. to 0.002 μfd.; it is not critical except where tapping the coils for control of excitation is not possible. Parallel plate feed to the driver and series grid feed to the amplifier may be substituted in any of these circuits (§ 3-7).

Capacity coupling — Fig. 410 shows several types of capacitive coupling. In each case, *C* is the coupling condenser. The coupling condenser serves also as a blocking condenser (§ 2-13) to isolate the d.c. plate voltage of the driver from the grid of the amplifier. The circuits of C and D are preferable when a balanced circuit is used in the output of the driver; instead of both tubes being in parallel across one side, the output capacity of the driver tube and the input capacity of the amplifier are across opposite sides of the tank circuit, thereby preserving a better circuit balance. The circuits of E and F are designed for coupling to a push-pull stage.

In A, B, E and F, excitation is adjusted by moving the tap on the coil to provide an optimum impedance match. In E and F, the two grid taps should be maintained equidistant from the center-tap on the coil.

While capacitive coupling is simplest from the viewpoint of construction, it has certain disadvantages. The input capacity of the amplifier is shunted across at least a portion of the driver tank coil. When added to the output capacity of the driver tube, this additional capacity may be sufficient, in many cases, to prevent use of a desirable L/C ratio in circuits for frequencies above about 7 Mc.

Link coupling — At the higher frequencies it is advantageous in reducing the effects of tube capacities on the L/C ratio to use separate tank circuits for the driver plate and amplifier grid, coupling the two circuits by means of a link (§ 2-11). This method of coupling also has some constructional advantages, in that separate parts of the transmitter may be constructed as separate units without the necessity for running long leads at high r.f. potential.

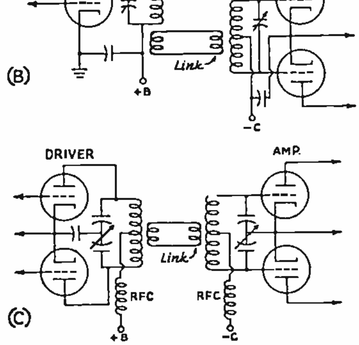


Fig. 411 — Link coupling between driver and amplifier.

Circuits for link coupling are shown in Fig. 411. The coupling ordinarily is by a turn or two of wire closely coupled to the tank inductance at a point of low r.f. potential, such as the center of the coil of a balanced tank circuit or the "ground" end of the coil in a single-ended circuit. The link line usually consists of two closely spaced parallel wires; occasionally the wires are twisted together, but this usually causes undue losses at high frequencies.

It is advisable to have some means of varying the coupling between link and tank coils. The link coil may be arranged to be swung in relation to the tank coil or, when it consists of a large turn around the outside of the tank coil, split into two parts which can be pulled apart or closed somewhat in the fashion of a pair of calipers. If the tank coils are wound on forms, the link may be wound close to the main coil.

With fixed coils, some adjustment of coupling usually can be obtained by varying the number of turns on the link. In general, the proper number of turns for the link must be found by experiment.

¶ 4-7 R.F. Power-Amplifier Circuits

Tetrode and pentode amplifiers — When the input and output circuits of an r.f. amplifier tube are tuned to the same frequency it will oscillate as a tuned-grid tuned-plate oscillator, unless some means is provided to eliminate the effects of feed-back through the plate-to-grid capacity of the tube (§ 3-5). In all transmitting r.f. tetrodes and pentodes, this capacity is reduced to a satisfactory degree by the internal shielding between grid and plate provided by the screen. Tetrodes and pentodes designed for audio use (such as the 6L6, 6V6, 6F6, etc.) are not sufficiently well screened for use as r.f. amplifiers without employing suitable means for nullifying the effect of the grid-plate capacity.

Typical circuits of tetrode and pentode r.f. amplifiers are shown in Fig. 412. The high power sensitivity (§ 3-3) of pentodes and tetrodes, makes them prone to self-oscillate with very small values of feed-back voltage, however, so that particular care must be used to prevent feed-back by means external to the tube itself. This calls for adequate isolation of plate and grid tank circuits to prevent undesired magnetic or capacity coupling between them. The requisite isolation can be secured either by keeping the circuits well separated and mounting the coils so that magnetic coupling is minimized, or by the use of interstage shielding (§ 2-11).

Triode amplifiers — The feed-back through the grid-plate capacity of a triode cannot be eliminated, and therefore special circuit means called *neutralization* must be used to prevent oscillation. A properly neutralized triode amplifier then behaves as though it were operating at very low frequencies, where the grid-plate capacity feed-back is negligible (§ 3-3).

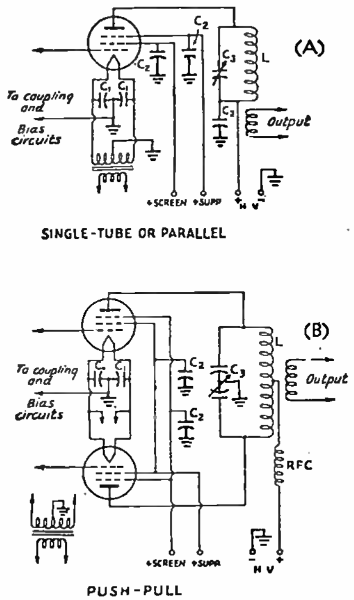


Fig. 412 — Typical tetrode-pentode r.f. amplifier circuits. C_1 — 0.01 μfd. C_2 — 0.001 μfd. C_3-L — See § 4-8. In circuits for tetrodes, the suppressor-grid connection and its associated by-pass condenser are omitted.

Neutralization — Neutralization amounts to taking some of the radio-frequency current from the output or input circuit of the amplifier and introducing it into the other circuit in such a way that it effectively cancels the current flowing through the grid-plate capacity of the tube, thus rendering it impossible for the tube to supply its own excitation. For full neutralization, the two currents must be opposite in phase (§ 2-7) and equal in amplitude.

The out-of-phase current (or voltage) can be obtained quite readily by using a balanced tank circuit for either grid or plate, taking the neutralizing voltage from the end of the tank opposite that to which the grid or plate is connected. The amplitude of the neutralizing voltage can be regulated by means of a small condenser, the *neutralizing condenser*, having the same order of capacity as the grid-plate capacity of the tube. Circuits in which the neutralizing voltage is obtained from a balanced grid tank and fed to the plate through the neutralizing condenser are termed *grid-neutralized* circuits, while if the neutralizing voltage is obtained from a balanced plate tank and fed to the grid of the tube the circuit is *plate-neutralized*.

Plate-neutralized circuits — The circuits for plate neutralization are shown in Fig. 413 at A, B and C. In A, voltage induced in the extension of the tank coil is fed back to the grid through the neutralizing condenser, C_n, to balance the voltage appearing between grid and plate. In this circuit, the capacity required at C_n increases as the tank coil extension is made smaller; in general, neutralization is satisfactory over only a small range of frequencies since the coupling between the two sections of the tank coil will vary with the amount of capacity in use at C.

In B the tank coil is center-tapped to give equal voltages on either side of the center tap, the tank condenser being across the whole coil. The neutralizing capacity is approximately equal to the grid-plate capacity of the tube, in this case. A disadvantage of the circuit, when used with the single tank condenser shown, is that the rotor of the condenser is above ground potential, and hence small capacity changes caused by bringing the hand near the tuning control (*hand capacity*) cause detuning. In general, neutralization is complete at only one

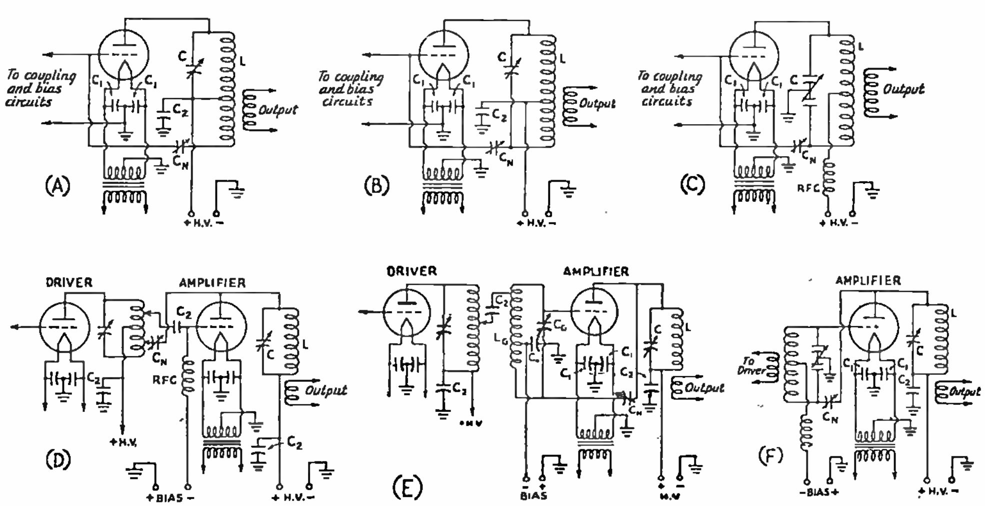


Fig. 413 — Neutralized triode amplifier circuits. Plate neutralization is shown in A, B and C, while D, E and F show types of grid neutralization. Either capacitive or link coupling may be used with the circuits of A, B or C. C-L — See § 4-8. C_g-L_g — Grid tank circuit. C_n — Neutralizing condensers. C_1 — 0.01 μfd. C_2 — 0.001 μfd.

frequency since the plate-cathode capacity of the tube is across only half the tank coil; also, it is difficult to secure an exact center-tap. Both of these factors cause unbalance, which in turn causes the voltages across the two halves of the coil to differ when the frequency is changed.

The circuit of C also uses a center-tapped tank circuit, the voltage division being secured by use of a balanced (split-stator) tank condenser, the two condenser sections being identical. C_n is approximately equal to the grid-plate capacity of the tube. In this circuit the upper section of the tank condenser is in parallel with the output capacity of the tube, hence the circuit can be completely neutralized at only one setting of the tank condenser unless a

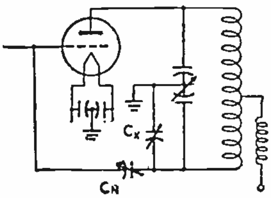


Fig. 414 — Compensating for unbalance in the single-tube neutralizing circuit. C_x, the balancing condenser, has a maximum capacity somewhat larger than the tube output capacity.

compensating capacity (Fig. 414) is connected across the lower section. It is adjusted so the neutralizing condenser need not be changed when frequency is shifted. In practice, if the capacity in use in the tank circuit is large compared to the plate-cathode capacity the unbalancing effect is not serious.

Grid-neutralized circuits — Typical circuits employing grid neutralization are shown in Fig. 413 at D, E and F. The principle of balancing out the feed-back voltage is the same as in plate neutralization. However, in these circuits the feed-back voltage may be either in phase or out of phase with the excitation voltage on the grid side of the input tank circuit (and the opposite on the other side) depending upon whether the tank is divided by means of a balanced condenser or a tapped coil. Circuits such as those at D and E, neutralized by ordinary procedure (described below), will be regenerative when the plate voltage is applied; the circuit at F will be degenerative. In addition the normal unbalancing effects previously described are present, so that grid neutralizing is less satisfactory than the plate method.

Inductive neutralization — With this type of neutralization, inductive coupling between the grid and plate circuits is provided in such a way that the voltage induced in the grid coil by magnetic coupling from the plate coil opposes the voltage fed back through the grid-plate capacity of the tube. A representative circuit arrangement, using a coupling link to provide the mutual inductance (§ 2-11), is shown in Fig. 415-A. The link coils are of one or two turns coupled to the grounded ends of the tank coils. Neutralization is adjusted by moving the link coils in relation to the tank coils. Reversal of connections to one coil may be required for proper phasing. Ordinary inductive coupling between the two coils also could be used, but it is less convenient. Inductive neutralization is complete only at one frequency since the effective mutual inductance changes to some extent with tuning, but is useful in cases where the grid-plate capacity of the tube is very small and suitable circuit balance cannot be obtained by using neutralizing condensers.

Another form of neutralization, known as "coil" or "shunt" neutralization, is shown at B. Its operation is based on making the inductance of L_n such that, together with the grid-plate capacity of the tube, it resonates at the operating frequency. C_2 is merely a plate-voltage blocking condenser. If the Q of the coil is sufficiently high, the parallel resonant impedance between grid and plate is much higher than the grid-cathode circuit impedance. Because the system is difficult to adjust and functions satisfactorily only at one frequency, it is used chiefly in fixed-frequency transmitters. The variation in Fig. 414-C is useful for v.h.f. In this arrangement the coil is replaced by a parallel line, the effective length of which is adjusted until it is resonant when loaded by the grid-plate capacity.

Push-pull neutralization — With push-pull circuits two neutralizing condensers are used, as shown in Fig. 416. In these circuits, the grid-plate capacities of the tubes and the neutralizing capacities form a capacity bridge (§ 2-11) which is independent of the grid and plate tank circuits. The neutralizing capacities are approximately the same as the tube grid-plate capacities. With electrically similar tubes and symmetrical construction (stray capacities to ground equal on both sides of the circuit), the neutralization is complete and independent of frequency. A circuit using a balanced condenser, as at B, is preferred, since it is an aid in obtaining good circuit balance.

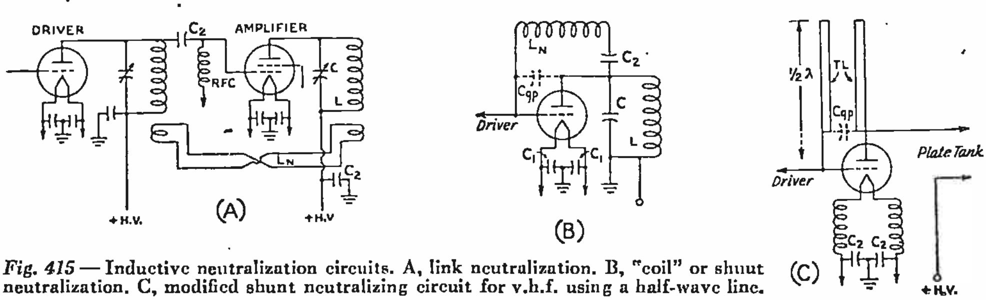


Fig. 415 — Inductive neutralization circuits. A, link neutralization. B, "coil" or shunt neutralization. C, modified shunt neutralizing circuit for v.h.f. using a half-wave line.

Frequency effects — The effects of slight dissymmetry in a neutralized circuit become more important as the frequency is raised, and may be sufficient at the very-high frequencies (or even lower) to prevent good neutralization. At these frequencies the inductances and stray capacities of even short leads become important elements in the circuit, while input loading effects (§ 7-6) may make it impossible to get proper phasing, particularly in single-tube circuits. In such cases the use of a push-pull amplifier, with its general freedom from the effects of dissymmetry, is not only much to be preferred but may be the only type of circuit which can be satisfactorily neutralized.

Neutralizing condensers — In most cases the neutralizing voltage will be equal to the r.f. voltage between the plate and grid of the tube, so that for perfect balance the capacity required in the neutralizing condenser theoretically will be equal to the grid-plate capacity. If, in the circuits having tapped tank coils, the tap is more than half the total number of turns from the plate end of the coil, the required neutralizing capacity will increase approximately in proportion to the relative number of turns in the two sections of the coil.

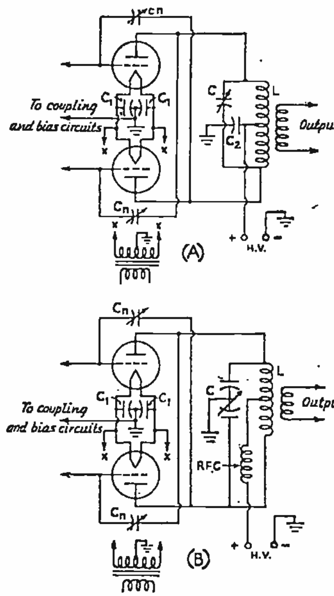


Fig. 416 — "Cross-neutralized" push-pull r.f. amplifier circuits. Either capacitive or link coupling may be used.
C-L — See § 4-8. C_n — Neutralizing condensers.
C_1 — 0.01 μfd. C_2 — 0.001 μfd. or larger.

With tubes having grid and plate connections brought out through the bulb, a condenser having at about half-scale or less a capacity equal to the grid-plate capacity of the tube should be chosen. If the grid and plate leads are brought through a common base the capacity needed is greater, because the tube socket and its associated wiring adds some capacity to the actual interelement capacities. When two or more tubes are connected in parallel, the neutralizing capacity required will be in proportion to the number of tubes.

The voltage rating of neutralizing condensers must at least equal the r.f. voltage across the condenser plus the sum of the d.c. plate voltage and the grid-bias voltage.

Neutralizing procedure — The procedure in neutralizing is essentially the same for all tubes and circuits. The filament of the tube should be lighted and excitation from the preceding stage fed to the grid circuit There should be no plate voltage on the amplifier.

The grid-circuit milliammeter makes a good neutralizing indicator. If the circuit is not completely neutralized, tuning of the plate tank circuit through resonance will change the tuning of the grid circuit and affect its loading, causing a change in the rectified d.c. grid current. The setting of the neutralizing condenser which leaves the grid current unaffected as the plate tank is tuned through resonance is the correct one. If the circuit is out of neutralization, the grid current will drop perceptibly as the plate tank is tuned through resonance. As the point of neutralization is approached, by adjusting the neutralizing capacity in small steps the dip in grid current as the plate condenser is swung through resonance will become less and less pronounced, until, at exact neutralization, there will be no dip at all. Further change of the neutralizing capacity in the same direction will bring the grid-current dip back. The neutralizing condenser should always be adjusted with a screwdriver of insulating material to avoid hand-capacity effects.

Adjustment of the neutralizing condenser may affect the tuning of the grid tank or driver plate tank, so both circuits should be retuned each time a change is made in neutralizing capacity. In neutralizing a push-pull amplifier the neutralizing condensers should be adjusted together, step by step, keeping their capacities as equal as possible.

With single-ended circuits having split-stator neutralizing, the behavior of the grid meter will depend somewhat upon the type of tube used. If the tube output capacity is not great enough to upset the balance, the action of the meter will be the same as in other circuits. With high-capacity tubes, however, the meter usually will show a gradual rise and fall as the plate tank is tuned through resonance, reaching a maximum right at resonance when the circuit is properly neutralized.

When an amplifier is not neutralized a neon bulb touched to the plate of the amplifier tube or to the plate side of the tuning condenser will glow when the tank circuit is tuned through resonance, providing the driver has sufficient power. The glow will disappear when the amplifier is neutralized. However, touching the neon bulb to such an ungrounded point in the circuit may introduce enough stray capacity to unbalance the circuit slightly, thus upsetting the neutralizing.

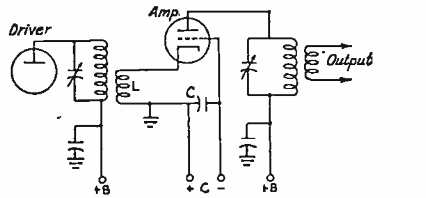


Fig. 417 — Inverted amplifier. The number of turns at *L* should be adjusted by experiment to give optimum grid excitation. By-pass condenser *C* is 0.001 μfd. or larger.

A flashlight bulb connected in series with a single-turn loop of wire 2½ or 3 inches in diameter, with the loop coupled to the tank coil, also will serve as a neutralizing indicator. Capacitive unbalance can be avoided by coupling the loop to the low-potential part of the tank coil.

Incomplete neutralization — If a setting of the neutralizing condenser can be found which gives minimum r.f. current in the plate tank circuit without completely eliminating it, there may be magnetic or capacity coupling between the input and output circuits external to the tube itself. Short leads in neutralizing circuits are highly desirable, and the input and output inductances should be so placed with respect to each other that magnetic coupling is minimized. Usually this requires that the axes of the coils must be at right angles to each other. In some cases it may be necessary to shield the input and output circuits from each other. Magnetic coupling can be detected by disconnecting the plate tank from the remainder of the circuit and testing for r.f. in it (by means of the flashlight lamp and loop) as the tank condenser is tuned through resonance. The driver stage must be operating while this is done, of course.

With single-ended amplifiers there are many stray capacities left uncompensated for in the neutralizing process. With large tubes, especially those having relatively high interelectrode capacities, these commonly neglected stray capacities can prevent perfect neutralization. Symmetrical arrangement of a push-pull stage is about the only way to obtain practically perfect balance throughout the amplifier.

The neutralization of tubes with extremely low grid-plate capacity, such as the 6L6, is often difficult, since it frequently happens that the wiring itself will introduce sufficient capacity between the right points to "over-neutralize" the grid-plate capacity. The use of a neutralizing condenser only aggravates the condition. Inductive or link neutralization, as shown in Fig. 415, has been used successfully with such tubes.

The inverted amplifier — The circuit of Fig. 417 avoids the necessity for neutralization by operating the control grid of the tube at ground potential, thus making it serve as a shield between the input and output circuits. It is particularly useful with tubes of low grid-plate capacity, which are difficult to neutralize by ordinary methods. Excitation is applied between grid and cathode through the coupling coil, *L*; since this coil is common to both the plate and grid circuits the amplifier is degenerative with the circuit constants normally used, hence more excitation voltage and power are required for a given output than is the case with a neutralized amplifier. The tube used must have low plate-cathode capacity (of the order of 1 μμfd. or less) since larger values will give sufficient feed-back to permit it to oscillate, the circuit then becoming the ultraudion (§ 3-7). Tubes having sufficiently low plate-cathode capacity (audio pentodes, for example) can be used without danger of oscillation at frequencies up to 30 Mc. or so.

4-8 Power Amplifier Operation

Efficiency — An r.f. power amplifier is usually operated Class-C (§ 3-4) to obtain a reasonably high value of plate efficiency (§ 3-3). The higher the plate efficiency the higher the power input that can be applied to the tube without exceeding the plate dissipation rating (§ 3-2), up to the limits of other tube ratings (plate voltage and plate current). Plate efficiencies of the order of 75 per cent are readily obtainable at frequencies up to the 30–60-Mc. region. The *over-all* efficiency of the amplifier will be lower by the power lost in the tank and coupling circuits, so that the actual efficiency is less than the plate efficiency.

Operating angle — The operating angle is the proportionate part of the exciting grid-voltage cycle (§ 2-7) during which plate current flows, as shown in Fig. 418. For Class-C operation, it is usually in the vicinity of 120–150 degrees. With other operating considerations, this angle results in an optimum relationship between plate efficiency and grid driving power.

Load impedance — The load impedance (§ 3-3) for an r.f. power amplifier is adjusted, by tuning the plate tank circuit to resonance, to represent a pure resistance at the operating frequency (§ 2-10). Its value, which usually is in the neighborhood of a few thousand ohms, is

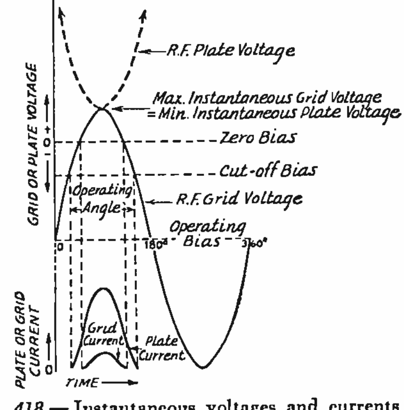


Fig. 418 — Instantaneous voltages and currents in a Class-C amplifier operating under optimum conditions.

adjusted by varying the loading on the tank circuit, closer coupling to the load giving lower values of load resistance and vice versa (§ 2-11). The load may be either the grid circuit of a following stage or the antenna circuit.

For highest efficiency the value of load resistance should be relatively high, but if only limited excitation voltage is available greater power output will be secured by using a lower value of load resistance. The latter adjustment is accompanied by a decrease in plate efficiency. The optimum load resistance is that which, for the maximum permissible peak plate current, causes the minimum instantaneous plate voltage (Fig. 418) to be equal to the maximum instantaneous grid voltage required to cause the peak plate current to flow; this gives the optimum ratio of plate efficiency to required grid driving power.

R.f. grid voltage and grid bias — For most tubes optimum operating conditions result when the minimum instantaneous plate voltage is 10 to 20 per cent of the d.c. plate voltage, so that the maximum instantaneous positive grid voltage must be approximately the same figure. Since plate current starts flowing when the instantaneous voltage reaches the cut-off value (§ 3-2), the d.c. grid voltage must be considerably higher than cut-off to confine the operating angle to 150 degrees or less (with grid bias at cut-off, the angle would be 180 degrees). For an angle of 120 degrees, the r.f. grid voltage must reach 50 per cent of its peak value (§ 2-7) at the cut-off point. The corresponding figure for an angle of 150 degrees is 25 per cent. Hence, the operating bias required is the cut-off value plus 25 to 50 per cent of the peak r.f. grid voltage. These relations are shown in Fig. 418. The grid bias should be at least twice cut-off if the amplifier is to be plate modulated, so that the operating angle will be not less than 180 degrees when the plate voltage rises to twice the steady d.c. value (§ 5-3). Because of their relatively high amplification factors, with most modern tubes Class-C operation requires considerably more than twice cut-off bias to make the operating angle fall in the region mentioned above. Suitable operating conditions are usually given in the data accompanying the type of tube used.

Grid bias may be secured either from a bias source (*fixed bias*), a grid leak (§ 3-6) of suitable value, or from a combination of both. When a bias supply is used, its voltage regulation should be taken into consideration (§ 8-9).

Driving power — As indicated in Fig. 418, grid current flows only during a small portion of the peak of the r.f. grid voltage cycle. The power consumed in the grid circuit therefore is approximately equal to the peak r.f. grid voltage multiplied by the average rectified grid current as read by a d.c. milliammeter. The peak r.f. grid voltage, if not included in the tube manufacturer's operating data, can be estimated roughly by adding 10 to 20 per cent of the plate voltage to the operating grid bias, assuming the operating conditions are as described above.

At frequencies up to 30 Mc. or so, the grid losses are practically entirely those resulting from grid-current flow. At the very-high frequencies, however, dielectric losses in the glass envelope and base materials become appreciable, together with losses caused by transit-time effects (§ 7-6), and may necessitate supplying several times the driving power indicated above. At any frequency, the driving stage should be capable of a power output two to three times the power it is expected the grid circuit of the amplifier will consume. This is necessary because losses in the tank and coupling circuits must also be supplied, and also to provide reasonably good regulation of the r.f. grid voltage. Good voltage regulation (see § 8-1 for general definition) insures that the waveform of the excitation voltage will not be distorted because of the changing load on the driver during the r.f. cycle.

Grid impedance — During most of the r.f. grid-voltage cycle no grid current is flowing, as

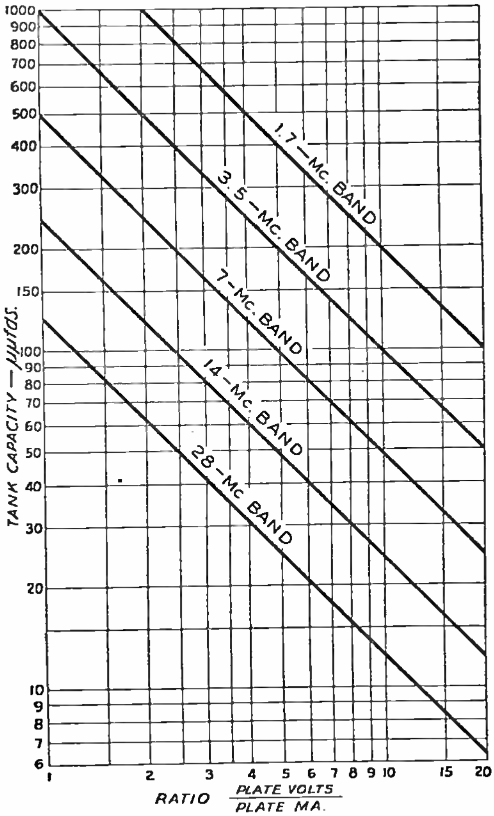


Fig. 419 — Chart showing tank capacities required for a Q of 12 with various ratios of plate voltage to plate current, for various frequencies. In circuits F, G, H (Fig. 420), the capacities shown in the graph may be divided by four. In circuits C, D, E, I, J and K, the capacity of each section of the split-stator condenser may be one-half that shown by the graph. The values given by the graph should be used for circuits A and B.

indicated in Fig. 418, hence the grid impedance is infinite. During the peak of the cycle, however, the impedance may drop to very low values (of the order of 1000 ohms), depending upon the type of tube. Both the minimum and average values of grid impedance depend to a considerable extent on the amplification factor of the tube, being lower with tubes having large amplification factors.

The average grid impedance is equal to E^2/P, where E is the r.m.s. (§ 2-7) value of r.f. grid voltage and P is the grid driving power. Under optimum operating conditions, values of average grid impedance ranging from 2000 ohms for high-μ tubes to four or five times as much for low-μ types are representative. Values in the vicinity of 4000 to 5000 ohms are typical of modern triodes with amplification factors of 20 to 30.

Because of the large change in impedance during the cycle, it is necessary that the tank circuit associated with the amplifier grid have fairly high Q. This is essential to provide sufficient storage capacity so that the voltage regulation over the cycle will be good. The requisite Q may be obtained by adjusting the L/C ratio or by tapping the grid circuit across only part of the tank (§ 4-6).

Tank-circuit Q — Besides serving as a means for transforming the actual load resistance to the required value of plate load impedance for the tube, the plate tank circuit also should suppress the harmonics present in the tube output as a result of the non-sinusoidal plate current (§ 2-7, 3-3). For satisfactory harmonic suppression, a Q of 12 or more (with the circuit fully loaded) is desirable. A Q of this order also is helpful from the standpoint of securing adequate coupling to the load or antenna circuit (§ 2-11). The proper Q can be obtained by suitable selection of L/C ratio in relation to the optimum plate load resistance for the tube (§ 2-10).

For a Class-C amplifier operated under optimum conditions as described above, the plate load impedance is approximately proportional to the ratio of d.c. plate voltage to d.c. plate current. For a given effective Q the tank capacity required at a given frequency will be inversely proportional to the parallel resistance (§ 2-10), so that it will also be inversely proportional to the plate-voltage/plate-current ratio.

The tank capacity required on various amateur bands for a Q of 12 is shown in Fig. 419 as a function of this ratio. The capacity given is for single-ended tank circuits, as shown in Fig. 420 at A and B. When a balanced tank circuit is used the total tank capacity required is reduced to one-fourth this value, because the tube is connected across only half the circuit (§ 2-9). Thus, if the plate-voltage/plate-current ratio calls for a capacity of 200 $\mu\mu$fd. in a single-ended circuit at the desired frequency, only 50 $\mu\mu$fd. would be needed in a balanced circuit. If a split-stator or balanced tank condenser is used each section should have a capacity of 100 $\mu\mu$fd., the total capacity of the two in series being 50 $\mu\mu$fd. These are "in use" capacities; not simply the rated maximum capacity of the condenser. Larger values may be used with an increase in the effective Q.

To reduce energy loss in the tank circuit, the inherent Q of the coil and condenser should be high. Since transmitting coils usually have Qs ranging from 100 to several hundred, the tank transfer efficiency generally is 90 per cent or more. An unduly large C/L ratio is not advisable since it will result in large circulating r.f. tank current and hence relatively large losses in the tank, with a consequent reduction in the power available for the load.

Tank constants — When the capacity necessary for a Q of 12 has been determined from Fig. 419, the inductance required to resonate at the given frequency can be found by means

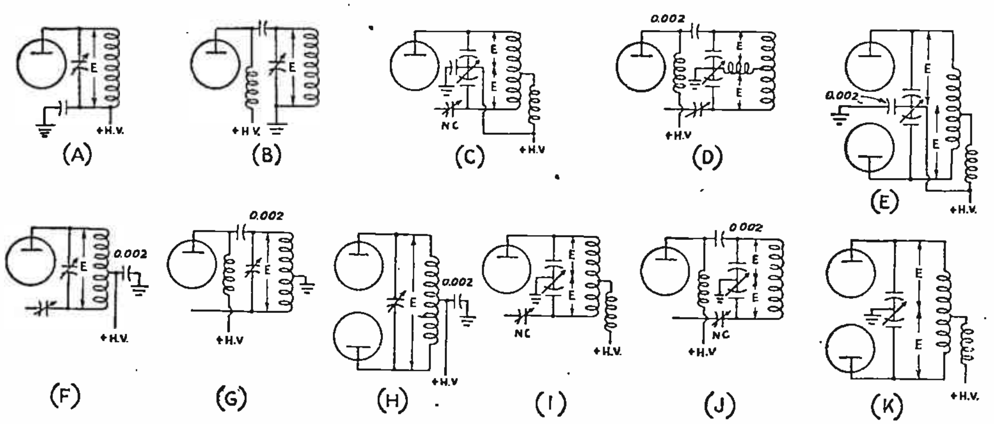


Fig. 420 — In circuits A, B, C, D and E, the peak voltage E will be approximately equal to the d.c. plate voltage applied for c.w. or twice this value for 'phone. In circuits F, G, H, I, J and K, E will be twice the d.c. plate voltage for c.w. or four times the plate voltage for 'phone. The circuit is assumed to be fully loaded. Tubes in parallel in any of the circuits will not affect the peak voltage. Circuits A, C, E, F, G and H require that the tank condenser be insulated from chassis or ground and that it be provided with a suitably insulated shaft coupling for tuning.

of the formula in § 2-10. Alternatively, the required number of turns on coils of various construction can be found from the charts of Figs. 421 and 422.

Fig. 421 is for coils wound on receiving-type forms having a diameter of 1½ inches and ceramic forms having a diameter of 1¾ inches and winding length of 3 inches. Such coils would be suitable for oscillator and buffer stages where the power is not over 50 watts. In all cases, the number of turns given must be wound to fit the length indicated and the turns should be evenly spaced.

Fig. 422 gives data on coils wound on transmitting-type ceramic forms. In the case of the smallest form, extra curves are given for double spacing (winding turns in alternate grooves). This is sometimes advisable in the case of 14- and 28-Mc. coils when only a few turns are required. In all other cases, the specified number of turns should be wound in the grooves without any additional spacing.

Ratings of components — The peak voltage to be expected between the plates of a tank condenser depends upon the arrangement of the tank circuit as well as the d.c. plate voltage. Peak voltage may be determined from Fig. 420, which shows all of the commonly used tank-circuit arrangements. These estimates assume that the amplifier is fully loaded; the voltage will rise considerably should the amplifier be

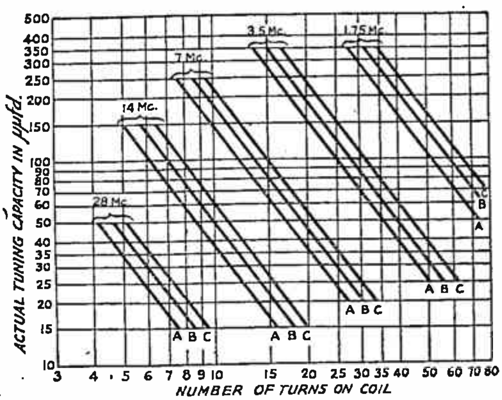


Fig. 421 — Coil-winding data for receiving-type forms, diameter 1½ inches. Curve A — winding length, 1 inch; Curve B — winding length, 1½ inches; Curve C — winding length, 2 inches. Curve C is also suitable for coils wound on 1¾-inch diameter transmitting-type ceramic forms with 3 inches of winding length.

operated without load. The figures include a reasonable factor of safety.

The condenser plate spacing required to withstand any particular voltage will vary with the construction. Most manufacturers specify peak-voltage ratings in describing their condensers.

Plate or screen by-pass condensers of 0.001 μfd. should be satisfactory for frequencies as low as 1.7 Mc. Cathode-resistor and filament by-passes in r.f. circuits should be not less than 0.01 μfd. Fixed condensers used for these pur-

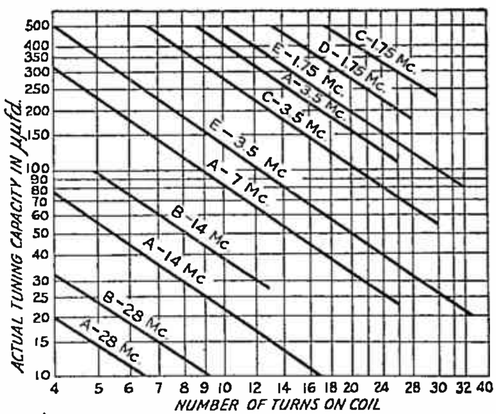


Fig. 422 — Coil-winding data for ceramic transmitting-type forms. Curve A — ceramic form 2½-inch effective diameter, 26 grooves, 7 per inch; Curve B — same as A, but with turns wound in alternate grooves; Curve C — ceramic form 2⅞-inch effective diameter, 32 grooves, 7.1 turns per inch, approximately; Curve D — ceramic form 4-inch effective diameter, 28 grooves, 5.85 turns per inch, approximately; Curve E — ceramic form 5-inch effective diameter, 26 grooves, 7 per inch. Coils may be wound with either No. 12 or No. 14 wire.

poses should have voltage ratings 25 to 50 per cent greater than the maximum d.c. or a.c. voltage across them.

Interstage coupling condensers should have voltage ratings 50 to 100 per cent greater than the sum of the driver plate and amplifier grid-biasing voltages.

4-9 Adjustment of Power Amplifiers

Excitation — The effectiveness of adjustments to the coupling between the driver plate and amplifier grid circuits can be gauged by the relative values of amplifier rectified grid current and driver plate current, the object being to obtain maximum grid current with minimum driver loading. The amplifier grid circuit represents the load on the driver stage, and the average grid impedance must therefore be transformed to the value for optimum driver operation (§ 4-8).

With capacity coupling, either the driver plate or amplifier grid must be tapped down on the driver tank coil, as shown in Fig. 410 at A and B, unless the grid impedance is approximately the right value for the driver plate load, when it will be satisfactory to connect both elements to the end of the tank. If the grid impedance is lower than the required driver plate load, Fig. 410-A is used; if higher, Fig. 410-B. In either case, the coupling which gives the desired grid current with minimum driver loading should be determined experimentally by moving the tap. Should both plate and grid be connected to the end of the circuit it is sometimes possible to control the loading, when the grid impedance is low, by varying the capacity of the coupling condenser, *C*, but this method is not altogether satisfactory since it is simply an expedient to prevent driver overloading without giving suitable impedance matching.

In push-pull circuits the method of adjustment is similar, except that the taps should be kept symmetrically located with respect to the center of the tank circuit.

With link coupling, Fig. 411, the object of adjustment is the same. The two tanks are first tuned to resonance, as indicated by maximum grid current, and the coupling adjusted by means of the links (§ 4-6) to give maximum grid current with minimum driver plate current. This usually will suffice to load the driver to its rated output, provided the driver plate and amplifier grid tank circuits have reasonable values of *Q*. If the *Q* of one or both of the circuits is too low, it may not be possible to load the driver fully with any adjustment of link turns or coupling at either tank. In such a case, the *Q*s of the tank circuits must be increased to the point where adequate coupling is secured. If the driver plate tank is designed to have a *Q* of 12, the difficulty almost invariably is in the amplifier grid tank. The *Q* can be increased to a suitable value either by adjustment of the *L/C* ratio or by tapping the load across part of the coil (§ 2-10).

Whatever the type of coupling, a preliminary adjustment should be made with the proper bias voltage and/or grid leak, but with the amplifier plate voltage off; then the amplifier should be carefully neutralized. After neutralization the driver-amplifier coupling should be readjusted for optimum power transfer, after which plate voltage may be applied and the amplifier plate circuit adjusted to resonance and coupled to its load. Under actual operating conditions the grid current decreases below the value obtained without plate voltage on the amplifier and the effective grid impedance rises, hence the final adjustment is to re-check the coupling to take care of this shift.

With recommended bias, the grid current obtained before plate voltage is applied to the amplifier should be 25 to 30 per cent higher than the value required for operating conditions. If this value is not obtained, and the driver plate input is up to rated value, the reason may be either improper matching of the amplifier grid to the driver plate or simply insufficient power output from the driver to take care of all losses. Driver operating voltages should be checked to assure they are up to rated values. If batteries are used for bias and are not strictly fresh, they should be replaced, since batteries which have been in use for some time often develop high internal resistance which effectively acts as additional grid-leak resistance. If a rectified a.c. bias supply is used, the bleeder or voltage-divider resistances should be checked to make certain that low grid current is not caused by greater grid-circuit resistance than is recommended. In this connection it is helpful to measure the actual bias when grid current is flowing, by means of a high-resistance d.c. voltmeter. There is also the possibility of loss of filament emission of the amplifier tube, either from prolonged service or from operating the filament under or over the rated voltage.

Plate tuning — In preliminary tuning, it is desirable to use low plate voltage to avoid possible damage to the tube. With excitation and plate voltage applied, rotate the plate tank condenser until the plate current dips. Then set the condenser at the minimum plate-current point (resonance). When the resonance point has been found, the plate voltage may be increased to its normal value.

With adequate excitation, the off-resonance plate current of a triode amplifier may be two or more times the normal operating value. With screen-grid tubes the off-resonance plate current may not be much higher than the normal operating value, since the plate current is principally determined by the screen rather than the plate voltage.

Under reasonably efficient operating conditions the minimum plate current with the amplifier unloaded will be a small fraction of the rated plate current for the tube (usually a fifth or less), since with no load the parallel impedance of the tank circuit is high. If the excitation is low the "dip" will not be very marked, but with adequate excitation the plate current at resonance without loading will be just high enough so that the d.c. plate power input supplies all the losses in the tube and circuit. As an indication of probable efficiency, the minimum plate current value should not be taken too seriously, because without load the *Q* of the circuit is high and the tank current relatively large. When the amplifier is delivering power to a load, the circulating current drops considerably and the tank losses correspondingly decrease. High minimum unloaded plate current is chiefly encountered at 28 Mc. and above, where tank losses are higher and the tank *L/C* ratio is usually lower than normal because of irreducible tube capacities. The effect is particularly noticeable with screen-grid tubes, which have relatively high output capacity. Because of the decrease in tank r.f. current with loading, however, the actual efficiency under load is reasonably good.

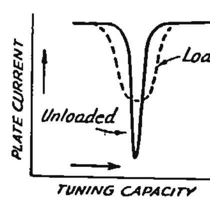


Fig. 423 — Typical behavior of d.c. plate current vs. tuning capacity in the plate circuit of an amplifier.

With the load (antenna or following amplifier grid circuit) connected, the coupling between plate tank and load should be adjusted to make the tube take rated plate current, keeping the tank always tuned to resonance. As the output coupling is increased the minimum plate current also will increase, about as shown in Fig. 423. Simultaneously the tuning becomes less sharp, because of the increase in effective resistance of the tank. If the load circuit simulates a resistance, the resonance setting of the tank condenser will be practically unchanged with loading; this is generally the case, since the load circuit usually is also tuned to resonance. A reactive load (such as an antenna or feeder system not tuned exactly to resonance) may cause the tank condenser setting to change with loading, since reactance as well as resistance is coupled into the tank (§ 2-11).

Power output — As a check on the operation of an amplifier, its power output may be measured by the use of a load of known resistance, coupled to the amplifier output as shown in Fig. 424. At A a thermoammeter, M, and a noninductive (ordinary wire-wound resistors are not satisfactory) resistance, R, are connected across a coil of a few turns coupled to the amplifier tank coil. The higher the resistance of R, the greater the number of turns required in the coupling coil. A resistor used in this way is generally called a "dummy antenna," since its use permits the transmitter to be adjusted without actually radiating power. The loading may readily be adjusted by varying the coupling between the two coils, so that the amplifier draws rated plate current when tuned to resonance. The power output is then calculated from Ohm's Law:

$$P \text{ (watts)} = I^2R$$

where I is the current indicated by the thermoammeter and R is the resistance of the noninductive resistor. Special resistance units are available for this purpose, ranging from 73 to 600 ohms (simulating antenna and transmission-line impedances) at power ratings up to 100 watts. For higher powers, the units may be connected in series-parallel. The meter scale required for any expected value of power output may also be determined from Ohm's Law:

$$I = \sqrt{\frac{P}{R}}$$

Incandescent light bulbs can be used to replace the special resistor and thermoammeter. The lamp should be equipped with a pair of leads, preferably soldered to the terminals on the lamp base. The coupling should be varied until the greatest brilliance is obtained for a given plate input. In using lamps as dummy antennas a size corresponding to the expected power output should be selected, so that the lamp will operate near its normal brilliancy. Then, when the adjustments have been completed, an approximation of the power output can be obtained by comparing the brightness of the lamp with the brightness of one of similar power rating in a 115-volt socket.

The circuit of Fig. 424-B is for resistors or lamps of relatively high resistance. In using this circuit, care should be taken to avoid accidental contact with the plate tank when the power is on. This danger is avoided by circuit C, in which a separate tank circuit, LC, tuned to the operating frequency, is coupled to the plate tank circuit. The loading is adjusted by varying the number of turns across which the dummy antenna is connected on L and by changing the coupling between the two coils. With push-pull amplifiers, the dummy antenna should be tapped equally on either side of the center of the tank when the circuit of Fig. 424-B is used.

Harmonic suppression — The most important step in the elimination of harmonic radiation (§ 4-8 2-12) is to use an output tank circuit having a Q of 12 or more. Beyond this it is desirable to avoid any considerable amount of over-excitation of a Class-C amplifier, since excitation in excess of that required for normal Class-C operation further distorts the plate-current pulse and increases the harmonic content in the output of the amplifier even though the proper tank Q is used. If the antenna system in use will accept harmonic frequencies they will be radiated when distortion is present, and consequently the antenna coupling system preferably should be selected with harmonic transfer in mind (§ 10-6).

Harmonic content can be reduced to some extent by preventing distortion of the r.f. grid-voltage waveshape. This can be done by using a grid tank circuit with high effective Q. Link coupling between the driver and final amplifier are helpful, since the two tank circuits provide more attenuation than one at the harmonic frequencies. However, the advantages of link coupling in this respect may be nullified unless the Q of the grid tank is high enough to give good voltage regulation, which minimizes harmonic transfer and thus prevents distortion in the grid circuit.

The stray capacity between the antenna coupling coil and the tank coil may be sufficient to couple harmonic energy into the antenna system. This coupling may be eliminated by the use of electrostatic shielding (*Faraday shield*) between the two coils. Fig. 425 shows the construction of such a shield, while Fig. 426 illustrates the manner in which it is installed. The construction shown in Fig. 425 prevents current flow in the shield, which would occur if the wires formed closed circuits since the shield is in the magnetic field of the tank coil.

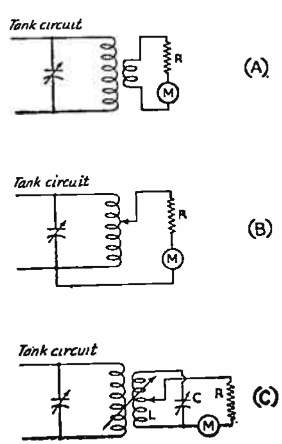


Fig. 424 — "Dummy antenna" circuits for checking power output and making operating adjustments under load without applying power to the actual antenna.

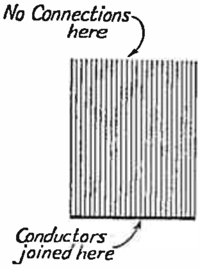


Fig. 425 — The Faraday electrostatic shield for eliminating capacitive transfer of harmonic energy. It is made of parallel conductors, insulated from each other except at one end where all are joined. Stiff wire or small diameter rod may be used, spaced about the diameter of the wire or rod. The shield should be larger than the diameter of the coil.

Should this occur, there would be magnetic shielding as well as electrostatic; in addition, there would be a power loss in the shield.

Improper operation — Inexact neutralization or stray coupling between plate and grid circuits may result in regeneration. This effect is most evident with low excitation, when the amplifier will show a sudden increase in output when the plate tank circuit is tuned slightly to the high-frequency side of resonance. It is accompanied by a pronounced increase in grid current.

Self-oscillation is apt to occur with tubes of high power sensitivity, such as the r.f. pentodes and tetrodes. In event of either regeneration or oscillation, circuit components should be arranged so that those in the plate circuit are well isolated from those of the grid circuit. Plate and grid leads should be made as short as possible and the screen should be by-passed as close to the socket terminal as possible. A cylindrical shield surrounding the lower portion of the tube up to the lower edge of the plate is sometimes required.

"Double resonance," or two tuning spots on the plate-tank condenser, one giving minimum plate current and the other maximum power output, may occur when the tank circuit Q is too low (§ 2-10). A similar effect also occurs at times with screen-grid amplifiers when the screen-voltage regulation (§ 8-1) is poor, as when the screen is supplied through a dropping resistor. The screen voltage decreases with an increase in plate current, because the screen current increases under the same conditions. Thus the minimum plate-current point causes the screen voltage, and hence the power output, to be less than when a slightly higher plate current is drawn.

A phenomenon known as "grid emission" may occur when the amplifier tube is operated at higher than rated power dissipation on either the plate or grid. It is particularly likely to occur with tubes having oxide-coated cathodes, such as the indirectly heated types. It is caused by the grid reaching a temperature high enough to cause electron emission (§ 2-4). The electrons so emitted are attracted to the plate, further increasing the power input and heating, so that grid emission is characterized by gradually increasing plate current and heat which eventually will ruin the tube if the power is not removed. Grid emission can be prevented by operating the tube within its ratings.

4-10 Parasitic Oscillations

Description — If the circuit conditions in an oscillator or amplifier are such that self-oscillation exists at some frequency other than that desired, the spurious oscillation is termed *parasitic*. The energy required to maintain a parasitic oscillation is wasted insofar as useful output is concerned, hence an oscillator or amplifier having parasitics will operate at reduced efficiency. In addition, its behavior at the operating frequency often will be erratic. Parasitic oscillations may be either higher or lower in frequency than the operating frequency.

The parasitic oscillation usually starts the instant plate voltage is applied, or, when the amplifier is biased beyond cut-off, at the instant excitation is applied. In the latter case, the oscillation frequently will be self-sustaining after the excitation has been removed. At other times the oscillation may not be self-sustaining, becoming active only in the presence of excitation. It may be apparent only by the production of abnormal key clicks (§ 6-1) over a wide frequency range, or by the presence of spurious side-bands (§ 5-2) with 'phone modulation.

Low-frequency parasitics — Parasitic oscillations at low frequencies (usually 500 kc. or less) are of the tuned-plate tuned-grid type, the tuned circuits being formed by r.f. chokes and associated by-pass and coupling condensers, with the regular tank tuning condensers having only a minor effect on the oscillation. The operating-frequency tank coil has negligible inductance for such low frequencies and may be short-circuited without affecting the oscillations. The oscillations do not occur when no r.f. chokes are used, hence whenever possible in series-fed circuits such chokes should be omitted. With single-ended amplifiers, it is usually possible to arrange the circuit so that either the grid or plate circuit needs no choke. In push-pull stages, where chokes must be used in both plate and grid circuits, it is helpful to connect an unby-passed grid leak from the choke to the bias supply or ground, thus placing the resistance in the parasitic circuit and tending to prevent oscillation. When the driver plate circuit has parallel feed and the amplifier grid circuit series feed (§ 3-7) this type of oscillation cannot occur if no choke is used in the series grid circuit, since the grid is grounded through the tank coil for the parasitic frequency.

Parasitics near operating frequency — In circuits utilizing a tap on the plate tank coil to establish a ground for a balanced neutralizing circuit, such as Fig. 413-B, a parasitic oscillation may be set up if the amplifier grid is tapped down on the grid (or driver plate) tank circuit for adjustment of driver-amplifier coupling (§ 4-6). In this case the turns between grid and ground and between plate and ground form, with the stray and other capacities present, a t.p.t.g. circuit (§ 3-7) which oscillates at a frequency somewhat higher than the nominal operating frequency. Such an oscillation can be prevented by dispensing with the taps in either the plate or grid circuit. Balancing the plate circuit by means of a split-stator condenser (Fig. 413-C) is recommended.

Very-high-frequency parasitics — Parasitics in the v.h.f. region are likely to occur with any amplifier having a balanced tank circuit, particularly when associated with neutralizing connections. The parasitic resonant circuit, formed by the leads connecting the various components, may be of either the t.p.t.g. or the ultraudion type.

The frequency of such oscillations may be determined by connecting a tuned circuit in series with the grid lead to the tube. A variable condenser (50 or 100 μμfd.) may be used, in conjunction with three or four self-supporting turns of heavy wire wound into a coil an inch or so in diameter. With the amplifier oscillating at the parasitic frequency, the condenser is slowly tuned through its range until oscillations cease. If this point is not found on the first trial, the turns of the coil may be spread apart or a turn removed and the process repeated. The use of such a tuned circuit as a trap is an almost certain remedy if the frequency can be determined, and introduces little if any loss at the operating frequency.

An alternative cure, which is feasible when the oscillation is of the t.p.t.g. type, is to detune the parasitic circuit in either the plate or grid circuit. Since this type of oscillation occurs most frequently with push-pull amplifiers, it may often be cured by making the grid and plate leads to their respective tank circuits of considerably different length. Similar considerations apply to neutralizing connections in push-pull circuits. The extra wire length may be coiled up in the form of a so-called "choke," which in this case is simply additional inductance for detuning the parasitic circuit.

Testing for parasitic oscillations — An amplifier always should be tested for parasitic oscillations before being considered ready for service. The preferable method is first to neutralize the amplifier, then apply sufficient fixed bias to permit a moderate value of plate current to flow without excitation. (The plate current should not be large enough to cause the power input to exceed the rated plate dissipation of the tube.) If the amplifier is free from self-starting parasitics, the plate current will remain steady as the tank condensers are varied; also, there will be no grid current and a neon bulb touched either to the plate or grid will show no glow. Extreme care must be

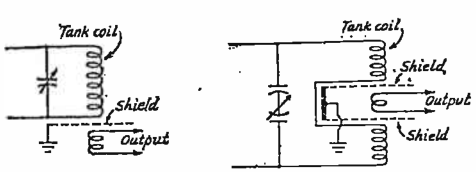


Fig. 426 — Methods of using Faraday shields. Two are required with a push-pull or balanced tank circuit.

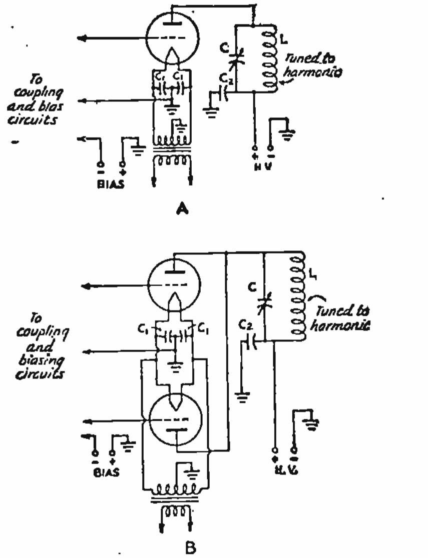


Fig. 427 — Frequency-multiplying circuits. A is for triodes, used either singly or in parallel. The push-push doubler is shown at B. Any type of coupling may be used between the grid circuit and the driver. C_1 should be 0.01 μfd. or larger; C_2, 0.001 μfd. or larger.

used not to let the hand come into contact with any metal parts of the transmitter when using the neon bulb.

If any of these effects are present, the frequency of the parasitic must first be determined. If r.f. chokes are used in both the plate and grid circuits, one of them should be short-circuited to determine if the oscillation is at a low frequency; if so, it may be eliminated by the methods outlined above. If the test indicates that the parasitic is not a low-frequency oscillation, the grid trap described above should be tried for the v.h.f. type. The type which occurs near the operating frequency will not exist unless the plate and grid tank coils are both tapped, hence may be eliminated from consideration if this is not the case in the circuit used. When such an oscillation is present its existence can be detected by moving the grid tap to include the whole tank circuit, whereupon the oscillation will cease.

Some indication of the frequency of the parasitic can be obtained from the color of the glow in the neon bulb. Usually it will be yellowish with low-frequency oscillations and violet with v.h.f. oscillations.

If the amplifier is stable under the conditions described above, excitation should be applied and then removed to ascertain if a self-sustaining oscillation is set up with excitation. If the plate current does not return to the previous value when the excitation is cut off, the same tests should be applied to determine the parasitic frequency.

As a final test, the transmitter should be put on the air and a near-by receiver tuned over as wide a frequency range as possible, to locate any off-frequency signals associated with the radiation. Parasitics usually can be recognized by their poor stability as contrasted to the normal harmonics of the signal, which will have the same stability as the fundamental signal as well as the usual harmonic relationship. Harmonics should be quite weak compared to the output at the fundamental frequency, whereas parasitic oscillations may have considerable strength.

4-11 Frequency Multiplication

Circuits — A frequency multiplier is an amplifier having its plate tank circuit tuned to a multiple (harmonic) of the frequency applied to its grid. The difference between a straight amplifier (§ 4-1) and a frequency multiplier is in the way in which it is operated, rather than in the circuit. However, since the grid and plate tank circuits are tuned to different frequencies a triode frequency multiplier will not self-oscillate, hence does not need neutralization. A typical circuit arrangement is shown in Fig. 427-A. For screen-grid multipliers, the circuit is the same as in Fig. 412-A. Under usual conditions the plate efficiency of a frequency multiplier drops off rapidly with an increase in the number of times the frequency is multiplied. For this reason most multipliers are used as *frequency doublers*, giving second harmonic output.

A special circuit for frequency doubling ("push-push" doubler) is shown in Fig. 427-B. The grids of the tubes are in push-pull and the plates in parallel, thus the plate tank receives two pulses of plate current for each cycle of excitation frequency. The circuit is similar to that of a full-wave rectifier (§ 8-3), where the output ripple frequency is twice the applied frequency.

Push-pull amplifiers are suitable for frequency multiplication at odd harmonics, particularly the third, but they are unsuited to even-harmonic multiplication because the even harmonics are largely balanced out in the push-pull tank circuit (§ 3-3).

Operating conditions and circuit constants — To obtain good efficiency the operating angle at the harmonic frequency must be 180 degrees or less, preferably in the vicinity of 150–120 degrees (§ 4-8). In a doubler, this means that plate current should flow during only half this angle of fundamental frequency. Consequently the r.f. grid voltage, operating bias, and grid driving power must be increased considerably beyond the values obtaining for normal Class-C amplification. For comparable plate efficiency the bias will ordinarily be four to five times the normal Class-C bias, and the r.f. grid voltage must be considerably larger to drive the tube to the same peak plate current. Since the plate and grid current pulses under these conditions have the same peak amplitudes but only half the time duration as in a straight amplifier, the average d.c. values should be one-half those for normal Class-C operation. That is, a tube operated in this way will have the same plate efficiency as a Class-C amplifier but can be operated at only half the plate input, so that the output power also is halved. The driving power required usually is about twice that necessary with straight-through amplification to obtain the same plate efficiency.

Greater output can be secured by using a larger operating angle (lower grid bias) or a lower plate load resistance, to increase the plate current; but this is accompanied by a decrease in efficiency. Since operation of the tube as described in the preceding paragraph is below its maximum plate dissipation rating, the decreased efficiency usually can be tolerated in the interests of securing more power output. In practice, an efficiency of 40 to 50 per cent is about average.

The tank circuit should have reasonably high Q (12 is satisfactory) to give good output voltage regulation (§ 4-9), since a plate-current pulse occurs only once for every two cycles of the output frequency. A low-Q circuit (high L/C ratio) is helpful chiefly when the operating angle is greater than 180 degrees at the second harmonic. Such a tank circuit will have relatively high impedance to the fundamental-frequency component of plate current which is present with large operating angles, and thus will aid in reducing the average d.c. plate current.

The grid impedance of a frequency multiplier is considerably higher than that of a straight-through amplifier, because of the high bias voltage. The average impedance can be calculated as previously described (§ 4-8). The L/C ratio of the grid tank circuit may be higher, therefore, for a given Q. Often it is advantageous to use a fairly high ratio, since a large r.f. voltage must be developed between grid and cathode. However, it must not be made too high (Q too low) to permit adequate coupling between the grid tank circuit and the preceding driver stage.

It may prove necessary to step up the driver output voltage to obtain sufficient r.f. grid voltage for the doubler; this can be done by tapping the driver plate on its tank circuit, when capacity coupling is used, or by similar tapping down or the use of a higher C/L ratio in the driver plate tank when the stages are link-coupled (§ 4-6).

Tubes for frequency multiplication — There is no essential difference between tubes of various characteristics in their performance as frequency doublers. Tubes having high amplification factors will require somewhat less bias for equivalent operation but the grid driving power needed is almost independent of the μ, assuming tubes of otherwise similar construction and characteristics. Pentodes and tetrodes will, as in normal amplifier operation, require less driving power than triodes for efficient doubling, although more power will be needed than for straight amplification.

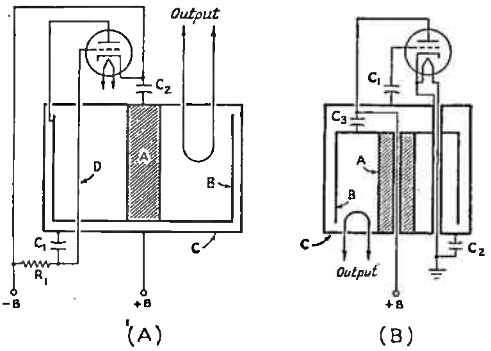


Fig. 428 — High-Q "pot"-type lumped-constant tank circuit as used in v.h.f. oscillators. The tank, shown in cross-section, is made of concentric closed cylinders.

4-12 Very-High-Frequency Oscillators

High-Q circuits with lumped constants — To obtain reasonable high effective Q when a low resistance is connected across the tank circuit, it is necessary to use a high C/L ratio and a tank of inherently high Q (§ 2-10). At low frequencies the inherent Q of any well-designed circuit will be high enough so that it may be neglected in comparison to the effective Q when loaded, so that no special precautions have to be taken with respect to the resistance of coils and condensers. At the very-high frequencies these internal resistances are too large to be ignored, however.

Reduction of the L/C ratio will not increase the effective Q unless the internal resistance of the tank can be made very small. This resistance can be reduced by use of large conducting surfaces and elimination of radiation. In such cases special lumped-constant tank circuits (§ 2-12) are used. The oscillator shown in Fig. 428-A uses a "pot"-type tank in a plate-tickler circuit (§ 3-7), with the feed-back coil in the grid circuit; this inductance is the wire D in the diagram. Output is taken from the tank by means of a hairpin coupling loop.

Fig. 428-B corresponds to the shunt-fed Hartley circuit. Such a tank also may be used in the ultraudion circuit. A variable condenser may be connected across the tank for tuning, although the Q may be reduced if a considerable portion of the tank r.f. current flows through it.

Linear Circuits — A quarter-wave or half-wave line, either of the parallel-conductor open type or of the coaxial type, is equivalent to a resonant circuit (§ 2-12) and can be used as the tank circuit (§ 3-7) in an oscillator.

The resonant line is usually constructed of thin-walled copper tubing to reduce resistance and provide a mechanically stable circuit, particularly at the lower frequencies. At frequencies above 100 Mc. flat copper strip conductor of equivalent cross-section may be used for parallel-line circuits with comparable efficiency. Frequency can be changed by moving a shorting bar or condenser to change the effective line length, or by reducing its length and loading it to resonance by connecting a low-capacity variable condenser across the open end of the line. The added capacity makes it necessary to shorten the line considerably for a given frequency. This, together with the additional loss in the condenser, causes a decrease in Q. These effects will be less if the condenser is connected down on the line. Tapping down also gives greater bandspread effect (§ 7-7).

At frequencies above 150 Mc., an adequate ground connection for the cathode circuit becomes a problem because of the inductance of the cathode lead. Special tubes are available

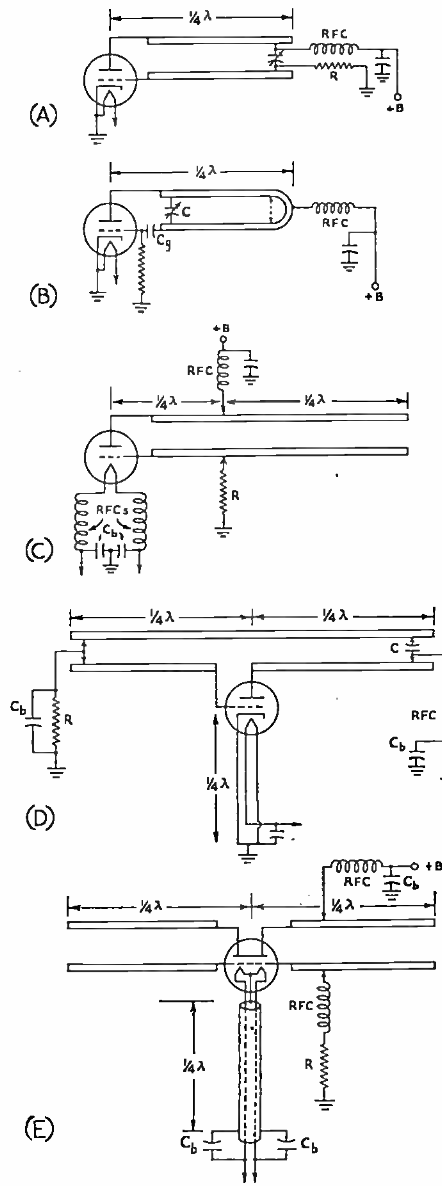


Fig. 429 — Typical single-tube parallel-line oscillators. Constants and applications are discussed in the text.

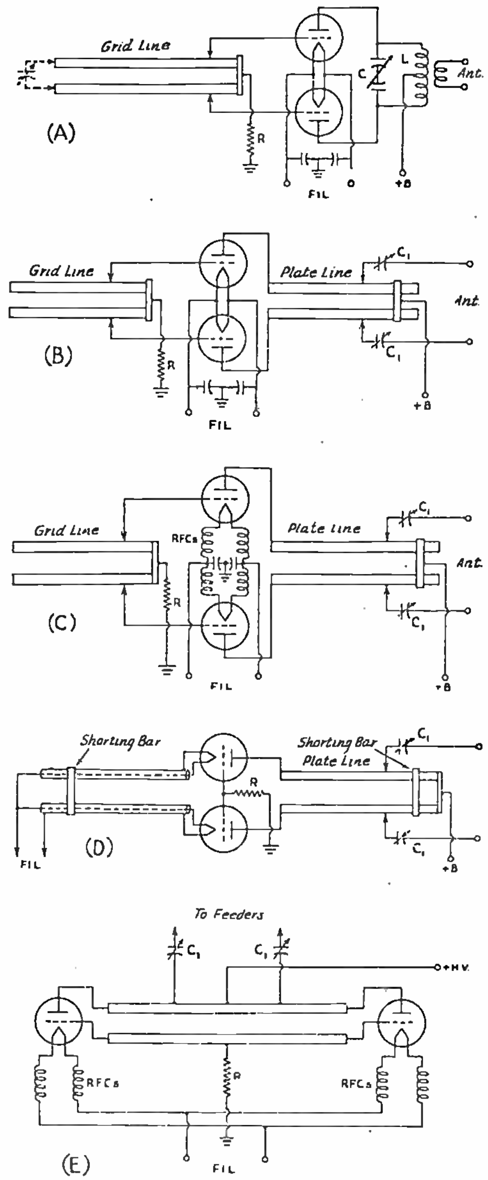


Fig. 430 — Push-pull parallel-line oscillator circuits.

with two or three cathode leads (§ 3-6); connected in parallel, these reduce the effective inductance. With ordinary tubes, r.f. chokes may be inserted in the filament circuit to compensate for the effects of the internal inductance. The effective length of the filament circuit should be one-half wavelength, to bring the cathode filament to the same potential as the shorted ends of the tank lines. The added inductance required must be determined by experiment, the coils being adjusted for optimum stability and power output.

Another method is to use a tuned line in the filament circuit, adjusting its length so that the electrical length of the line plus that of the filament is one-half wavelength. A convenient arrangement is the use of a coaxial (or trough) line with an initial length of about 3/8 wavelength. A shorting disc in the form of a movable plunger equipped with an extension handle may be provided for ease of adjustment. With filament-type tubes one such line will be required for each filament lead. In the case of cathode-type tubes only one line is necessary, the cathode and one side of the filament being connected to the outer conductor and the other filament connection being made by an insulated lead running through a hollow-tubing inner conductor. The return lead should be by-passed where it emerges from the line.

The antenna or other load may be connected through blocking condensers direct to the line (the correct point being determined experimentally). Alternatively, a hair-pin coupling link or, in the case of an oscillator-amplifier system, direct inductive coupling to the grid line of the amplifier may be used.

For highest-frequency operation separate lines must be used for each electrode — grid, plate and cathode. This places all of the interelectrode capacities in series, reducing the loading effect. Still higher frequencies can be reached by using double-lead tubes (§ 3-5), in which case the leads form an integral part of the line and the interelectrode capacities are divided between the two quarter-wave sections.

Parallel-line oscillators — Typical parallel-line oscillator circuits are shown in Fig. 429. In A, a shorting condenser (which may be either a fixed blocking condenser or a small variable which will provide a limited tuning range) is used to bridge the line at the voltage antinode; the frequency can also be changed by sliding the shorting condenser along the line.

The circuit at B eliminates the need for a blocking condenser at the voltage antinode, where the r.f. current reaches its maximum value. An r.f. choke may be inserted between the grid and the associated grid resistor, *R*. This circuit also can be resonated either by a variable condenser, *C*, or by a sliding bar as indicated by the dashed line.

Fig. 429-C uses a half-wave open-ended line. The grid and plate feed connections are made at nodal points on the line. As indicated on the diagram, these do not occur at the physical center of the line because of the loading effect of the tube. In practice, the position of the taps, as well as the over-all length of the line, are adjusted to obtain maximum grid current. Using this circuit, a 955 acorn or a 9002 can be made to oscillate up to 600 or 700 Mc.

Fig. 429-D is a variation of the above preferable for use with tubes having grid and plate terminals at opposite ends of the envelope. The circuit of Fig. 429-E is most useful with double-lead tubes. To attain high output at the maximum operating frequency, the desirable arrangement is to use two or more double-lead tubes, each in a circuit such as this, with the lines connected end to end.

Push-pull parallel-line oscillators — It is often advantageous to use push-pull oscillator circuits at the very-high frequencies, not only as a means to secure more power output but also for better circuit symmetry.

Fig. 430 shows typical push-pull circuits of this type. Figs. 430-A, -B and -C all employ the same circuit — the t.p.t.g. type (§ 3-7). The grid line is usually operated as the frequency-controlling circuit, since it is not associated with the load and hence its Q can be kept high. The same adjustment considerations apply as in the case of single-tube oscillators. Grid taps in particular should be tapped down as far as possible, to improve the frequency stability.

In Fig. 430-A, a conventional coil-and-condenser tank is used in the plate circuit where the lower Q does not have so great an effect on frequency stability. For maximum efficiency the use of a linear output circuit is desirable at the higher frequencies, however. This is shown at B, and at C with isolating r.f. chokes in the filament circuit.

Fig. 430-B shows a push-pull oscillator having tuned plate and cathode lines, the cathode circuit being tuned with a quarter-wave line which controls excitation and, to some extent, tuning. The grids are connected together and grounded through the grid leak, R_1; ordinarily no by-pass condenser is needed across R_1. This circuit gives good power output at very-high frequencies, but is not especially stable unless the plates are tapped down on the plate tank circuit to avoid too great a reduction in Q. Tapping on the cathode line is not feasible for mechanical reasons. With ordinary tubes this oscillator is capable of higher-frequency operation than the conventional t.g.t.p. type, and it has been found particularly useful on 224 Mc.

The symmetrical circuit at E is preferable above 200 Mc. because it reduces the end effect — foreshortening of the line by virtue of the fact that the electrical termination occurs at the midpoint of the balanced system rather than at the wire ends, causing additional loss. Coaxial or equivalent lines may be used instead of r.f. chokes for ultra high-frequency operation. With this modification, and assuming the use of double-lead tubes, by the addition of additional quarter-wave sections at each end this circuit may be considered equivalent to the center section of a double linear oscillator as discussed in connection with Fig. 429-E.

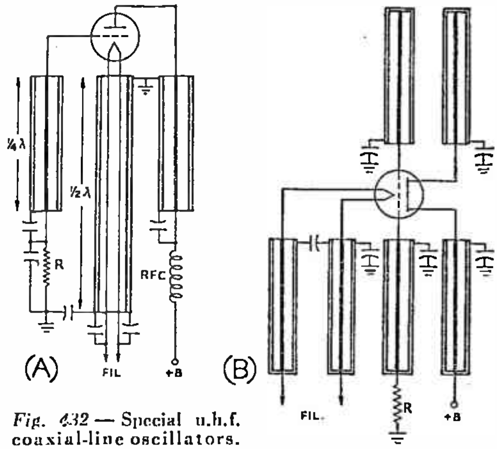


Fig. 432 — Special u.h.f. coaxial-line oscillators.

Coaxial-line circuits — At frequencies in the neighborhood of 300 Mc. the radiation loss (§ 2-12) from open lines greatly reduces the Q, because the conductor spacing comes to represent an appreciable fraction of a wavelength. Consequently, at these frequencies and higher coaxial lines, in which the field is confined inside the line so that radiation is negligible, are used. A further advantage is that the outside of the line is "cold"; that is, no r.f. potentials develop between points on the outer surface. While the coaxial line is also advantageous at lower frequencies, it is more complicated to construct and adjust than parallel lines.

For ease of construction, the coaxial line sometimes is modified into a "trough," in which the cross-section of the outer conductor is in the shape of a square U, one side being left open for tapping and adjustment of the inner conductor. Some radiation takes place with this type of construction, although not so much as with open lines.

The conventional coaxial-line oscillator circuits shown in Fig. 431 illustrate the application of two basic circuits — the Hartley and the t.g.t.p. — to both cathode-type and filamentary tubes. The tube loads the line, as previously described; hence the actual length is always shorter than a quarter wavelength. The length can be adjusted by a short-circuiting sliding plunger, a close-fitting low-resistance contact being necessary to avoid losses. The inner conductor may also have a short tight-

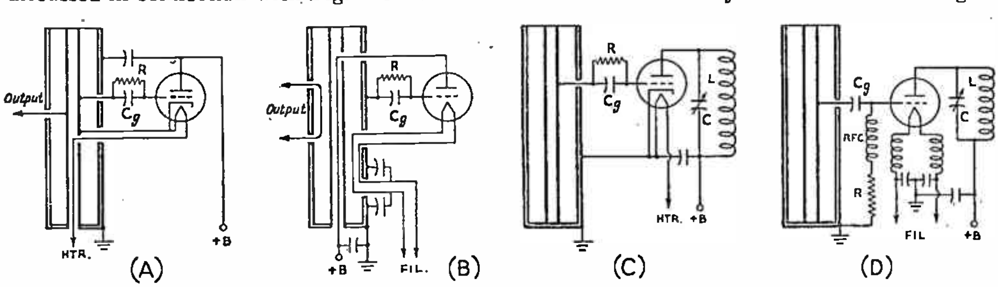


Fig. 431 — **Single-tube v.h.f. coaxial-line oscillators. A and B use Hartley circuits; C and D are t.g.t.p. equivalents.**

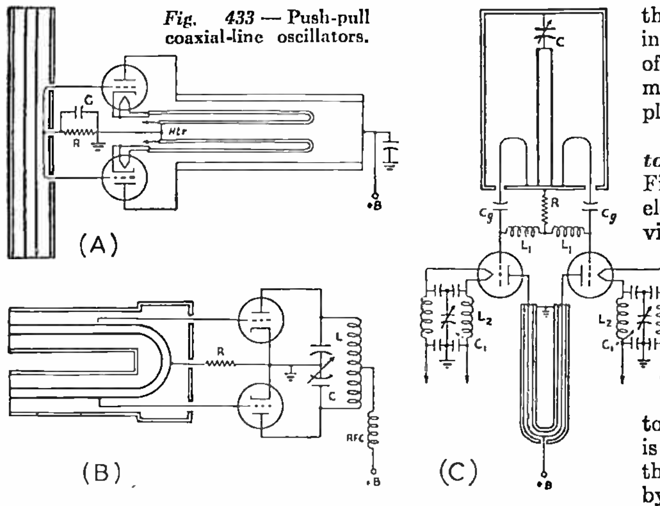


Fig. 433 — Push-pull coaxial-line oscillators.

fitting extension tube which is slid in or out to change the effective conductor length.

The t.g.t.p. circuits are somewhat easier to adjust and load as well as to construct, but are not as satisfactory from the standpoint of frequency stability because of reaction on the frequency-controlling grid line by the tuning of the output circuit. The grid tap should be as far down on the line as will permit reliable oscillation under load. Under some conditions the addition of a small adjustable feed-back capacity between grid and plate not only permits a lower tap location but also increases the upper frequency limit obtainable by advancing the phase of the grid excitation to compensate partially for transit-time lag in the tube.

In the Hartley circuit at A, an output tap is provided on the inner conductor. At B inductive output coupling by means of a half-turn "hairpin" is shown; loading can be changed to some extent by varying its position.

Fig. 432 shows two types of coaxial-line oscillator circuits designed particularly for operation near the upper frequency limits for negative-grid tubes. The circuit at A, with quarter-wave grid and plate lines and a half-wave filament line, is convenient for use with single-lead tubes such as the 955 and 316-A. With the three lines arranged in the form of a triangle, so that their inner conductors attach directly to the tube terminals for minimum lead length, this oscillator will function satisfactorily up to 700–800 Mc.

The circuit of Fig. 432-B is designed to take maximum advantage of the u.h.f. capabilities of double-lead and ring-electrode tube types. Interelectrode capacities are divided between each pair of grid and plate lines, and separate parallel-resonant filament lines complete the isolation. Frequencies as high as 1500–1700 Mc. have been attained with this arrangement.

The by-pass condensers shown in the two circuits of Fig. 432 are made of copper plates insulated by sheet mica. Flanges soldered to the ends of the outer conductor in each line constitute one plate of the condenser; a grounded metal sheet serves as the other plate.

Push-pull coaxial-line oscillators — The push-pull circuits of Fig. 433 employ the same basic elements as the arrangements previously described. At A, a half-wave open-ended line is used in the grid circuit, the grids of the tubes being "tapped" down on the line by coupling them inductively through a small balanced loop running inside the outer conductor. A conventional parallel line is used in the plate circuit, with the cathodes balanced to ground by means of closed half-wave lines. The cathode lines may be small-diameter copper tubing, folded to conserve space, through which rubber-insulated wire is run for the return circuit. These lines may be shielded from the plate line by running them underneath the chassis or separated by a shielding partition.

A folded half-wave grid line is used at B. The copper-tubing inner conductor is bent into the shape of a U. The outer conductor may be either a square-section double trough of sheet copper or two short sections of pipe soldered to a rectangular box of sheet copper which forms the "closed" end. Where even more compact construction is required, the dimensions of the grid line may be still further reduced by using sections of folded coaxial line (§ 2-12). A conventional coil-and-condenser output circuit is shown; at the comparatively low frequencies where this type of construction would be advantageous in the interest of compactness, such an output circuit should be satisfactory.

The arrangement at C has certain modifications which make it particularly suitable for use with higher-powered tubes. The quarter-wave capacity-loaded coaxial line in the grid circuit is of relatively large dimensions and consequently has high Q. Coupling to the tube grids, which is made very loose to preserve the Q of the line, is by means of twin hairpin loops. The inductance of the shunt choke coils, L_1, is adjusted for maximum grid current.

To minimize radiation loss and preserve circuit symmetry, a coaxial line is used in the plate tank circuit. If desired this line may be tuned by a balanced split-stator condenser of the type which has the rotor connection at the center, connected across the plate terminals.

Parallel resonant circuits in the filament leads, tuned to resonance at the operating frequency by the variable condensers, C_1, isolate the filament from ground. The fixed by-pass condensers must have low reactance at the operating frequency. The filament coils, which are in parallel for r.f., are of copper tubing.

Chapter Five

Radiotelephony

5-1 Modulation

The carrier — The steady radio-frequency power generated by transmitting circuits cannot alone result in the transmission of an intelligible message to a receiving point. The continuous wave from the transmitter itself serves only as a "carrier" for the message; the intelligence is conveyed by *modulation* (a change) of the carrier. In radiotelephony, this modulation reproduces electrically the sounds it is intended to convey in a form which can be correctly interpreted or demodulated at the receiving end.

Sound and alternating currents — Sounds are caused by vibrations of air particles. The pitch of the sound depends upon the rate of vibration; the more rapid the vibration, the higher the pitch. Most sounds consist of complex combinations of vibrations of differing rates or frequencies; the human voice, for instance, generates frequencies from about 100 cycles per second to several thousand per second. The problem of transmitting speech by radio, therefore, is one of varying the r.f. carrier in a way which corresponds to the air-particle vibrations. The first step in doing this is to change the sound vibrations into alternating electrical currents of the same frequency and relative intensity; the electromechanical device which achieves this translation is the *microphone*. These audio-frequency currents then may be amplified and used to vary or modulate the normally steady r.f. output of the transmitter.

Methods of modulation — The carrier may be made to vary in accordance with the speech current by using the current to change the phase (§ 2-7), frequency or amplitude of the carrier. *Amplitude modulation* of a constant-frequency carrier is by far the most common system, and is used exclusively on all frequencies below the very-high-frequency region (§ 2-7). *Frequency modulation* of a constant-amplitude carrier, which has special characteristics which make its use desirable under certain conditions, is used to a considerable extent on the very-high frequencies. *Phase modulation*, which is closely related to frequency modulation, has had little or no direct application in practical communication.

Other specialized varieties of modulation, developed for other applications of radio transmission, have been proposed for voice communication. Thus far none of these has achieved practical utilization, however.

5-2 Amplitude Modulation

Carrier requirements — For proper amplitude modulation, the carrier should be completely free from inherent amplitude variations such as might be caused by insufficient filtering of a rectified-a.c. power supply (§ 8-4). It is also essential that the carrier *frequency* be entirely unaffected by the application of modulation. If modulating the amplitude of the carrier also causes a change in the carrier frequency the signal wobbles back and forth with the modulation, introducing distortion and widening the channel taken by the signal. This causes unnecessary interference to other transmissions. In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage which is isolated from the frequency-controlling oscillator by a "buffer" amplifier. Amplitude modulation of an oscillator almost always is accompanied by frequency modulation. Under existing regulations it is permitted, therefore, only on frequencies above 112 Mc., because the

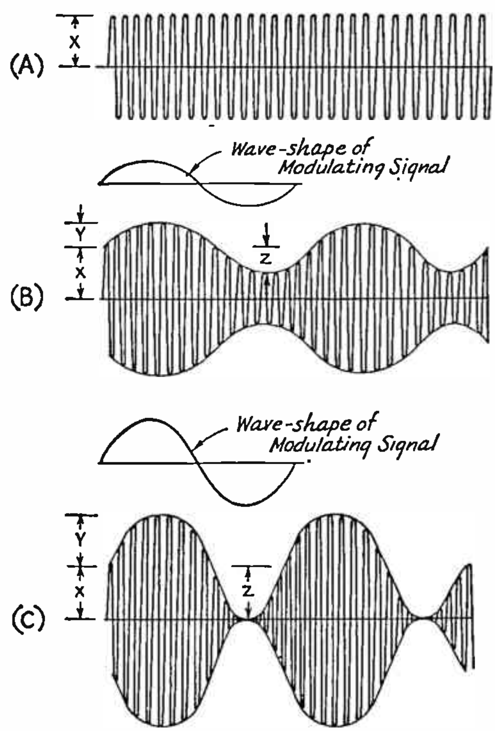


Fig. 501 — **Graphical representation of (A) carrier unmodulated, (B) modulated 50%, (C) modulated 100%.**

problem of interference is less ar region than on lower frequencies.

Percentage of modulation — In the amplitude-modulation system the audible output at the receiver depends entirely upon the amount of variation — termed *depth of modulation* — in the carrier wave, and not upon the strength of the carrier alone. It is desirable therefore to obtain the largest permissible variations in the carrier wave. This condition is reached when the carrier amplitude during modulation is at times reduced to zero and at other times increased to twice its unmodulated value. Such a wave is said to be *fully modulated*, or *100 per cent modulated*. Any desired degree of modulation can be expressed as a percentage, using the unmodulated carrier as a base. Fig. 501 shows, at A, an unmodulated carrier wave; at B, the same wave modulated 50 per cent, and at C, the wave with 100 per cent modulation, using a sine-wave (§ 2-7) modulating signal. The outline of the modulated r.f. wave is called the *modulation envelope*.

The percentage modulation can be found by dividing either Y or Z by X and multiplying the result by 100. If the modulating signal is not symmetrical, the larger of the two (Y or Z) should be used.

Power in modulated wave — The amplitude values correspond to current or voltage, so that the drawings may be taken to represent instantaneous values of either. Since power varies as the square of either the current or voltage (so long as the resistance in the circuit is unchanged), at the peak of the modulation up-swing the instantaneous power in the wave of Fig. 501-C is four times the unmodulated carrier power. At the peak of the down-swing the power is zero, since the amplitude is zero. With a sine-wave modulating signal, the *average* power in a 100 per cent modulated wave is one and one-half times the unmodulated carrier power; that is, the power output of the transmitter increases 50 per cent with 100 per cent modulation.

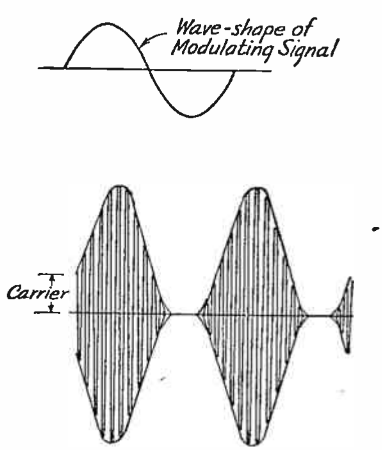


Fig. 502 — **An overmodulated r.f. carrier wave.**

Linearity — Up to the limit of 100 per cent modulation, the amplitude of the carrier should follow faithfully the amplitude variations of the modulating signal. When the modulated r.f. amplifier is incapable of meeting this condition, it is said to be *non-linear*. The amplifier may not, for instance, be capable of quadrupling its power output at the peak of 100 per cent modulation. A non-linear modulated amplifier causes distortion of the modulation envelope.

Modulation characteristic — A graph showing the relationship between r.f. amplitude and instantaneous modulating voltage is called the *modulation characteristic* of the modulated amplifier. This graph should be a straight line (linear) between the limits of zero and twice carrier amplitude. Curvature of the line between these limits indicates non-linearity in the amplifier.

Modulation capability — The *modulation capability* of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from non-linearity. The maximum capability is, of course, 100 per cent. The modulation capability should be as high as possible, so that the most effective signal can be transmitted for a given carrier power.

Overmodulation — If the carrier is modulated more than 100 per cent, a condition such as is shown in Fig. 502 occurs. Not only does the peak amplitude exceed twice the carrier amplitude, but actually there may be a considerable period during which the output is entirely cut off. The modulated wave is therefore distorted (§ 3-3), with the result that harmonics of the audio modulating frequency appear. The carrier should never be modulated more than 100 per cent.

Sidebands — The combining of the audio frequency with the r.f. carrier is essentially a heterodyne process, and therefore gives rise to beat frequencies equal to the sum and difference of the a.f. and r.f. frequencies involved (§ 2-13). Therefore, for each audio frequency appearing in the modulating signal, two new radio frequencies appear, one equal to the carrier frequency plus the audio frequency, the other equal to the carrier minus the audio frequency. These new frequencies are called *side frequencies*, since they appear on each side of the carrier, and the groups of side frequencies representing a band or group of modulation frequencies are called *sidebands*. Hence a modulated signal occupies a group of radio frequencies, or *channel*, rather than a single frequency as in the case of the unmodulated carrier. The *channel width* is twice the highest modulation frequency.

To accommodate the largest number of transmitters in a given part of the r.f. spectrum it is apparent that the channel width should be as small as possible. On the other hand it is necessary, for speech transmission of reasonably good quality, to use modulating frequencies up to a minimum of about 3000 or 4000 cycles. This calls for a channel width of 6 to 8 kilocycles.

Spurious sidebands — Besides the normal sidebands required by speech frequencies, unwanted sidebands may be generated by the transmitter. These usually lie outside the normally required channel, and hence cause it to be wider without increasing the useful modulation. By increasing the channel width, these spurious sidebands cause unnecessary interference to other transmitters. The quality of transmission also is adversely affected when spurious sidebands are generated.

The chief causes of spurious sidebands are harmonic distortion in the audio system, overmodulation, unnecessary frequency modulation, and lack of linearity in the modulated r.f. system.

Types of amplitude modulation — The most widely used type of amplitude-modulation system is that in which the modulating signal is applied in the plate circuit of a radio-frequency power amplifier (*plate modulation*). In a second type the audio signal is applied to a control-grid (*grid-bias modulation*). A third system, involving variation of both plate and grid voltages, is called *cathode modulation*.

¶ 5-3 Plate Modulation

Transformer coupling — In Fig. 503 is shown the most widely used system of plate modulation. A balanced (push-pull Class-A, Class-AB or Class-B) modulator is transformer-coupled to the plate circuit of the modulated r.f. amplifier. The audio-frequency power generated in the modulator plate circuit is combined with the d.c. power in the modulated-amplifier plate circuit by transfer through the coupling transformer, *T*. For 100 per cent modulation the audio-frequency output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the modulated amplifier varies between zero and twice the d.c. operating plate voltage, thus causing corresponding variations in the amplitude of the r.f. output.

Modulator power — The average power output of the modulated stage must increase 50 per cent for 100 per cent modulation (§ 5-2), so that the modulator must supply to the modulated r.f. stage audio power equal to 50 per cent of the d.c. plate input. For example, if the d.c. plate power input to the r.f. stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Modulating impedance; linearity — The *modulating impedance*, or load resistance presented to the modulator by the modulated r.f. amplifier, is equal to

$$\frac{E_b}{I_p} \times 1000$$

where E_b is the d.c. plate voltage and I_p the d.c. plate current in milliamperes, both measured without modulation.

Since the power output of the r.f. amplifier must vary as the square of the plate voltage (the r.f. voltage must be proportional to the applied plate voltage) in order for the modulation to be linear, the amplifier must operate under Class-C conditions (§ 3-4). The linearity then depends upon having sufficient grid excitation and proper bias, and upon the adjustment of circuit constants to the proper values (§ 4-8).

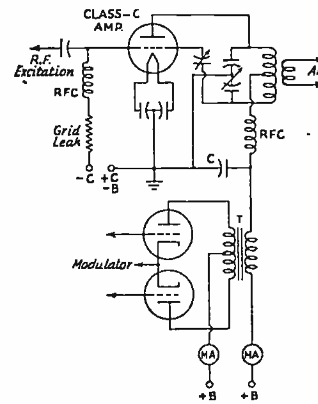


Fig. 503 — Plate modulation of a Class-C r.f. amplifier. The r.f. plate by-pass condenser, *C*, in the amplifier stage should have high reactance at audio frequencies. A capacity of 0.002 μfd. or less usually is satisfactory.

Power in speech waves — The complex waveform of a speech sound translated into alternating current does not contain as much power, on the average, as there is in a pure tone or sine wave of the same peak (§ 2-7) amplitude. That is, with speech waveforms the ratio of peak to average amplitude is higher than in the sine wave. For this reason, the previous statement that the power output of the transmitter increases 50 per cent with 100 per cent modulation, while true for tone modulation, is not true for speech. On the average, speech waveforms will contain only about half as much power as a sine wave, both having the same peak amplitude. The average power output of the transmitter therefore increases only about 25 per cent with 100 per cent speech modulation. However, the *instantaneous* power output must quadruple on the peak of 100 per cent modulation (§ 5-2) regardless of the modulating waveform. Therefore, the peak output power capacity of the transmitter must be the same for any type of modulating signal.

Adjustment of plate-modulated amplifiers — The general operating conditions for Class-C operation have been described (§ 3-4, 4-8). The grid bias and grid current required for plate modulation usually are given in the operating data supplied by the tube manufacturer; in general, the bias should be such as to give an operating angle (§ 4-8) of about 120 degrees at carrier plate voltage, and the excitation should be sufficient to maintain the plate efficiency constant when the plate voltage is varied over the range from zero to twice the d.c. plate voltage applied to the amplifier. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value, supplemented by grid-leak bias to supply the remainder of the required operating bias.

The maximum permissible d.c. plate power input for 100 per cent modulation is twice the sine-wave audio-frequency power output of the modulator. This input is obtained by varying the loading on the amplifier (keeping its tank circuit tuned to resonance) until the product of d.c. plate voltage and plate current is the desired power. The modulating impedance under these conditions will be the proper value for the modulator, if the proper output-transformer turns ratio (§ 2-9) is used.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause non-linearity. The amplifier also should be free from parasitic oscillations (§ 4-10).

Although the *effective* value (§ 2-7) of power input increases with modulation, as described above, the *average* plate input to a plate-modulated amplifier does not change, since each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current. Consequently, the d.c. plate current to a properly modulated amplifier is always constant, with or without modulation.

Screen-grid amplifiers — Screen-grid tubes of the pentode or beam tetrode type can be used as Class-C plate-modulated amplifiers provided the modulation is applied to both the plate and screen grid. The method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 504. The dropping resistor, R, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

The modulating impedance is found by dividing the d.c. plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents is the power-input figure which is used as the basis for determining the audio power required from the modulator.

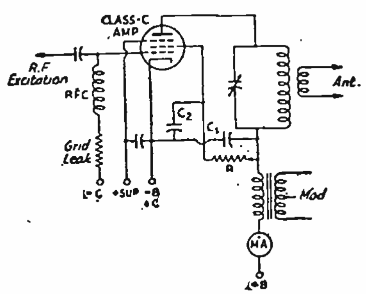


Fig. 504 — Plate and screen modulation of a Class-C r.f. amplifier using a pentode tube. The plate and screen r.f. by-pass condensers, C_1 and C_2, should have high reactance at all audio frequencies (0.002 μfd. or less).

Choke coupling — In Fig. 505 is shown the circuit of the choke-coupled system of plate modulation. The plate power for the modulator tube and modulated amplifier is furnished from a common source through the modulation choke, L, which has high impedance for audio frequencies. The modulator operates as a power amplifier with the plate circuit of the r.f. amplifier as its load, the audio output of the modulator being superimposed on the d.c. power supplied to the amplifier. For 100 per cent modulation, the audio voltage applied to the r.f. amplifier plate circuit across the choke, L, must have a peak value equal to the d.c. voltage on the modulated amplifier. To obtain this without distortion the r.f. amplifier must be operated at a d.c. plate voltage less than the modulator plate voltage, the extent of the voltage difference being determined by the type of modulator tube used. The necessary drop in voltage is provided by the resistor, R_1, which is by-passed for audio frequencies by the by-pass condenser, C_1.

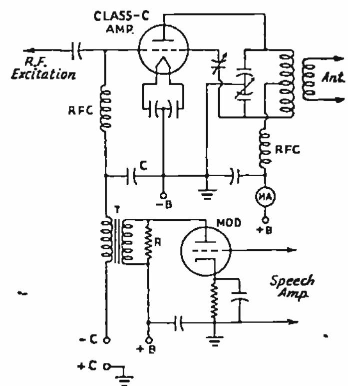


Fig. 505 — Choke-coupled plate modulation.

This type of modulation seldom is used except in very low-power portable sets, because a single-tube Class-A (§ 3-4) modulator is required. The output of a Class-A modulator is very low compared to that obtainable from a pair of tubes of the same size operated Class B, hence only a small amount of r.f. power can be modulated.

Absorption modulation — Absorption or "loss" modulation, in its basic form the oldest and simplest method of all, recently has been revived for ultrahigh-frequency use, where the inertia of a high-Q tuned circuit may make the usual plate-voltage modulation unsatisfactory. In the system shown in Fig. 506, the modulating tubes are connected to the antenna feed line through a quarter-wave stub line, located a quarter-wavelength from the transmitter tank circuit. With no modulation (i.e., no

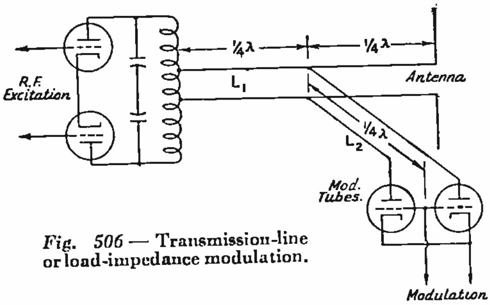


Fig. 506 — Transmission-line or load-impedance modulation.

conduction through the modulating tubes) the stub appears as a short circuit across the line and little or no power reaches the antenna. When modulating voltage is applied to the grids of the modulator tubes, however, their conductance serves to increase the effective impedance of the quarter-wave shunt, permitting a proportionate amount of energy to reach the antenna. At maximum modulation the stub becomes practically an open circuit, allowing the full r.f. output to reach the antenna.

¶ 5-4 Grid-Bias Modulation

Circuit — Fig. 507 is the diagram of a typical arrangement for grid-bias modulation. In this system, the secondary of an audio-frequency output transformer, the primary of which is connected in the plate circuit of the modulator tube, is connected in series with the grid-bias supply for the modulated amplifier. The audio voltage thus introduced varies the grid bias, and thus the power output of the r.f. stage, when suitable operating conditions are chosen. The r.f. stage is operated as a Class-C amplifier, with the d.c. grid bias considerably beyond cut-off.

Operating principles — In this system the plate voltage is constant, and the increase in power output with modulation is obtained by making the plate current and plate efficiency vary with the modulating signal. For 100 per cent modulation, both plate current and efficiency must, at the peak of the modulation upswing, be twice their carrier values, so that the peak power will be four times the carrier power. Since the peak efficiency in practicable circuits is of the order of 70 to 80 per cent, the carrier efficiency ordinarily cannot exceed about 35 to 40 per cent. For a given r.f. tube, the carrier output is about one-fourth the power obtainable from the same tube plate-modulated. Grid bias, r.f. excitation, plate loading and the audio voltage in series with the grid must be adjusted to give a linear modulation characteristic.

Modulator power — Since the increase in average carrier power with modulation is secured by varying the plate efficiency and d.c. plate input of the amplifier, the modulator need supply only such power losses as may be occasioned by connecting it in the grid circuit. These are quite small, hence a modulator capable of only a few watts output will suffice for transmitters of considerable power. Since the load on the modulator varies over the a.f. cycle as the rectified grid current of the modulated amplifier changes, the modulator should have good voltage regulation (§ 5-6).

Grid-bias source — The change in bias voltage with modulation causes the rectified grid current of the amplifier also to vary, the r.f. excitation being fixed. If the bias source has appreciable resistance, the change in grid current also will cause a change in bias in a direction opposite to that caused by the modulation. It is necessary, therefore, to use a grid-bias source having low resistance, so that these bias variations will be negligible. Battery bias is satisfactory. If a rectified a.c. bias supply is used, the type having regulated output (§ 8-9) should be chosen. Grid-leak bias for a grid-modulated amplifier is unsatisfactory, and its use should not be attempted.

Driver regulation — The load on the driving stage varies with modulation, and a linear modulation characteristic may not be obtained if the r.f. voltage from the driver does not stay constant with changes in load. Driver regulation (ability to maintain constant output voltage with changes in load) may be improved by using a driving stage having two or three times the power output necessary for excitation of the amplifier (this is somewhat less than the power required for ordinary Class-C operation), and by dissipating the extra power in a constant load such as a resistor. The load variations are thereby reduced in proportion to the total load.

Adjustment of grid-bias modulated amplifiers — This type of amplifier should be adjusted with the aid of an oscilloscope, to obtain optimum operating conditions. The oscilloscope should be connected as described in § 5-10, the wedge pattern being preferable. A tone source for modulating the transmitter will be convenient. The fixed grid bias should be two or three times the cut-off value (§ 3-2). The d.c. input to the amplifier, assuming 33

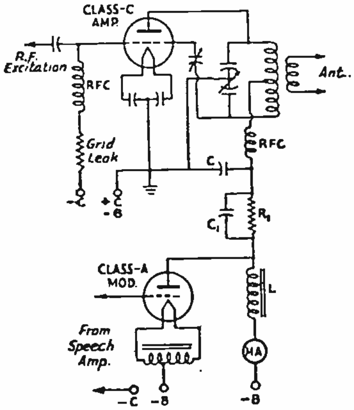


Fig. 507 — Grid-bias modulation of a Class-C amplifier. The r.f. grid by-pass condenser, *C*, should have high reactance at audio frequencies (0.002 μfd. or less).

per cent carrier efficiency, will be 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage. The plate current for this input (in milliamperes, 1000 P/E, where P is the power and E the d.c. plate voltage) must be determined. Apply r.f. excitation

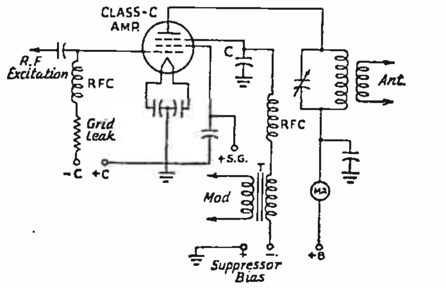


Fig. 508 — Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, C, should be 0.002 μfd. or less.

and, without modulation, adjust the plate loading to give the required plate current (keeping the plate tank circuit tuned to resonance). Next, apply modulation and increase the modulating signal until the modulation characteristic shows curvature (§ 5-10). This probably will occur well below 100 per cent modulation, indicating that the plate efficiency is too high. Increase the plate loading and reduce the excitation to maintain the same plate current; then apply modulation and check the characteristic again. Continue this process until the characteristic is linear from the axis to twice the carrier amplitude. It is advantageous to use the maximum permissible plate voltage on the tube, since it is usually easier to obtain a more linear characteristic with high plate voltage and low current (carrier conditions) than with relatively low plate voltage and high plate current.

The amplifier can be adjusted without an oscilloscope by determining the plate current as described above, then setting the bias to the cut-off value (or slightly beyond) for the d.c. plate voltage used and applying maximum excitation. Adjust the plate loading, keeping the tank circuit at resonance, until the amplifier draws twice the carrier plate current, and note the antenna current. Decrease the excitation until the output and plate current just start to drop. Then increase the bias, leaving the excitation and plate loading unchanged, until the plate current drops to the proper carrier value. The antenna current should be just half the previous value; if it is larger, try somewhat more loading and less excitation; if smaller, less loading and more excitation. Repeat until the antenna current drops to half its maximum value when the plate current is biased down to the carrier value. Under these conditions the amplifier should modulate properly, provided the plate supply has good voltage regulation (§ 8-1) so that the plate voltage is practically the same at both values of plate current during the initial testing. The d.c. plate current should be substantially constant with or without modulation (§ 5-3).

Suppressor modulation — The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 508. The operating principles are the same as for grid-bias modulation. However, the r.f. excitation and modulating signals are applied to separate grids, which gives the system a simpler operating technique since best adjustment for proper excitation requirements and proper modulating circuit requirements are more or less independent. The carrier plate efficiency is approximately the same as for grid-bias modulation, and the modulator power requirements are similarly small. With tubes having suitable suppressor-grid characteristics, linear modulation up to practically 100 per cent can be obtained with negligible distortion.

The method of adjustment is essentially the same as that described in the preceding paragraph. Apply normal excitation and bias to the control grid and, with the suppressor bias at zero or the positive value recommended for c.w. telegraph operation with the particular tube used, adjust the plate loading to obtain twice the carrier plate current (on the basis of 33 per cent carrier efficiency). Then apply sufficient negative bias to the suppressor to bring the plate current to the carrier value, leaving the loading unchanged. Simultaneously, the antenna current also should drop to half its maximum value. The amplifier is then ready for modulation. Should the plate current not follow the antenna current in the same proportion when the suppressor bias is made negative, the loading and excitation should be readjusted to make them coincide.

¶ 5-5 Cathode Modulation

Circuit — The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 509. This type of modulation is a com-

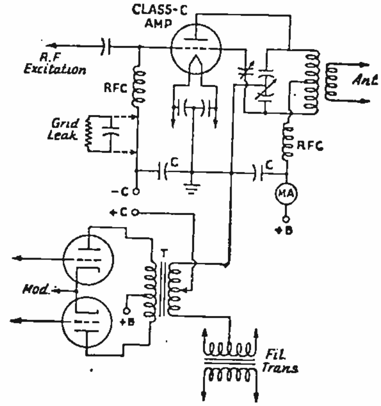


Fig. 509 — Cathode modulation of a Class-C r.f. amplifier. The grid and plate r.f. by-pass condensers, C, should be 0.002 μfd. or less (for high a.f. reactance).

bination of the plate and grid-bias methods, and permits a carrier efficiency midway between the two. The audio power is introduced in the cathode circuit, and both grid bias and plate voltage vary during modulation.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter; that is, when filament-type tubes are modulated they must be supplied from a separate filament transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio-frequency modulation.

Operating principles — Because part of the modulation is by the grid-bias method, the plate efficiency of the modulated amplifier must vary during modulation. The carrier efficiency therefore must be lower than the efficiency at the modulation peak. The required reduction in carrier efficiency depends upon the proportion of grid modulation to plate modulation; the higher the percentage of plate modulation, the higher the permissible carrier efficiency, and vice versa. The audio power required from the modulator also varies with the percentage of plate modulation, being greater as this percentage is increased.

The way in which the various quantities vary is illustrated by the curves of Fig. 509. In these curves the performance of the cathode-modulated r.f. amplifier is plotted in terms of the tube ratings for plate-modulated telephony, with the percentage of plate modulation as a base. As the percentage of plate modulation is decreased, it is assumed that the grid-bias modulation is increased to make the over-all percentage of modulation reach 100 per cent. The limiting condition, 100 per cent plate modulation and no grid-bias modulation, is at the right (A); pure grid-bias modulation is represented by the left-hand ordinate (B and C).

As an example, assume that 40 per cent plate modulation is to be used. Then the modulated r.f. amplifier must be adjusted for a carrier plate efficiency of 56 per cent, the permissible plate input will be 65 per cent of the ratings of the same tube with pure plate modulation, the power output will be 48 per cent of the rated output of the tube with plate modulation, and the audio power required from the modulator will be 20 per cent of the d.c. input to the modulated amplifier.

Modulating impedance — The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m is the percentage of plate modulation expressed as a decimal, E_b is the plate voltage and I_b the plate current of the modulated r.f. amplifier. This figure for the modulating impedance is used in the same way as the corresponding figure for pure plate modulation, in determining the proper modulator operating conditions (§ 5-6).

Conditions for linearity — R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and grid-bias modulation. More excitation is required as the percentage of plate modulation is increased. Grid bias should be considerably beyond cut-off; fixed bias from a supply having good voltage regulation (§ 8-9) is preferred, especially when the percentage of plate modulation is small and the amplifier is operating more nearly like a grid-bias modulated stage. At the higher percentages of plate modulation a combination of fixed and grid-leak bias can be used, since the variation in rectified grid current is smaller. The grid leak should be by-passed for audio frequencies. The percentage of grid modulation may be regulated by choice of a suitable tap on the modulation transformer secondary.

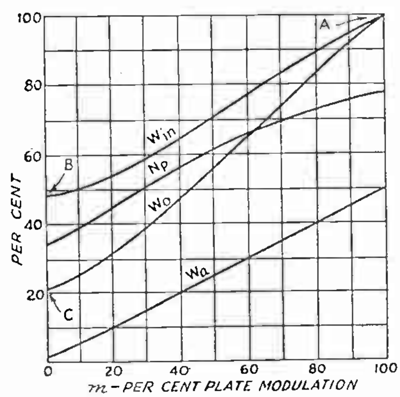


Fig. 510 — Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class-C telephony tube ratings.

W_{in} — D.c. plate input watts in terms of percentage of plate-modulation rating.

W_o — Carrier output watts in per cent of plate-modulation rating (based on plate efficiency of 77.5%).

W_a — Audio power in per cent of d.c. watts input.

N_p — Plate efficiency of the amplifier in percentage.

Adjustment of cathode-modulated amplifiers — In most respects, the adjustment procedure is similar to that for grid-bias modulation (§ 5-4). The critical adjustments are those of antenna loading, grid bias, and excitation. The proportion of grid-bias to plate modulation will determine the operating conditions.

Adjustments should be made with the aid of an oscilloscope (§ 5-10). With proper antenna loading and excitation, the normal wedge-shaped pattern will be obtained at 100 per cent modulation. As in the case of grid-bias modulation, too-light antenna loading will cause flattening of the upward-peaks of modulation (indicating downward modulation), as also will too-high excitation (§ 5-10). The cathode current will be practically constant with or without modulation when the proper operating conditions have been established (§ 5-3).

5-6 Class-B Modulators

Modulator tubes — In the case of plate modulation, the relatively large audio power needed (§ 5-3) practically dictates the use of a Class-B (§ 3-4) modulator, since the power can be obtained most economically with this type of amplifier. A typical circuit is given in Fig. 511. A pair of tubes must be chosen which is capable of delivering sine-wave audio power equal to half the d.c. input to the modulated Class-C amplifier. It is sometimes convenient to use tubes which will operate at the same plate voltage as that applied to the Class-C stage, since one power supply of adequate current capacity may then suffice for both stages. Available components do not always permit this, however, and better over-all performance and economy may result from the use of separate power supplies.

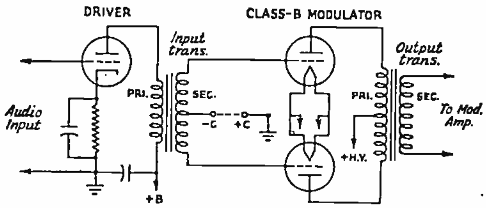


Fig. 511 — Class-B audio modulator and driver circuit.

Matching to load — In giving Class-B ratings on power tubes, manufacturers specify the plate-to-plate load impedance (§ 3-3) into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance (§ 5-3) of the Class-C r.f. stage, so that a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

$$\sqrt{\frac{Z_p}{Z_m}}$$

where Z_m is the Class-C modulating impedance and Z_p is the plate-to-plate load impedance specified for the Class-B tubes.

Commercial Class-B output transformers usually are rated to work between specified primary and secondary impedances and are designed for specific Class-B tubes. In such a case, the turns ratio can be found by substituting the given impedances in the formula above. Many transformers are provided with primary and secondary taps, so that various turns ratios can be obtained to meet the requirements of various tube combinations.

Driving power — Class-B amplifiers are driven into the grid-current region, so that power is consumed in the grid circuit (§ 3-3). The preceding stage (*driver*) must be capable of supplying this power at the required peak audio-frequency grid-to-grid voltage. Both of these quantities are given in the manufacturer's tube ratings. The grids of the Class-B tubes represent a variable load resistance over the audio-frequency cycle, since the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source which has good *regulation* — that is, which will maintain the waveform of the signal without distortion even though the load varies. This can be brought about by using a driver capable of delivering two or three times the actual power consumed by the Class-B grids, and by using an input coupling transformer having a turns ratio giving the largest step-down in the voltage between the driver plate or plates and the Class-B grids that will permit obtaining the specified grid-to-grid a.f. voltage.

Driver coupling — A Class-A or Class-AB (§ 3-4) driver is used to excite a Class-B stage. Tubes for the driver preferably should be triodes having low plate resistance, since these will have the best regulation. Having chosen a tube or tubes capable of ample power output from tube data sheets, the peak output voltage will be, approximately,

$$E_o = 1.4\sqrt{PR}$$

where P is the power output and R the load resistance. The input transformer ratio, primary to secondary, will be

$$\frac{E_o}{E_g}$$

where E_o is as given above and E_g is the peak grid-to-grid voltage required by the modulator tubes.

Commercial transformers normally are designed for specific driver-modulator combinations, and usually are adjusted to give as good driver regulation as the conditions will permit.

Grid bias — Modern Class-B audio tubes are intended for operation without fixed bias. This lessens the variable grid-circuit loading effect and eliminates the need for a grid-bias supply.

When a grid-bias supply is required, it must have low internal resistance so that the flow of grid current with excitation of the Class-B tubes does not cause a continual shift in the actual grid bias and thus cause distortion. Batteries or a regulated bias supply (§ 8-9) should be used.

Plate supply — The plate supply for a Class-B modulator should be sufficiently well filtered (§ 8-3) to prevent hum modulation of the r.f. stage (§ 5-2). An additional requirement is that the output condenser of the supply should have low reactance (§ 2-8) at 100 cycles or less compared to the load into which each tube is working, which is one-fourth the plate-to-plate load resistance. A 4-μfd. output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply, usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

Overexcitation — When a Class-B amplifier is overdriven in an attempt to secure more than the rated power, distortion in the output waveshape increases rapidly. The high-frequency harmonics which result from the distortion (§ 3-3) modulate the transmitter, producing spurious sidebands (§ 5-2) which readily can cause serious interference over a band of frequencies several times the channel width required for speech. This may happen even though the transmitter is not being overmodulated, as in the case where the modulator is incapable of delivering the power required to modulate the transmitter fully, or when the Class-C amplifier is not adjusted to give the proper modulating impedance (§ 5-3).

The tubes used in the Class-B modulator should be capable of somewhat more than the power output nominally required (50 per cent of the d.c. input to the modulated amplifier) to take care of losses in the output transformer. These usually run from 10 per cent to 20 per cent of the tube output. In addition, the Class-C amplifier should be adjusted to give the proper modulating impedance and the correct output transformer turns ratio should be used. Such high-frequency harmonics as may be generated in these circumstances can be reduced by connecting condensers across the primary and secondary of the output transformer (about 0.002 μfd. in the average case), to form, with the transformer leakage inductance (§ 2-9) a low-pass filter (§ 2-11) which cuts off just above the maximum audio frequency required for speech transmission (about 4000 cycles). The condenser voltage ratings should be adequate for the peak a.f. voltages appearing across them.

Operation without load — Excitation should never be applied to a Class-B modulator until after the Class-C amplifier is turned on and is drawing the value of plate current required to present the rated load to the modulator. With no load to absorb the power, the primary impedance of the transformer rises to a high value and excessive audio voltages are developed across it — frequently high enough to break down the transformer insulation. If the modulator is to be tested separately from the transmitter, a load resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the transformer secondary.

¶ 5-7 Low-Level Modulators

Selection of tubes — Modulators for grid-bias and suppressor modulation can be small audio power tubes, since the audio power required usually is small. A triode such as the 2A3 is preferable because of its low plate resistance, but pentodes will work satisfactorily.

Matching to load — Since the ordinary Class-A receiving power tube will develop about 200 to 250 peak volts in its plate circuit, which is ample for most low-level modulator applications, a 1:1 coupling transformer is generally used. If more voltage is required, a step-up ratio must be provided in the transformer. It is usual practice to load the primary of the output-coupling transformer with a resistance equal to or slightly higher than the rated load resistance for the tube, to stabilize the voltage output and thus improve the regulation. This is indicated in Figs. 507 and 508.

¶ 5-8 Microphones

Sensitivity — The *level* of a microphone is its electrical output for a given speech intensity input. Level varies greatly with microphones of different basic types, and also varies between different models of the same type. The output is also greatly dependent on the character of the individual voice (that is, the audio frequencies present in the voice) and the distance of the speaker's lips from the microphone, decreasing approximately as the square of the distance. Hence, only approximate values based on averages of "normal" speaking voices can be attempted. The values given in the following paragraphs are based on close talking; that is, with the microphone less than an inch from the speaker's lips.

Frequency response — The *frequency response* or *fidelity* of a microphone is its relative ability to convert sounds of different frequencies into alternating current. With fixed sound intensity at the microphone, the electrical output may vary considerably as the sound frequency is varied. For understandable speech transmission only a limited frequency range is necessary, and natural-sounding speech can be obtained if the output of the microphone does not vary more than a few decibels (§ 3-3) at any frequency within a range of about 200 cycles to 4000 cycles. When the variation expressed in terms of decibels is small between two frequency limits, the microphone is said to be *flat* between those limits.

Carbon microphones — Fig. 512-A and B show connections for single- and double-button carbon microphones, with a rheostat included in each circuit for adjusting the button current to the correct value as specified with each microphone. The single-button microphone consists of a metal diaphragm placed against an insulating cup containing loosely packed carbon granules (*microphone button.*) Current from a battery flows through the granules, the diaphragm being one connection and the metal back-plate the other. The primary of a transformer is connected in series with the battery and microphone. As the diaphragm vibrates its pressure on the granules alternately increases and decreases, causing a corresponding increase and decrease of current flow through the circuit, since the pressure changes the resistance of the mass of granules. The resulting change in the current flowing through the transformer primary causes an alternating voltage, of corresponding frequency and intensity, to be set up in the transformer secondary (§ 2-9). The double-button type is similar, but with two buttons in push-pull.

Good quality single-button carbon microphones give outputs ranging from 0.1 to 0.3 volt across 50 to 100 ohms; that is, across the primary winding of the microphone transformer. With the step-up of the transformer, a peak voltage of between 3 and 10 volts across 100,000 ohms or so can be assumed available at the grid of the first tube. The usual button current is 50 to 100 ma.

The level of good-quality double-button microphones is considerably less, ranging from 0.02 volt to 0.07 volt across 200 ohms. With this type of microphone and the usual push-pull input transformer, a peak voltage of 0.4 to 0.5 across 100,000 ohms or so can be assumed available at the first speech-amplifier grid. The button current with this type of microphone ranges from 5 to 50 ma. per button.

Crystal microphones — The input circuit for a piezoelectric or crystal type of microphone is shown in Fig. 512-F. The element in this type consists of a pair of Rochelle salts crystals cemented together, with plated electrodes. In the more sensitive types, the crystal is mechanically coupled to a diaphragm. Sound waves actuating the diaphragm cause the crystal to vibrate mechanically and, by piezoelectric action (§ 2-10), to generate a corresponding alternating voltage between the electrodes, which are connected to the grid circuit of a vacuum-tube amplifier, as shown. The crystal type requires no separate source of current or voltage.

Although the level of crystal microphones varies with different models, an output of 0.01 to 0.03 volt is representative for communication types. The level is affected by the length of the cable connecting the microphone to the first amplifier stage; the above figure is for lengths of 6 or 7 feet. The frequency characteristic is unaffected by the cable, but the load resistance (amplifier grid resistor) does affect it, the lower frequencies being attenuated as the shunt resistance becomes less. A grid-resistor value of 1 megohm or more should be used for reasonably flat response, 5 megohms being a customary figure.

Condenser microphones — The condenser microphone of Fig. 512-C consists of a two-plate capacity, with one plate stationary. The other, which is separated from the first by about a thousandth of an inch, is a thin metal membrane serving as a diaphragm. This condenser is connected in series with a resistor and a d.c. voltage source. When the diaphragm vibrates, the change in capacity causes a small charging current to flow through the circuit. The resulting audio voltage, which appears across the resistor is fed to the grid of the tube through the coupling condenser.

The output of condenser microphones varies with different models, the high-quality type being about one-hundredth to one-fiftieth as sensitive as the double-button carbon microphone. The first speech-amplifier stage must be built into the microphone, since the capacity of a connecting cable would impair both output and frequency range.

Velocity and dynamic microphones — In a velocity or "ribbon" microphone, the element acted upon by the sound waves is a thin corrugated metallic ribbon suspended between the poles of a magnet. When vibrating, the ribbon cuts the lines of force between the poles, first in one direction and then the other, thus generating an alternating voltage. The movement of the ribbon is proportional to the velocity of the sound-energized air particles. Velocity microphones are built in two types, high impedance and low impedance, the former being used in most applications. A high-impedance microphone can be directly connected to the grid of an amplifier tube, shunted by a resistance of 0.5 to 5 megohms (Fig. 512-E). Low-impedance microphones are used when a long connecting cable (75 feet or more) must be employed. In such a case the output of the microphone is coupled to the first amplifier stage through a suitable step-up transformer, as shown in Fig. 512-D.

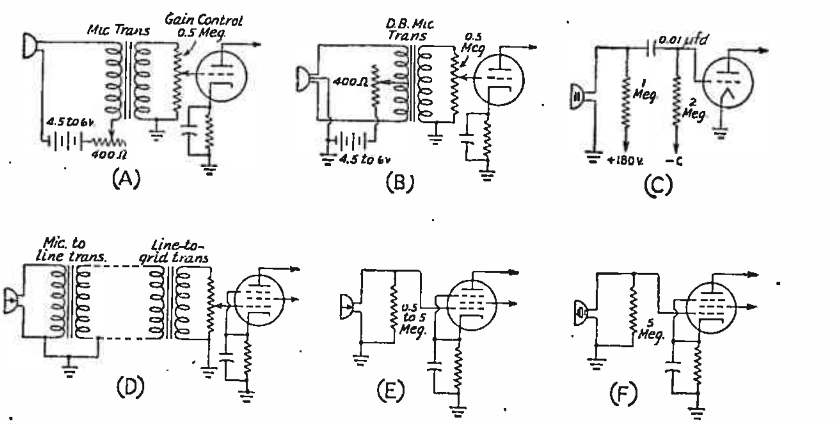


Fig. 512 — Speech input circuits of five commonly used types of microphones. A, single-button carbon; B, double-button carbon; C, condenser; D, low-impedance velocity; E, high-impedance velocity; F, crystal.

The level of the velocity microphone is about 0.03 to 0.05 volt. This figure applies directly to the high-impedance type, and to the low-impedance type when the voltage is measured across the coupling transformer secondary.

The dynamic microphone somewhat resembles a dynamic loud speaker in principle. A light-weight voice coil is rigidly attached to a diaphragm, the coil being placed between the poles of a permanent magnet. Sound causes the diaphragm to vibrate, thus moving the coil back and forth between the magnet poles and generating an alternating voltage the frequency of which is proportional to the frequency of the impinging sound and the amplitude proportional to the sound pressure. The dynamic microphone usually is built with high-impedance output, suitable for working directly into the grid of an amplifier tube. If the connecting cable must be unusually long a low-impedance type should be used, with a step-up transformer at the end of the cable. A small permanent-magnet speaker can be used as a dynamic microphone, although the fidelity is not as good as is obtainable with a properly designed microphone.

5-9 The Speech Amplifier

Description — The function of the speech amplifier is to build up the weak microphone voltage to a value sufficient to excite the modulator to the required output. It may have from one to several stages. The last stage nearly always must deliver a certain amount of audio power, especially when it is used to excite a Class-B modulator. Speech amplifiers for grid-bias modulation usually end in a power stage which also functions as the modulator.

The speech amplifier frequently is built as a unit separate from the modulator, and in such a case may be provided with a step-down transformer designed to work into a low impedance, such as 200 or 500 ohms (tube-to-line transformer). When this is done, a step-up input transformer intended to work between the same impedance and the modulator grids (line-to-grid transformer) is provided in the modulator circuit. The line which connects the two transformers may be made of any convenient length.

General design considerations — The last stage of the speech amplifier must be selected on the basis of the power output required from it; for instance, the power necessary to drive a Class-B modulator (§ 5-6). It may be either single-ended or push-pull (§ 3-3), the latter generally being preferable because of the higher power output and lower harmonic distortion. Push-pull amplifiers may be either Class A, Class AB_1 or Class AB_2 (§ 3-4), as the power requirements dictate. If a Class-A or AB_1 amplifier is used, the preceding stages all may be voltage amplifiers, but when a Class-AB_2 amplifier is used the stage immediately preceding it must be capable of furnishing the power consumed by its grids at full output. The requirements in this case are much the same as those which must be met by a driver for a Class-B stage (§ 5-6), but the actual power needed is considerably smaller and usually can be supplied by one or two small receiving triodes. All lower-level speech amplifier stages invariably are worked purely as voltage amplifiers.

The minimum amplification which must be provided ahead of the last stage is equal to the peak audio-frequency grid voltage required by the last stage for full output (peak grid-to-grid voltage in the case of a push-pull stage), divided by the output voltage of the microphone or secondary of the microphone transformer if one is used (§ 5-8). The peak a.f. grid voltage required by the output tube or tubes is equal to the d.c. grid bias in the case of a single-tube Class-A amplifier, and approximately twice the grid bias for a push-pull Class-A stage. The requisite information for Class-AB_1 and AB_2 amplifiers can be obtained from the manufacturer's data on the type considered. If the gain is not obtainable in one stage, several stages must be used in cascade. When the output stage is operated Class AB_2, due allowance must be made for the fact that the next-to-the-last stage must deliver power as well as voltage. In such cases, suitable driver combinations usually are recommended by manufacturers of tubes and interstage transformers. The coupling transformer must be designed especially for the purpose.

The total gain provided by a multi-stage amplifier is equal to the product of the individual stage gains. For example, when three stages are used, the first having a gain of 100, the second 20 and the third 15, the total gain is $100 \times 20 \times 15$, or 30,000. It is good practice to provide two or three times the minimum required gain in designing the speech amplifier. This will insure having ample gain available to cope with varying conditions.

When the gain must be fairly high, as when a crystal microphone is used, the speech amplifier frequently has four stages, including the power output stage. The first generally is a pentode, because of the high gain attainable with this type of tube. The second and third stages usually are triodes, the third frequently having two tubes in push-pull when it drives a Class-AB_2 output stage. Two pentode stages seldom are used consecutively, because of the difficulty of getting stable operation when the gain per stage is very high. With carbon microphones less amplification is needed and hence the pentode first stage usually is omitted, one or two triode stages being ample to obtain full output from the power stage.

Stage gain and voltage output — In voltage amplifiers, the *stage gain* is the ratio of a.c. output voltage to a.c. voltage applied to the grid. It will vary with the applied audio frequency, but for speech the variation should be small over the range of 100–4000 cycles. This condition is easily met in practice.

The output voltage is the maximum value which can be taken from the plate circuit without distortion. It is usually expressed in terms of the peak value of the a.c. wave (§ 2-7), since this value is independent of the waveform. The peak output voltage usually is of interest only when the stage drives a power amplifier, since only in this case is the stage called upon to work near its maximum capabilities. Low-level stages very seldom are worked near their full capacity, hence harmonic distortion is negligible and the voltage gain of the stage is the primary consideration.

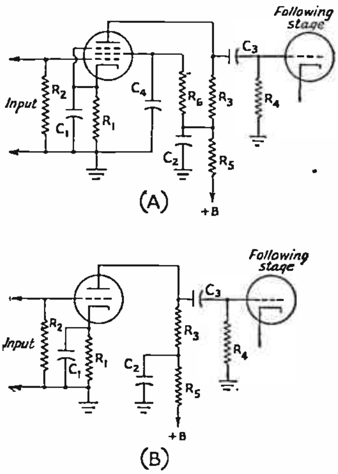


Fig. 513 — Resistance-coupled voltage amplifier circuits. A, pentode; B, triode. Designations are as follows:
C_1 — Cathode by-pass condenser.
C_2 — Plate by-pass condenser.
C_3 — Output coupling condenser (blocking condenser).
C_4 — Screen by-pass condenser.
R_1 — Cathode resistor.
R_2 — Grid resistor.
R_3 — Plate resistor.
R_4 — Next-stage grid resistor.
R_5 — Plate decoupling resistor.
R_6 — Screen resistor.
Values for suitable tubes are given in Chapter Fourteen.

Resistance coupling — Resistance coupling generally is used in voltage amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there is little danger of hum pick-up from stray magnetic fields associated with heater wiring. It is the only type of coupling suitable for the output circuits of pentodes and high-μ triodes, since with transformers a sufficiently high load impedance (§3-3) cannot be obtained without considerable frequency distortion. Typical circuits are given in Fig. 512 and design data in § 3-6.

Transformer coupling — Transformer coupling between stages ordinarily is used only when power is to be transferred (in such a case resistance coupling is very inefficient), or when it is necessary to couple between a single-ended and a push-pull stage. Triodes having an amplification factor of 20 or less are used in transformer-coupled voltage amplifiers.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 514. That at A uses a combination of resistance and transformer coupling, and may be used for exciting the grids of a Class-A or AB_1 following stage. The resistance coupling is used to keep the d.c. plate current from flowing through the transformer primary, thereby preventing a reduction in primary inductance below its no-current value (§ 8-4). This improves the low-frequency response. With low-μ triodes (6C5, 6J5, etc.), the gain is equal to that with resistance coupling multiplied by the secondary-to-primary turns ratio of the transformer.

In B the transformer primary is in series with the plate of the tube, and thus must carry the tube plate current. When the following amplifier operates without grid current, the voltage gain of the stage is practically equal to the μ of the tube multiplied by the transformer ratio. This circuit also is suitable for transferring power (within the capabilities of the tube) as in the case of a following Class-AB_2 stage used as a driver for a Class-B modulator.

Gain control — The over-all gain of the amplifier may be changed to suit the output level of the microphone, which will vary with voice intensity and distance of the speaker from the microphone, by varying the proportion of a.c. voltage applied to the grid of one of the stages.

The gain-control potentiometer should be near the input end of the amplifier, so that there will be no danger of overloading the stages ahead of the gain control. With carbon microphones the gain control may be placed directly across the microphone transformer secondary, but with other types the gain control usually will affect the frequency response of the microphone when connected directly across it. The control therefore usually is placed in the grid circuit of the second stage.

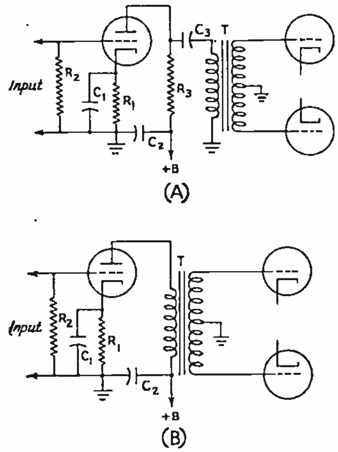


Fig. 514 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B, for transformer coupling. Designations correspond to those in Fig. 513. In A, values can be taken from Table I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given for the particular type of tube used (§ 3-6).

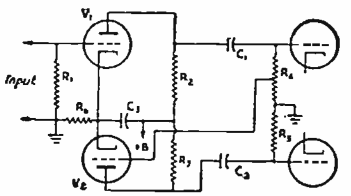


Fig. 515 — Phase-inverter circuit for resistance-coupled push-pull output. With a double-triode tube (such as the 6N7) the following values are typical:
R_1, R_4, R_5 — 0.5 megohm. R_2, R_3 — 0.1 megohm.
R_6 — 1500 ohms. C_1, C_2 — 0.1 μfd.
R_4 should be tapped as described in the text. The voltage gain of a stage using these constants is 22.

Phase inversion — Push-pull output may be secured with resistance coupling by using an extra tube, as shown in Fig. 515. There is a phase shift of 180 degrees through any normally operating resistance-coupled stage (§ 3-3), and the extra tube is used purely to provide this phase shift without additional gain. The outputs of the two tubes are then added to provide push-pull excitation for the following amplifier.

Output limiting — It is desirable to modulate as heavily as possible without overmodulating, yet it is difficult to speak into the microphone at a constant intensity. To maintain reasonably constant output from the modulator in spite of variations in speech intensity, it is possible to use automatic gain control which follows the *average* (not instantaneous) variations in speech amplitude. This is accomplished by rectifying and filtering (§ 8-2, 8-3) some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier.

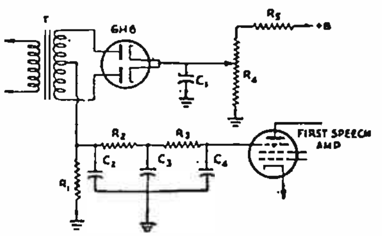


Fig. 516 — Speech amplifier output-limiting circuit. C_1, C_2, C_3, C_4 — 0.1-μfd. R_1, R_2, R_3 — 0.25 megohm. R_4 — 25,000-ohm pot. R_5 — 0.1 megohm. T — See text.

A practical circuit for this purpose is shown in Fig. 516. The rectifier must be connected, through the transformer, to a tube capable of delivering some power output (a small part of the output of the power stage may be used) or else a separate amplifier for the rectifier circuit alone may have its grid connected in parallel with that of the last voltage amplifier. Resistor R_4 in series with R_5 across the plate supply provides variable bias on the rectifier plates, so that the limiting action can be delayed until a desired microphone input level is reached. R_2, R_3, C_2, C_3, and C_4 form the filter (§ 2-11), and the output of the rectifier is connected to the suppressor grid of the pentode first stage of the speech amplifier.

A step-down transformer with a turns ratio such as to give about 50 volts when its primary is connected to the output circuit of the power stage should be used. A half-wave rectifier may be used instead of the full-wave circuit shown, although satisfactory filtering will be more difficult to achieve.

Noise — It is important that the noise level in a speech amplifier be low compared to the level of the desired signal. Noise in the speech amplifier is caused chiefly by hum, which may be the result of insufficient power-supply filtering or may be introduced into the grid circuit of a tube by magnetic or electrostatic means from heater wiring. The plate voltage for the amplifier should be free from ripple (§ 8-4), particularly the voltage applied to the low-level stages. A two-section condenser-input filter (§ 8-5) usually is satisfactory. The decoupling circuits mentioned in the preceding paragraphs also are helpful in reducing plate-supply hum.

Hum from heater wiring may be reduced by keeping the wiring well away from ungrounded components or wiring, particularly in the vicinity of the grid of the first tube. Complete shielding of the microphone jack is advisable, and when tubes with grid caps instead of the single-ended types are used the caps and the exposed wiring to them should be shielded. Heater wiring preferably should run in the corners of a metal chassis, to reduce the magnetic field. A ground should be made either on one side of the heater circuit or to the center-tap of the heater winding. The shells of metal tubes should be grounded; glass tubes require separate shields, especially when used in low-level stages. Heater connections to the tube sockets should be kept as far as possible from the plate and grid prongs, and the heater wiring to the sockets should be kept close to the chassis. A connection to a good ground (such as a cold water pipe) also is advisable. The speech amplifier always should be constructed on a metal chassis, with all ground connections made directly to the metal chassis.

When the power supply is mounted on the same chassis with the speech amplifier, the power transformer and filter chokes should be well separated from audio transformers in the amplifier proper to reduce magnetic coupling, which would cause hum and raise the residual noise level.

ℂ 5-10 Checking 'Phone Transmitter Operation

Modulation percentage — The most reliable method of determining percentage of modulation is by means of the cathode-ray oscilloscope (§ 3-9). The oscilloscope gives a direct picture of the modulated output of the transmitter, and by its use the waveform errors inherent in other types of measurements are eliminated.

Two types of oscilloscope patterns may be obtained, known as the "wave envelope" and "trapezoid." The former shows the shape of the modulation envelope (§ 5-2) directly, while the latter in effect plots the modulation characteristic (§ 5-2) of the modulated stage on the cathode-ray tube screen. To obtain the wave-envelope pattern, the oscilloscope must have a horizontal sweep circuit. The trapezoidal pattern requires only the oscilloscope, the sweep circuit being supplied by the transmitter itself. Fig. 517 shows methods of connecting the oscilloscope to the transmitter for both types of patterns. The oscilloscope connections for the wave-envelope pattern, Fig. 517-A, are usually simpler than those for the trapezoidal figure. The vertical-deflection plates are coupled to the amplifier tank coil or an antenna coil by means of a pick-up coil of a few turns connected to the oscilloscope through a twisted-pair line. The position of the pick-up coil is varied until a carrier pattern, Fig. 518-B, of suitable height is obtained. The sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. It is frequently helpful in eliminating r.f. harmonics from the pattern to connect a resonant circuit, tuned to the operating frequency, between the vertical deflection plates, using link coupling between this circuit and the transmitter tank circuit.

With the application of voice modulation, a rapidly changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100 per cent (§ 5-2). This is illustrated by Fig. 518-D, where the point X represents the sweep line

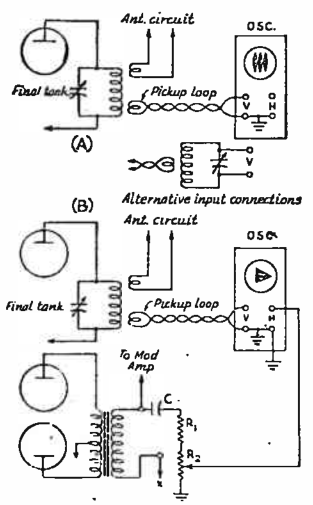


Fig. 517 — Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation.

(reference line) alone, YZ is the carrier height, and PQ is the maximum height of the modulated wave. If the height is greater than the distance PQ, as illustrated in E, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 per cent. Assuming that the modulation is symmetrical, however, any modulation percentage can be measured directly from the screen by measuring the maximum height with modulation and the height of the carrier alone; calling these two heights h_1 and h_2 respectively, the modulation percentage is

$$\frac{h_1 - h_2}{h_2} \times 100$$

Connections for the trapezoidal pattern are shown in Fig. 517-B. The vertical plates are similarly coupled to the transmitter tank circuit through a pick-up loop; the tuned input circuit to the oscilloscope may also be used. The horizontal plates are coupled to the output of the modulator through a voltage divider (§ 2-6), R_1R_2, the resistance of R_2 being variable to permit adjustment of the audio voltage to a suitable value to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control resistor. The value of R_1 will depend upon the audio output voltage of the modulator. This voltage is equal to $\sqrt{PR}$, where P is the audio power output of the modulator and R is the modulating impedance of the modulated r.f. amplifier. In the case of grid-bias modulation with a 1:1 output transformer, it will be satisfactory to assume that the a.c. output voltage of the modulator is equal to $0.7E$ for a single tube or $1.4E$ for a push-pull stage, where E is the d.c. plate voltage on the modulator. If the transformer ratio is other than 1:1, the voltage so calculated should be multiplied by the actual secondary-to-primary turns ratio.

The total resistance of R_1 and R_2 in series should be 0.25 megohm for every 150 volts of modulator output; for example, if the modulator output voltage is 600, the total resistance should be four (600/150) times 0.25 megohm, or 1 megohm. Then, with 0.25 megohm at R_2, R_1 should be 0.75 megohm. The blocking condenser, C, should be 0.1 μfd or more, and its voltage rating should be greater than the maximum voltage appearing in the circuit. With plate modulation, this is twice the d.c. voltage applied to the plate of the modulated amplifier.

The trapezoidal patterns are shown in Fig. 518 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line

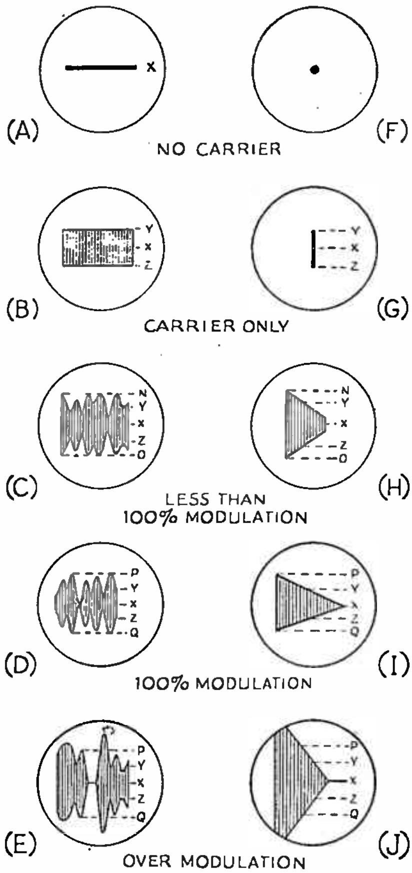


Fig. 518 — Wave-envelope and trapezoidal patterns encountered under different conditions of modulation.

should be adjusted, by means of the pick-up coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher the modulation percentage, the wider and more pointed the wedge becomes. At 100 per cent modulation it just makes a point on the axis, *A*, at one end, and the height, *PQ*, at the other end is equal to twice the carrier height, *YZ*. Overmodulation in the upward direction is indicated by increased height over *PQ*, and in the downward direction by an extension along the axis *X* at the pointed end. The modulation percentage may be found by measuring the modulated and unmodulated carrier heights, in the same way as with the wave-envelope pattern.

Non-symmetrical waveforms — In voice waveforms the average maximum amplitude in one direction from the axis frequently is greater than in the other direction, although the average *energy* on both sides is the same. For this reason the percentage of modulation in the "up" direction frequently differs from that in the "down" direction. With a given voice and microphone, this difference in modulation percentage is usually always in the same direction. Since overmodulation in the downward direction causes more out-of-channel interference than overmodulation upward because of the steeper wavefront (§ 6-1), it is advisable to "phase" the modulation so that the side of the voice waveform having the larger excursions causes the instantaneous carrier power to increase and the smaller excursions to cause a power decrease. This reduces the likelihood of overmodulation on the "down" peak. The direction of the larger excursions can readily be found by careful observation of the oscilloscope pattern. The phase can be reversed by reversing the connections of one winding of any transformer in the speech amplifier or modulator.

Modulation monitoring — While it is desirable to modulate as fully as possible, 100 per cent modulation should not be exceeded, particularly in the downward direction, because harmonic distortion will be introduced and the channel width increased (§ 5-2), thus causing unnecessary interference to other stations. The oscilloscope may be used to provide a continuous check on the modulation, but simpler indicators may be used for the purpose, once calibrated. A convenient indicator, when a Class-B modulator (§ 5-6) is used, is the plate milliammeter in the Class-B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity which gives 100 per cent modulation on voice peaks, and simultaneously observe the maximum Class-B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice in regular operation to adjust the gain so that it is not exceeded.

A sensitive rectifier-type voltmeter (copper-oxide type) also can be used for modulation monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading which represents 100 per cent modulation.

The plate milliammeter of the modulated r.f. stage may also be used as an indicator of overmodulation. Since the average plate current is constant (§ 5-3, 5-4, 5-5) when the amplifier is linear, the reading will be the same with or without modulation. When the amplifier is overmodulated, especially in the downward direction, the operation is no longer linear and the average plate current will change. A flicker of the pointer may therefore be taken as an indication of overmodulation or non-linearity. However, it is possible that the average plate current will remain constant with considerable overmodulation under some operating conditions, so that an indicator of this type is not wholly reliable unless it has been checked previously against an oscilloscope.

Linearity — The linearity (§ 5-2) of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave envelope pattern, and less auxiliary equipment is required. The connections are the same as for measuring modulation percentage (Fig. 517). If the amplifier is perfectly linear, the sloping sides of the trapezoid will be perfectly straight from the point at the axis up to at least 100 per cent modulation in the upward direction. Non-linearity will be shown by curvature of the sides. Curvature near the point, extending the point farther along the axis than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may also be caused by imperfect neutralization (a push-pull amplifier is recommended because better neutralization is possible than with single-ended amplifiers) or by r.f. leakage from the exciter through the final stage. The latter condition can be checked by removing the plate voltage from the modulated stage, when the carrier should disappear, leaving only the beam spot remaining on the screen (Fig. 518-F). If a small vertical line remains, the amplifier should be re-neutralized; if this does not eliminate the line, it is an indication that r.f. is being picked up from lower-power stages, either by coupling through the final tank or via the oscilloscope pick-up loop.

Inward curvature at the large end of the pattern is caused by improper operating conditions of the modulated amplifier, usually improper bias or insufficient excitation, or both, with plate modulation. In grid-bias and cathode-modulated systems, the bias, excitation and plate loading are not correctly proportioned when such curvature occurs, usually because the amplifier has been adjusted to have too-high carrier efficiency without modulation (§ 5-4, 5-5).

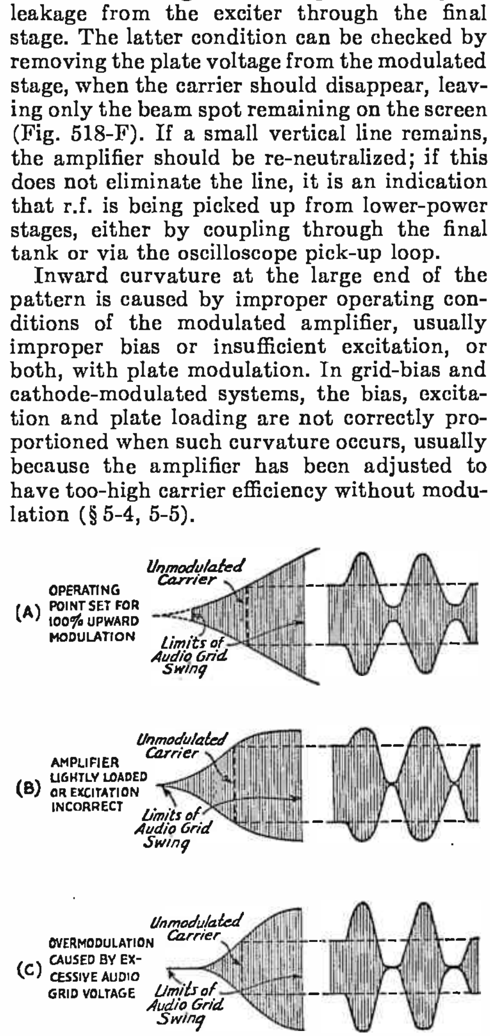


Fig. 519 — Oscilloscope patterns representing proper and improper adjustments for grid-bias or cathode modulation. The pattern obtained with a correctly adjusted amplifier is shown at A. The other drawings indicate non-linear modulation from typical causes.

For the wave-envelope pattern, it is necessary to have a linear horizontal-sweep circuit in the oscilloscope and a source of sine-wave audio signal voltage (such as an audio oscillator or signal generator) which can be synchronized with the sweep circuit. The linearity can be judged by comparing the wave envelope with a true sine wave. Distortion in the audio circuits will affect the pattern in this case (such distortion has no effect on the trapezoidal pattern, which shows the modulation characteristic of the r.f. amplifier alone), and it is also readily possible to misjudge the shape of the modulation envelope, so that the wave envelope is less useful than the trapezoid for checking linearity of the modulated amplifier.

Fig. 519 shows typical patterns of both types. The cause of the distortion is indicated for grid-bias and suppressor modulation. The patterns at A, although not truly linear, are representative of properly operated grid-bias modulation systems. Better linearity can be obtained with plate modulation of a Class-C amplifier.

Faulty patterns — The drawings of Figs. 518 and 519 show what is normally to be expected in the way of pattern shapes when the oscilloscope is used to check modulation. If the actual patterns differ considerably from those shown, it is probable that the pattern is faulty rather than the transmitter. It is important that only r.f. from the modulated stage be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding paragraph. If r.f. is present also on the horizontal plates, the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a spot where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$fd.) should be connected across the horizontal plates as close to the cathode-ray tube as possible. An r.f. choke (2.5 mh. or smaller) may also be connected in series with the ungrounded horizontal plate.

"Folded" trapezoidal patterns occur when the audio sweep voltage is taken from some point in the audio system other than that where the a.f. power is applied to the modulated stage. Such patterns are caused by a phase difference between the sweep voltage and the modulating voltage. The connections should always be as shown in Fig. 517-B.

Plate-current shift — As mentioned above, the d.c. plate current of a modulated amplifier will be the same with and without modulation so long as the amplifier operation is perfectly linear and other conditions remain unchanged. This also assumes that the modulator is working within its capabilities. Because there is usually some curvature of the modulation characteristic with grid-bias modulation there is normally a slight upward change in plate current of a stage so modulated, but this occurs only at high modulation percentages and is barely detectable under the usual conditions of voice modulation.

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1) Insufficient excitation to the modulated r.f. amplifier.
2) Insufficient grid bias on the modulated stage.
3) Wrong load resistance for the Class-C r.f. amplifier.
4) Insufficient output capacity in the filter of the modulated-amplifier plate supply.
5) Heavy overloading of the Class-C r.f. amplifier tube or tubes.

Any of the following may cause an upward shift in plate current:

1) Overmodulation (excessive audio power, audio gain too great).
2) Incomplete neutralization of the modulated amplifier.
3) Parasitic oscillation in the modulated amplifier.

When a common plate supply is used for both a Class-B (or Class-AB) modulator and a modulated r.f. amplifier, the plate current of the latter may "kick" downward because of poor power-supply voltage regulation (§ 8-1) with the varying additional load of the modulator on the supply. The same effect may occur with high-power transmitters because of poor regulation of the a.c. supply mains, even when a separate power-supply unit is used for the Class-B modulator. Either condition may be detected by measuring the plate voltage applied to the modulated stage; in addition, poor line regulation also may be detected by observing if there is any downward shift in filament or line voltage.

With grid-bias modulation, any of the following may be the cause of a plate current shift greater than the normal mentioned above:

Downward kick: Too much r.f. excitation; insufficient operating bias; distortion in modulator or speech amplifier; too-high resistance in bias supply; insufficient output capacity in plate-supply filter to modulated amplifier; amplifier plate circuit not loaded heavily enough; plate-circuit efficiency too high under carrier conditions.

Upward kick: Overmodulation (excessive audio voltage); distortion in audio system; regeneration because of incomplete neutralization; operating grid bias too high.

A downward kick in plate current will accompany an oscilloscope pattern like that of Fig. 519-B; the pattern with an upward kick will look like Fig. 519-A, with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and hum on carrier — These may be detected by listening to the signal on a receiver sufficiently removed from the transmitter to avoid overloading. The hum level should be low compared to the voice at 100 per cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing (§ 8-4). With a hum-free carrier, hum introduced by the modulator can be checked by turning on the modulator but leaving the speech amplifier off; power-supply filtering is the likely source of such hum. If carrier and modulator are both clean, connect the speech amplifier and observe the increase in hum level. If the hum disappears with the gain control at minimum, the hum is being introduced in the stage or stages preceding the gain control. The microphone also may pick up hum, a condition which can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground on the microphone and speech system usually is essential to hum-free operation.

Hum can be checked with the oscilloscope, where it appears as modulation on the carrier in the same way as the normal modulation. While the percentage usually is rather small, if the carrier shows modulation with no speech input hum is the likely cause. The various parts of the transmitter may be checked through as described above.

Spurious sidebands — A superheterodyne receiver having a crystal filter (§ 7-8, 7-11) is needed for checking spurious sidebands outside the normal communication channel (§ 5-2). The r.f. input to the receiver must be kept low enough, by removing the antenna or by adequate separation from the transmitter, to avoid overloading and consequent spurious receiver responses (§ 7-8). With the crystal filter in its sharpest position and the beat oscillator turned on, tune through the region outside the normal channel limits (3 to 4 kilocycles each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent beat notes coinciding with voice peaks, or, in bad cases of distortion or overmodulation, as "clicks" or crackles well away from the carrier frequency. Sidebands more than 4 kilocycles from the carrier should be of negligible strength in a properly modulated 'phone transmitter. The causes are overmodulation or non-linear operation (§ 5-3).

R.f. in speech amplifier — A small amount of r.f. current in the speech amplifier — particularly in the first stage, which is most susceptible to such r.f. pick-up — will cause overloading and distortion in the low-level stages. Frequently also there is a regenerative effect which causes an audio-frequency oscillation or "howl" to be set up in the audio system. In such cases the gain control cannot be advanced very far before the howl builds up, even though the amplifier may be perfectly stable when the r.f. section of the transmitter is not turned on.

Complete shielding of the microphone, microphone cord, and speech amplifier are necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Unsymmetrical or capacity coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible in that these systems sometimes cause the transmitter chassis to take an r.f. potential above ground. Inductive coupling to a two-wire transmission line is advisable. This antenna effect can be checked by disconnecting the antenna and dissipating the power in a dummy antenna (§ 4-9), when it usually will be found that the r.f. feed-back disappears. If it does not, the speech amplifier and microphone shielding are at fault.

ℂ 5-11 Frequency Modulation

Principles — In frequency modulation the carrier amplitude is constant and the output frequency of the transmitter is mad. to vary about the carrier or mean frequency at a rate corresponding to the audio frequencies of the speech currents. The extent to which the frequency changes in one direction from the unmodulated or carrier frequency is called the *frequency deviation.* It corresponds to the change of carrier amplitude in the amplitude-modulation system (§ 5-2). Deviation is usually expressed in kilocycles, and is equal to the difference between the carrier frequency and either the highest or lowest frequency reached by the carrier in its excursions with modulation. There is no modulation percentage, in the usual sense; with suitable circuit design the deviation may be made as large as desired without encountering any effect equivalent to overmodulation in the amplitude-modulated system.

Deviation ratio — The ratio of the maximum frequency deviation to the audio frequency of the modulation is called the *deviation ratio.* This ratio is also called the modulation index. Unless otherwise specified, it is taken as the ratio of the maximum frequency deviation to the *highest* audio frequency to be transmitted.

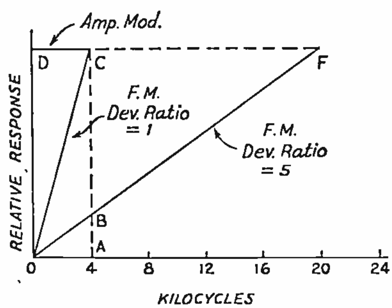


Fig. 520 — Triangular spectrum showing the noise response in an f.m. receiver compared with amplitude modulation. Deviation ratios of 1 and 5 are shown.

Advantages of f.m. — The chief advantage of frequency modulation over amplitude modulation is noise reduction at the receiver. All electrical noises in the radio spectrum, including those originating in the receiver, are r.f. oscillations which vary in amplitude, this variation causing the noise response in amplitude-modulation receivers. If the receiver does not respond to amplitude variations but only to frequency changes, noise can affect it only by causing a phase shift which appears as frequency modulation on the signal. The effect of such frequency modulation by the noise can be made small by making the frequency change (deviation) in the signal large.

A second advantage is that the power required for modulation is inconsequential, since there is no power variation in the modulated output of the transmitter.

Triangular spectrum — The way in which noise is reduced by a large deviation ratio is illustrated by Fig. 520. In this figure the noise is assumed to be evenly distributed over the channel used, an assumption which is almost always true. It is also assumed that audio frequencies above 4000 cycles (4 kc.) are not necessary to voice communication, and that the audio system in the receiver has no response above this frequency. Then, if an amplitude modulation receiver is used and its selectivity is such that there is no attenuation of side-bands (§ 5-2) below 4000 cycles, the noise components of all frequencies within the channel will produce equal response when they beat with a carrier centered in the channel. The response under these conditions is shown by the line *DC*.

In the f.m. receiver the output amplitude is proportional to the frequency deviation, and noise components in the channel can be considered to frequency-modulate the steady carrier with a deviation proportional to the difference between the actual frequency of the component and the frequency of the carrier, and also to give an audio-frequency beat of the same frequency difference. This leads to a rising response characteristic, such as the line *OC*, where the noise amplitude is proportional to the audio beat frequency. The average noise power output is proportional to the square root of the sum of the squares of all the amplitude values (§ 2-7), so that the noise power with frequency modulation having a deviation ratio of 1 is only one-third that with amplitude modulation, or an improvement of 4.75 db.

If the deviation ratio is increased to 5, the noise response is represented by the line *OF*. Since only frequencies up to 4000 cycles are reproduced in the output, however, the audible noise is confined to the triangle *OAB*. These relations hold only when the carrier is strong compared to the noise. For reception of stations with weak signal strength, the signal-to-noise ratio is better with a deviation ratio of 1.

Linearity — A transmitter in which frequency deviation is directly proportional to the amplitude of the modulating signal is said to be *linear.* It is essential also that the carrier amplitude remain constant under modulation, which in turn requires that the transmitter tuned circuits, as well as the antenna, have broad enough response to handle without discrimination the entire range of audio frequencies transmitted. This requirement is easily met under ordinary conditions.

Sidebands — In frequency modulation there is a series of sidebands on either side of the carrier frequency for each audio-frequency component in the modulation. In addition to the usual sum and difference frequencies (§ 5-2) there are also beats at harmonics of the fundamental modulating frequency, even though the latter may be a pure tone. This occurs because of the necessity for maintaining the proper phase relationships between the carrier and sidebands to keep the power output constant. Hence, a frequency-modulated signal inherently occupies a wider channel than an amplitude-modulated signal. Because of the necessity for conserving space in the usual communication spectrum, the use of f.m. by amateurs is confined to the very-high frequencies in the region above 28 Mc.

The number of sidebands for a single modulating frequency increases with the frequency deviation. When the deviation ratio is of the order of 5 the sidebands beyond the maximum frequency deviation are usually negligible, so that the channel required is approximately twice the frequency deviation.

5-12 Methods of Frequency Modulation

Requirements and methods — At present there are no fixed standards of frequency deviation in amateur work. Since a deviation ratio of 5 is considered high enough in any case, the maximum deviation necessary is 15 to 20 kc. for an upper audio-frequency limit of 3000 or 4000 cycles (§ 5-2), or a channel width of 30 to 40 kc. The permissible deviation is determined by the receiver (§ 7-18), since deviation beyond the limits of the receiver pass-band causes distortion. If the transmitter is designed to be linear (§ 5-11) with a deviation of about 15 kc., it can be used at a lower deviation ratio simply by reducing the gain in the speech amplifier. Thereby it can be made to conform to the requirements of the particular receiver in use.

The several possible methods of frequency modulation include mechanical modulation (for instance, varying condenser plate spacing in accordance with voice vibrations), initial phase-shift modulation which later is transformed into frequency modulation, and direct frequency modulation of an oscillator by electronic means. The latter, in the form of the *reactance-tube modulator,* is the simplest and most satisfactory system for amateur use.

The reactance modulator — The reactance modulator consists of a vacuum tube connected to the r.f. tank circuit of an oscillator in such a way as to act as a variable inductance or capacity, of a value dependent upon the instantaneous a.f. voltage applied to its grid. Fig. 521 is a representative circuit. The control

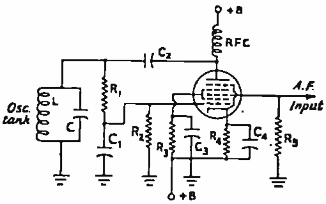


Fig. 521 — Reactance modulator circuit using a 6L7 tube.
C — Tank capacity. C_1 — 3–10 μμfd. C_2 — 250 μμfd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 0.01 μfd. L — Oscillator tank inductance.
R_1 — 50,000 ohms. R_2, R_5 — 0.5 megohm.
R_3 — 30,000 ohms. R_4 — 300 ohms.

grid circuit of the 6L7 tube is connected across the small capacity, C_1, which is in series with the resistor, R_1, across the oscillator tank circuit. Any type of oscillator circuit (§ 3-7) may be used. R_1 is large compared to the reactance (§ 2-8) of C_1, so the r.f. current through R_1C_1 will be practically in phase (§ 2-7) with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_1 will lag the current by 90 degrees (§ 2-8). The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage (§ 3-3), and consequently is 90 degrees behind the current through C_1, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank (in an inductance the current lags the voltage by 90 degrees — § 2-8). The frequency increased in proportion to the lagging plate current of the modulator, as determined by the a.f. voltage applied to the No. 3 grid of the 6L7; hence the oscillator frequency varies with the audio signal voltage.

If, on the other hand, C_1 and R_1 are reversed and the reactance of C_1 is made large compared to the resistance of R_1 the r.f. current in the 6L7 plate circuit will lead the oscillator tank r.f. voltage, making the reactance capacitive rather than inductive. In either case, the value of the applied reactance will be inversely proportional to the transconductance of the modulator tube.

Other circuit arrangements to produce the same effect may be employed. It is convenient to use a tube (such as the 6L7) in which the r.f. and a.f. voltages can be applied to separate control grids; however, both voltages may be applied to the same grid provided precautions are taken to prevent r.f. from flowing in the external audio circuit, and vice versa (§ 2-13).

The modulated oscillator usually is operated on a relatively low frequency, so that a high-order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired. The frequency deviation increases with the number of times the initial frequency is multiplied; for instance, if the oscillator is operated on 7 Mc. and the output frequency is to be 112 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 16,000 cycles at the output frequency.

Design considerations — The sensitivity of the modulator (frequency change per unit change in grid voltage) increases when C_1 is made smaller, for a fixed value of R_1, and also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier stability of the oscillator depends on the L/C ratio (§ 3-7), it is desirable to use the highest tank capacity which will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 521 is used in connection with a 7-Mc. oscillator, a linear deviation of 2000 cycles above and below the carrier frequency can be secured when the oscillator tank capacity is approximately 200 $\mu\mu$fd. A peak a.f. input of two volts is required for full deviation. At 56 Mc. the maximum deviation would be 8×2000, or 16 kc.

Since a change in any of the voltages on the modulator tube will cause a change in r.f. plate current, and consequently a frequency change, it is advisable to use a regulated plate power supply for both modulator and oscillator. At the low voltages used (250 volts), the required stabilization can be secured by means of gaseous regulator tubes (§ 8-8).

Speech amplification — The speech amplifier preceding the modulator follows ordinary design (§ 5-9), except that no power is required from it and the a.f. voltage taken by the modulator grid usually is small — not more than 10 or 15 volts, even with large modulator tubes. Because of these modest requirements, only a few speech-amplifier stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance-coupled, will suffice for crystal microphones (§ 5-8).

R.f. amplifier stages — The frequency multiplier and output stages following the modulated oscillator may be designed and adjusted in accordance with ordinary principles. No special excitation requirements are imposed, since the amplitude of the output is constant. Enough frequency multiplication must be used to give the desired maximum deviation at the final frequency; this depends upon the maximum linear deviation available from the modulator-oscillator. All stages in the transmitter should be tuned to resonance, and careful neutralization (§ 4-7) of any straight amplifier stages is necessary to prevent r.f. phase shifts which might cause distortion.

Checking operation — The two quantities to be checked in the f.m. transmitter are linearity and frequency deviation. With a modulator of the type shown in Fig. 521, both the r.f. and a.f. voltages are small enough to make the operation Class A (§ 3-4), so that the plate current of the modulator is constant so long as operation is over the linear portions of the No. 1 and No. 3 grid characteristics. Hence, non-linearity will be indicated by a change in plate current as the a.f. modulating voltage is increased. The distortion will be within acceptable limits, with the tube and constants given in Fig. 521, when the plate current does not change more than 5 per cent with signal.

Non-linearity is accompanied by a shift in the carrier frequency, so it also can be checked by means of a selective receiver such as one with a crystal filter (§ 7-11). A tone source is convenient for the test. Set the receiver for high selectivity, switch on the beat oscillator, and tune to the oscillator carrier frequency. (The check does not need to be made at the output frequency and the oscillator frequency usually is more convenient, since it will fall within the tuning range of a communications receiver.) Increase the modulating signal until a definite shift in carrier frequency is observed; this indicates the point at which non-linearity starts. The modulating signal should be kept below the level at which carrier shift is observed, for minimum distortion.

A selective receiver also can be used to check frequency deviation, again at the oscillator frequency. A source of tone of known frequency is required, preferably a continuously variable calibrated audio oscillator or signal generator. Tune in the carrier as described above, using the beat oscillator and high selectivity, and adjust the modulating signal to the maximum level at which linear operation is secured. Starting with the lowest frequency available, slowly raise the tone frequency while listening closely to the carrier beat note. As the tone frequency is raised the beat note first will decrease in intensity, then disappear entirely at a definite frequency, and finally come back and increase in intensity as the tone frequency is raised still more. The frequency at which the beat note disappears, multiplied by 2.4, is the frequency deviation at that level of modulating signal; for example, if the beat note disappears with an 800-cycle tone, the deviation is 2.4×800, or 1920 cycles. The deviation at the output frequency is the oscillator deviation multiplied by the number of times the frequency is multiplied; in this example, if the oscillator is on 7 Mc. and the output on 56 Mc., the final deviation is 1920×8, or 15.36 kc.

The output of the transmitter can be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated carrier value when the transmitter is modulated. Where there is no antenna ammeter in the transmitter, a flashlight lamp and loop can be coupled to the final tank coil to serve as a current indicator. If the carrier amplitude is constant, the lamp brilliance will not change with modulation.

Chapter Six

Keying

6-1 Keying Principles and Characteristics

Requirements — The keying of a transmitter can be considered satisfactory if the method employed reduces the power output to zero when the key is open, or "up," and permits full power to reach the antenna when the key is closed, or "down." Furthermore, the keying system should accomplish this without producing keying transients or "clicks," which cause interference with other amateur stations and with local broadcast reception, and the keying process should not affect the frequency of the emitted wave.

Back-wave — From various causes, some energy may get through to the antenna during keying spaces. The effect then is as though the dots and dashes were only louder portions of a continuous carrier; in some cases, in fact, the *back-wave*, or signal heard during the keying spaces, may seem to be almost as loud as the keyed signal. Under these conditions the keying is hard to read. A pronounced back-wave often results when the amplifier stage feeding the antenna is keyed; it may be present because of incomplete neutralization (§ 4-7) of the final stage, allowing some energy to get to the antenna through the grid-plate capacity of the tube, or because of magnetic coupling between antenna coupling coils and one of the low-power stages.

A back-wave also may be radiated if the keying system does not reduce the input to the keyed stage to zero during keying spaces. This trouble will not occur in keying systems which cut off the plate voltage when the key is open, but may be present in grid-blocking systems (§ 6-3) if the blocking voltage is not great enough and in power-supply primary keying systems (§6-3) if only the final-stage power-supply primary is keyed.

Keying waveform and sidebands — A keyed c.w. signal can be considered equivalent to a modulated signal (§ 5-1), except that, instead of being modulated by sinusoidal waves and their harmonics, it is modulated by a rectangular wave, as in Fig. 601-A. If it were modulated by a sinusoidal wave of single frequency, as in Fig. 601-B, the only sidebands would be those equal to the carrier frequency plus and minus the modulation frequency (§ 5-2). A keying speed of 50 words per minute, sending sinusoidal dots, would give sidebands only 20 cycles either side of the carrier. However, when harmonics are present in the modulation the sidebands will extend out on both sides of the signal as far as the frequency of the highest harmonic. The rectangular wave form contains an infinite number of harmonics of the keying frequency, so a carrier modulated by truly rectangular dots would have sidebands covering the entire spectrum. Actually, the high-order harmonics are eliminated because of the selectivity of the tuned circuits (§ 2-10) in the transmitter, but there still is enough energy in the lower harmonics to extend the sidebands considerably. Considered from another viewpoint, whenever a pulse of current has a steep front (or back) high frequencies are certain to be present. If the pulse can be slowed down, or caused to *lag*, through a suitable filter circuit, the highest-order harmonics are filtered out.

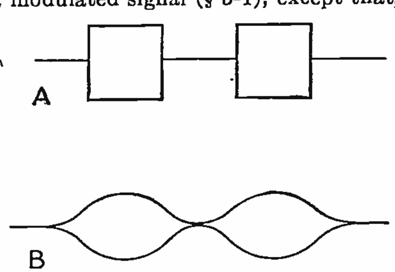


Fig. 601 — Extremes of possible keying waveshapes A, rectangular characters; B, sine-wave characters.

Key clicks — Because the high-order harmonics exist only during the brief interval when the keying character is started or ended (when the amplitude of the keying wave is building up or dying down), their effects outside the normal communication channel are observed as pulses of very short duration. These pulses are called *key clicks*.

Tests have shown that practically all operators prefer to copy a signal which is "solid" on the "make" end of each dot or dash; i.e., one that does not build up too slowly but just slowly enough to have a slight click when the key is closed. The same tests indicate that the most pleasing and least difficult signal to copy, particularly at high speeds, is one that has a fairly soft "break" characteristic; i.e., one that has practically no click as the key is opened. A signal with heavy clicks on both make and break is difficult to copy at high speeds (and also causes considerable interference), but if it is too "soft" the dots and dashes will tend to run together. It is relatively simple to adjust the keying of a transmitter so that for all normal hand speeds (15 to 40 w.p.m.) the readability will be satisfactory while the keying still will not cause interference to reception of other signals near the frequency of the transmitter.

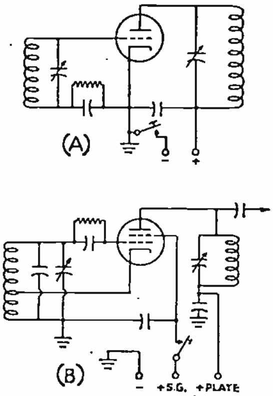


Fig. 602 — A, shows plate keying; B, screen grid keying. Oscillator circuits are shown in both cases, but the same keying methods can be used with amplifier circuits.

Break-in keying — Since, in code transmission, there are definite intervals between dots and dashes and between words, when no power is being radiated by the transmitter it is possible, with suitable keying methods, to allow the receiver to operate continuously and thus be capable of receiving incoming signals during the keying intervals. This practice facilitates communication, because the receiving operator can signal the transmitting operator, by holding down the key of his transmitter, whenever he has failed to copy part of the message, and thus obtain a repetition of the missing part without loss of time. This is called *break-in* operation.

Frequency stability — Keying should have no effect upon the output frequency of a properly designed and adjusted transmitter. However, in many instances keying will cause a "chirp," or small frequency change, at the instant of closing or opening the key, which makes the signal difficult to read. Multi-stage transmitters keyed in a stage subsequent to the oscillator usually are free from this condition, unless the keying causes line-voltage changes which in turn affect the frequency of the oscillator. When the oscillator is keyed for break-in operation, special care must be taken to insure that the signal does not have keying chirps.

Selecting the stage to key — It is advantageous from an operating standpoint to design the c.w. transmitter for break-in operation. In ordinary cases this dictates that the oscillator be keyed, since a continuously running oscillator will create interference in the receiver and thus prevent break-in operation on or near the transmitter frequency. On the other hand, it is easier to avoid a chirpy signal by keying a buffer or amplifier stage. In either case, the tubes following the keyed stage must be provided with sufficient fixed bias to limit the plate currents to safe values when the key is up and the tubes are not being excited (§ 8-9). Complete cut-off reduces the possibility of a back-wave if a stage other than the oscillator is keyed, but the keying waveform is not as well preserved and some clicks can be introduced even though the keyed stage itself produces no clicks. It is a good general rule to bias the tubes so that they draw a key-up plate current equal to about 5 per cent of the normal key-down value.

Keyed power — The power broken by the key is an important consideration, both from the standpoint of safety for the operator and that of arcing at the key contacts. Keying the oscillator or a low-power stage is favorable in both respects. The use of a keying relay is highly recommended when a high-power circuit is keyed.

ℂ 6-2 Keying Circuits

Plate-circuit keying — Any stage of the transmitter can be keyed by opening and closing the plate power circuit. Two methods are shown in Fig. 602. In A the key is in series with the negative lead from the plate power supply to the keyed stage. It could also be placed in the positive lead, although this is to be avoided whenever possible because the key is necessarily at the plate voltage above ground, and there is danger of shock unless a keying relay is used.

Fig. 602-B shows the key in the screen-supply lead of an electron-coupled oscillator. This can be considered to be a variation of plate keying.

Both the plate and screen-grid keying circuits, A and B of Fig. 602, respond well to the use of key-click filters, and are particularly suitable for use with crystal and self-controlled oscillators which are operated at low plate voltage and power input.

Power-supply keying — A variation of plate keying, in which the keying is introduced in the power-supply system itself, rather than in

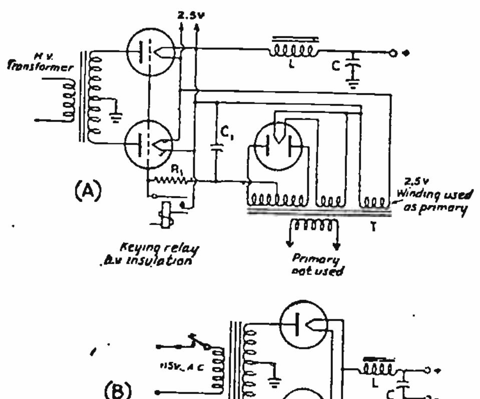


Fig. 603 — Power-supply keying. Grid-control rectifiers are used in A. Transformer *T* is a small multiple-secondary unit of the type used in receiver power supplies, and is used in conjunction with the full-wave rectifier tube to develop bias voltage for the grids of the high-voltage rectifiers. R_1 limits the load on the bias supply when the keying relay is closed; 50,000 ohms is a suitable value. C_1 may be 0.1 μfd. or larger. *L* and *C* constitute the smoothing filter for the high-voltage supply in both circuits. B shows direct keying of the transformer primary.

the connections between the power supply and transmitter, is illustrated by the diagrams in Fig. 603.

Fig. 603-A shows the use of grid-controlled rectifier tubes (§ 3-5) in the power supply. Keying is accomplished by applying suitable bias to the grids to cut off plate current flow when the key is open, and by removing the bias when the key is closed. Since in practice this circuit is used only with high-powered high-voltage supplies, a well-insulated keying relay is a necessity.

Direct keying of the primary of the plate power transformer for the keyed stage or stages is shown in Fig. 603-B. This and the method at A inherently have a keying lag because of the time constant (§ 2-6) of the smoothing filter. If enough filter is provided to reduce ripple to a low percentage (§ 8-4) the lag (§ 6-1) is too great to permit crisp keying at speeds above about 25 words per minute, although this type of keying is very effective in eliminating key clicks. A single-section plate-supply filter (§ 8-6) is about the most elaborate type that can be used if a reasonably good keying characteristic is to be achieved.

Blocked-grid keying — Keying may be accomplished by applying sufficient negative bias voltage to a control or suppressor grid to cut off plate current flow when the key is open, and by removing this *blocking* bias when the key is closed. The blocking bias voltage must be sufficient to overcome the r.f. grid voltage, in the case where the bias is applied to the control grid, and hence must be considerably higher than the nominal cut-off value for the tube at the operating d.c. plate voltage. The fundamental circuits are shown in Fig. 604.

In both circuits the key is connected in series with a resistor, R_1, which limits the current drain on the blocking-bias source when the key is closed. R_2C_1 is a resistance-capacity filter (§ 2-11) for controlling the lag on make and break of the key circuit. The lag increases as the time constant (§ 2-6) of this circuit is made larger. Since grid current flows through R_2 when the key is closed in Fig. 604-A, additional

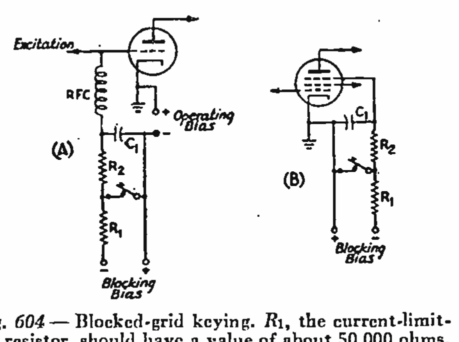


Fig. 604 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 also depends on the performance characteristic desired, values being of the order of 5000 to 10,000 ohms in most cases.

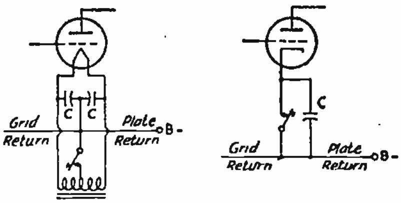


Fig. 605 — Center-tap and cathode keying. The condensers, *C*, are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

operating bias is developed, hence somewhat less bias is needed from the regular bias supply. The operating and blocking biases can be obtained from the same supply, if desired, by utilizing suitable taps on a voltage divider (§ 8-10). For circuits in which no fixed bias is used R_2 can be the regular grid leak (§ 3-6) for the stage.

With blocked-grid keying a relatively small direct current is broken as compared to other systems. Thus any sparking at the key is reduced. The keying characteristic (lag) readily can be controlled by a suitable choice of values for C_1 and R_2.

Cathode keying — Opening the d.c. circuits of both plate and grid simultaneously is called *cathode keying*. It is usually called *center-tap* keying with a directly heated filament-type tube, since in this case the key is placed in the filament-transformer center-tap lead. Typical circuits for this type of keying are shown in Fig. 605.

Cathode keying results in less sparking at the key contacts, for the same plate power, as compared with keying in the plate-supply lead. When used with an oscillator it does not respond as readily to key-click filtering (§ 6-3) as does plate keying, but there is little difference in this respect between the two systems when an amplifier is keyed.

ℂ 6-3 Key-Click Reduction

R.f. filters — A spark at the key contacts, even though minute, will cause a damped oscillation to be set up in the keying circuit which may modulate the transmitter output or may simply be radiated by the wiring in the keying circuit. Interference from the latter source is usually confined to the immediate vicinity of the transmitter, and is similar in nature and effects to the click which is frequently heard in a receiver when an electric light is turned on or off. It can be minimized by isolating the key from the wiring by means of a low-pass filter (§ 2-11), which usually consists of an r.f. choke in each key lead, placed as close as possible to the key, and by-passed on the keying-line side by a condenser, as shown in Fig. 606. Suitable values must be determined by experiment. Choke values may range from 2.5 to 80 millihenrys, and condenser capacities from 0.001 to 0.1 μfd.

This type of r.f. filter is required in nearly every keying installation, in addition to the lag circuits which are discussed in the next paragraph.

Lag circuits — A filter used to give a desired shape to the keying character, to eliminate unnecessary sidebands and consequent interference, is called a *lag circuit*. In one form, suitable for the circuits of Figs. 602 and 605, it consists of a condenser across the key terminals and an inductance in series with one of the leads. This is shown in Fig. 607. The optimum values of capacity and inductance must be found by experiment, but are not especially critical. If a high-voltage low-current circuit is being keyed a small condenser and large inductance will be necessary, while if a low-voltage high-current circuit is keyed the capacity required will be high and the inductance

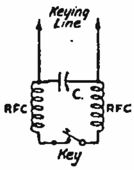


Fig. 606 — R.f. filter used for eliminating the effects of sparking at key contacts. Suitable values for best results with individual transmitters must be determined by experiment. Values for *RFC* range from 2.5 to 80 millihenries and for *C* from 0.001 to 0.1 μfd.

small. For example, a 300-volt 6-ma. circuit will require about 30 henrys and 0.05 μfd., while a 300-volt 50-ma. circuit needs about 1 henry and 0.5 μfd. For any given circuit and fixed values of current and voltage, increasing the inductance will reduce the clicks on "make" and increasing the capacity will reduce the clicks on "break."

Blocked-grid keying is adjusted by changing the values of resistors and condensers in the circuit. In Fig. 604, the click on "make" is reduced by increasing the capacity of C_1, and the click on break is reduced by increasing C_1 and/or R_2. The values required for individual installations will vary with the amount of blocking voltage and the grid current. The constants given in Fig. 604 will serve as a first approximation.

Tube keying — A tube keyer is a convenient adjunct to the transmitter, because it allows the keying characteristic to be adjusted easily without necessitating condenser and inductance values which may not be readily available. It uses the plate resistance of a tube (or tubes in parallel) to replace the key in a plate or cathode circuit, the keyer tube (or tubes) being keyed by the blocked-grid method (§ 6-2). A typical circuit is shown in Fig. 608. Type 45 tubes are suitable because of their low plate resistance and consequent small voltage drop between plate and cathode. When a tube keyer is used to replace the key in a plate or cathode circuit, the power output of the stage will be somewhat reduced because of the voltage drop across the keyer tube, but this can be compensated for by a slight increase in the supply voltage. The use of a tube keyer makes the key itself entirely safe to handle, since the high resistance in series with the key and blocking voltage prevents possible danger of shock through contact with high-voltage circuits.

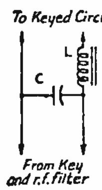
¶ 6-4 Checking Transmitter Keying

Clicks — Transmitter keying can be checked by listening to the signal on a superheterodyne receiver. The antenna should be disconnected, so that the receiver does not overload, and, if necessary, the r.f. gain may be reduced as well. Listening with the beat oscillator and a.v.c. off, the keying should be adjusted so that a slight click is heard as the key is closed but practically none can be heard when the key is released. When the keying constants have been adjusted to meet this condition, the clicks will be about optimum for all normal amateur work. If the clicks are too pronounced, they will cause interference with other amateur transmissions, and possibly to nearby broadcast receivers.

Chirps — Keying chirps (instability) may be checked by tuning in the signal or one of its harmonics on the highest frequency range of the receiver and listening with the b.f.o. on and the a.v.c. off. The gain should be sufficient to give moderate signal strength, but it should be low enough to preclude the possibility of overloading. Adjust the tuning to give a low-frequency beat note and key the transmitter. Any chirp introduced by the keying adjustment will be readily apparent. Listening to a harmonic will magnify the effect of any instability by the order of the harmonic, and thus make more perceptible.

Oscillator keying — The keying of an amplifier is relatively straightforward and requires no special treatment, but a few additional pre-

Fig. 607 — Lag circuit used for shaping the keying character to eliminate unnecessary sidebands. Actual values for any given circuit must be determined by experiment, and may range from 1 to 30 henries for *L* and from 0.05 to 0.5 μfd. for *C*, depending on the plate current.



cautions will be found necessary with oscillator keying. Any oscillator, either self-excited or crystal, will key well if it will oscillate at low plate voltages (of the order of one or two volts) and if its change in frequency with plate-voltage change is negligible. A crystal oscillator will oscillate at low plate voltages if a regenerative type of circuit such as the Tri-tet or grid-plate (§ 4–5) is used and if an r.f. choke is connected in series with the grid leak, to reduce loading on the crystal. Crystal oscillators of this type generally are free from chirp unless there is a relatively large air-gap between the crystal and top plate of the crystal holder, as is the case with a variable-frequency crystal set at the high-frequency end of its range.

Self-controlled oscillators can be made to meet the same requirements by using a high C/L ratio in the tank circuit, low plate and screen currents, and judicious feed-back adjustment (§ 3-7). A self-controlled oscillator intended to be keyed should be designed for good keying rather than maximum output.

Stages following keying — When a keying filter is being adjusted, the stages following the keyed tube should be made inoperative by removing the plate voltage. This facilitates monitoring the keying without the introduction of additional effects. The following stages should then be added, one at a time, checking the keying after each addition. An increase in click intensity (for the same carrier strength) indicates that the clicks are being added in the stages following the one being keyed. The fixed bias on such stages should be sufficient to reduce the idling plate current (no excitation) to a low value, but not to zero. Under these conditions, any instability or tendency toward parasitic oscillations, either of which can adversely affect the keying characteristic, usually will evidence itself.

Monitoring of keying — Most operators find a keying monitor helpful in developing and maintaining a good "fist," especially if a "bug" or semi-automatic key is used. While several types have been devised, the most popular consists of an audio oscillator the output of which is coupled to the receiver loud speaker or headphones, and which is keyed simultaneously with the transmitter. Fig. 609 shows the circuit diagram of a simple keying-monitor oscillator. The plate voltage, as well as the heater voltage, is supplied by a 6.3-volt filament transformer. One section of the 6F8G dual triode is used as the rectifier to supply d.c. for the plate of the second section, which is used as the oscillator. A change in the value of R_1 will alter the output tone. The output terminal labeled *Gnd* should be connected directly to the receiver chassis, while P_1 should be connected to the "hot" side of the headphones. Shunting of the 'phones by the oscillator may cause some loss of volume on received signals, unless the coupling capacity, C_3, is made sufficiently small. However, the capacity should be made large enough to provide good transfer of the oscillator signal.

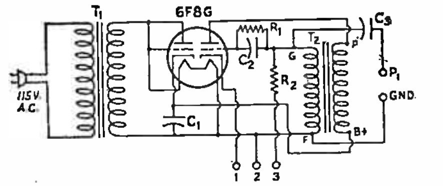


Fig. 609 — Circuit diagram of a keying monitor of the audio-oscillator type, with self-contained power supply.
C_1 — 25-μfd. 25-volt electrolytic.
C_2 — 250-μμfd. mica.
C_3 — Approximately 0.01 μfd. (see text).
R_1 — 0.15 megohm, ½ watt.
R_2 — Approximately 0.1 megohm, 1 watt (see text).
T_1 — 6.3-volt 1-ampere filament transformer.
T_2 — Small audio transformer, interstage type.

If the transmitter oscillator is keyed for break-in, the keying terminals of the oscillator may be connected in parallel with those of the transmitter. With cathode keying, terminals 1 and 2 will be connected across the key, with terminal 2 going to the ground side of the key. With blocked-grid keying, terminals 2 and 3 go to the key and a resistance of 0.1 megohm or so is inserted in series with terminal 3.

Electronic Keys—Several electronic circuits have been devised for producing automatic dots and dashes. A typical example is shown in Fig. 610. The values provide for a maximum speed of 60 w.p.m. with a 300-volt supply. R_1 and R_2 should be of the same type and ganged to form the speed control. To adjust for proper operation, ground the right cathode and adjust R_7 until the left plate current is zero. Do the same thing with the sections reversed, biasing the right section to cut-off temporarily. Adjust R_5 until the plate voltages are equal. Return the circuit to normal and check the average plate voltages with the key on the "dot" side. If they are unequal, adjust a fixed resistor connected in series with R_1 or R_2 until they are equal. On dashes, the plate voltage of the right section should drop one-third and that of the left section should increase by one-third. Adjust the size of C_3 until this condition is met. (See *QST* for March, 1944.)

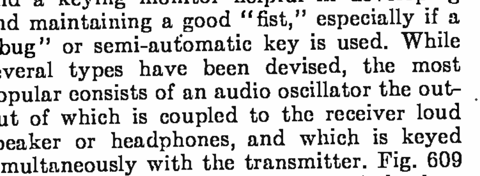
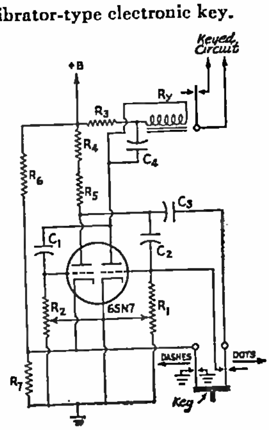


Fig. 608 — Vacuum-tube keyer circuit. The voltage drop across the tubes will be approximately 90 volts with the two Type 45 tubes shown, when the keyed current is 100 milliamperes. More tubes can be connected in parallel to reduce the drop. Suggested values are as follows:
C_1 — 2-μfd. 600-volt paper.
C_2 — 0.003-μfd. mica.
C_3 — 0.005-μfd. mica.
R_1 — 0.25 megohm, 2 watt.
R_2 — 50,000 ohms, 10 watt.
R_3, R_4 — 5 megohms, ½ watt.
R_5 — 0.5 megohm, ½ watt.
Sw_1, Sw_2 — 1-circuit 3-position rotary switch.
T_1 — Power transformer, 325 volts each side of center-tap, with 5-volt and 2.5-volt filament windings.

A wider range of lag adjustment can be obtained by using additional resistors and condensers. Suggested values of capacity, in addition to C_2 and C_3, are 0.001 and 0.002 μfd. Resistors in addition to R_2 could be 2, 2, 3 and 5 megohms. More switch positions will be required.

Fig. 610 — A multi-vibrator-type electronic key.
C_1, C_2 — 0.005 μfd. mica.
C_3 — 0.01 μfd. 400-volt paper.
C_4 — 0.01 μfd., approximately.
R_1, R_2 — 2-megohm variable (see text).
R_3, R_4 — 50,000 ohms, ½ watt.
R_5 — 3000 ohms (or resistance equal to resistance of Ry).
R_6 — 0.25 megohm, ½ watt.
R_7 — 75,000 ohms, ½ watt.
Ry — Sensitive relay (Eby).



Chapter Seven

Receiver Principles and Design

7-1 Elements of Receiving Systems

Basic requirements — The purpose of a radio receiving system is to abstract energy from passing radio waves and convert it into a form which conveys the intelligence contained in the transmitted signal. The receiver also must be able to select a desired signal and eliminate those not wanted. The fundamental processes involved are those of amplification and detection.

Detection — The high frequencies used for radio signaling are well beyond the audio-frequency range (§ 2-7), and therefore cannot be used to actuate a loudspeaker directly. Neither can they be used to operate other devices, such as relays, by means of which a message might be transmitted. The process of converting a modulated radio-frequency wave to a usable low frequency, called *detection* or *demodulation*, is essentially that of rectification (§ 3-1). The modulated carrier (§ 5-1) is thereby converted to a unidirectional current, the amplitude of which will vary at the same rate as the modulation. These low-frequency variations are readily amplified, and can be applied to the headphones, loudspeaker or other form of electromechanical device.

Code signals — The dots and dashes of code (c.w.) transmissions are rectified as described, but in themselves can produce no audible tone in the headphones or loudspeaker because they are of constant amplitude. For aural reception it is necessary to introduce a second radio frequency, differing from the signal frequency by a suitable audio frequency, into the detector circuit to produce an audible beat (§ 2-13). The frequency difference, and hence the *beat note*, is generally of the order of 500 to 1000 cycles, since these tones are within the range of optimum response of both the ear and the headset. If the source of the second radio frequency is a separate oscillator, the system is known as *heterodyne* reception; if the detector itself is made to oscillate and produce the second frequency, it is known as an *autodyne* detector.

Amplification — To build up weak signals to usable output level, modern receivers employ considerable amplification — often of the order of hundreds of thousands of times. Amplifiers are used at the frequency of the incoming signal (*r.f. amplifiers*), after detection (*a.f. amplifiers*), and, in superheterodyne receivers, at one or more intermediate radio frequencies (*i.f. amplifiers*). R.f. and i.f. amplifiers practically always employ tuned circuits.

Types of receivers — Receivers may vary in complexity from a simple detector with no amplification to multi-tube arrangements having amplification at several different radio frequencies as well as at audio frequency. A regenerative detector (§ 7-4) with or without audio-frequency amplification (§ 7-5) is known as a *regenerative receiver;* if the detector is preceded by one or more tuned r.f. amplifier stages (§ 7-6), the combination is known as a *t.r.f. (tuned radio frequency) receiver*. The *superheterodyne receiver* (§ 7-8) employs r.f. amplification at a fixed intermediate frequency as well as at the frequency of the signal itself, the latter being converted by the heterodyne process to the intermediate frequency.

At very-high frequencies the superregenerative detector (§ 7-4), usually with audio amplification, is used in the *superregenerative receiver* or *superregenerator*, providing large amplification of weak signals with simple circuit arrangements.

7-2 Receiver Characteristics

Sensitivity — Sensitivity is defined as the strength of the signal (usually expressed in microvolts) which must be applied to the input terminals of the receiver to produce a specified audio-frequency power output at the loudspeaker or headphones (§ 7-5). It is a measure of the amplification or gain of the receiver.

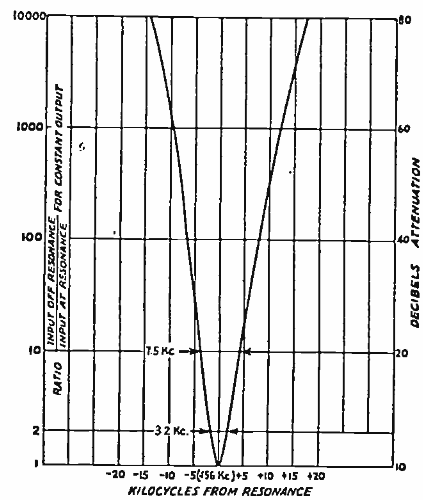


Fig. 701 — Selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios; the corresponding decibel steps are shown at the right.

Signal-to-noise ratio — Every receiver generates some noise of a hiss-like character, and signals weaker than the noise cannot be separated from it no matter how much amplification is used. This relation between noise and a weak signal is expressed by the term *signal-to-noise* ratio. It can be defined in various ways, one simple way being to give it as the ratio of signal power output to noise output from the receiver at a specified value of modulated carrier voltage applied to the input terminals.

The hiss-like noise mentioned above is inherent in the circuits and tubes of the receiver, and its amplitude depends upon the selectivity of the receiver. The greater the selectivity the smaller the noise, other things being equal (§ 7-6). In addition to inherent receiver noise, atmospheric electricity (natural "static") and electrical devices in the vicinity of the receiver also cause noise which adversely affects the signal-to-noise ratio.

Selectivity — Selectivity is the ability of a receiver to discriminate against signals of frequencies differing from that of the desired signal. The over-all selectivity will depend upon the selectivity of the individual tuned circuits and the number of such circuits.

The selectivity of a receiver is shown graphically by drawing a curve which gives the ratio of signal strength required at various frequencies off resonance to the signal strength at resonance, to give constant output. A *resonance curve* of this type (taken on a typical communications-type superheterodyne receiver) is shown in Fig. 701. The *band-width* is the width of the resonance curve (in cycles or kilocycles) of a receiver at a specified ratio; in Fig. 701, the band-widths are indicated for ratios of response of 2 and 10 ("2 times down" and "10 times down").

Selectivity for signals within a few kilocycles of the desired-signal frequency is called *adjacent-channel* selectivity, to distinguish it from the discrimination against signals considerably removed from the desired frequency.

Stability — The stability of a receiver is its ability to give constant output, over a period of time, from a signal of constant strength and frequency. Primarily, it means the ability to stay tuned to a given signal. However, a receiver which at some settings of its controls has a tendency to break into oscillation, or "howl," also is said to be unstable.

The stability of a receiver is affected principally by temperature variations, supply-voltage changes, and constructional features of a mechanical nature.

Fidelity — Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, 'phone, etc.) carried by the incoming signal. For exact reproduction the band-width must be great enough to accommodate the highest modulation frequency transmitted, and the relative amplitudes of the various frequency components within the band must not be changed in the output.

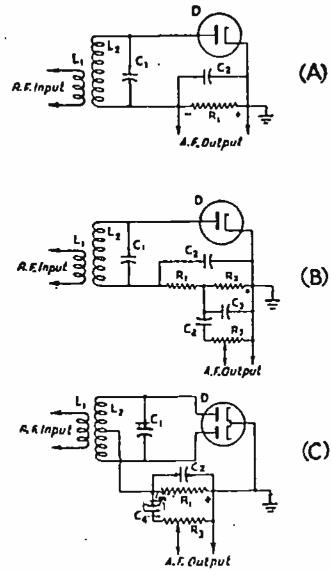


Fig. 702 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values for C_2 and R_1 in A and B are 250 $\mu\mu$fd. and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 $\mu\mu$fd. each; R_1, 50,000 ohms; and R_2, 250,000 ohms. C_4 is 0.1 μfd. and R_3 may be 0.5 to 1 megohm.

¶ 7-3 Detectors

Characteristics — The important characteristics of a detector are its sensitivity, fidelity or linearity, resistance or impedance, and signal-handling capability.

Detector *sensitivity* is the ratio of audio-frequency output to radio-frequency input. *Linearity* is a measure of the ability of the detector to reproduce, as an audio frequency, the exact form of the modulation on the incoming signal. The *resistance* or *impedance* of the detector is important in circuit design, since a relatively low resistance means that power is consumed in the detector. The *signal-handling capability* means the ability of the detector to accept signals of a specified amplitude without overloading.

Diode detectors — The simplest detector is the diode rectifier. Circuits for both half-wave and full-wave (§ 8-3) diodes are given in Fig. 702. The simplified half-wave circuit at 702-A includes the r.f. tuned circuit, L_2C_1, a coupling coil, L_1, from which the r.f. energy is fed to L_2C_1, and the diode, D, with its load resistance, R_1, and by-pass condenser, C_2. The flow of rectified r.f. current through R_1 causes a d.c. voltage to develop across its terminals, and this voltage varies with the modulation on the signal. The − and + signs show the polarity of the voltage. The variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across R_1. The load resistor, R_1, usually

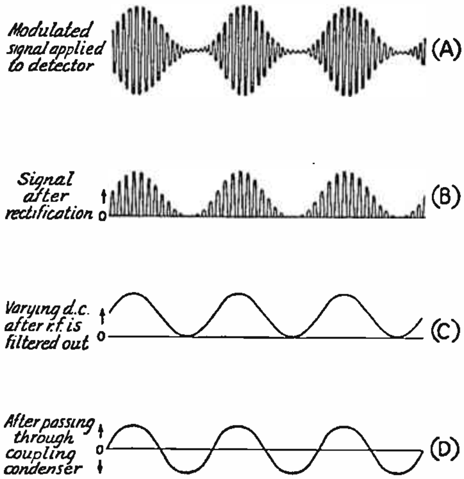


Fig. 703 — Diagrams showing the detection process.

has a rather high value of resistance, so that a fairly large voltage will develop from a small rectified-current flow.

The progress of the signal through the detector or rectifier is shown in Fig. 703. A typical modulated signal as it exists in the tuned circuit is shown at A. When applied to the rectifier tube, current flows from plate to cathode only during the part of the r.f. cycle when the plate is positive with respect to the cathode, so that the output of the rectifier consists of half-cycles of r.f. still modulated as in the original signal. These current "pulses" flow in the load circuit comprised of R_1 and C_2, the resistance of R_1 and the capacity of C_2 being so proportioned that C_2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R_1 is smoothed out, as shown in C. C_2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component which varies in the same way as the modulation on the original signal. When this varying d.c. voltage is applied to a following amplifier through a coupling condenser (C_4 in Fig. 702-B), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 702-B, R_1 and C_2 have been divided for the purpose of providing a more effective filter for r.f. It is important to prevent the appearance of any r.f. voltage in the output of the detector, because it may cause overloading of a succeeding amplifier tube. The audio-frequency variations can be transferred to another circuit through a coupling condenser, C_4 in Fig. 702, to a load resistor, R_3, which usually is a "potentiometer" (§ 8-10) so that the volume can be adjusted to a desired level.

The full-wave diode circuit at 702-C differs in operation from the half-wave circuit only in that both halves of the r.f. cycle are utilized. The full-wave circuit has the advantage that very little r.f. voltage appears across the load resistor, R_1, because the midpoint of L_2 is at the same potential as the cathode, or "ground" for r.f.

The reactance of C_2 must be small compared to the resistance of R_1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R_1 (§ 2-8, 2-13). This condition is satisfied by the values shown. If the capacity of C_2 is too large, response at the higher audio frequencies will be lowered.

Compared with other detectors, the sensitivity of the diode is low. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity (§ 2-10). The linearity is good, however, and the signal-handling capability is high.

Grid-leak detectors — The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 704-A, the grid corresponds to the diode plate and the rectifying action is exactly the same as just described. The d.c. voltage from rectified-current flow through the grid leak, R_1, biases the grid negatively with respect to cathode, and the audio-frequency variations in voltage across R_1 are amplified through the tube just as in a normal a.f. amplifier. In the plate circuit, R_2 is the plate load resistance (§ 3-3) and C_3 is a by-pass condenser to elim-

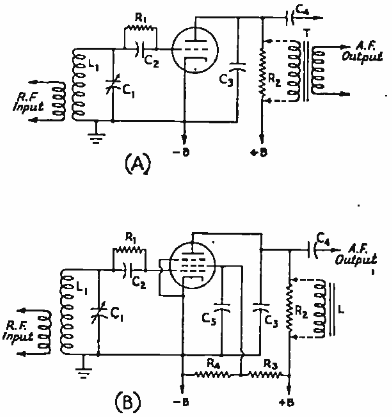


Fig. 704 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit tuned to the signal frequency. The grid leak, R_1, may be connected directly from grid to cathode instead of across the grid condenser as shown. The operation with either connection will be the same. Representative values for components are:

Component	*Circuit A*	*Circuit B*
C_2	100 to 250 μμfd.	100 to 250 μμfd.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	1 to 2 megohms.	1 to 5 megohms.
R_2	50,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.
T	Audio transformer.	
L		500-henry choke.

The plate voltage in A should be about 50 volts for best sensitivity. In B, the screen voltage should be about 30 volts and the plate voltage from 100 to 250.

inate r.f. in the output circuit. C_4 is the output coupling condenser. With a triode, the load resistor, R_2, may be replaced by an audio transformer, T, in which case C_4 is not used.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity can be further increased by using a screen-grid tube instead of a triode, as at 704-B. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, C_5, should have low reactance (§ 2-8, 2-13) for both radio and audio frequencies. R_3 and R_4 constitute a voltage divider (§ 8-10) from the plate supply to furnish the proper d.c. voltage to the screen. In both circuits, C_2 must have low r.f. reactance and high a.f. reactance compared to the resistance of R_1; the same applies to C_3 with respect to R_2.

Because of the high plate resistance of the screen-grid tube (§ 3-5), transformer coupling from the plate circuit of a screen-grid detector is not satisfactory. An impedance (L in Fig. 704-B) can be used in place of a resistor, with a gain in sensitivity because a high value of load impedance can be developed with little loss of plate voltage as compared to the voltage drop through a resistor. The coupling coil, L_2, for a screen-grid detector should have an inductance of the order of 300 to 500 henrys.

The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited.

Plate detectors — The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cut-off point, so that the application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 705. C_3 is the plate by-pass condenser, R_1 is the cathode resistor which provides the operating grid bias (§ 3-6), and C_2 is a by-pass for both radio and audio frequencies across R_1 (§ 2-13). R_2 is the plate load resistance (§ 3-3), across which a voltage appears as a result of the rectifying action described above. C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper potential (about 30 volts) to the screen, and C_5 is a by-pass condenser between screen and cathode. C_5 must have low reactance for both radio and audio frequencies.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance even of a triode is very high when the bias is set near the plate-current cut-off point (§ 3-2, 3-3). Impedance coupling may be used in place of the resistance coupling shown in Fig. 705. The same order of inductance is required as with the screen-grid detector described previously.

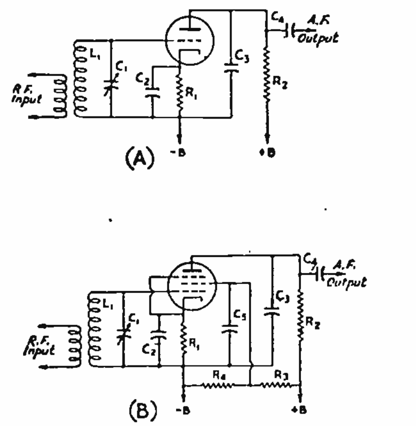


Fig. 705 — Circuits for plate detection. A, triode; B, pentode. The input circuit, L_1C_1, is tuned to the signal frequency. Typical values for the other constants are:

Component	*Circuit A*	*Circuit B*
C_2	0.5 μfd. or larger.	0.5 μfd. or larger.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	25,000 to 150,000 ohms.	10,000 to 20,000 ohms.
R_2	50,000 to 100,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.

Plate voltages from 100 to 250 volts may be used. Effective screen voltage in B should be about 30 volts.

The plate detector is more sensitive than the diode since there is some amplifying action in the tube, but less so than the grid-leak detector. It will handle considerably larger signals than the grid-leak detector, but is not quite so tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point the detector takes no power from the tuned circuit, and so does not affect its Q and selectivity (§ 2-10).

Infinite-impedance detector — The circuit of Fig. 706 combines the high signal-handling capabilities of the diode detector with low distortion (good linearity), and, like the plate detector, does not load the tuned circuit to which it is connected. The circuit resembles that of the plate detector, except that the load resistance, R_1, is connected between cathode and ground and thus is common to both grid and plate circuits, giving negative feed-back for the audio frequencies. The cathode resistor is by-passed for r.f. (C_1) but not for audio (§ 2-13), while the plate circuit is by-passed to ground for both audio and radio frequencies. R_2 forms, with C_2, an RC filter (§ 2-11) to isolate the plate from the "B" supply at a.f.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across R_1 similarly increases with signal, because of the increased plate current. Because of this and the fact that the initial drop across R_1 is large, the grid cannot be driven positive with respect to the cathode by the signal, hence no grid current can be drawn.

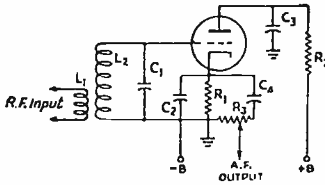


Fig. 706 — The infinite-impedance or linear detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other constants are:

C_2 — 250 μμfd. R_1 — 0.15 megohm.
C_3 — 0.5 μfd. R_2 — 25,000 ohms.
C_4 — 0.1 μfd. R_3 — 0.25-megohm volume control.

A tube having a medium amplification factor (about 20) should be used. Plate voltage should be 250 volts.

ℂ 7-4 Regenerative Detectors

Circuits — By providing controllable r.f. feed-back or regeneration (§ 3-3) in a triode or pentode detector circuit, the incoming signal can be amplified many times, thereby greatly increasing the sensitivity of the detector. Regeneration also increases the effective Q of the circuit, and hence increases the selectivity (§ 2-10) by virtue of the fact that the maximum regenerative amplification takes place only at the frequency to which the circuit is tuned. The grid-leak type of detector is most suitable for the purpose. Except for the regenerative connection, the circuit values are identical with those previously described for this type of detector, and the same considerations apply. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate (§ 3-7) and the critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned.

Fig. 707 shows the circuits of regenerative detectors of various types. The circuit of A is for a triode tube, with a variable by-pass condenser, C_3, in the plate circuit to control regeneration. When the capacity is small the tube does not regenerate, but as it increases toward maximum its reactance (§ 2-8) becomes smaller until a critical value is reached where there is sufficient feed-back to cause oscillation. If L_2 and L_3 are wound end-to-end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, L_3, when the grid connection is to the outside of L_2.

The circuit of B is for a screen-grid tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, L_3, is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5 μfd. or more) to filter out scratching noise when the arm is rotated (§ 2-11). The feed-back is adjusted by varying the number of turns on L_3 or the coupling (§ 2-11) between L_2 and L_3, until the tube just goes into oscillation at a screen voltage of approximately 30 volts.

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type (§ 3-7). Since the screen and plate are in parallel for r.f. in this circuit, only a small amount of "tickler" — that is, relatively few turns between the cathode tap and ground — is required for oscillation.

Adjustment for smooth regeneration — The ideal regeneration control would permit the detector to go into and out of oscillation smoothly, would have no effect on the frequency of oscillation, and would give the same value of regeneration regardless of frequency and the loading on the circuit. In practice, the effects of loading, particularly the loading that occurs when the detector circuit is coupled to an antenna, are difficult to overcome. Likewise, the regeneration is affected by the frequency to which the grid circuit is tuned.

In all circuits it is best to wind the tickler at the ground or cathode end of the grid coil, and to use as few turns on the tickler as will allow the detector to oscillate easily over the whole tuning range at the plate (and screen, if a pentode) voltage which gives maximum sensitivity. Should the tube break into oscillation suddenly as the regeneration control is advanced, making a click, the operation often can be made smoother by changing the grid-leak resistance to a higher or lower value. The wrong grid leak plus too-high plate and screen voltage are the most frequent causes of lack of smoothness in going into oscillation.

Antenna coupling — If the detector is coupled to an antenna, slight changes in the antenna constants (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feed-back required or the higher will be the voltage necessary to make the detector oscillate. The antenna coupling should be the maximum that will allow the detector to go into oscillation smoothly with the correct voltages on the tube. If capacity coupling (§ 2-11) to the grid end of the coil is used, only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

At frequencies where the antenna system is resonant the absorption of energy from the oscillating detector circuit will be greater, with the consequence that more regeneration is needed. In extreme cases it may not be possible to make the detector oscillate with normal voltages, causing so-called "dead spots." The remedy for this is to loosen the antenna coupling to the point which permits normal oscillation and smooth regeneration control.

Body capacity — A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (*body capacity*) can be caused by poor design of the receiver, or by the antenna if the detector is coupled directly to it. If body capacity is present when the antenna is disconnected, it can be eliminated by better shielding, and sometimes by r.f. filtering of the 'phone leads. Body capacity which is present only when the antenna is connected is caused by resonance effects in the antenna, which tend to cause a portion of a standing wave (§ 2-12) of r.f. voltage to appear on the ground lead and thus raise the whole detector circuit above ground potential. A good, short ground connection should be made to the receiver and the length of the antenna varied electrically (by adding a small coil or variable condenser in the antenna lead) until the effect is minimized. Loosening the coupling to the antenna circuit also will help.

Hum — Hum at the power-supply frequency may be present in a regenerative detector, especially when it is used in an oscillating condition for c.w. reception, even though the plate supply itself is free from ripple (§ 8-4). The hum may result from the use of a.c. on the tube heater, but effects of this type normally are troublesome only when the circuit of Fig. 707-C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, and grid condenser and leak are not electrostatically shielded. This type of hum is easily recognizable because of its rather high pitch, a result of harmonics (§ 2-7) in the power-supply system. The hum is caused by a species of grid modulation (§ 5-4).

Antenna resonance effects frequently cause a hum of the same nature as that just described which is most intense at the various resonance points, and hence varies with tuning. For this reason it is called tunable hum. It is prone to occur with a rectified a.c. plate supply (§ 8-1) when a standing wave effect of the type described in the preceding paragraph occurs, and is associated with the non-linearity of the rectifier tube in the plate supply. Elimination of antenna resonance effects as described and by-passing the rectifier plates to cathode (using by-pass condensers of the order of 0.001 μfd.) usually will cure it.

Tuning — For c.w. reception, the regeneration control is advanced until the detector breaks into a "hiss," which indicates that the detector is oscillating. Further advancing the regeneration control after the detector starts oscillating will result in a slight decrease in the strength of the hiss, indicating that the sensitivity of the detector is decreasing.

The proper adjustment of the regeneration control for best reception of c.w. signals is where the detector just starts to oscillate, when it will be found that c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal the tone first will be heard as a very high pitch, then will go down through "zero beat" (the region where the frequencies of the incoming signal and the oscillating detector are so nearly alike that the difference or beat is less than the lowest audible tone) and rise again on the other side, finally disappearing at a very high pitch. This behavior is shown in Fig. 708. It will be found that a low-pitched beat-note cannot be obtained from a strong signal because the detector "pulls in" or "blocks"; that is, the signal tends to control the detector in such a way that the latter oscillates at the signal frequency, despite the fact that the circuit may not be tuned exactly to resonance. This phenomenon, commonly observed when an oscillator is coupled to a source of a.c. voltage of approximately the

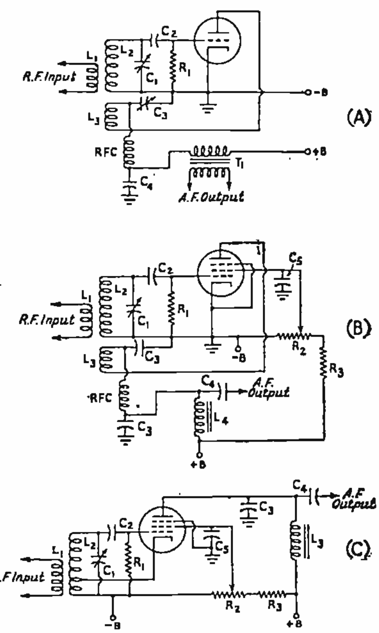


Fig. 707 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 μμfd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily will have from 10 to 25 per cent of the number of turns on L_2; in C, the cathode tap is about 10 per cent of the number of turns on L_2 above ground. Regeneration control condenser C_3 in A should have a maximum capacity of 100 μμfd. or more; by-pass condensers C_3 in B and C are likewise 100 μμfd. C_5 is ordinarily 1 μfd. or more; R_2, a 50,000-ohm potentiometer; R_3, 50,000 to 100,000 ohms. L_4 in B (L_3 in C) is a 500-henry inductance, C_4 is 0.1 μfd. in both circuits. T_1 in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. *RFC* is 2.5 mh. In A, the plate voltage should be about 50 volts for best sensitivity. Pentode circuits require about 30 volts on the screen; plate voltage may be 100 to 250 volts.

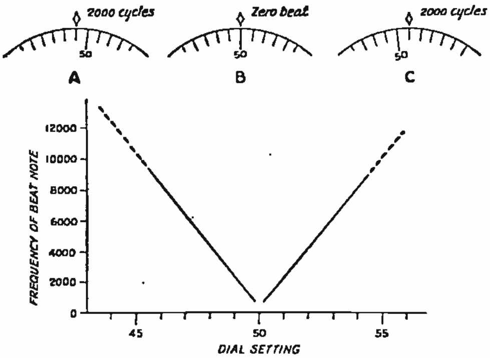


Fig. 708 — As the tuning dial of a receiver is turned past a c.w. signal, the beat note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.

frequency at which the oscillator is operating, is called "locking-in"; the more stable of the two frequencies assumes control over the other. "Blocking" usually can be corrected by advancing the regeneration control until the beat-note occurs again. If the regenerative detector is preceded by an r.f. amplifier stage, the blocking can be eliminated by reducing the gain of the r.f. stage. If the detector is coupled to an antenna, the blocking condition can be eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the receiver starts oscillating is the most sensitive condition for c.w. reception. Further advancing the regeneration control makes the receiver less prone to blocking by strong signals, but also less capable of receiving weak signals.

If the receiver is in the oscillating condition and a 'phone signal is tuned in, a steady audible beat-note will result. While it is possible to listen to 'phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is also the most sensitive operating point.

Superregeneration — The limit to which ordinary regenerative amplification can be carried is the point at which oscillations commence, since at that point further amplification ceases. The *superregenerative* detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range (of the order of 20 to 200 kilocycles), in such a way as to vary the detector's operating point (§ 3-3). As a consequence of the introduction of this *quench* or *interruption* frequency, the detector can oscillate only when the varying operating point is in a region suitable for the production of oscillations. Because the oscillations are constantly being interrupted, the regeneration can be greatly increased, and the amplified signal will build up to tremendous proportions. A one-tube superregenerative detector is capable of an inherent sensitivity approaching the thermal-agitation noise level of the tuned circuit, and may have an antenna input sensitivity of two microvolts or better.

Because of its inherent characteristics, the superregenerative circuit is suitable only for the reception of modulated signals, and operates best on the very-high frequencies. Typical superregenerative circuits for the very-high frequencies are shown in Fig. 709.

The basic regenerative detector circuit is the ultraudion oscillator (§ 3-7). In Fig. 709-A the quench frequency is obtained from a separate oscillator and introduced into the plate circuit of the detector. The quench oscillator, operating at a low radio frequency, alternately allows oscillations to build up in the regenerative circuit and then causes them to die out. In the absence of a signal, the thermal agitation noise in the input circuit produces the voltage that initiates the build-up process. However, when an incoming signal provides the initiating pulse, it has the effect of advancing the starting time of the oscillations. This causes the area within the envelope to increase, as indicated in Fig. 710-C.

If regeneration in an ordinary regenerative circuit is carried sufficiently far, the circuit will break into a low-frequency oscillation simultaneously with that at the operating radio frequency. This low-frequency oscillation has much the same quenching effect as that from a separate oscillator, hence a circuit so operated is called a *self-quenching* superregenerative detector. The frequency of the quench oscillation depends upon the feed-back and upon the time constant of the grid leak and condenser, the oscillation being a "blocking" or "squegging" in which the grid accumulates a strong negative charge which does not leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

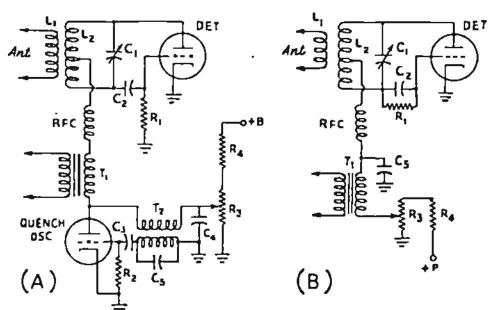


Fig. 709 — (A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:

C_2 — 50 μμfd.
C_3 — 500 μμfd.
C_4 — 0.1 μfd.
C_5 — 0.001–0.005 μfd.
R_1 — 2-10 megohms.
R_2 — 50,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 50,000 ohms.
T_1 — Audio transformer, plate-to-grid type.
RFC — R.f. choke, value depending upon frequency. Small low-capacity chokes are required for v.h.f. operation.

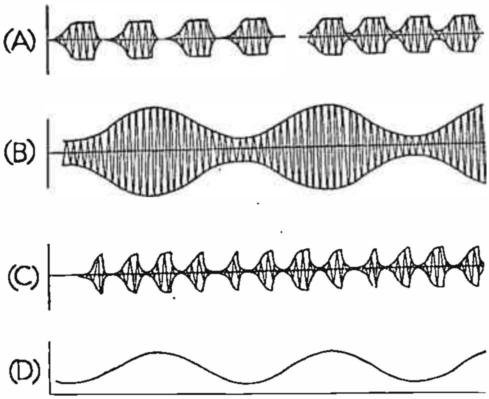


Fig. 710 — R.f. oscillation envelopes in a self-quenched superregenerative detector. Without signal (A at left) oscillations are completely quenched after each period, resuming in random phase depending on momentary noise voltages. At right, when the initiating pulses are supplied by a received signal the starting time of the oscillations is advanced causing the build-up period to begin before damping is complete. This advance is proportional to the carrier amplitude when modulated (B). Since the building-up period varies in accordance with modulation (C), when these wave trains are rectified the average rectified current is proportional to the amplitude of the signal. Amplitude modulation is therefore reproduced as an audio wave in the output circuit (D).

The greater the difference between the quenching and signal frequencies the greater the amplification, because the signal then has a longer period in which to build up during the nonquenching half-cycle when the resistance of the circuit is negative. This ratio should not exceed a certain limit, however, for during the quenched or nonregenerative intervals the input selectivity is merely that of the Q of the tuned circuit alone. The optimum quench frequency is in the neighborhood of 150 kc. for the 60-Mc. band and 250 kc. for 112 Mc.

The superregenerative detector has relatively little selectivity as compared to a regular regenerative detector, but discriminates against noise such as ignition interference. It also has marked a.v.c. action, strong signals being amplified much less than weak signals.

Adjustment of superregenerative detectors — Because of the greater amplification, the hiss noise when a superregenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in, the hiss will disappear to a degree which depends upon the signal strength.

Lack of hiss indicates insufficient feed-back at the signal frequency, or inadequate quench voltage. Antenna loading effects will cause dead spots which are similar to those in regenerative detectors and can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid leak and grid condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

7-5 Audio-Frequency Amplifiers

General — The ordinary detector does not produce very much audio-frequency power output — usually not enough to give satisfactory sound volume, even in headphone reception. Consequently, audio-frequency amplifiers are used after the detector to increase the power level. One amplifier usually is sufficient for headphones, but two stages generally are used where the receiver is to operate a loudspeaker. A few milliwatts of a.f. power is sufficient for headphones, but a loudspeaker requires a watt or more for good room volume.

In all except battery-operated receivers, the negative grid bias of audio amplifiers usually is secured from the voltage drop in a cathode resistor (§ 3-6). The cathode resistor must be by-passed by a condenser having low reactance at the lowest audio frequency to be amplified, compared to the resistance of the cathode resistor (10 per cent or less) (§ 2-8, 2-13). In battery-operated receivers, a separate grid-bias battery generally is used.

Headset and voltage amplifiers — The circuits shown in Fig. 711 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones (§ 3-3). Triodes usually are preferred to pentodes because they are better suited to working into an audio transformer or headset, the input impedances of which are of the order of 20,000 ohms.

In these circuits, R_2 is the cathode bias resistor and C_1 the cathode by-pass condenser. The grid resistor, R_1, gives volume control action (§ 5-9). Its value ordinarily is from 0.25 to 1 megohm. C_2 is the input coupling condenser, already discussed under detectors; it is, in fact, identical to C_4 in Figs. 704 and 705, if the amplifier is coupled to a detector.

Power amplifiers — A popular type of power amplifier is the single pentode, operated Class A or AB; the circuit diagram is given in Fig. 711-A. The grid resistor, R_1, may be a potentiometer for volume control, as shown at

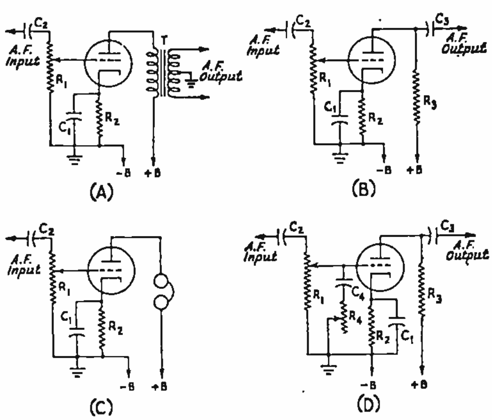


Fig. 711 — Audio amplifier circuits used for voltage amplification and to provide power for headphone output. The tubes are operated as Class-A amplifiers (§ 3-4).

R_1 in Fig. 711. The output transformer, T, should have a turns ratio (§ 2-9) suitable for the loudspeaker used; many of the small loudspeakers now available are furnished complete with output transformer.

When greater volume is needed, a pair of pentodes or tetrodes may be connected in push-pull (§ 3-3), as shown in Fig. 712-B. Transformer coupling to the voltage-amplifier stage is the simplest method of obtaining push-pull input for the amplifier grids. The interstage transformer, T_1, has a center-tapped secondary with a secondary-to-primary turns ratio of about 2 to 1. An output transformer, T_2, with a center-tapped primary must be used. No by-pass condenser is needed across the cathode resistor, R, since the a.f. current does not flow through the resistor as it does in single-tube circuits (§ 3-3).

Tone control — A tone control is a device for changing the frequency response (§ 3-3) of an audio amplifier; usually it is simply a method for reducing high-frequency response. This is helpful in reducing hissing and crackling noises without disturbing the intelligibility of the signal. R_4 and C_4, in Fig. 711-D, together form an effective tone control of this type. The maximum effect is secured when the resistance of R_4 is entirely out of the circuit, leaving C_4 connected directly between grid and ground. R_4 should be large compared to the reactance of C_4 (§ 2-8) so that when its resistance is all in circuit the effect of C_4 on the frequency response is negligible.

Headphones and loudspeakers — Two types of headphones are in general use, the *magnetic* and *crystal* types. They are shown in cross-section in Fig. 713. In the magnetic type the signal is applied to a coil or pair of coils having a great many turns of fine wire wound on a permanent magnet. (Headphones having one coil are known as the "single-pole" type, while those having two coils, as shown in Fig. 713, are called "double-pole.") A thin circular diaphragm of iron is placed close to the open ends of the magnet. It is tightly clamped by the earpiece assembly around its circumference, and the center is drawn toward the permanent magnet under some tension. When an alternating current flows through the windings the field set up by the current alternately aids and opposes the steady field of the permanent magnet, so that the diaphragm alternately is drawn nearer to and allowed to spring farther away from the magnet. Its motion sets the air into corresponding vibration. Although the d.c. resistance of the coils may be of the order of 2000 ohms, the a.c. impedance of a magnetic type headset will be of the order of 20,000 ohms at 1000 cycles.

In the crystal headphone, two piezoelectric crystals (§ 2-10) of Rochelle salts are cemented together in such a way that the pair tends to be bent in one direction when a voltage of a certain polarity is applied and to bend in the other direction when the polarity is reversed. The crystal unit is rigidly mounted to the earpiece, with the free end coupled to a diaphragm. When an alternating voltage is applied, the alternate bending as the polarity of the applied voltage reverses makes the diaphragm vibrate back and forth. The impedance is several times that of the magnetic type.

Magnetic-type headsets tend to give maximum response at frequencies of the order of 500 to 1000 cycles, with a considerable reduction of response (for constant applied voltage) at frequencies both above and below this region. The crystal type has a "flatter" frequency-response curve, and is particularly good at reproducing the higher audio frequencies. The peaked response curve of the magnetic type is advantageous in code reception, since it tends to reduce interference from signals having beat tones lying outside the region of maximum response, while the crystal type is better for the reception of voice and music. Magnetic headsets can be used in circuits in which d.c. is flowing, such as the plate circuit of a vacuum tube, providing the current is not too large to be carried safely by the wire in the coils; the limit is a few milliamperes. Crystal headsets must be used only on a.c. (since a steady d.c. voltage will damage the crystal unit), and consequently must be coupled to the tube through a device, such as a condenser, which isolates the d.c. voltage but permits the passage of an alternating current.

The most common type of loudspeaker is the *dynamic* type, shown in cross-section in Fig. 713. The signal is applied to a small coil (the *voice coil*) which is free to move in the gap between the ends of a magnet. The magnet is made in the form of a cylindrical coil slightly smaller than the form on which the voice coil is wound, with the magnetic circuit completed through a pole piece which fits around the outside of the voice coil leaving just enough clearance for free movement of the coil. The path of the flux through the magnet is as shown by the dotted lines in the figure.

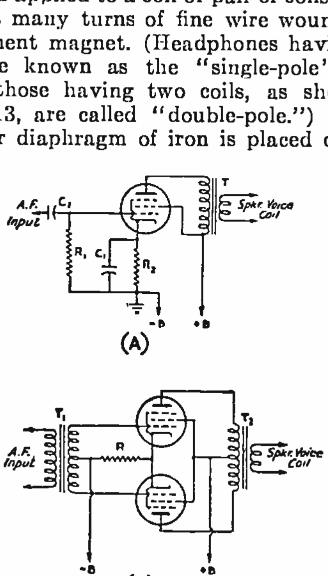


Fig. 712 — Power-output audio amplifier circuits. Either Class A or AB amplification (§ 3-4) may be used.

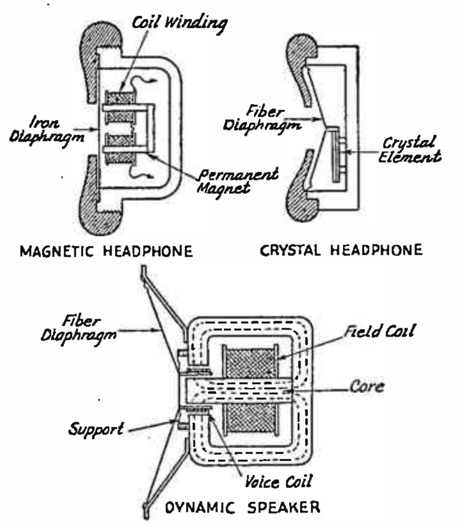


Fig. 713 — Headphone and loudspeaker construction.

The voice coil is supported so that it is free to move along its axis but not in other directions, and is fastened to a fiber or paper conical diaphragm. When current is sent through the coil it moves in a direction determined by the polarity of the current (§ 2-5), and thus moves back and forth when an alternating voltage is applied. The motion is transmitted by the diaphragm to the air, setting up sound waves.

The type of speaker shown in Fig. 713 obtains its fixed magnetic field by electromagnetic means, direct current being sent through the *field coil* for this purpose. Other types use permanent magnets to replace the electromagnet, and hence do not require a source of d.c. power. The voice coils of dynamic speakers have few turns and therefore low impedance, values of 3 to 15 ohms being representative.

ℂ 7-6 Radio-Frequency Amplifiers

Circuits — Although there may be variations in detail, practically all r.f. amplifiers conform to the basic circuit shown in Fig. 714. A screen-grid tube, usually a pentode, is used, since a triode will oscillate when its grid and plate circuits are tuned to the same frequency (§ 3-5). The amplifier operates Class A, without grid current (§ 3-4). The tuned grid circuit, L_1C_1, is coupled through L_2 to the antenna (or, in some cases, to a preceding stage). R_1 and C_2 are the cathode bias resistor and by-pass condenser, C_3 is the screen by-pass condenser, and R_2 is the screen dropping resistor. L_3 is the primary of the output transformer (§ 2-11), tightly coupled to L_4, which, with C_5, constitutes the tuned circuit feeding the detector or following amplifier. The input and output circuits, L_1C_1 and L_4C_5, are both tuned to the signal frequency.

Shielding — The screen-grid construction of the amplifier tube prevents feed-back (§ 3-3) from plate to grid inside the tube, but in addition it is necessary to prevent transfer of energy from the plate circuit to the grid circuit external to the tube. This is accomplished by enclosing the coils in grounded shielding containers and by keeping the plate and grid leads well separated. With "single-ended" tubes, care in laying out the wiring to obtain the maximum possible physical separation between plate and grid leads is necessary to prevent capacity coupling.

The shield around a coil will reduce the inductance and Q of the coil (§ 2-11) to an extent which depends upon the shielding material and the distance it is placed from the coil. Adjustments therefore must be made with the shield in place.

By-passing — In addition to shielding, good by-passing (§ 2-13) is imperative. This is not simply a matter of choosing the proper type and capacity of by-pass condenser. Short *separate* leads from C_3 and C_4 to cathode or ground are a prime necessity. At the higher radio frequencies even an inch of wire will have enough inductance to provide feed-back coupling, and hence cause oscillation, if the wire happens to be common to both the plate and grid circuits.

Gain control — The gain of an r.f. amplifier usually is varied by varying the grid bias. This method works best with variable-μ type tubes (§ 3-5), hence this type usually is found in r.f. amplifiers. In Fig. 714, R_3 and R_4 comprise the gain-control circuit. R_3 is the control resistor (§ 3-6) and R_4 a dropping resistor of such value as to make the voltage across the outside terminals of R_3 about 50 volts (§ 8-10). The gain is maximum with the variable arm on R_3 all the way to the left (grounded), and minimum at the right. R_3 could simply be placed in series with R_1, omitting R_4 entirely, but the range of control with this connection is limited because it depends on the cathode current alone.

In a multi-tube receiver the gain of several stages may be varied simultaneously, a single control sufficing for all. The lower ends of the several cathode resistors (R_1) are then connected together and to the movable contact on R_3 in Fig. 714.

Circuit values — The value of the cathode resistor, R_1, should be calculated for the *minimum* recommended bias for the tube used. The capacities of C_2, C_3 and C_4 must be such that the reactance is low at radio frequencies; this condition is easily met by using 0.01-μfd. condensers at communication frequencies, or 0.001 to 0.002 mica units at very-high fre-

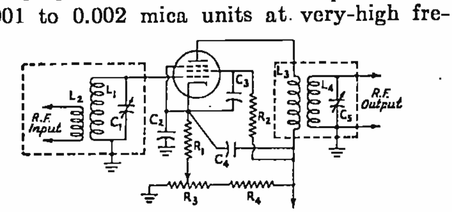


Fig. 714 — Basic circuit of a tuned radio-frequency amplifier. Component values are discussed in the text.

quencies up to 112 Mc. R_2 is found by taking the difference between the recommended plate and screen voltages, then substituting this and the rated screen current in Ohm's Law (§ 2-6). R_3 must be selected on the basis of the number of tubes to be controlled; a resistor must be chosen which is capable of carrying, at its low-resistance end, the sum of all the tube currents plus the bleeder current. A resistor of suitable current-carrying capacity being found, the bleeder current necessary to produce a drop through it of about 50 volts can be calculated by Ohm's Law. The same formula will give R_4, using the plate voltage less 50 volts for E and the bleeder current previously found for I.

The constants of the tuned circuits will depend upon the frequency range, or band, to be covered. A fairly high L/C ratio (§ 2-10) should be used on each band; this is limited, however, by the irreducible minimum capacities. To an allowance of 10 to 20 $\mu\mu$fd. for tube and stray capacities should be added the minimum capacity of the tuning condenser.

If the input circuit of the amplifier is connected to an antenna, the coupling coil, L_2, should be adjusted to provide critical coupling (§ 2-11) between the antenna and grid circuit. This will give maximum energy transfer. The turns ratio of L_1/L_2 will depend upon the frequency, the type of tube used, the Q of the tuned circuit and the constants of the antenna system, and in general is best determined experimentally. The selectivity will increase as the coupling is reduced below this "optimum" value, a consideration which it is well to keep in mind if selectivity is of more importance than maximum gain.

The output-circuit coupling depends upon the plate resistance (§ 3-2) of the tube, the input resistance of the succeeding stage, and the Q of the tuned circuit, L_4C_5. L_3 usually is coupled as closely as possible to L_4 (avoiding the necessity for an additional tuning condenser across L_3) and the energy transfer is maximum when L_3 has ⅔ to ⅘ as many turns as L_4, with ordinary receiving pentodes.

Tube and circuit noise — In any conductor electrons will be moving in random directions simultaneously and, as a result, small irregular voltages are developed across the conductor terminals. The voltage is larger the greater the resistance of the conductor and the higher its temperature. This is known as the *thermal-agitation* effect, and it produces a hiss-like noise voltage distributed uniformly throughout the radio-frequency spectrum. The thermal-agitation noise voltage appearing across the terminals of a tuned circuit will be the same as in a resistor of a value equal to the parallel impedance (§ 2-10) of the tuned circuit, even though the actual circuit resistance is low. Hence, the higher the Q of the circuit, the greater the thermal agitation noise.

Another component of hiss noise is developed in the tube because the rain of electrons on the plate is not entirely uniform. Small irregularities caused by gas in the tube also contribute to the effect. Tube noise varies with the type of tube; in general, the higher the cathode current and the lower the mutual conductance of the tube, the more internal noise it will generate.

To obtain the best signal-to-noise ratio, the signal must be made as large as possible at the grid of the tube, which means that the antenna coupling must be adjusted to that end and also that the Q of the grid tuned circuit must be high. A tube with low inherent noise obviously should be chosen. In an amplifier having good signal-to-noise ratio, the thermal-agitation noise will be greater than the tube noise. This can easily be checked by disconnecting the antenna so that no outside noise is being introduced into the receiver, then grounding the grid through a 0.01-μfd. condenser and observing whether there is a decrease in noise. If there is no change the tube noise is greatly predominant, indicating a poor signal-to-noise ratio in the stage. The test is valid only if there is no regeneration in the amplifier. The signal-to-noise ratio will decrease as the frequency is raised, because it becomes increasingly difficult to obtain a tuned circuit of high effective Q (§ 7-7).

The first stage of the receiver is the important one from the standpoint of signal-to-noise ratio. Noise generated in the second and subsequent stages, while comparable in magnitude to that generated in the first, is masked by the amplified noise and signal from the first stage. After the second stage, further contributions by tubes and circuits to the total noise are inconsequential in any normal receiver.

Tube input resistance — At high radio frequencies the tube may consume power from the tuned grid circuit, even though the grid is not driven positive by the signal. Above 7 Mc. all tubes "load" the tuned circuit to some extent, the amount of loading varying with the type of tube. This effect comes about because of the transit time necessary for electrons to travel from the cathode to the grid becomes comparable to the time of one r.f. cycle, and because of the degenerative effect (§ 3-3) of the cathode lead inductance. It becomes more pronounced as the frequency is increased. Certain types of tubes may have an input resistance of only a few thousand ohms at 28 Mc. and as little as a few hundred ohms at very-high frequencies. The input resistance of the same tubes at 7 Mc. and lower frequencies may be so high as to be considered infinite.

This *input-loading* effect is in addition to the normal decrease in the Q of the tuned circuit alone, because of increased losses in the coil and condenser at the higher frequencies. Thus the selectivity and gain of the circuit both are affected adversely by increasing frequency.

Comparison of tubes — At 7 Mc. and lower frequencies, the signal-to-noise ratio, gain, and selectivity of an r.f.-amplifier stage are sufficiently high with any of the standard receiving tubes. At 14 Mc. and higher, however, this is no longer true, and the choice of a tube must be based on several conflicting considerations.

Gain is highest with high mutual-conductance pentodes, the 6AB7 and 6AC7 being examples of this type. These tubes also develop less noise than any of the others. The input-loading effect is greatest with them, however, so that selectivity is decreased and the tuned-circuit gain is lowered.

Pentodes, such as the 6K7, 6J7 and corresponding types in glass, have lesser input-loading effects at high frequencies, moderate gain, and relatively high inherent noise.

"Acorn" and equivalent miniature pentodes are excellent from the input-loading standpoint; gain is about the same as with standard types, and the inherent noise is somewhat lower.

Where selectivity is paramount the acorns are best, the standard pentodes second, and the 6AB7–6AC7 types worst. On signal-to-noise ratio the latter tubes are first, acorns are second and standard pentodes third. The same order of precedence holds for over-all gain.

At 56 Mc. the standard types are usable, but acorns are capable of better performance because of lesser loading. The 954 and 956 and the corresponding types, 9001 and 9003, are practically the only usable types for r.f. amplification at 112 Mc. and higher.

¶ 7-7 Tuning and Band-Changing Methods

Band-changing — The resonant circuits which are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers, since the amateur frequency assignments consist of groups or bands of frequencies at widely spaced intervals. The same *LC* combination cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for changing the circuit constants for various frequency bands. As a matter of convenience the same tuning condenser usually is retained, but new coils are inserted in the circuit for each band.

There are two favorite methods of changing inductances. One is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. The second is to use coils wound on forms with contacts (usually pins) which can be plugged in and removed from a socket.

Bandspreading — The tuning range of a given coil and variable condenser will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called *bandspreading*. Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 715.

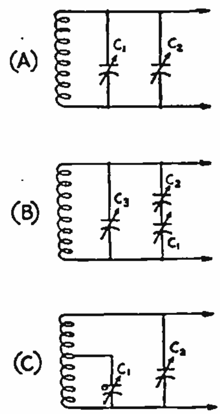


Fig. 715 — Essentials of the three basic bandspread tuning systems.

In A, a small *bandspread condenser*, C_1 (15 to 25 μμfd. maximum capacity), is used in parallel with a condenser, C_2, which is usually large enough (140 to 175 μμfd.) to cover a 2-to-1 frequency range. The setting of C_2 will determine the minimum capacity of the circuit, and the maximum capacity for bandspread tuning will be the maximum capacity of C_1 plus the setting of C_2. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. In practicable circuits it is almost impossible, because of the non-harmonic relation of the various bands, to get full bandspread on all bands with the same pair of condensers, especially when the coils are wound to give continuous frequency coverage on C_2, which is variously called the *band-setting* or *main-tuning* condenser. C_2 must be reset each time the band is changed.

The method shown at B makes use of condensers in series. The tuning condenser, C_1, may have a maximum capacity of 100 μμfd. or more. The minimum capacity is determined principally by the setting of C_3, which usually has low capacity, and the maximum capacity by the setting of C_2, which is of the order of 25 to 50 μμfd. This method is capable of close adjustment to practically any desired degree of bandspread. Either C_2 and C_3 must be adjusted for each band or separate pre-adjusted condensers must be switched in.

The circuit at C also gives complete spread on each band. C_1, the bandspread condenser, may have any convenient value of capacity; 50 μμfd. is satisfactory. C_2 may be used for continuous frequency coverage ("general coverage") and as a band-setting condenser. The effective maximum-minimum capacity ratio depends upon the capacity of C_2 and the point at which C_1 is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the bandspread, and vice versa. For a given coil and tap, the bandspread will be greater if C_2 is set at larger capacity. C_2 may be mounted in the plug-in coil form and pre-set, if desired. This requires a separate condenser for each band, but eliminates the necessity for resetting C_2 each time the band is changed.

Ganged tuning — The tuning condensers of the several r.f. circuits may be coupled together mechanically and operated by a single control. However, this operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits *track* — that is, tune to the same frequency at each setting of the tuning control.

True tracking can be obtained only when the inductance, tuning condensers, and circuit minimum and maximum capacities are identical in all "ganged" stages. A small *trimmer* or *padding* condenser may be connected across the coil, so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 716, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer necessarily increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget condensers having maximum capacities of 15 to 30 μμfd. are commonly used.

The same methods are applied to bandspread circuits which must be tracked. The circuits are identical with those of Fig. 715. If both general-coverage and bandspread tuning are to be available, an additional trimmer condenser must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then C_3 in Fig. 715-B, and C_2 in Fig. 715-C serve as trimmers.

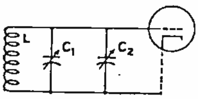


Fig. 716 — Showing the use of a trimmer condenser, to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

The coil inductance can be adjusted by starting with a larger number of turns than necessary and removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method, provided the inductance is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole coil, or to use a single short-circuited turn the position of which can be varied with respect to the coil. The application of these methods is shown in Fig. 717.

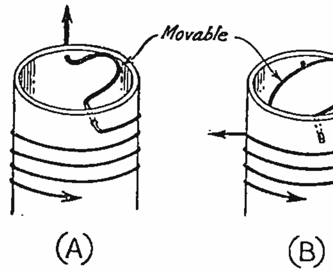


Fig. 717 — Methods of adjusting the inductance for ganging. The half turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the plane of the loop is parallel to the axis of the coil, and will give maximum reduction of the coil inductance when perpendicular to the coil axis.

V.h.f. circuits — Interelectrode capacities are practically constant for a given tube regardless of the operating frequency, and the same is approximately true of stray circuit capacities. Hence, at very-high frequencies these capacities become an increasingly larger part of the usable tuning capacity, and reasonably high L/C ratios (§ 2-10) are more difficult to secure as the frequency is raised. Because of this irreducible minimum capacity, standard types of tubes cannot be tuned to frequencies higher than about 200 Mc., even when the inductance in the circuit is simply that of a straight wire between the tube elements.

Along with these capacity effects, the input loading (§ 7-6) increases rapidly at very-high frequencies, so that ordinary tuned circuits have very low effective Qs when connected to the grid circuit of a tube. The effect is still further aggravated by the fact that losses in the tuned circuit itself are higher, causing a still further reduction in Q. For these reasons, the frequency limit at which an r.f. amplifier will give any gain is in the vicinity of 60 Mc. with standard tubes. At higher frequencies there will be a loss, instead of amplification. This condition can be mitigated somewhat by taking steps to improve the effective Q of the circuit, either by tapping the grid down on the coil, as shown in Fig. 718-A, or by using a lower L/C ratio (§ 2-10). The Q of the tuned circuit alone can be greatly improved by using a linear circuit (§ 2-12), which when properly constructed will give Qs much higher than those attainable at lower frequencies with conventional coils and condensers. The concentric type of line, Fig. 718-B, is best both from the standpoint of Q and of adaptability to nonsymmetrical circuits such as are used in receivers. Since the capacity and resistance loading effects of the tube are still present, the Q of such a circuit will be destroyed if the grid-cathode circuit of the tube is connected directly across it. Hence, tapping down on the line, as shown, is necessary.

Very-high-frequency amplifiers employ tubes of the acorn or miniature type, which have the least loading effect as well as low interelectrode capacities. The smaller loading effect means higher input resistance, and, for a given loaded Q of the tuned circuit, a higher voltage is developed between the grid and cathode. Thus the amplification of the stage is higher and the noise level lower.

A concentric circuit may be tuned by varying the length of the inner conductor (usually by using close-fitting tubes, one sliding inside the other) or by connecting an ordinary tuning condenser across the line. Tapping the condenser down, as shown in Fig. 718-B, gives a bandspread effect, which is advantageous. It also helps to keep the Q of the circuit higher than it would be with the condenser connected directly across the open end of the line, since at very-high frequencies most condensers have losses which cannot be neglected.

Ordinary bakelite-based receiving-type tubes will function quite satisfactorily as oscillators and superregenerative detectors at frequencies where r.f. amplification is impossible with standard tubes (as in the 112-Mc. band), since tube losses are compensated for by energy taken from the power supply. Ordinary coil and condenser circuits are practicable with such tubes at 112 Mc. At higher frequencies, however, the special v.h.f. tubes are essential.

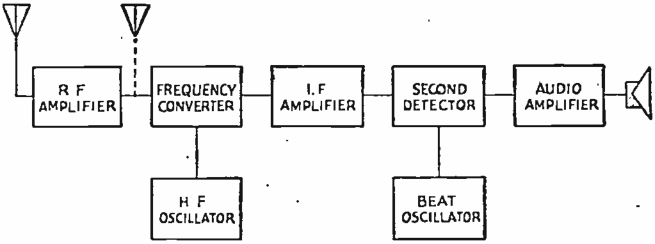


Fig. 719 — Block diagram of the basic elements of the superheterodyne

7-8 The Superheterodyne

Principles — In the *superheterodyne,* or *superhet,* receiver the frequency of the incoming signal is changed to a new radio frequency, the *intermediate frequency* (*i.f.*), then amplified, and finally detected. The frequency is changed by means of the heterodyne process (§ 7-1), the output of an adjustable *local oscillator* (the *h.f. oscillator*) being combined with the incoming signal in a *mixer* or *converter* stage (*first detector*) to produce a beat frequency equal to the intermediate frequency.

Fig. 719 gives the essentials of the superheterodyne in block form. C.w. signals are made audible by heterodyning the signal at the second detector by the *beat-frequency oscillator* (*b.f.o.*) or *beat oscillator*, set to differ from the i.f. by a suitable audio frequency.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen and that the incoming signal is on 7000 kc. Then the h.f. oscillator frequency may be set to 7455 kc., in order that the beat frequency (7455 minus 7000) will be 455 kc. The h.f. oscillator also could be set to 6545 kc., which will give the same frequency difference. To produce an audible c.w. signal of, say, 1000 cycles at the second detector, the beat oscillator would be set to either 454 kc. or 456 kc.

Characteristics — The frequency-conversion process permits r.f. amplification at a relatively low frequency. Thus high selectivity can be obtained, and this selectivity is constant regardless of the signal frequency. Higher gain also is possible at the lower frequency. The separate oscillators can be designed for

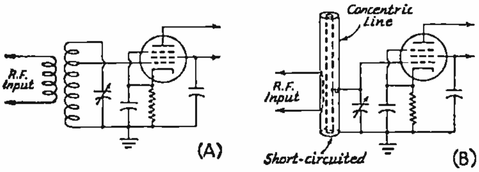


Fig. 718 — Circuits of improved *Q* for very-high frequencies. A, reducing tube loading by tapping down on the resonant circuit; B, use of a concentric-line circuit, with the tube similarly tapped down. The line should be a quarter-wave long, electrically; because of the additional shunt capacity represented by the tube, the physical length will be somewhat less than given by the formula (§ 10-5). In general, this reduction in length will be greater the higher the grid tap on the inner conductor. The coupling turn should be parallel to the axis of the line and must be insulated from the outer conductor.

stability, and, since the h.f. oscillator is working at a frequency considerably removed from the signal frequency, its stability is practically unaffected by the incoming signal.

Images — Each h.f. oscillator frequency will cause i.f. response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kc. to respond to a 7000-kc. signal, for example, it will respond also to a signal on 7910 kc., which likewise gives a 455-kc. beat. The undesired signal of the two is called the *image.*

The radio-frequency circuits of the receiver (those used before the frequency is converted to the i.f.) normally are tuned to the desired signal, so that the selectivity of the circuits reduces the response to the image signal. If the desired signal and image have equal strengths at the input terminals of the receiver, the ratio of the receiver voltage output from the desired signal to that from the image is called the *signal-to-image ratio,* or *image ratio.*

The image ratio depends upon the selectivity of the r.f. tuned circuits preceding the mixer tube. Also, the higher the intermediate frequency, the higher the image ratio, since raising the i.f. increases the frequency separation between the signal and the image and places the latter farther away from the peak of the resonance curve (§ 2-10) of the signal-frequency input circuits.

Other spurious responses — In addition to images, other signals to which the receiver is not ostensibly tuned may be heard. Harmonics of the high-frequency oscillator may beat with signals far removed from the desired frequency to produce output at the intermediate frequency; such spurious responses can be reduced by adequate selectivity before the mixer stage, and by using sufficient shielding to prevent signal pick-up by any means other than the antenna. When a strong signal is received, the harmonics (§ 2-7) generated by rectification in the second detector may, by stray coupling, be introduced into the r.f. or mixer circuit and converted to the intermediate frequency, to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal, and are principally bothersome when the frequency of the incoming signal is not greatly different from the intermediate frequency. The cure is proper circuit isolation and shielding.

Harmonics of the beat oscillator also may be converted in similar fashion and amplified through the receiver; these responses can be reduced by shielding the beat oscillator and operating it at low output level.

The double superheterodyne — At high and very-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kc. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kc.), and then — sometimes after further amplification — reconverted to a lower i.f. where higher adjacent-channel selectivity can be obtained. Such a receiver is called a *double superheterodyne.*

ℂ 7-9 Frequency Converters

Characteristics — The first detector or mixer resembles an ordinary detector. A circuit tuned to the intermediate frequency is placed in the plate circuit of the mixer, so that the highest possible i.f. voltage will be developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are bypassed to ground, since they are not wanted in the output. The i.f. tuned circuit should have low impedance for these frequencies, a condition easily met if they do not approach the intermediate frequency.

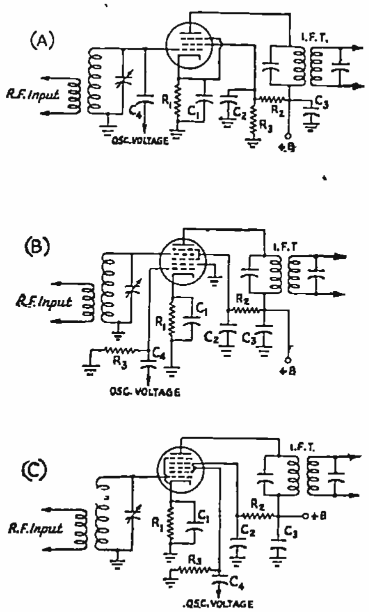


Fig. 720 — Mixer or converter circuits. A, grid injection with a pentode plate detector; B and C, separate injection circuits for converter tubes. Circuit values are:

Component	*Circuit A*	*Circuit B*	*Circuit C*
C_1, C_2, C_3 —	0.01–0.1 μfd.	0.01–0.1 μfd.	0.01–0.1 μfd.
C_4 —	Approx. 1 μμfd.	50–100 μμfd.	50–100 μμfd.
R_1 —	10,000 ohms.	300 ohms.	500 ohms.
R_2 —	0.1 megohm.	50,000 ohms.	15,000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.	50,000 ohms.

Plate voltage should be 250 in all circuits. If a 6AB7 or 6AC7 tube is used in Circuit A, R_1 should be 500 ohms.

The *conversion efficiency* of the mixer is the ratio of i.f. output voltage from the plate circuit to r.f. signal voltage applied to the grid. High conversion efficiency is desirable. The mixer tube noise also should be low if a good signal-to-noise ratio is wanted, particularly if the mixer is the first tube in the receiver.

The mixer should not require too much r.f. power from the h.f. oscillator, since it may be difficult to supply the power and yet maintain good oscillator stability (§ 3-7). Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain), since it is difficult to maintain constant output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called *pulling.* If the mixer and oscillator could be completely isolated, mixer tuning would have no effect on the oscillator frequency; but in practice this is a difficult condition to attain. Pulling should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. Pulling decreases with separation of the signal and h.f. oscillator frequencies, being less with high i.f.s.

Circuits — Typical frequency-conversion circuits are given in Fig. 720. The variations are chiefly in the way in which the oscillator voltage is introduced. In Fig. 720-A, the screen-grid pentode functions as a plate detector; the oscillator is capacity-coupled to the grid of the tube, in parallel with the tuned input circuit. Inductive coupling may be used instead. The conversion gain and input selectivity generally are good, so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator power required is negligible.

A pentagrid-converter tube is used in the circuit at B. Although intended for combination oscillator-mixer use, this type of tube usually will give more satisfactory performance when used in conjunction with a separate oscillator, the output of which is coupled in as shown. The circuit gives good conversion efficiency, and, because of the electron coupling, affords desirable isolation between the mixer and oscillator circuits. A small amount of power is required from the oscillator.

Circuit C is for the 6L7 mixer tube. The oscillator voltage can vary over a considerable range without affecting the conversion gain. There are no critical adjustments, and the oscillator-mixer isolation is good. The oscillator must supply somewhat more power than in B.

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. Practically the same number of circuit components is required whether or not a combination tube is used, so that there is little difference from the cost standpoint.

Tubes for frequency conversion — Any sharp cut-off pentode may be used in the circuit of Fig. 720-A. The 6AB7 and 6AC7 give high conversion gain and excellent signal-to-noise ratio — comparable, in fact, to the gain and signal-to-noise ratio obtainable with r.f. amplifiers — and in these respects are far superior to any other tubes used as mixers, particularly between 14 and 100 Mc. However, this type of tube loads the circuit more (§ 7-6) and thus decreases the selectivity.

The 6K8 is a good tube for the circuit at B; its oscillator plate connection may be ignored. The 6SA7 also is excellent in this circuit, although it has no anode grid (No. 2 grid, in the diagram). In addition to these two types, any pentagrid converter tube may be used.

V.H.F. and U.H.F. converters. — At frequencies above 50 Mc. the performance of the special mixer and converter tubes employed on the lower frequencies falls off because of greatly reduced input resistance and hence increased thermal noise (§ 3-5). Up to 100 Mc. the high-transconductance "television" pentodes (6AB7 and 6AC7) perform effectively in the circuit of Fig. 720-A. For the high-frequency oscillator the preferred tube in this range is the 6J5.

In the very-high-frequency region above 100 Mc. the greater input capacity and large plate-current requirements of the high-transconductance pentodes, which result in increased equivalent noise resistance, render them inferior to the special high-frequency pentodes such as the 9000 and acorn series. These tubes perform successfully up to 400 Mc. or so.

At still higher frequencies — or, for that matter, anywhere above 200 Mc. — other types of converters are preferred. At these frequencies triode mixers, when operated as plate-rectifier detectors in suitable circuits, give the least noise and maximum conversion transconductance.

Fig. 721-A shows the elementary circuit for a single triode with cathode oscillator-voltage injection. In such an arrangement the cathode connection usually terminates (with as short a lead as possible) in a small link near the oscillator tank, one end of which is grounded. Alternatively, direct capacity-coupled grid injection may be used in an arrangement similar to that of Fig. 720-A, C_4 being a very small coupling condenser of perhaps 1 or 2 $\mu\mu$fd. — often merely the free end of the coupling lead placed within the field of the oscillator coil.

The balanced triode circuit of Fig. 721-B affords the added advantages of symmetry to ground and complete cancellation of both the received-signal and oscillator voltages in the plate circuit. This serves further to improve the signal/noise ratio as well as to stabilize operation. For optimum performance the oscillator-voltage input should be carefully adjusted to give maximum converter gain. The balanced converter circuit, operated as a balanced square-law detector, is most frequently used with miniature dual triodes such as the 6J6, with which it performs effectively up to 600 Mc. or higher. The oscillator may be operated either on its fundamental or a harmonic. At frequencies above 200 Mc. coaxial or "trough"-line circuits are chiefly used.

At still higher frequencies converters employing conventional tubes are inferior to other, basically different types, including highly specialized versions of velocity-modulation tubes of various types. These techniques, however, are beyond the scope of the present treatment; information concerning practical tubes and circuits is largely classified.

For amateur work on these higher frequencies the use of special small u.h.f. diodes with extremely close element spacing as converters is a logical solution. Crystal detectors have also been used extensively because of their ready availability and independence of frequency limitations. Crystal detectors are not susceptible to the transit time limitations of electronic tubes. Silicon is the most popular material for such applications; the crystals are ground to minute dimensions and permanently mounted in fixed miniature holders with tungsten contacts. Fig. 722-A shows a typical crystal mixer circuit with inductive coupling to a triode oscillator (955 or 9002).

Because of the inherent insensitiveness of the crystal detector to changes in temperature and also because stability can be achieved only at the expense of sensitivity, diode detectors are preferred for other than purely experimental purposes. They have the further advantage that they will function as mixers by using a harmonic of the oscillator voltage, making possible the use of conventional triode oscillators for receivers operating up to the 2000-Mc.

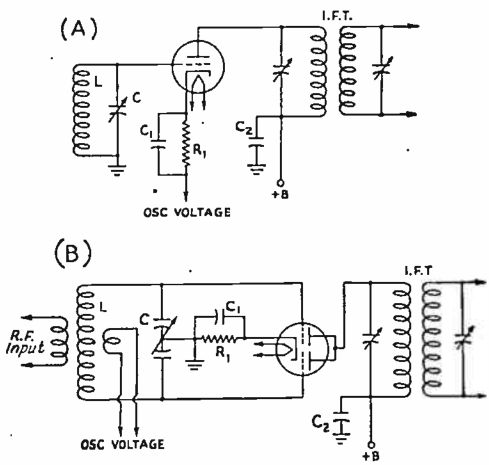


Fig. 721 — V.h.f. frequency converter circuits. A, triode mixer with separate oscillator tube; B, balanced square-law mixer using a dual triode tube with push-pull input circuit. *L* and *C* are tuned to the signal frequency.
C_1 — 100-$\mu\mu$fd. silvered mica.
C_2 — 0.005-$\mu\mu$fd.
R_1 — 10,000-50,000 ohms.

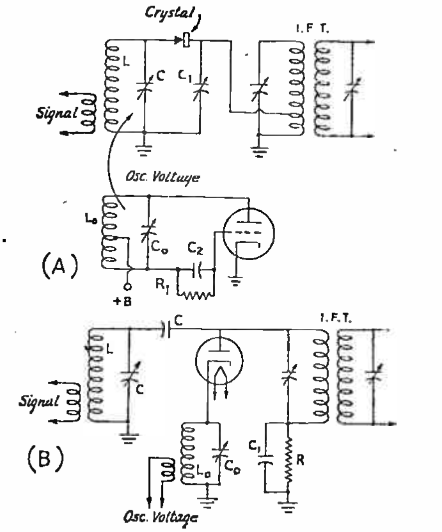


Fig. 722 — U.h.f. frequency converter circuits. A, crystal-detector mixer with an inductively coupled triode oscillator; B, diode mixer with cathode-link coupling to the oscillator circuit. L and C are tuned to the signal frequency; L_o and C_o to the oscillator frequency.

C_1 — 3–30-μμfd. mica trimmer.
C_2 — 25-μμfd. silvered mica.
C_3 — 10-μμfd. silvered mica.
C_4 — 0.005-μμfd.
R_1 — 50,000 ohms (metallized carbon).
R_2 — 5000–20,000 ohms.

region or higher. While operation of the oscillator on a fundamental is the more efficient method, the loss in conversion efficiency does not exceed 2 to 1 even with third harmonic operation provided the oscillator input is sufficient to establish a diode current of 0.2 to 0.5 ma. Diode mixers are considerably more tolerant as concerns oscillator voltage and other circuit conditions than the crystal type.

In the circuit of Fig. 722-B the cathode tuned circuit, L_oC_o, is tuned to the oscillator fundamental, C_o is being made large enough so that it is effectively a cathode by-pass condenser for the signal frequency.

ℭ 7-10 The High-Frequency Oscillator

Design considerations — Stability of the receiver (§ 7-2) is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to changes in voltage, loading, and mechanical shock. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. These ends can be attained by the use of good insulating materials and circuit components, suitable electrical design, and careful mechanical construction.

In addition, the oscillator must be capable of furnishing sufficient r.f. voltage and power for the particular mixer circuit chosen, at all frequencies within the range of the receiver, and its harmonic output should be as low as possible to reduce spurious response (§ 7-8).

It is desirable to make the L/C ratio in the oscillator tuned circuit low (high-C), since this results in increased stability (§ 3-7). Particular care should be taken to insure that no part of the oscillator circuit can vibrate mechanically. This calls for short leads and "solid" mechanical construction. The chassis and panel material should be heavy and rigid enough so that pressure on the tuning dial will not cause torsion and a shift in the frequency. Care in mechanical construction is well repaid by increased frequency stability.

Circuits — Several oscillator circuits are shown in Fig. 723. The point at which output voltage is taken for the mixer is indicated in each case by X or Y. Circuits A and B will give about the same results, and require only one coil. However, in these two circuits the cathode is above ground potential for r.f., which often is a cause of hum modulation of the oscillator output at 14 Mc. and higher frequencies when 6.3-volt heater tubes are used. Hum usually is not bothersome with 2.5-volt tubes, nor, of course, with tubes which are heated by direct current. The circuit of Fig. 723-C overcomes hum, since the cathode is

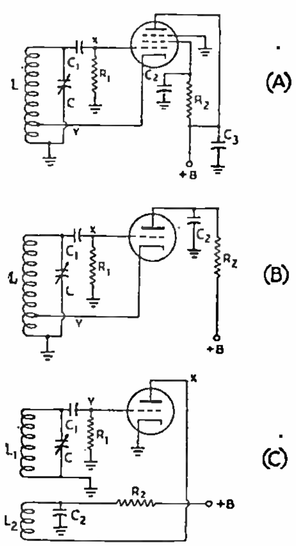


Fig. 723 — High-frequency oscillator circuits. A, screen-grid grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage than from X; this type is best adapted to mixer circuits with small oscillator-voltage requirements. Typical values for components are as follows:

	Circuit A	*Circuit B*	*Circuit C*
C_1 —	100 μμfd.	100 μμfd.	100 μμfd.
C_2 —	0.1 μfd.	0.1 μfd.	0.1 μfd.
C_3 —	0.1 μfd.		
R_1 —	50,000 ohms.	50,000 ohms.	50,000 ohms.
R_2 —	50,000 ohms.	10,000 to 25,000 ohms.	10,000 to 25,000 ohms.

The plate-supply voltage should be 250 volts. In circuits B and C, R_2 is used to drop the supply voltage to 100–150 volts; it may be omitted if voltage is obtained from a voltage divider in the power supply (§ 8-10).

grounded. The two-coil arrangement is advantageous in construction, since the feed-back adjustment (altering the number of turns on L_2 or the coupling between L_1 and L_2) is simple mechanically.

Besides the use of a fairly high C/L ratio in the tuned circuit, it is necessary to adjust the feed-back to obtain optimum results. Too much feed-back will cause the oscillator to "squeg," or operate at several frequencies simultaneously (§ 7-4); too little feed-back will cause the output to be low. In the tapped-coil circuits (A, B), the feed-back is increased by moving the tap toward the grid end of the coil; in C, by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will cause reduced tube heating and thereby reduce frequency drift. The oscillator and mixer circuits should be well isolated, preferably by shielding, since coupling other than by the means intended may result in pulling.

To avoid plate-voltage changes which may cause the oscillator frequency to change, it is good practice to use a voltage-regulated plate supply employing a gaseous VR tube (§ 8-8).

Tracking — For ganged tuning, there must be a *constant difference* in frequency between the oscillator and mixer circuits. This difference must be exactly equal to the intermediate frequency (§ 7-8).

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 724. The tracking capacity, C_5, commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. In practice, the trimmer, C_4, is first set for the high-frequency end of the tuning range, and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the tables under Fig. 724.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit which gives practically straight-line-frequency tuning (equal frequency change for each dial division), and then adjusting the oscillator and mixer tuned circuits so that both cover the same total number of kilocycles. For example, if the i.f. is 455 kc. and the mixer circuit tunes from 7000 to 7300 kc. between two given points on the dial, then the oscillator must tune from 7455 to 7755 kc. between the same two dial readings. With the bandspread arrangement of Fig. 715-C, the tuning will be practically straight-line-frequency if the capacity actually in use at C_2 is not too small; the same is true of 715-A if C_1 is small compared to C_2.

7-11 The Intermediate-Frequency Amplifier

Choice of frequency — The selection of an intermediate frequency is a compromise between various conflicting factors. The lower the i.f. the higher the selectivity and gain, but a low i.f. brings the image nearer the desired signal and hence decreases the image ratio (§ 7-8). A low i.f. also increases pulling of the oscillator frequency (§ 7-9). On the other hand, a high i.f. is beneficial to both image ratio and pulling, but the selectivity and gain are lowered. The difference in gain is least important.

An i.f. of the order of 455 kc. gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 Mc. The image ratio is poor at 14 Mc. when the mixer is connected to the antenna, but adequate when there is a

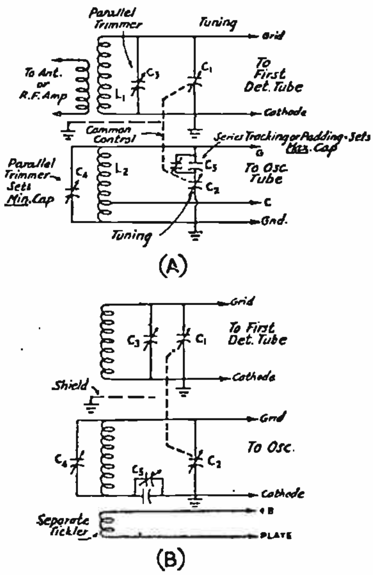


Fig. 724 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140 $\mu\mu$fd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 35 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
1.7–4 Mc.	50 μh.	40 μh.	0.0013 μfd.
3.7–7.5 Mc.	14 μh.	12.2 μh.	0.0022 μfd.
7–15 Mc.	3.5 μh.	3 μh.	0.0045 μfd.
14–30 Mc.	0.8 μh.	0.78 μh.	None used

Approximate values for 450- to 465-kc. i.f.s with a 2.5-to-1 tuning range, C_1 and C_2 being 350-$\mu\mu$fd. maximum, minimum including C_3 and C_4 being 40 to 50 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
0.5–1.5 Mc.	240 μh.	130 μh.	425 $\mu\mu$fd.
1.5–4 Mc.	32 μh.	25 μh.	0.00115 μfd.
4–10 Mc.	4.5 μh.	4 μh.	0.0028 μfd.
10–25 Mc.	0.8 μh.	0.75 μh.	None used

tuned r.f. amplifier between antenna and mixer. At 28 Mc. and on the very-high frequencies, the image ratio is very poor unless several r.f. stages are used. Above 14 Mc., pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 28 and 56 Mc., and pulling can be reduced to negligible proportions. However, the i.f. selectivity is considerably lower, so that more tuned circuits must be used to increase the selectivity. For very-high frequencies, including 28 Mc., the best solution is to use a double superheterodyne (§ 7-8), choosing one high i.f. for image reduction (5 and 10 Mc. are frequently used) and a lower one for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. The frequencies mentioned are fairly free of such interference.

Fidelity, sideband cutting — As described in § 5-2, modulation of a carrier causes the generation of sideband frequencies numerically equal to the carrier frequency plus and minus the highest modulation frequency present. If the receiver is to give a faithful reproduction of modulation which contains, for instance, audio frequencies up to 5000 cycles, it must be capable of amplifying equally all frequencies contained in a band extending from 5000 cycles above to 5000 cycles below the carrier frequency. In a superheterodyne, where all carrier frequencies are changed to the fixed intermediate frequency, this means that the i.f. amplifier should amplify equally well all frequencies within that band. In other words, the amplification must be uniform over a band 10 kc. wide, with the i.f. at its center. The signal-frequency circuits usually do not have enough over-all selectivity to affect materially the "adjacent channel" selectivity (§ 7-2), so that only the i.f. amplifier selectivity need be considered.

A 10-kc. band is considered sufficient for reasonably faithful reproduction of music, but much narrower band-widths can be used for communication work where intelligibility rather than fidelity is the primary objective.

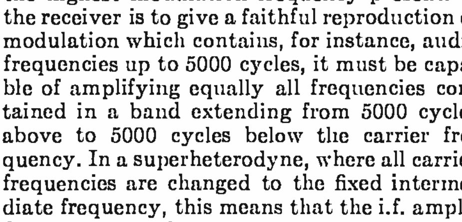


Fig. 725 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:
C_1 — 0.1 μfd. at 455 kc.; 0.01 μfd. at 1600 kc. and higher.
C_2 — 0.01 μfd.
C_3, C_4, C_5 — 0.1 μfd. at 455 kc.; 0.01 μfd. above 1600 kc.
R_1 — 300 ohms. R_3 — 2000 ohms.
R_2 — 0.1 megohm. R_4 — 0.25 megohm.

If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, the higher modulating frequencies are attenuated as compared to the lower frequencies; that is, the upper-frequency sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of greater selectivity.

The selectivity of an i.f. amplifier, and hence the tendency to cut sidebands, increases with the number of amplifier stages and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is not serious with two-stage amplifiers at frequencies as low as 455 kc.

Circuits — I.f. amplifiers usually consist of one or two stages. Two stages at 455 kc. give all the gain usable, in view of the minimum receiver noise level, and also give suitable selectivity for good-quality 'phone reception.

A typical circuit arrangement is shown in Fig. 725. A second stage would simply duplicate the circuit of the first. In principle, the i.f. amplifier is the same as the tuned r.f. amplifier (§ 7-6). However, since a fixed frequency is used, the primary as well as the secondary of the coupling transformer is tuned, giving higher selectivity than is obtainable with a closely coupled untuned primary. The cathode resistor, R_1, is connected to a gain control circuit of the type previously described (§ 7-6); usually both stages, if two are used, are controlled by a single variable resistor. The decoupling resistor, R_3 (§ 2-11), helps isolate the amplifier, and thus prevents stray feed-back. C_2 and R_4 are part of the automatic volume-control circuit (§ 7-13); if no a.v.c. is used, the lower end of the i.f. transformer secondary is simply connected to ground.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor (R_2) as shown, the screens being connected in parallel, or from a voltage divider (§ 8-10) across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

When two stages are used the high gain will tend to cause instability and oscillation, so that good shielding, by-passing, and careful circuit arrangement to prevent stray coupling, with exposed r.f. leads well separated, is necessary.

I.f. transformers — The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a metal-shield container in which the coils and tuning condensers are mounted. Both air-core and powdered-iron-core universal-wound coils are used, the latter having somewhat higher Qs and, hence, greater selectivity and gain per unit. In universal windings the coil is wound in layers but with each turn traversing the length of the coil, back and forth, rather than being wound in a plane perpendicular to the axis as it is in ordinary single-layer coils. With straight layer winding, the turns on adjacent layers at the edges of the coil have a rather large potential difference between them as compared to the difference between any two adjacent turns in the same layer; hence a fairly large capacity current can flow between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects (§ 2-8).

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning), in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f. transformer construction is shown in Fig. 726.

Besides the type of i.f. transformer shown in Fig. 726, special units to give desired selectivity characteristics are available. For higher than ordinary adjacent-channel selectivity (§ 7-2) *triple-tuned* transformers, with a third tuned circuit inserted between the input and output windings, are used. The energy is transferred from the input to the output windings via this *tertiary* winding, thus adding its selectivity to the over-all selectivity of the transformer. *Variable-selectivity* transformers also can be obtained. These usually are provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the Q and selectivity (§ 2-10) to broaden the selectivity curve. The variation in selectivity is brought about by switching the resistor in and out of the circuit. Another method is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve and undercoupling to sharpen it (§ 2-11).

Selectivity — The over-all selectivity of the i.f. amplifier will depend on the frequency and the number of stages. The following figures are indicative of the band-widths (§ 7-2) to be expected with good-quality transformers in amplifiers so constructed as to keep regeneration to a minimum:

	Band-width in kilocycles		
Intermediate frequency	*2 times down*	*10 times down*	*100 times down*
One stage, 455 kc. (air core) . . .	8.7	17.8	32.3
One stage, 455 kc. (iron core) . .	4.3	10.3	20.4
Two stages, 455 kc. (iron core) .	2.9	6.4	10.8
Two stages, 1600 kc.	11.0	16.6	27.4
Two stages, 5000 kc.	25.8	46.0	100.0

Tubes for i.f. amplifiers — Variable-μ pentodes (§ 3-5) are almost invariably used in i.f. amplifier stages, since grid-bias gain control (§ 7-6) is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i.f. tubes has practically no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier (if the latter is used).

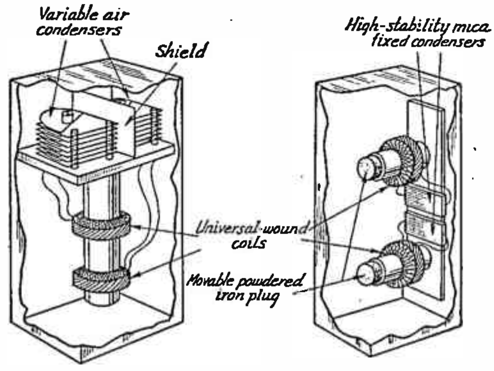


Fig. 726 — Representative i.f. transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores consist of finely divided iron particles supported in an insulating binder, formed into cylindrical "plugs." The tuning capacity is fixed, and the inductances of the coils are varied by moving the iron plugs in and out.

When single-ended tubes (§ 3-5) are used, care should be taken to keep the plate and grid leads well separated. With these tubes it is advisable to mount the screen by-pass condenser directly on the bottom of the socket, cross-wise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be connected to ground.

Single-signal effect — In heterodyne c.w. reception with a superheterodyne receiver, the beat oscillator is set to give a suitable audio-frequency beat note when the incoming signal is converted to the intermediate frequency. For example, the beat oscillator may be set to 456 kc. (the i.f. being 455 kc.) to give a 1000-cycle beat note. Now, if an interfering signal appears at 457 kc., it will also be heterodyned by the beat oscillator to produce a 1000-cycle beat. This *audio-frequency image* corresponds to the high-frequency images already discussed (§ 7-8). It can be reduced by providing enough i.f. selectivity, since the image signal is off the peak of the i.f. resonance curve.

When this is done, tuning through a given signal will show a strong response at the desired beat note on one side of zero beat only, instead of the two beat notes on either side of zero beat characteristic of less-selective reception; hence the name, "single-signal" reception.

The necessary selectivity is difficult to obtain with non-regenerative amplifiers using ordinary tuned circuits unless a very low intermediate frequency or a large number of circuits is used. In practice it is secured either by regenerative amplification or by a crystal filter.

Regeneration — Regeneration can be used to give a pronounced single-signal effect, particularly when the i.f. is 455 kc. or lower. The resonance curve of an i.f. stage at critical regeneration (just below the oscillating point) is extremely sharp, a band-width of 1 kc. at 10 times down and 5 kc. at 100 times down being obtainable in one stage. The audio-frequency image of a given signal thus can be reduced by a factor of nearly 100 for a 1000-cycle beat note (image 2000 cycles from resonance).

Regeneration is easily introduced into an i.f. amplifier by providing a small amount of capacity coupling between grid and plate. Bringing a short length of wire, connected to the grid, into the vicinity of the plate lead usually will suffice. The feed-back may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength This prevents overloading and increases selectivity.

The higher selectivity with regeneration reduces the over-all response to noise generated in the earlier stages of the receiver, just as does high selectivity produced by other means, and therefore improves the signal-to-noise ratio. The disadvantage is that the regenerative gain varies with signal strength, being less on strong signals, and the selectivity varies accordingly.

Crystal filters — The most satisfactory method of obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i.f. amplifier (§ 2-10). Compared to a good tuned circuit, the Q of such a crystal is extremely high. The dimensions of the crystal are made such that it is resonant at the desired intermediate frequency. It is then used as a selective coupler between i.f. stages.

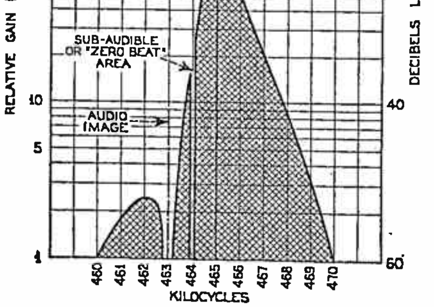


Fig. 727 — Graphical representation of single-signal selectivity. The shaded area indicates the overall band-width, or region in which response is obtainable.

Fig. 727 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides great discrimination against signals very close to the desired signal in frequency, and, by reducing the band-width, reduces the response of the receiver to noise both from sources external to the receiver and in the r.f. stages of the receiver itself.

Crystal filter circuits; phasing — Several crystal filter circuits are shown in Fig. 728. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit (§ 2-11), with the secondary side of T_1, the input transformer, balanced to ground either through a pair of condensers, C-C, (A) or by a center-tap on the secondary, L_2 (B). The bridge is completed by the crystal, X, and the *phasing condenser*, C_2, which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When C_2 is set to balance the crystal-holder capacity, the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and thus allows signals of the desired frequency to be fed through C_3 to L_3L_4, the output transformer. Without C_2, the holder capacity (with the crystal acting as a dielectric) would pass signals of undesired frequencies.

The phasing control has an additional function besides neutralization of the crystal-holder capacity. The holder capacity becomes a part of the crystal circuit and causes it to act as a parallel-tuned resonant circuit at a frequency slightly higher than its series-resonant frequency. Signals at the parallel-resonant frequency thus are prevented from reaching the output circuit. The phasing control, by varying the effect of the holder capacity, permits shifting the parallel-resonant frequency over a considerable range, providing adjustable rejection of interfering signals. The effect of rejection is illustrated in Fig. 727, where the audio image is reduced, by proper setting of the phasing control, far below the value that would be expected if the resonance curve were symmetrical.

Variable selectivity — In circuits such as A and B, Fig. 728, variable selectivity is obtained by adjustment of the variable input impedance, which is effectively in series with the crystal resonator. This is accomplished by varying C_1 (the *selectivity control*), which tunes the balanced secondary circuit of T_1. When the secondary is tuned to i.f. resonance the parallel impedance of the L_2C_1 combination is maximum and is purely resistive (§ 2-10). Since the secondary circuit is center-tapped, approximately one-fourth of this resistive impedance is in series with the crystal through C_3 and L_4. This lowers the Q of the crystal circuit and makes its selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

When the input circuit is detuned from the crystal resonant frequency the resistance component of the input impedance decreases, and so does the total parallel impedance. Accordingly, the selectivity of the crystal circuit becomes higher and the applied voltage falls off. At first the resistance decreases faster than the applied voltage, with the result that the c.w. output from the filter *increases* as the selectivity is increased. The output falls off gradually as the input circuit is detuned further from resonance, however, and the selectivity becomes still higher.

In the circuits of A and B in Fig. 728, the minimum selectivity is still much greater than that of a normal two-stage 455-kc. amplifier and it is desirable to provide a wider range of selectivity, particularly for 'phone reception. A circuit which does this is shown at Fig. 728-C. The principle of operation is similar, but a much higher value of resistance can be introduced in the crystal circuit to reduce the selectivity. The output tuned circuit, L_3C_3, must have high Q. A compensated condenser is used at C_2 (phasing) to maintain circuit balance, so that the phasing control does not affect the resonant frequency. The output circuit functions as a voltage divider in such a way that the amplitude of the carrier delivered to the next grid does not vary appreciably with the selectivity setting. The variable resistor, R, may consist of a series of separate fixed resistors selected by a tap switch.

7-12 The Second Detector and Beat Oscillator

Detector circuits — The second detector of a superheterodyne receiver performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great r.f. amplification. Therefore, the ability to handle large signals without distortion is preferable to high sensitivity. Plate detection is used to some extent, but the diode detector is most popular. It is especially adapted to furnishing automatic gain or volume control (§ 7-13). The basic circuits are as described in § 7-3, although in many cases the diode elements are incorporated in a multi-purpose tube which contains an amplifier section in addition to the diode unit.

The beat oscillator — Any standard oscillator circuit (§ 3-7) may be used for the beat oscillator. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with circuits such as those shown at Fig. 723-A and -B, with the output taken from Y. A variable condenser of about 25-$\mu\mu$fd. capacity may be connected between cathode and ground to provide fine adjustment. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed condenser of a few $\mu\mu$fd. capacity.

The beat oscillator should be well shielded, to prevent coupling to any part of the circuit except the second detector and to prevent its harmonics from getting into the front end of the receiver and being amplified like regular signals. To this end, the plate voltage should be as low as is consistent with sufficient audio-frequency output. If the beat oscillator output is too low, strong signals will not give a proportionately strong audio response.

An oscillating second detector may be used to give the audio beat note, but, since the detector must be detuned from the i.f., the selectivity and signal strength will be reduced, while blocking (§ 7-4) will be pronounced because of the high signal level at the second detector.

7-13 Automatic Volume Control

Principles — Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is a great advantage, especially in 'phone reception, since it tends to keep the output level of the receiver constant regardless of input signal strength. It is readily accomplished in superheterodyne receivers by using the average rectified d.c. voltage, developed by the received signal across a resistance in a detector circuit (§ 7-3), to vary the bias on the r.f. and i.f. amplifier tubes.

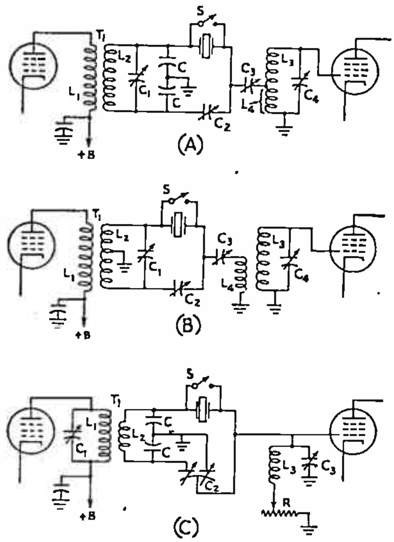


Fig. 728 — Crystal filter circuits of three types. All give variable band-width, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely coupled to tuned secondary, L_2; C_1, 50-$\mu\mu$fd. variable; C, each 100-$\mu\mu$fd. fixed (mica); C_2, 10- to 15-$\mu\mu$fd. (max.) variable; C_3, 50-$\mu\mu$fd. trimmer; L_3C_4, i.f. tuned circuit, with L_3 tapped to match crystal-circuit impedance. In circuit B, T_1 is the same as in circuit A except that the secondary is center-tapped; C_1 is 100-$\mu\mu$fd. variable; C_2, C_3 and C_4, same as for circuit A; L_3L_4 is a transformer with primary, L_4, corresponding to tap on L_3 in A. In circuit C, T_1 is a special i.f. input transformer with tuned primary and low-impedance secondary; C, each 100-$\mu\mu$fd. fixed (mica); C_2, opposed stator phasing condenser, approximately 8 $\mu\mu$fd. maximum capacity each side; L_3C_3, high-Q i.f. tuned circuit; R, 0 to 3000 ohms (selectivity control).

Since this voltage is proportional to the average amplitude of the signal, the gain is reduced as the signal strength becomes greater. The control will be more complete as the number of stages to which the a.v.c. bias is applied is increased. Control of at least two stages is advisable.

Circuits — A typical circuit using a diode-triode type tube as a combined a.v.c. rectifier, detector and first audio amplifier is shown in Fig. 729. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate is fed from the detector diode through the small coupling condenser, C_3. A negative bias voltage resulting from the flow of rectified carrier current is developed across R_4, the diode load resistor. This negative bias is applied to the grids of the controlled stages through the filtering resistors (§ 2-11), R_5, R_6, R_7 and R_8. When S_1 is closed the a.v.c. line is grounded, thereby removing the a.v.c. bias from the amplifier without disturbing the detector circuit.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector and a.v.c. rectifier. This could be done in Fig. 729. The a.v.c. filter and line would connect to the junction of R_2 and C_2, while C_3 and R_4 would be omitted from the circuit.

Delayed a.v.c. — In Fig. 729 the audio diode return is made directly to the cathode and the a.v.c. diode return to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay the a.v.c. would start working even with a very small signal. This is undesirable, because the full amplification of the receiver then could not be realized on weak signals. In the audio diode circuit this fixed bias would cause distortion, and must be avoided; hence, the return is made directly to the cathode.

Time constant — The time constant (§ 2-6) of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal, and in practice would cause frequency distortion. On the other hand, the time constant must not be too great or the a.v.c. would be unable to follow rapid fading. The capacity and resistance values indicated in Fig. 729 will give a time constant which is satisfactory for high-frequency reception.

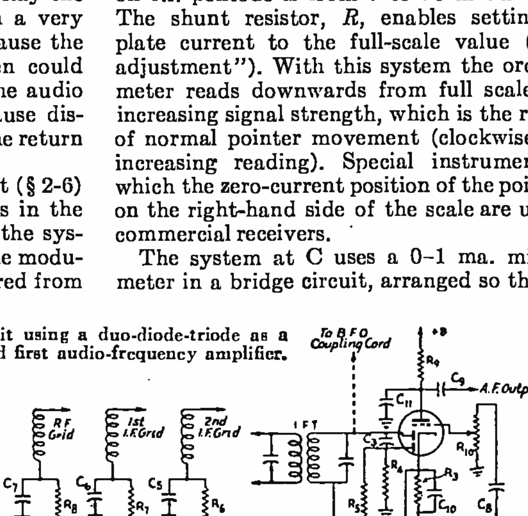
Signal-strength and tuning indicators — A useful accessory to the receiver is an indicator which will show relative signal strength. Not only is it an aid in giving reports to transmitting stations, but it is helpful also in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

Three types of indicators are shown in Fig. 730. That at A uses an electron-ray tube (§ 3-5), several types of which are available. The grid of the triode section usually is connected to the a.v.c. line. The particular type of tube used depends upon the voltage available for its grid; where the a.v.c. voltage is large, a remote cut-off type (6G5 or 6N5) should be used in preference to the more sensitive sharp cut-off type (6E5).

In B, a milliammeter is connected in series with the d.c. plate lead to one or more r.f. and i.f. tubes, the grids of which are controlled by a.v.c. voltage. Since the plate current of such tubes varies with the strength of the incoming signal, the meter will indicate relative signal intensity and may be calibrated in "S" points. The scale range of the meter should be chosen to fit the number of tubes in use; the maximum plate current of the average remote cut-off r.f. pentode is from 7 to 10 milliamperes. The shunt resistor, R, enables setting the plate current to the full-scale value ("zero adjustment"). With this system the ordinary meter reads downwards from full scale with increasing signal strength, which is the reverse of normal pointer movement (clockwise with increasing reading). Special instruments in which the zero-current position of the pointer is on the right-hand side of the scale are used in commercial receivers.

The system at C uses a 0–1 ma. milliammeter in a bridge circuit, arranged so that the

Fig. 729 — Automatic volume control circuit using a duo-diode-triode as a combined a.v.c. rectifier, second detector and first audio-frequency amplifier.



R_1 — 0.25 megohm.
R_2 — 50,000 to 250,000 ohms.
R_3 — 2000 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.5 to 1 megohm.
R_6, R_7, R_8, R_9 — 0.25 megohm.
R_{10} — 0.5-megohm variable.
C_1, C_2, C_3 — 100 μμfd.
C_4 — 0.1 μfd.
C_5, C_6, C_7 — 0.01 μfd.
C_8, C_9 — 0.01 to 0.1 μfd.
C_{10} — 5- to 10-μfd. electrolytic. C_{11} — 250 μμfd.

meter reading and the signal strength increase together. The current through the branch containing R_1 should be approximately equal to the current through that containing R_2. In some manufactured receivers this is brought about by draining the screen voltage-divider current and the current to the screens of three r.f. pentodes (r.f. and i.f. stages) through R_2, the sum of these currents being about equal to the maximum plate current of one a.v.c.-controlled tube. Typical values for this type of circuit are given. The sensitivity can be increased by increasing the resistance of R_1, R_2 and R_3. The initial setting is made with the manual gain control set near maximum, when R_3 should be adjusted to make the meter read zero with no signal.

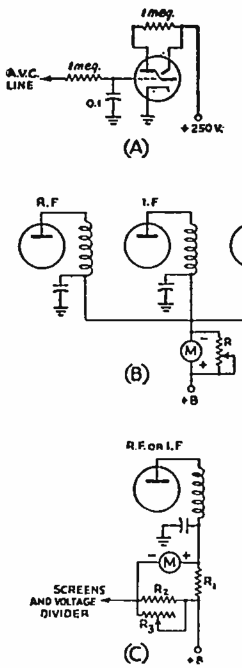


Fig. 730 — Tuning indicator or "S"-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor R should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: R_1, 250 ohms; R_2, 350 ohms; R_3, 1000-ohm variable.

ℂ 7-14 Preselection

Purpose — Preselection is added signal-frequency selectivity incorporated before the mixer stage is reached. An r.f. amplifier preceding the mixer generally is called a *preselector*, its purpose, in part at least, being to discriminate in favor of the signal against the image. The preselector may consist of one or more r.f. amplifier stages. When its tuning control is ganged with those of the mixer and oscillator, its circuits must track with the mixer circuit.

The circuit is the same as discussed earlier (§ 7-6). An external preselector stage may be used with receivers having inadequate image ratios. In this case it is built as a separate unit, often with a tuned output circuit which gives a further improvement in selectivity. The output circuit usually is link-coupled (§ 2-11) to the receiver.

Signal/noise ratio — An r.f. amplifier will have a better signal-to-noise ratio (§ 7-2) than a mixer because the gain is higher and because the mixer-tube electrode arrangement results in higher internal tube noise than does the ordinary pentode structure. Hence, a preselector is advantageous in increasing the signal-to-noise ratio over that obtainable when the mixer is fed directly from the antenna.

Image suppression — The image ratios (§ 7-8) obtainable at frequencies up to and including 7 Mc. with a single preselector stage are high enough, when the intermediate frequency is 455 kc., so that for all practical purposes there is no appreciable image response. Average image ratios on 14-Mc. and 28 Mc. are 50–75 and 10–15, respectively. This is the overall selectivity of the r.f. and mixer tuned circuits. A second preselector stage, adding another tuned circuit, will increase the ratios to several hundred at 14 Mc. and to 30–40 at 28 Mc.

On very-high frequencies, it is impracticable to attempt to secure a good image ratio with a 455-kc. i.f. Good performance can be secured only by using a high i.f. or a double superheterodyne (§ 7-8) with a high-frequency first i.f.

Regeneration — Regeneration may be used in a preselector stage to increase both gain and selectivity. Since its use makes tuning more critical and increases ganging problems, regeneration is seldom employed except at 14 Mc. and above, where adequate image suppression is difficult to obtain with non-regenerative circuits. The same disadvantages exist as in the case of a regenerative i.f. amplifier (§ 7-11). The effect of regeneration is roughly equivalent to adding another non-regenerative preselector stage.

Regeneration may be introduced by the same method as used in regenerative i.f. amplifiers (§ 7-11). The manual gain control of the stage will serve as a volume control.

Regeneration in a preselector does not improve the signal-to-noise ratio, since the tube noise is fed back to the grid circuit along with the signal to add to the thermal-agitation noise originally present. This noise also is amplified.

ℂ 7-15 Noise Reduction

Types of noise — In addition to tube and circuit noise (§ 7-6), much of the noise interference experienced in reception of high-frequency signals is caused by domestic electrical equipment and by automobile ignition systems. The interference is of two types in its effects. The first is the "hiss" type, consisting of overlapping pulses similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol-shot" or "machine-gun" type, consisting of separated impulses of high amplitude. The "hiss"

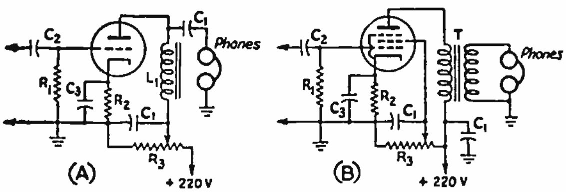


Fig. 731 — Audio output-circuit amplitude-limiting noise-reducing circuits for c.w. reception.
C_1 — 0.25 μfd.
C_2 — 0.01 μfd.
C_3 — 5 μfd.
R_1 — 0.5 megohm.
R_2 — 2000 ohms.
R_3 — 50,000-ohm potentiometer.
T — Output transformer.
L_1 — 15-henry choke.

type of interference usually is caused by commutator sparking in d.c. and series-wound a.c. motors, while the "shot" type results from separated spark discharges (a.c. power leaks, switch and key clicks, ignition sparks, and the like).

Impulse noise — Impulse noise, because of the extremely short duration of the pulses as compared to the time between them, must have high pulse amplitude to contain much average energy. Hence, noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general principle of devices intended to reduce such noise is that of allowing the signal amplitude to pass through the receiver unaffected, but making the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared to its time of duration the more successful the noise reduction, since more of the constituent energy can be suppressed.

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the Q or flywheel effect (§ 2-10) of the circuits. Hence, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good noise suppression.

Audio limiting — A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio output circuit of a receiver. Such limiters also maintain the signal output nearly constant with fading. Diagrams of typical output-limiter circuits are shown in Fig. 731. Circuit A employs a triode tube operated at reduced plate voltage (approximately 10 volts), so that it saturates at a low signal level. The arrangement of B has better limiting characteristics. A pentode audio tube is operated at reduced screen voltage (35 volts or so), so that the output power remains practically constant over a grid excitation-voltage range of more than 100 to 1. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous circuits.

Second-detector circuits — The circuit of Fig. 732 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes non-conducting above a predetermined signal level. The audio output of the detector must pass through the diode to the grid of the amplifier tube. The diode normally would be non-conducting with the connections shown were it not for the fact that it is given positive bias from a 30-volt source through the adjustable potentiometer, R_3. Resistors R_1 and R_2 must be fairly large in value to prevent loss of audio signal.

The audio signal from the detector can be considered to modulate (§ 5-1) the steady diode current, and conduction will take place so long as the diode plate is positive with respect to the cathode. When the signal is sufficiently large to swing the cathode positive with respect to the plate, however, conduction ceases, and that portion of the signal is cut off from the audio amplifier. The point at which cut-off occurs can be selected by adjustment of R_3. By setting R_3 so that the signal just passes through the "valve," noise pulses higher in amplitude than the signal will be cut off. The circuit of Fig 732-A, using an infinite-impedance detector (§ 7-3), gives a positive voltage on rectification. When the rectified voltage is negative, as it is from the usual diode detector (§ 7-3), the circuit arrangement shown in Fig. 732-B must be used.

An audio signal of about ten volts is required for good limiting action. When a beat oscillator is used for c.w. reception the b.f.o. voltage should be small, so that incoming noise will not have a strong carrier to beat against and so produce large audio output.

A second-detector noise-limiting circuit which automatically adjusts itself to the received carrier level is shown in Fig. 733. The diode load circuit (§ 7-3) consists of R_6, R_7, R_8 (shunted by the high-resistance audio volume control, R_4) and R_5 in series. The cathode of the 6N7 noise limiter is tapped on the load resistor at a point such that the average rectified carrier voltage (negative) at its grid is approximately twice the negative voltage at the cathode, both measured with reference to ground. A filter network, R_1C_1, is inserted in the grid circuit, so that the audio modulation on the carrier does not reach the grid; hence, the grid potential is maintained at substantially the rectified carrier voltage alone. The cathode, however, is free to follow the modulation, and when the modulation is 100 per cent the peak cathode voltage will just equal the steady grid voltage.

At all modulation percentages below 100 per cent the grid is negative with respect to cathode, and current cannot flow in the 6N7 plate-cathode circuit. A noise pulse exceeding the peak voltage which represents 100 per cent modulation will, however, make the grid positive with respect to cathode. The relatively low plate-cathode resistance of the 6N7 then shunts the high-resistance audio output circuit, effectively short-circuiting it, so that there is practically no response for the duration of the peak over the 100 per cent modulation limit.

R_5 is used to make the noise-limiting tube more sensitive by applying to the plate an audio voltage out of phase with the cathode voltage, so that, at the instant the grid goes positive with respect to cathode, the highest positive potential also is applied to the plate, thus further lowering the effective plate-cathode resistance.

I.f. noise silencer — In the circuit shown in Fig. 734, noise pulses are made to decrease the gain of an i.f. stage momentarily and thus silence the receiver for the duration of the pulse. Any noise voltage in excess of the desired signal's maximum i.f. voltage is taken off at the grid of the i.f. amplifier, amplified by the noise amplifier stage, and rectified by the full-wave diode noise rectifier. The noise circuits are tuned to the i.f. The rectified noise voltage is applied as a pulse of negative bias to the No. 3 grid of the 6L7 i.f. amplifier, wholly or partially disabling this stage for the duration of the individual noise pulse, depending on the amplitude of the noise voltage. The noise amplifier-rectifier circuit is biased by means of the "threshold control," R_2, so that rectification will not start until the noise voltage exceeds the desired-signal amplitude. With automatic volume control the a.v.c. voltage can be applied to the grid of the noise amplifier, to augment this threshold bias. This system improved the signal-to-noise ratio some 30 db. (power ratio of 1000) with heavy ignition interference, raising the signal-to-noise ratio from − 10 db. without the silencer to + 20 db. with the silencer in a typical instance.

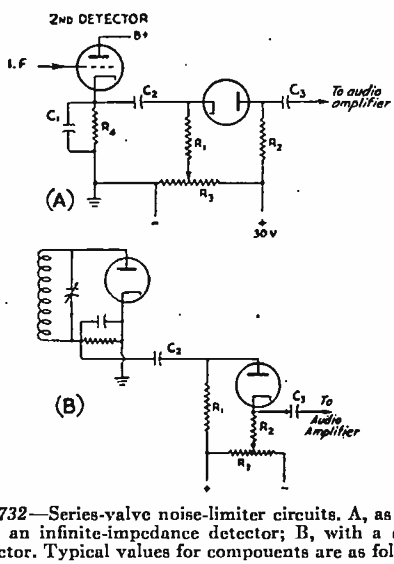


Fig. 732—Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for compouents are as follows:

R_1 — 0.25 megohm.	R_4 — 20,000 to 50,000 ohms.
R_2 — 50,000 ohms.	C_1 — 250 μμfd.
R_3 — 10,000 ohms.	C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

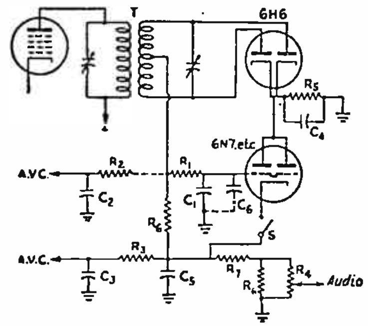


Fig. 733 — Automatic noise-limiter for superheterodynes.

T — I.f. transformer with a balanced secondary for working into a diode rectifier.

R_1, R_2, R_3 — 1 megohm.	C_1 — 0.1-μfd. paper.
R_4 — 1-megohm variable.	C_2, C_3 — 0.05-μfd. paper.
R_5 — 250,000 ohms.	C_4, C_5 — 50-μμfd. mica.
R_6, R_8 — 100,000 ohms.	C_6 — 0.001-μfd. mica (for r.f. filtering, if needed).
R_7 — 25,000 ohms.	
Sw — S.p.s.t. toggle (on-off switch).	

The switch should be mounted close to the circuit elements and controlled by an extension shaft if necessary.

Circuit values are normal for i.f. amplifiers (§ 7-11), except as indicated. The noise-rectifier transformer, T_1, has an untuned secondary closely coupled to the primary and center-tapped for full-wave rectification. The center-tap rectifier (§ 8-3) is used to reduce the possibility of r.f. feed-back into the i.f. amplifier (noise-silencer) stage. The time constant (§ 2-6) of the noise-rectifier load circuit, $R_1C_1C_2$, must be small, to prevent disabling the noise-silencer stage for a longer period than the duration of the noise pulse. The r.f. choke, *RFC*, must be effective at the intermediate frequency.

Adequate shielding and isolation of the noise-amplifier and rectifier circuits from the noise-silencer stage must be provided to prevent possible self-oscillation and instability. This circuit should be applied to the first i.f. stage of the receiver, before the high-selectivity circuits are reached. On the other hand, it is most effective when the signal and noise levels are fairly high (meaning one or two r.f. stages before the mixer) since several volts must be obtained from the noise rectifier for good silencing.

¶ 7-16 Operating Superheterodyne Receivers

C.w. reception — For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency (§ 7-8). To adjust the beat-oscillator frequency, first tune in a moderately weak but steady carrier with the beat oscillator turned off. Adjust the receiver tuning for maximum signal strength, as indicated by maximum hiss. Then turn on the beat oscillator and adjust its frequency (leaving the receiver tuning unchanged) to give a suitable beat note. The beat oscillator need not subsequently be touched, except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The use of a.v.c. (§ 7-13) is not generally satisfactory in c.w. reception because the receiver gain rises in the spaces between the dots and dashes, giving an increase in noise in the same intervals, and because the rectified beat-oscillator voltage in the second detector circuit also operates the a.v.c. circuit. This gives a constant reduction in gain and prevents utilization of the full sensitivity of the receiver. Hence, the gain preferably should be manually adjusted to give suitable audio-frequency output.

To avoid overloading in the i.f. circuits, it is usually better to control the i.f. and r.f. gain and keep the audio gain at a fixed value than to use the a.f. gain control as a volume control and leave the r.f. gain fixed at its highest level.

Tuning with the crystal filter — If the receiver is equipped with a crystal filter the tuning instructions in the preceding paragraph still apply, but more care must be used both in the initial adjustment of the beat oscillator and in tuning. The beat oscillator is set as described above, but with the crystal filter in operation and adjusted to its sharpest position, if variable selectivity is available. The initial adjustment should be made with the phasing control (§ 7-11) in the intermediate position. After it is completed, the beat oscillator should be left set and the receiver tuned to the other side of zero beat (audio-frequency image) on the same carrier to give a beat note of the same tone. This beat will be considerably weaker than the first, and may be "phased out" almost completely by careful adjustment of the phasing control. This is the adjustment for normal operation; it will be found that one side of zero beat has practically disappeared, leaving maximum response on the desired side.

An interfering signal having a beat note differing from that of the a.f. image can be

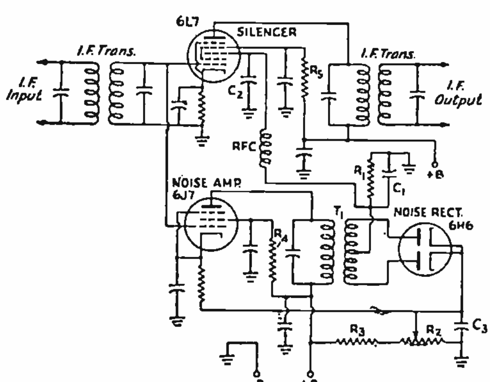


Fig. 734 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:
C_1 — 50–250 μμfd. (use smallest value possible without r.f. feedback).
C_2 — 50 μμfd. R_2 — 5000-ohm variable.
C_3 — 0.1 μfd. R_3 — 20,000 ohms.
R_1 — 0.1 megohm. R_4, R_5 — 0.1 megohm.
T_1 — Special i.f. transformer for noise rectifier.

similarly phased out, provided its carrier frequency is not too near the desired carrier.

Depending upon the filter design, maximum selectivity may cause the dots and dashes to lengthen out so that they seem to "run together." This, plus the fact that tuning is quite critical with extremely high selectivity, may make it desirable to use somewhat less selectivity in ordinary operation. However, it must be emphasized that, to realize the benefits of the crystal filter in reducing interference, it is necessary to do *all* tuning with it in the circuit. Its selectivity is so high that it is almost impossible to find the desired station quickly, should the filter be switched in only when interference is present.

'Phone reception — In reception of 'phone signals, the normal procedure is to set the r.f. and i.f. gain at maximum, switch on the a.v.c., and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.v.c. system in compensating for fading and maintaining constant audio output on either strong or weak signals. On occasion a strong signal close to the frequency of a weaker desired station may take control of the a.v.c., in which case the weaker station will practically disappear because of the reduced gain. In this case better reception may result if the a.v.c. is switched off, using the manual r.f. gain control to set the gain at a point which prevents "blocking" by the stronger signal.

A crystal filter will do much toward reducing interference in 'phone reception. Although the high selectivity cuts sidebands (§ 7-11) and thereby reduces the audio output, especially at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility even though the "quality" of the transmission may suffer. As in the case of c.w. reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control in the crystal filter. It cannot be prevented in a "straight" superheterodyne having no crystal filter.

A tone control often will be of help in reducing the effects of high-pitched heterodynes, sideband splatter (§ 5-2) and noise, by cutting off the higher audio frequencies. This, like sideband cutting with high selectivity, causes some reduction in naturalness.

Spurious responses — Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of zero beat opposite that on which the desired signal peaks.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat note. Signals getting into the i.f. via high-frequency oscillator harmonics tune more rapidly (less dial movement) through a given change in beat note than do signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat-oscillator pitch control. A smaller movement of the control will suffice for a given change in beat note than is necessary with legitimate signals.

7-17 Servicing Superheterodyne Receivers

Troubleshooting — Two basic methods are employed. One is the "point-by-point" system of static analysis, requiring chiefly a multirange volt-ohm-milliammeter. Beginning at the power transformer, the operating voltages at each point in the circuit are measured. Abnormally low or high voltages, or the absence of indication at a given point in the circuit, presumably indicate a defective component at that point. The analysis may then be completed with the aid of the ohmmeter and a little deduction, ending with repair or replacement of unserviceable components.

An alternative method, commonly employed by professional radio servicemen, is that of "dynamic" or "channel" analysis. The principle is that of applying a test signal to the r.f. input and tracing it stage-by-stage through the receiver. The r.f. and i.f. stages are checked by tuned amplifiers feeding a linear detector which operates an indicator such as vacuum-tube voltmeter, electron-ray voltmeter, or cathode-ray tube. A probe on the end of a shielded lead with a very small condenser (1–2 $\mu\mu$fd.) in series is used to pick up the signal in the output of any stage, and the tuned amplifiers are adjusted to the frequency of the stage. Thus the presence or absence of the signal at any point in the receiver may be determined, as well as the relative level.

I.f. alignment — A calibrated signal generator or test oscillator is a practical necessity for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver also is needed. If the receiver has a tuning meter, its indications will serve for this purpose. Alternatively, if the signal generator is of the modulated type, an a.c. output meter (high-resistance voltmeter with copper-oxide rectifier) can be connected across the primary of the output transformer, or from the plate of the last audio amplifier through a 0.1-μfd. blocking condenser (§ 2-13) to the receiver chassis. The intensity of sound from the loudspeaker can be judged by ear, if no output meter is available, but this method is not as accurate as those using instruments.

The procedure is as follows: The test oscillator is adjusted to the desired intermediate frequency, and the "hot" or ungrounded output lead is clipped on the grid terminal of the last i.f. amplifier tube. The grounded lead is connected to the receiver chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum signal output. The hot lead from the generator is next clipped on the grid of the next-to-last i.f. tube, and the second from last i.f. transformer is brought into alignment by adjusting its trimmers for maximum output. This process is continued, working back from the second detector, until all of the i.f. transformers have been aligned. It will be necessary to reduce the output of the signal generator as more of the i.f. amplifier is brought into use, because the increased gain otherwise may cause overloading and consequent inaccurate results. It is desirable always to use the minimum signal strength which gives useful output readings.

The i.f. transformer in the plate circuit of the mixer is aligned with the signal-generator output lead connected to the mixer grid. Since the tuned circuit feeding the mixer grid is tuned to a considerably higher frequency, it can effectively short-circuit the signal-generator output, and therefore it may be necessary to disconnect this circuit. With tubes having a top grid-cap connection, this can be done by simply removing the grid clip from the tube cap.

If the tuning indicator is used as an output meter the a.v.c. should be on; if the audio-output method is used, the a.v.c. should be off. The beat oscillator should be off in either case.

If the i.f. amplifier has a crystal filter, the filter should be switched out. Alignment is then carried out as described above, setting the signal generator as closely as possible to the frequency of the crystal. After alignment, the crystal should be switched in and the oscillator frequency varied back and forth over a small range either side of the crystal frequency to find its exact frequency, which will be indicated by a sharp rise in output. Leaving the signal generator set on the crystal peak, the i.f. trimmers may be realigned for maximum output. The necessary readjustment should be small. The signal generator frequency should be checked frequently, to make sure it has not drifted from the crystal peak.

A modulated signal is not of much value for aligning a crystal-filter i.f. amplifier, since the high selectivity cuts sidebands and the results may be inaccurate if the audio output of the receiver is used as a criterion of alignment. Lacking an a.v.c. tuning meter the transformers may be aligned by ear, using a weak unmodulated signal adjusted to the crystal peak. Switch on the beat oscillator, adjust to a suitable tone, and align the transformers for maximum audio output.

An amplifier which is only slightly out of alignment, as a result of normal drift from temperature, humidity or aging effects, can be realigned by using any steady signal, such as a local broadcasting station, in lieu of a test oscillator. Allow the receiver to warm up thoroughly (an hour or so), tune in the signal as usual, and "touch up" the i.f. trimmers.

R.f. alignment — The objective in aligning the r.f. circuits in a gang-tuned receiver is to secure adequate tracking over each tuning range. The adjustment may be carried out with a test oscillator of suitable frequency range, or even on noise or such signals as may be heard. First set the tuning dial at the high-frequency end of the range in use. Then set the test oscillator to the frequency indicated by the receiver dial. The test-oscillator output may be connected to the antenna terminals of the receiver for this test. Adjust the oscillator trimmer condenser in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and carefully tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal as indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacity in the tracking condenser) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscillator. Most commercial receivers provide some means for varying the inductances of the coils or the capacity of the tracking condenser, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacity or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacity is secured. In many cases, better over-all tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. Adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacity in any circuit, more inductance is needed; if less capacity resonates the circuit, less inductance is required.

Tracking seldom is perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

Visual alignment — More accurate and efficient alignment of receiver circuits may be performed with the aid of a visual curve-tracer or "wobbulator" which traces out the response curve visually on a cathode-ray oscilloscope. This is accomplished by using a special signal generator in which the oscillator frequency is varied over a suitable range at a low audio rate. The horizontal sweep of the oscilloscope is synchronized with the rate of variation of the test frequency, so that the horizontal deflection is a function of frequency. The rectified output of the second detector is connected to the vertical deflection plates of the oscilloscope. The spot on the screen therefore traces a curve proportional to the receiver response in terms of the instantaneous value of the oscillator frequency. This visual response curve, which may be that of the entire receiver or of any stage, is continually visible as a whole. Thus the effect of any adjustment of the circuits may be made much more rapidly and with greater accuracy than is possible with an ordinary signal generator and output meter, particularly in the case of wide-band i.f. circuits.

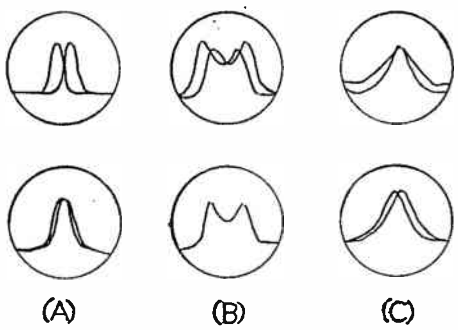


Fig. 735 — Oscilloscope patterns of response characteristics on a visual curve tracer. The upper row illustrates various kinds of misalignment; the lower row shows the same stages properly aligned. A, i.f. curve of a selective communications-type receiver, with all transformers mistuned on one side of resonance (top). Below, the peaks coincide when properly aligned, even though skirts do not precisely match. B, at the top, a broad-band f.m. receiver curve taken after alignment by the fixed-frequency and output-meter method; the lower curve shows the improvement after careful visual alignment. C, the pattern of a medium-selectivity receiver with transformers misaligned symmetrically on either side of resonance (top); below, the same i.f. correctly aligned but with the test oscillator tuned slightly off frequency to displace the return trace for better examination.

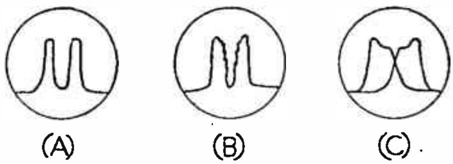


Fig. 736 — A, a typical single-trace response curve of a selective high-fidelity i.f. system. B, pattern of the amplifier in A made highly regenerative, illustrating instability. C, double trace of a single overcoupled i.f. stage with the return trace displaced. A similar "knee" located lower on the skirts would indicate regeneration.

Apparatus and methods for obtaining visual curve traces are described in Chapter Twenty. The simplest arrangement is that which employs a reactance-tubemodulated oscillator on 1000 kc., the output of which is combined with that of an unmodulated, variable tuned oscillator in a mixer tube to provide a heterodyned signal at the desired center frequency.

Either "double trace" and "single trace" patterns may be used. The double trace pattern is obtained by applying a triangular sweep to the f.m. oscillator at a frequency half that of the sawtooth sweep on the horizontal plates of the cathode-ray tube. The return trace produces two superimposed patterns, useful for checking symmetry and frequency calibration. The single-trace pattern shows the same two opposite-sequence resonance curves, but with the return trace displaced by the r.f. sweep frequency. It is useful in displaying irregularities in the pattern which might be obscured by superposition of the traces.

The alignment procedure follows that described for the oscillator–output-meter method. Assuming a diode second detector, run a shielded lead to the vertical input terminals of the oscilloscope from the "high" side of the diode load resistor — usually the audio volume control. With a triode biased detector, the bias resistor and by-pass condenser circuit should be opened and the vertical terminal connected to the cathode of the detector tube across a 0.5-megohm leak to ground, by-passed with a 250-$\mu\mu$fd. condenser. The plate load should be shorted out. This will make the resonance patterns appear upside down, but does not change their interpretation.

The r.f. output from the mixer should connect directly to the grid of the first i.f. tube. Add the i.f. frequency to 1000 kc. and set the unmodulated signal generator to this frequency. For example, if the i.f. is 465 kc., set the a.m. signal generator to 1465 kc. At the usual bandwidth of 30 kc., the signal at the grid of the last i.f. stage will swing from 450 kc. to 480 kc. and back. If the signal generator is set to the exact i.f., a double-trace pattern should appear on the screen. Center this pattern with the oscilloscope sweep vernier. Adjust the i.f. trimmers until these peaks coincide. For single-trace analysis, the sweep frequency should be reduced one half.

To align the next i.f. stage, move the r.f. output lead to the grid of the tube and adjust the next i.f. transformer. It may be necessary to readjust the output transformer after this operation. When aligning triple-tuned or high-fidelity i.f. circuits, it is most important that the peaks in the double pattern coincide and have nearly equal amplitude.

To align the r.f. and mixer input circuits, the variable-frequency signal generator should be set to a frequency which, by addition to 1000 kc., produces the desired r.f. signal frequency. As each stage is added, the output level must be reduced to keep the pattern on the screen. Only enough signal should be used to overcome local interference to avoid overloading any stage. Adjust the r.f. trimmers for maximum vertical amplitude of the pattern, as with an output meter. Dial calibration can be checked by setting the calibrated oscillator on frequency and adjusting the h.f. oscillator trimmer in the receiver to center the pattern on the screen.

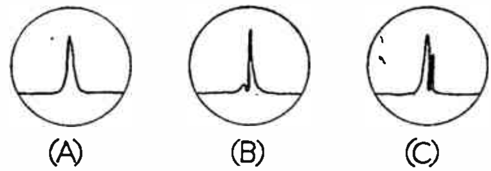


Fig. 737 — Response curves of a superheterodyne with crystal filter (made at a very low repetition rate). A, crystal in "broad" position, phasing control at center. B, phasing control set to place the rejection slot on low-frequency side. C, with slot on high-frequency side.

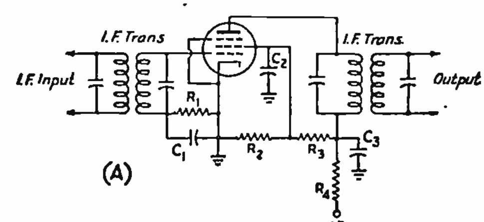
Oscillation in r.f. or i.f. amplifiers — Oscillation in high-frequency amplifier and mixer circuits may be evidenced by squeals or "birdies" as the tuning is varied, or by complete lack of audible output if the oscillation is strong enough to cause the a.v.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits, especially to the tuning-condenser rotors. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. In some cases it may be advisable to provide a shield between the stators of pre-r.f. amplifier and first-detector ganged tuning condensers, in addition to the usual tube and interstage shielding. A metal tube with an ungrounded shell will cause trouble. Improper screen-grid voltage, resulting from a shorted or too-low screen-grid series resistor, also may be responsible for such instability.

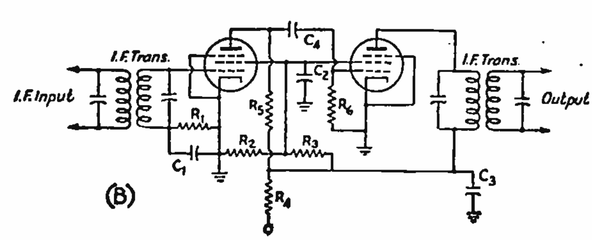
Oscillation in the i.f. circuits is independent of high-frequency tuning, and is indicated by a continuous squeal which appears when the gain is advanced with the c.w. beat oscillator on. It can result from similar defects in i.f. amplifier circuits. Inadequate cathode by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1 to 0.25 μfd. usually will remedy the trouble. Similar treatment can be applied to the screen-grid and plate by-pass filters of i.f. stages.

Instability — "Birdies" or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is "squegging" or oscillating simultaneously at high and low frequencies (§ 7-4). This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feedback, or too-high grid-leak resistance.

A varying beat note in c.w. reception indicates instability in either the h.f. oscillator or beat oscillator, usually the former. The stability of the beat oscillator can be checked by introducing a signal of intermediate frequency (from a test oscillator) into the i.f. amplifier; if the beat note is unstable, the trouble is in the beat oscillator. Poor connections or defective parts are the likely cause. Instability in the high-frequency oscillator may be the result of poor circuit design (§ 7-10), loose connections, defective tubes or circuit components, or poor voltage regulation in the oscillator plate and/or screen supply circuits. Mixer pulling of the oscillator circuit (§ 7-9) also will cause the beat-note to "chirp" on strong c.w. signals because the oscillator load changes slightly.

In 'phone reception with a.v.c., a peculiar type of instability ("motorboating") may appear if the h.f. oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the electrode currents of the controlled tubes decrease, decreasing the load on the power supply and causing its output voltage to rise. Since this increases the voltage applied to the oscillator, its frequency changes correspondingly, throwing the signal off the peak of the i.f. resonance curve and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself, at a rate determined by the signal strength and the time constant of the power-supply circuits. This effect is most pronounced with high i.f. selectivity, as when a crystal filter is used, and can be cured by making the oscillator relatively insensitive to voltage changes and by regulating the plate voltage supply (§7-10).





ℂ 7-18 Reception of Frequency-Modulated Signals

F.m. receivers — A frequency-modulation receiver differs in circuit design from one designed for amplitude modulation chiefly in the arrangement used for detecting the signal. Detectors for amplitude-modulated signals do not respond to frequency modulation. It is also necessary, for full realization of the noise-reducing benefits of the f.m. system, that the signal applied to the detector be completely free from amplitude modulation. In practice, this is attained by preventing the signal from rising above a given amplitude by means of a limiter (§ 3-10, 7-15). Since the weakest signal must be amplitude-limited, high gain must be provided ahead of the limiter; the superheterodyne type of circuit almost invariably is used to provide the necessary gain.

The r.f. and i.f. stages in a superheterodyne for f.m. reception are practically identical in circuit arrangement with those in an a.m. receiver. Since the use of f.m. is confined to the very-high frequencies (above 28 Mc.) a high intermediate frequency is employed, usually between 4 and 5 Mc. This not only reduces image response but also provides the greater band-width necessary to accommodate wide-band frequency-modulated signals.

Receiver requirements — The primary requirements are sufficient r.f. and i.f. gain to "saturate" the limiter even with a weak signal, sufficient band-width (§ 7-2) to accommodate the full frequency deviation either side of the carrier frequency without undue attenuation at the edges of the band, a limiter circuit which functions properly on both rapid and slow variations in amplitude, and a detector which gives a linear relationship between frequency deviation and *amplitude* output. The audio circuits are the same as in other receivers (§ 7-5), except that in communications-type receivers it is desirable to cut off the upper audio range by a low-pass filter (§ 2-11) because higher-frequency noise components have the greatest amplitude in an f.m. receiver.

The limiter — Limiter circuits generally are of the plate-saturation type (§ 7-15), where low plate and screen voltage are used to limit the plate-current flow at high signal amplitudes. Fig. 738-A is a typical circuit. The tube is self-biased (§ 3-6) by a grid leak, R_1, and condenser, C_1. R_2, R_3 and R_4 form a voltage divider

Fig. 738 — **F.m. limiter circuits. A, single-tube plate-saturation limiter; B, cascade limiter. Typical values are:**

	Circuit A	*Circuit B*
C_1 —	100 μμfd.	100 μμfd.
C_2, C_3 —	0.1 μfd.	0.1 μfd.
C_4 —		250 μμfd.
R_1 —	0.1 megohm.	50,000 ohms.
R_2 —	2000 ohms.	2000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.
R_4 —	0–50,000 ohms.	0–50,000 ohms.
R_5 —		4000 ohms.
R_6 —		0.2 megohm.

Plate-supply voltage is 250 in both circuits.

(§ 8-10) which puts the desired voltages on the screen and plate. The lower the voltages the lower the signal level at which limiting occurs, but the r.f. output voltage of the limiter also is lower. C_2 and C_3 are the plate and screen by-pass condensers, of conventional value for the intermediate frequency used. The time constant (§ 2-6) of R_1C_1 determines the behavior of the limiter with respect to rapid and slow amplitude variations. For best operation on impulse noise (§ 7-15) the time constant should be small, but a too-small time constant limits the range of signal strengths the limiter can handle without departing from the constant-output condition. A larger time constant is better in this respect but is not so effective for rapid variations, compromise constants are shown in Fig. 738.

The cascade limiter, Fig. 738-B, overcomes this by making the time constant in the first grid circuit suitable for effective operation on impulse noise, and that in the second grid (C_4R_6) optimum for a wide range of input signal strengths. This results, in addition, in more constant output over a very wide range of input signal amplitudes because the voltage at the grid of the second stage already is partially amplitude-limited. Resistance coupling ($R_5C_4R_6$) is used for simplicity and to prevent unwanted regeneration, additional gain at this point being unnecessary.

The rectified voltage developed across R_1 in either circuit may be applied to the i.f. amplifier for a.v.c. (§ 7-13).

Discriminating circuits and operation — The f.m. detector commonly is called a *discriminator*, because of its ability to discriminate between frequency deviations above and those below the carrier frequency.

The fundamental requirement of an f.m. detector is a non-linear response to the modulation variations of the applied signal. An ordinary tuned circuit, adjusted so that the signal frequency falls on one side of its response curve, constitutes an elementary discriminator. If two such circuits are used, one tuned above and the other below the signal frequency, amplitude variations are balanced out and the combined rectified current is directly proportional to the frequency deviation.

The circuit most widely used is the "series" or center-tuned discriminator shown in Fig. 739-A. A special i.f. coupling transformer is used between the limiter and detector. Its secondary, L_1, is center-tapped and is connected back to the plate side of the primary circuit, which otherwise is conventional. C_4 is the tuning condenser. The load circuits of the two diode rectifiers ($R_1C_1R_2C_2$) are connected in series; constants are the same as in ordinary diode detector circuits (§ 7-3). Audio output is taken from across the two load resistances.

The primary and secondary circuits are both adjusted to resonance in the center of the i.f. pass-band. The voltage applied to the rectifiers consists of two components, that induced in the secondary by the inductive coupling and that fed to the center of the secondary through C_2. The phase relations between the two are such that at resonance the rectified load currents are equal in amplitude but flow in opposite directions through R_1 and R_2, hence the net voltage across the terminals marked "audio output" is zero. When the carrier deviates from resonance the induced secondary current either lags or leads, depending upon whether the deviation is to the high- or low-frequency side, and this phase shift causes the induced current to combine with that fed through C_2 in such a way that one diode gets more voltage than the other when the frequency is below resonance, while the second diode gets the larger voltage when the frequency is higher than resonance. The voltage appearing across the output terminals is the difference between the two diode voltages. Thus a characteristic like that of Fig. 740 results, where the net rectified output voltage has opposite polarity for frequencies on either side of resonance, and up to a certain point becomes greater in amplitude as the frequency deviation is greater. The straight-line portion of the curve is the useful detector characteristic. The separation between the peaks which mark the ends of the linear portion of the curve depends upon the Qs of the primary and secondary circuits and the degree of coupling. The separation becomes greater with low Qs and close coupling. The circuit ordinarily is designed so that the peaks fall just outside the limits of the pass-band, thus utilizing most of the straight portion of the curve. Since the audio output is proportional to the change in d.c. voltage with deviation, it is advantageous for maximum output to keep the separation between peaks the minimum necessary for a linear characteristic.

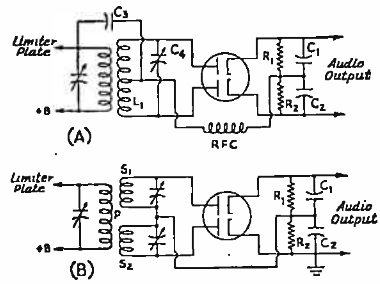


Fig. 739 — F.m. discriminator circuits. In both circuits typical values for C_1 and C_2 are 100 μμfd. each; R_1 and R_2, 0.1 megohm each. C_3 in A is approximately 50 μμfd., depending upon the intermediate frequency; *RFC* should be of a type designed for the i.f. in use (2.5 mh. is satisfactory for i.f.s of 4 to 5 Mc.). In either circuit the ground may be moved from the lower end of C_2 to the junction of C_1 and C_2, for push-pull audio output.

A second type of discriminator, known as the center-tuned circuit, is shown in Fig. 739-B. Two secondary circuits, S_1 and S_2, are used, one tuned above the center frequency of the i.f. pass-band and the other below. They are coupled equally to the primary, which is tuned to the center frequency. As the carrier fre-

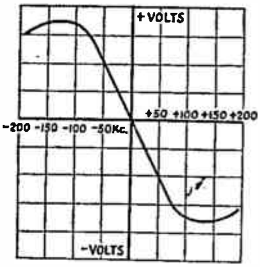


Fig. 740 — Characteristic of a typical f.m. detector. The vertical axis represents the voltage developed across the load resistor as the frequency varies from the exact resonance frequency. This detector would handle f.m. signals up to a band-width of 150 kc. over the linear portion of the curve.

quency deviates the voltages induced in the secondaries will change in amplitude, the larger voltage appearing across the secondary being nearer resonance with the instantaneous frequency. The detection characteristic is similar to that of the center-tuned discriminator. The peak separation is determined by the *Qs* of the circuits, the coefficient of coupling, and the tuning of the secondaries. High *Qs* and loose coupling are required for close peak separation.

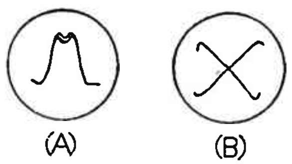
A simple self-quenched superregenerative receiver may be used as a frequency detector if it is tuned so that the carrier frequency falls along the slope of the resonance curve. Two such detectors, off-tuned on either side of the carrier, may be used in push-pull. An alternative arrangement employing a superregenerative stage as a first i.f. amplifier at 75 Mc., following a converter unit, provides high gain and linear response with relatively few stages.

F.m. receiver alignment — Alignment of f.m. receivers up to the limiter is carried out as described in § 7-17. For output measurement, a 0–1 milliammeter or 0–500 microammeter should be connected in series with the limiter grid resistor (R_1 in Fig. 738) at the grounded end; or, if the voltage drop across R_1 is used for a.v.c. and the receiver is provided with a tuning meter (§ 7-13), the tuning meter may be used as an output meter. An accurately calibrated signal generator or test oscillator is desirable, since the i.f. should be aligned to be as symmetrical as possible; that is, the output reading should be the same for any two test oscillator settings the same number of kilocycles above or below resonance. It is not necessary to have uniform response over the whole band to be received, although the output at the edges of the band (limit of deviation (§ 5-11) of the transmitted signals) should not be less than 25 per cent of the voltage at resonance. In communications work, a band-width of 30 kc. or less (15 kc. or less deviation) is commonly used. Output readings should be taken with the oscillator set at intervals of a few kilocycles either side of resonance up to the band limits.

After the i.f. (and front-end) alignment, the limiter operation should be checked. This can be done by temporarily disconnecting C_3, if the discriminator circuit of Fig. 739-A is used, disconnecting R_1 and C_1 on the cathode side, and inserting the milliammeter or microammeter in series with R_2 at the grounded end. This converts the discriminator to an ordinary diode rectifier. Varying the signal-generator frequency over the channel, with the discriminator transformer adjusted to resonance, should show no change in output (at the bandwidths used for communications purposes) as indicated by the rectified current read by the meter. At this point various plate and screen voltages can be tried on the limiter tube or tubes, to determine the set of conditions which gives maximum output with adequate limiting (no change in rectified current).

When the limiter has been checked the discriminator connections can be restored, leaving the meter connected in series with R_1. Provision should be made for reversing the connections to the meter terminals, to take care of the reversal in polarity of the net rectified current. Set the signal generator to the center frequency of the band and adjust the discriminator transformer trimmer condensers to resonance, which will be indicated by zero rectified current. Then set the test oscillator at the deviation limit (§ 5-11) on one side of the center frequency, and note the meter reading. Reverse the meter terminals and set the test oscillator at the deviation limit on the other side. The two readings should be the same. If they are not, they can be made so by a slight adjustment of the primary trimmer. This will necessitate rechecking the response at resonance to make sure it is still zero. Generally, the secondary trimmer will chiefly affect the zero-response frequency, while the primary trimmer will have most effect on the symmetry of the discriminator peaks. A detector curve having satisfactory linearity can be obtained by cut-and-try adjustment of both trimmers.

Fig. 741 — Oscilloscope patterns in f.m. i.f. alignment. A — I.f. response. B — Over-all characteristic through the f.m. detector.



A visual curve tracer is particularly advantageous in aligning the wide-band i.f. amplifiers of f.m. receivers. The i.f. is first aligned with the discriminator circuit converted into an a.m. diode detector, as described above, the pattern appearing as in Fig. 742-A. The over-all characteristic, including the f.m. detector, is shown in Fig. 741-B.

Tuning and operation — An f.m. receiver gives greatest noise reduction when the carrier is tuned exactly to the center of the receiver pass-band and to the point of zero response in the discriminator. Because of the decrease in noise, this point is readily recognized.

When an amplitude-modulated signal is tuned in its modulation practically disappears at exact resonance, only those nonsymmetrical modulation components which may be present being detected. If the signal is to one side or the other of resonance, however, it is capable of causing interference to an f.m. signal.

Chapter Eight

Power Supply

8-1 Power-Supply Requirements

Filament supply — Except for tubes designed for battery operation, the filaments or heaters of vacuum tubes used in both transmitters and receivers are universally operated on alternating current obtained from the power line through a step-down transformer (§ 2-9) delivering a secondary voltage equal to the rated voltage of the tubes used. The transformer should be designed to carry the current taken by the number of tubes which may be connected in parallel (§ 2-6) across it. The filament or heater transformer generally is center-tapped, to provide a balanced circuit for eliminating hum (§ 3-6).

For medium- and high-power r.f. stages of transmitters, and for high-power audio stages, it is desirable to use a separate filament transformer for each section of the transmitter, installed near the tube sockets. This avoids the necessity for abnormally large wires to carry the total filament current for all stages without appreciable voltage drop. Maintenance of rated filament voltage is highly important, especially with thoriated-filament tubes, since under- or over-voltage may reduce filament life.

Plate supply — Direct current must be used for the plates of tubes, since any variation in plate current arising from power-supply causes will be superimposed on the signal being received or transmitted, giving an undesirable type of modulation (§ 5-1) if the variations occur at an audio-frequency (§ 2-7) rate. Unvarying direct current is called *pure d.c.*, to distinguish it from current which may be unidirectional but of pulsating character. The use of pure d.c. on the plates of transmitting tubes is required by FCC regulations on all frequencies below 60 Mc.

Sources of plate power — D.c. plate power is usually obtained from rectified and filtered alternating current, but in low-power and portable installations may be secured from batteries. Dry batteries may be used for very low-power portable equipment, but in many cases a storage battery is used as the primary power source, in conjunction with an interrupter giving pulsating d.c. which is applied to the primary of a step-up transformer (§8-10).

Rectified-a.c. supplies — Since the power-line voltage ordinarily is 115 or 230 volts, a step-up transformer (§ 2-9) is used to obtain the desired voltage for the plates of the tubes in the equipment. The alternating secondary current is changed to unidirectional current by means of diode rectifier tubes (§ 3-1), and then passed through an inductance-capacity filter (§ 2-11) to the load circuit. The *load resistance* in ohms is equal to the d.c. output voltage of the power supply divided by the current in amperes (Ohm's Law, § 2-6).

Voltage regulation — Since there is always some resistance in power-supply circuits, and since the filter normally depends to a considerable extent upon the energy storage of inductance and capacity (§ 2-3, 2-5), the output voltage will depend upon the current drain on the supply. The change in output voltage with change in load current is called the *voltage regulation.* It is expressed as a percentage:

$$\% \text{ Regulation} = \frac{100\,(E_1 - E_2)}{E_2}$$

where E_1 is the no-load voltage (no current in the load circuit) and E_2 the full-load voltage (rated current in load circuit).

8-2 Rectifiers

Purpose and ratings — A rectifier is a device which will conduct current only in one direction. The diode tube (§ 3-1) is used almost exclusively for rectification in d.c. power supplies used with radio equipment. The important characteristics of tubes used as power-supply rectifiers are the voltage drop between plate and cathode at rated current, the maximum permissible inverse peak voltage, and the permissible peak plate current.

Voltage drop — Tube voltage drop depends upon the type of tube. In vacuum-type rectifiers it increases with the current flowing because of space-charge effect (§ 3-1), but can be minimized by using very small spacing between plate and cathode as is done in some rectifiers for receiver power supplies. Mercury-vapor rectifiers (§ 3-5) have a constant drop of about 15 volts, regardless of current. This is much smaller than the voltage drops encountered in vacuum-type rectifiers.

Inverse peak voltage — This is the maximum voltage developed between the plate and cathode of the rectifier when the tube is not conducting; i.e., when the plate is negative with respect to the cathode.

Peak plate current — This is the maximum *instantaneous* current through the rectifier. It can never be smaller than the load current in ordinary circuits, and may be several times higher.

Operation of mercury-vapor rectifiers — Because of its constant voltage drop, the mercury-vapor rectifier is more susceptible to damage than the vacuum type. With the latter, the increase in voltage drop tends to limit current flow on heavy overloads, but the mercury-vapor rectifier does not have this limiting action and the cathode may be damaged under similar conditions.

In mercury-vapor rectifiers a phenomenon known as "arc-back," or breakdown of the mercury vapor and conduction in the opposite direction to normal, occurs at high inverse peak voltages, hence such tubes always should be operated within their inverse-peak voltage ratings. Arc-back also may occur if the cathode temperature is below normal; therefore the heater or filament voltage should be checked to make sure that the rated voltage is applied. This check should be made at the tube socket, to avoid errors caused by voltage drop in the leads. For the same reason, the cathode should be allowed to come up to its final temperature before plate voltage is applied; the time required for this is of the order of 15 to 30 seconds. When a tube is first installed, or is put into service after a long period of idleness, the cathode should be heated for a period of 10 minutes or so before application of plate voltage.

ℂ 8-3 Rectifier Circuits

Half-wave rectifiers — The simple diode rectifier (§ 3-1) is called a *half-wave rectifier*, because it can pass only half of each cycle of alternating current. Its circuit is shown in Fig. 801-A. At the top of the figure is a representation of the applied a.c. voltage, with positive and negative alternations (§ 2-7) marked.

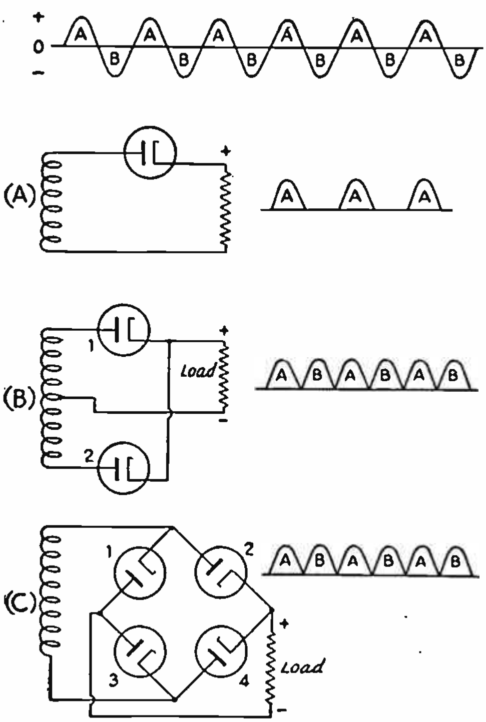


Fig. 801 — Fundamental vacuum-tube rectifier circuits.

When the plate is positive with respect to cathode, plate current flows through the load as indicated in the drawing at the right, but when the plate is negative with respect to cathode no current flows. This is indicated by the gaps in the output drawing. The output current is unidirectional but pulsating.

In this circuit the inverse peak voltage is equal to the maximum transformer voltage, which in the case of a sine wave is 1.41 times the r.m.s. voltage (§ 2-7).

Full-wave center-tap rectifier — Fig. 801-B shows the "full-wave center-tap" rectifier circuit, so called because both halves of the a.c. cycle are rectified and because the transformer secondary winding must consist of two equal parts with a connection brought out from the center. When the upper end of the winding is positive, current can flow through rectifier No. 1 to the load; this current cannot pass through rectifier No. 2 because its cathode is positive with respect to its plate. The circuit is completed through the transformer center-tap. When the polarity reverses the upper end of the winding is negative and no current can flow through No. 1, but the lower end is positive and therefore No. 2 passes current to the load, the return connection again being the center-tap. The resulting waveshape is shown at the right.

Since the two rectifiers are working alternately in this circuit, each half of the transformer secondary must be wound to deliver the full-load voltage; hence the total voltage across the transformer terminals is twice that required with the half-wave rectifier. Assuming negligible voltage drop in the particular rectifier which may be conducting at any instant, the inverse peak voltage on the other rectifier is equal to the maximum voltage between the outside terminals of the transformer. In the case of a sine wave, this is 1.41 times the total secondary r.m.s. voltage (§ 2-7).

Because energy is delivered to the load at twice the average rate as in the case of a half-wave rectifier, each tube carries only half the load current.

The bridge rectifier — The "bridge" type of full-wave rectifier is shown in Fig. 801-C. Its operation is as follows: When the upper end of the winding is positive, current can flow through No. 2 to the load but not through No. 1. On the return circuit, current flows through No. 3 by way of the lower end of the transformer winding. When the polarity reverses and the lower end of the winding becomes positive, current flows through No. 4 and the load and through No. 1 by way of the upper side of the transformer. The output waveshape is shown at the right.

The inverse peak voltage is equal to the maximum transformer voltage, or 1.41 times the r.m.s. secondary voltage in the case of a sine wave (§ 2-7). Energy is delivered to the load at the same average rate as in the case of the full-wave center-tap rectifier, each *pair* of tubes in series carrying half the load current.

8-4 Filters

Purpose of filter — As shown in Fig. 801, the output of a rectifier is pulsating d.c., which would be unsuitable for most vacuum-tube applications (§ 8-1). A *filter* is used to smooth out the pulsations so that practically unvarying direct current flows through the load circuit. The filter utilizes the energy-storage properties of inductance and capacity (§ 2-3, 2-5), by virtue of which energy stored in electromagnetic and electrostatic fields when the voltage and current are rising is restored to the circuit when the voltage and current fall, thus filling in the "gaps" or "valleys" in the rectified output.

Ripple voltage and frequency — The pulsations in the output of the rectifier can be considered to be caused by an alternating current superimposed on a steady direct current (§ 2-13). Viewed from this standpoint, the filter may be considered to consist of by-pass condensers which short-circuit the a.c. while not interfering with the flow of d.c., and *chokes* or inductances which permit d.c. to flow through them but which have high reactance for the a.c. (§ 2-13). The alternating component is called the *ripple*. The effectiveness of the filter may be measured by the *per cent ripple*, which is the r.m.s. value of the a.c. ripple voltage expressed as a percentage of the d.c. output voltage. With an effective filter, the ripple percentage will be low. Five per cent ripple is considered satisfactory for c.w. transmitters, but lower values (of the order of 0.25 per cent) are necessary for hum-free speech transmission and receiver plate supplies.

The ripple frequency depends upon the line frequency and the type of rectifier. In general, it consists of a fundamental plus a series of harmonics (§ 2-7), the latter being relatively unimportant since the fundamental is hardest to smooth out. With a half-wave rectifier, the fundamental is equal to the line frequency; with a full-wave rectifier, the fundamental is equal to twice the line frequency, or 120 cycles in the case of a 60-cycle supply.

Types of filters — Inductance-capacity filters are of the low-pass type (§ 2-11), using series inductances and shunt capacitances. Practical filters are identified as *condenser-input* and *choke-input*, depending upon whether a capacity or inductance is used as the first element in the filter. Resistance-capacity filters (§ 2–11) are used in applications where the current is very low and the voltage drop in the resistor can be tolerated.

Bleeder resistance — Since the condensers in a filter will retain their charge for a considerable time after power is removed (provided the load circuit is open at the time), it is good practice to connect a resistor across the output of the filter to discharge the condensers when the power supply is not in use. The resistance usually is high enough so that only a relatively small percentage of the total output current is consumed in it during normal operation.

Components — Filter condensers are made in several different types. Electrolytic condensers, which are available for voltages up to about 800, combine high capacity with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Condensers for higher voltages usually are made with a dielectric of thin paper impregnated with oil. The *working voltage* of a condenser is the voltage which it will withstand continuously.

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability (§ 2-5) decreases, consequently the inductance also decreases. Despite the air-gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding; hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding may be considerably higher than the load value.

8-5 Condenser-Input Filters

Ripple voltage — The conventional condenser-input filter is shown in Fig. 802-A. No simple formulas are available for computing

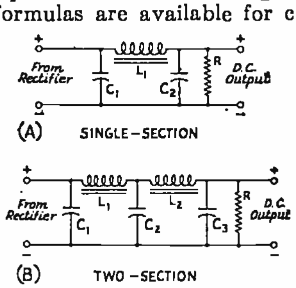


Fig. 802 — Condenser-input filter circuits.

the ripple voltage, but it will be smaller as both capacity and inductance are made larger. Adequate smoothing for transmitting purposes can be secured by using 4 to 8 μfd. at C_1 and C_2 and 20 to 30 henrys at L_1, for full-wave rectifiers with 120-cycle ripple (§ 8-4). A higher ratio of inductance to capacity may be used at higher load resistances (§ 8-1).

For receivers, as shown in Fig. 802-B, an additional choke, L_2, and condenser, C_3, of the same approximate values, are used to give additional smoothing. In such supplies the three condensers generally are 8 μfd. each, although the input condenser, C_1, sometimes is reduced to 4 μfd. Inductances of 10 to 20 henrys each will give satisfactory filtering with these capacity values.

For ripple frequencies other than 120 cycles, the inductance and capacity values should be multiplied by the ratio $120/F$, where F is the actual ripple frequency.

The bleeder resistance, R, should be chosen to draw 10 per cent or less of the rated output current of the supply. Its value is equal to $1000E/I$, where E is the output voltage and I the load current in milliamperes.

Rectifier peak current — The ratio of rectifier peak current to average load current is high with a condenser-input filter. Small rectifier tubes designed for low-voltage supplies (type 80, etc.) generally carry load-current ratings based on the use of condenser-input filters. With rectifiers for higher power, such as the 866/866-A, the load current should not exceed 25 per cent of the rated peak plate current for one tube when a full-wave rectifier is used, or one-eighth the half-wave rating.

Output voltage — The d.c. output voltage from a condenser-input supply will, with light loads or no load, approach the peak transformer voltage. This is 1.41 times the r.m.s. voltage (§ 2-7) of the transformer secondary, in the case of Figs. 801-A and C, or 1.41 times the voltage from the center-tap to one end of the secondary in Fig. 801-B. At heavy loads, it may decrease to the *average* value of secondary voltage or about 90 per cent of the r.m.s. voltage, or even less. Because of this wide range of output voltage with load current, the voltage regulation (§ 8-1) is inherently poor.

The output voltage obtainable from a given supply cannot readily be calculated, since it depends critically upon the load current and filter constants. Under average conditions it will be approximately equal to or somewhat less than the r.m.s. voltage between the center-tap and one end of the secondary in the full-wave center-tap rectifier circuit (§ 8-3).

Ratings of components — Because the output voltage may rise to the peak transformer voltage at light loads, the condensers should have a working-voltage rating (§ 8-4) at least as high and preferably somewhat higher, as a safety factor. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center-tap, the minimum safe condenser voltage rating will be 550 × 1.41 or 775 volts. An 800-volt, or preferably a 1000-volt, condenser should be used. Filter chokes should have the inductance specified at full-load current, and must have insulation between the winding and the core adequate to withstand the maximum output voltage.

ℂ 8-6 Choke-Input Filters

Ripple voltage — The circuit of a single-section choke-input filter is shown in Fig. 803-A. For 120-cycle ripple, a close approximation of the ripple to be expected at the output of the filter is given by the formula:

$$\left.\begin{matrix}\text{Single}\\ \text{Section}\\ \text{Filter}\end{matrix}\right\}\ \%\ \text{Ripple} = \frac{100}{LC}$$

where L is in henrys and C in μfd. The product, LC, must be equal to or greater than 20 to reduce the ripple to 5 per cent or less. This figure represents in most cases, the economical limit for the single-section filter. Smaller percentages of ripple usually are more economically obtained with the two-section filter of Fig. 803-B. The ripple percentage (120-cycle ripple) with this arrangement is given by the formula:

$$\left.\begin{matrix}\text{Two}\\ \text{Section}\\ \text{Filter}\end{matrix}\right\}\ \%\ \text{Ripple} = \frac{650}{L_1L_2\ (C_1 + C_2)^2}$$

For a ripple of 0.25 per cent or less, the denominator should be 2600 or greater.

These formulas can be used for other ripple frequencies by multiplying each inductance and capacity value in the filter by the ratio $120/F$, where F is the actual ripple frequency.

The distribution of inductance and capacity in the filter will be determined by the value of input-choke inductance required (next paragraph), and the permissible a.c. output impedance. If the supply is intended for use with an audio-frequency amplifier, the reactance (§ 2-8) of the last filter condenser should be small (20 per cent or less) compared to the other a.f. resistance or impedance in the circuit, usually the tube plate resistance and load resistance (§3-2, 3-3). On the basis of a lower a.f. limit of 100 cycles for speech amplification (§ 5-9), this condition is usually satisfied when the output capacity (last filter capacity) of the filter is 4 to 8 μfd., the higher value being used for the lower tube and load resistances.

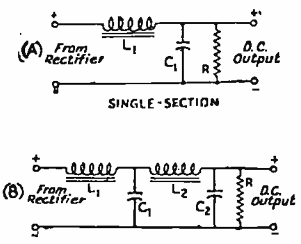


Fig. 803 — Choke-input filter circuits.

The input choke — The rectifier peak current and the power-supply voltage regulation depend almost entirely upon the inductance of the input choke in relation to the load resistance (§ 8-1). The function of the choke is to raise the ratio of average to peak current (by its energy storage), and to prevent the d.c. output voltage from rising above the average value (§ 2-7) of the a.c. voltage applied to the rectifier. For both purposes, its impedance (§ 2-8) to the flow of the a.c. component (§ 8-4) must be high.

The value of input-choke inductance which prevents the d.c. output voltage from rising above the average of the rectified a.c. wave is the *critical inductance*. For 120-cycle ripple, it is given by the approximate formula:

$$L_{\text{crit.}} = \frac{\text{Load resistance (ohms)}}{1000}$$

For other ripple frequencies, the inductance required will be the above value multiplied by the ratio of 120 to the actual ripple frequency.

With inductance values less than critical, the d.c. output voltage will rise because the filter tends to act as a condenser-input filter (§ 8-5). With critical inductance, the peak plate current of one tube in a center-tap rectifier will be approximately 10 per cent higher than the d.c. load current taken from the supply.

An inductance of twice the critical value is called the *optimum* value. This value gives a further reduction in the ratio of peak to average plate current, and represents the point at which further increase in inductance does not give correspondingly improved operating characteristics.

Swinging chokes — The formula for critical inductance indicates that the inductance required varies widely with the load resistance. In the case where there is no load except the bleeder (§ 8-4) on the power supply, the critical inductance required is highest; much lower values are satisfactory when the full-load current is being delivered. Since the inductance of a choke tends to rise as the direct current flowing through it is decreased (§ 8-4), it is possible to effect an economy in materials by designing the choke to have a "swinging" characteristic such that it has the required critical inductance value with the bleeder load only, and about the optimum inductance value at full load. If the bleeder resistance is 20,000 ohms and the full-load resistance (including the bleeder) is 2500 ohms, a choke which swings from 20 henrys to 5 henrys over the full output-current range will fulfill the requirements.

Resonance — Resonance effects in the series circuit across the output of the rectifier which is formed by the first choke (L_1) and first filter condenser (C_1) must be avoided, since the ripple voltage would build up to large values (§ 2-10). This not only is the opposite action to that for which the filter is intended, but also may cause excessive rectifier peak currents and abnormally high inverse peak voltages. For full-wave rectification the ripple frequency will be 120 cycles for a 60-cycle supply (§ 8-4), and resonance will occur when the product of choke inductance in henrys times condenser capacity in microfarads is equal to 1.77. The corresponding figure for 50-cycle supply (100-cycle ripple frequency) is 2.53, and for 25-cycle supply (50-cycle ripple frequency), 13.5. At least twice these products should be used to ensure against resonance effects.

Output voltage — Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by the equation:

$$E_o = 0.9E_t - \frac{(I_b + I_L)(R_1 + R_2)}{1000} - E_r$$

where E_o is the output voltage; E_t is the r.m.s. voltage applied to the rectifier (r.m.s. voltage between center-tap and one end of the secondary in the case of the center-tap rectifier); I_b and I_L are the bleeder and load currents, respectively, in milliamperes; R_1 and R_2 are the resistances of the first and second filter chokes; and E_r is the drop between rectifier plate and cathode (§ 8-2). These voltage drops are shown in Fig. 804.

At no load I_L is zero, hence the no-load voltage may be calculated on the basis of bleeder current only. The voltage regulation may be determined from the no-load and full-load voltages (§ 8-1).

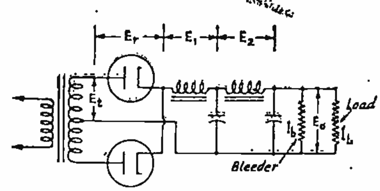


Fig. 804 — Voltage drops in the power-supply circuit.

Ratings of components — Because of better voltage regulation, filter condensers are subjected to smaller variations in d.c. voltage than in the condenser-input filter (§ 8-5). However, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no external load on the power supply, since the voltage then will rise to the same maximum value as with a condenser-input filter.

The input choke may be of the swinging type, the required no-load and full-load inductance values being calculated as described above. The second choke (*smoothing choke*) should have constant inductance with varying d.c. load currents. Values of 10 to 20 henrys ordinarily are used. Since chokes usually are placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply.

8-7 The Plate Transformer

Output voltage — The output voltage of the plate transformer depends upon the required d.c. load voltage and the type of rectifier circuit. With condenser-input filters, the r.m.s. secondary voltage usually is made equal to or slightly more than the d.c. output voltage, allowing for voltage drops in the rectifier tubes and filter chokes as well as in the transformer itself. The full-wave center-tap rectifier requires a transformer giving this voltage each side of the secondary center-tap (§ 8-3).

With a choke-input filter, the required r.m.s. secondary voltage (each side of center-tap for a center-tap rectifier) can be calculated by the equation:

$$E_t = 1.1\left[E_o + \frac{I(R_1 + R_2)}{1000} + E_r\right]$$

where E_o is the required d.c. output voltage, I is the load current (including bleeder current) in milliamperes, R_1 and R_2 are the resistances of the filter chokes, and E_r is the voltage drop in the rectifier. E_t is the full-load r.m.s. (§ 2-7) secondary voltage; the open-circuit voltage usually will be 5 to 10 per cent higher.

Volt-ampere rating — The volt-ampere rating (§ 2-8) of the transformer depends upon the type of filter (condenser or choke input).

With a condenser-input filter the heating effect in the secondary is higher because of the high ratio of peak to average current, consequently the volt-amperes consumed by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance (§ 8-6), the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{Sec. V.A.} = 0.00075\ EI$$

where E is the *total* r.m.s. voltage of the secondary (between the outside ends in the case of a center-tapped winding) and I is the d.c. output current in milliamperes (load current plus bleeder current). The primary volt-amperes will be 10 to 20 per cent higher because of transformer losses.

ℂ 8-8 Voltage Stabilization

Gaseous regulator tubes — There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit (such as the oscillator in a superhet receiver or the frequency-controlling oscillator in a transmitter) at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current. In such applications, gaseous regulator tubes (VR105–30, VR150–30, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately wide current range. The first number in the tube designation indicates the terminal voltage, the second the maximum permissible tube current.

The fundamental circuit for a gaseous regulator is shown in Fig. 805-A. The tube is connected in series with a *limiting resistor*, R_1, across a source of voltage which must be higher than the *starting* voltage, or voltage required for ionization of the gas in the tube. The starting voltage is about 30 per cent higher than the operating voltage. The load is connected in parallel with the tube. For stable operation, a minimum tube current of 5 to 10 ma. is required. The maximum permissible current with most types is 30 ma.; consequently, the load current cannot exceed 20 to 25 ma. if the voltage is to be stabilized over a range from zero to maximum load current.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when there is no load current. The latter value is generally used. It is given by the equation:

$$R = \frac{1000\ (E_s - E_r)}{I}$$

Where R is the limiting resistance in ohms, E_s is the voltage of the source across which the tube and resistor are connected, E_r is the rated voltage drop across the regulator tube, and I is the maximum tube current in milliamperes (usually 30 ma.).

Fig. 805-B shows how two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. The limiting resistor may be calculated as above, using the sum of the voltage drops across the two tubes for E_r. Since the upper tube must carry more current than the lower, the load connected to the low-voltage tap must take small current. The total current taken by the loads on both the high and low taps should not exceed 20 to 25 milliamperes.

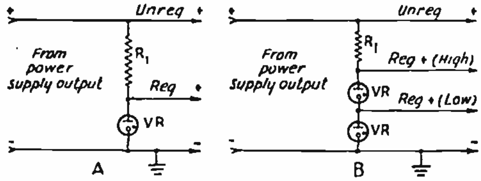


Fig. 805 — Voltage-stabilizing circuits using VR tubes.

Voltage regulation of the order of 1 per cent can be obtained with circuits of this type.

Electronic voltage regulation — A voltage regulator circuit suitable for higher voltages and currents than the gaseous tubes, and also having the feature that the output voltage can be varied over a rather wide range, is shown in Fig. 806. A high-gain voltage amplifier tube (§ 3-3), usually a sharp cut-off pentode (§ 3-5) is connected in such a way that a small change in the output voltage of the power supply causes a change in grid bias, and thereby a corresponding change in plate current. Its plate current flows through a resistor (R_5), the voltage drop across which is used to bias a second tube — the "regulator" tube — whose plate-cathode circuit is connected in series with the load circuit. The regulator tube therefore functions as an automatically variable series resistor. Should the output voltage increase slightly the bias on the control tube will become more positive, causing the plate current of the control tube to increase and the drop across R_5 to increase correspondingly. The bias on the regulator tube therefore becomes more negative and the effective resistance of the regulator tube increases, causing the terminal voltage to drop. A decrease in output voltage causes the reverse action. The time lag in the action of the system is negligible, and with proper circuit constants the output voltage can be held within a fraction of a per cent throughout the useful range of load currents and over a wide range of supply voltages.

An essential in this system is the use of a constant-voltage bias source for the control tube. The voltage change which appears at the grid of the tube is the *difference* between a fixed negative bias and a positive voltage which is taken from the voltage divider across the output. To get the most effective control, the negative bias must not vary with plate current. The most satisfactory type of bias is a dry battery of 45 to 90 volts, but a gaseous regulator tube (VR75–30) or a neon bulb of the type without a resistor in the base may be used instead. If the gas tube or neon bulb is used, a negative-resistance type of oscillation (§ 3-7) may take place at audio frequencies or higher, in which case a condenser of 0.1 μfd. or more should be connected across the tube. A similar condenser between the control-tube grid and cathode also is frequently helpful in this respect.

The variable resistor, R_3, is used to adjust the bias on the control tube to the proper operating value. It also serves as an output voltage control, setting the value of regulated voltage within the existing operating limits.

The maximum output voltage obtainable is equal to the power-supply voltage minus the minimum drop through the regulator tube. This drop is of the order of 50 volts with the tubes ordinarily used. The maximum current also is limited by the regulator tube; 100 milliamperes is a safe value for the 2A3. Two or more regulator tubes may be connected in parallel to increase the current-carrying capacity, with no change in the circuit.

ℭ 8-9 Bias Supplies

Requirements — A bias supply is not called upon to deliver current to a load circuit, but simply to furnish a fixed grid voltage to set the operating point of a tube (§ 3-3). However, in most applications it is nevertheless true that current flows through the bias supply, because such supplies are used chiefly in connection with power amplifiers of the Class-B and Class-C type, where grid-current flow is a feature of operation (§ 3-4). In circuit design a bias supply resembles the rectified-a.c. plate supply (§ 8-1), having a transformer-rectifier-filter system employing similar circuits. Bias supplies may be classified in two types, those furnishing only *protective* bias, intended to prevent excessive plate current flow in a power tube in case of loss of grid leak bias (§ 3-6) from excitation failure, and those which furnish the actual *operating* bias for the tubes. In the former type, voltage regulation (§ 8-1) is relatively unimportant; in the latter it may be of considerable importance.

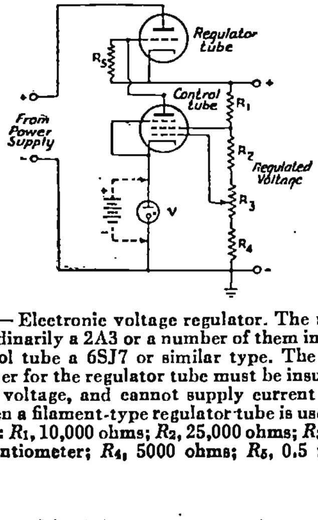


Fig. 806 — Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type regulator tube is used. Typical values: R_1, 10,000 ohms; R_2, 25,000 ohms; R_3, 10,000-ohm potentiometer; R_4, 5000 ohms; R_5, 0.5 megohm.

In general, a bias supply should have well-filtered d.c. output, especially if it furnishes the operating bias for the stage, since ripple voltage may modulate the signal on the grid of the amplifier tube (§ 5-1). Condenser-input filters are generally used, since the regulation of the supply is not a function of the filter. The constants given in § 8-5 are applicable.

Voltage regulation — A bias supply must always have a bleeder resistance (§ 8-4) connected across its output terminals, to provide a d.c. path from grid to cathode of the tube being biased. Although the grid circuit takes no current from the supply, grid current flows through the bleeder resistor and the voltage across the resistor therefore varies with grid current. This variation in voltage is practically independent of the bias-supply design unless special voltage-regulating means are used.

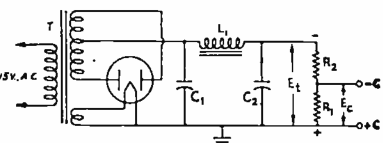


Fig. 807 — Supply for furnishing protective bias to a power amplifier. The transformer, *T*, should furnish peak voltage at least equal to the protective bias required.

Protective bias — This type of bias supply is designed to give an output voltage sufficient to bias the tube to which it is applied at or near the plate-current cut-off point (§ 3-2). A typical circuit is given in Fig. 807. The resistance, R_1, is the grid-leak resistor (§ 3-6) for the amplifier tube with which the supply is used, and the normal operating bias is developed by the flow of grid current through this resistor. R_2 is connected in series with R_1 across the output of the supply, to reduce the voltage across R_1, when there is no grid-current flow, to the cut-off value for the tube being biased. The value of R_2 is given by the formula:

$$R_2 = \frac{E_t - E_c}{E_c} \times R_1$$

where E_t is the output voltage of the supply with R_2 and R_1 in series as a load, E_c is the cut-off bias, and R_1 is as described above.

When such a supply is used with a Class-C amplifier, the voltage across R_1 from grid-current flow will normally be higher than that from the bias supply itself, since the latter is adjusted to cut-off while the operating bias will be twice cut-off or higher (§ 3-4). In some cases the grid-leak voltage may even exceed the peak output voltage of the transformer (1.41 times half the total secondary voltage, in the circuit shown). The filter condensers in such a bias supply must, therefore, be rated to stand the maximum operating bias voltage on the Class-C amplifier, if this voltage exceeds the nominal output voltage of the supply.

Voltage stabilization — When the bias supply furnishes operating rather than simply protective bias, the value of bias voltage should be as constant as possible even when the grid current of the biased tube varies. A simple method of improving bias voltage regulation is to make the bleeder resistance low enough so that the current through it from the supply is several times the maximum grid current to be expected. By this means, the percentage variation in current is reduced. This method requires, however, that a considerable amount of power be dissipated in the bleeder, which in turn calls for a relatively large power transformer and filter choke.

Bias-voltage variation may also be reduced by means of a regulator tube, as shown in Fig. 808. The regulator tube usually is a triode

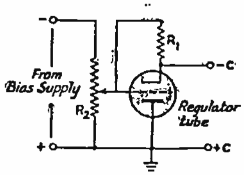


Fig. 808 — Automatic voltage regulator for bias supplies. For best operation the tube used should be one having high mutual conductance (§ 3-2).

having a plate-current rating adequate to carry the expected grid current. It is cathode-biased (§ 3-6) by the resistor, R_1, which is of the order of several hundred thousand ohms or a few megohms, so that with no grid current the tube is biased practically to cut-off. Because of this high resistance, the grid current will flow through the plate resistance of the regulator tube, which is comparatively low, rather than through R_1 and R_2; hence the voltage from the supply, across R_1 and the cathode-plate circuit of the regulator tube in series, can be considered constant. The bias voltage is equal to the voltage across the tube alone. When grid current flows, the voltage across the tube will tend to increase; hence the drop across R_1 decreases, lowering the bias on the regulator and reducing its plate resistance. This, in turn, reduces the tube voltage drop, and the bias voltage tends to remain constant over a fairly wide range of grid current values.

At low bias voltages it may be necessary to use a number of tubes in parallel to get sufficient variation of plate resistance for good regulating action. The bias supply must furnish the required bias voltage plus the voltage required to bias the regulator tube to cut-off, considering the output bias voltage as the plate voltage applied to the regulator. The current taken from the bias supply is negligible. R_2 may be tapped to provide a range of bias voltages to meet different tube requirements.

Multi-stage bias supplies — Where several power amplifier tubes are to be biased from a single supply, the various bias circuits must be isolated by some means. If the grid currents of all stages should flow through a single bleeder resistor, a variation in grid current in one stage would change the bias on all, a condition which would interfere with effective adjustment and operation of the transmitter.

When protective bias is to be furnished several stages, the circuit arrangement of Fig. 809, using rectifier tubes to isolate the individual grid-leaks of the various stages, may be employed. In the diagram, two type 80 rectifiers are used to furnish bias to four stages. Each pair of resistors (R_1R_2) constitutes a separate bleeder across the bias supply. R_1 is the grid-leak for the biased stage; R_2 is a dropping resistor to adjust the voltage across R_1 to the cut-off value (without grid-current flow) for the biased tube. The values of R_1 and R_2 may be calculated as described in the paragraph on protective bias. In this case, the bias supply should be designed to have inherently good voltage regulation; i.e., a choke-input filter with appropriate filter and bleeder constants (§ 8-6) should be used, the bleeder being separate from those associated with the rectifier tubes. When the voltage across R_1R_2 rises because of grid-current flow through R_1, the load on the supply will vary (hence the necessity for good voltage regulation in the supply), but there is no interaction of grid currents in the separate bleeders because the rectifiers can pass current only in one direction.

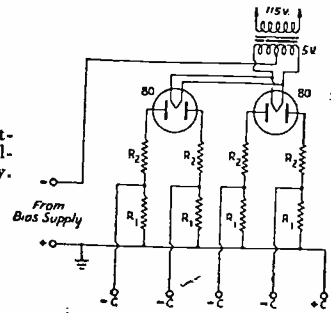


Fig. 809 — Isolating circuit for multiple bias supply.

When a single supply is to furnish operating bias for several stages, a separate regulator-tube circuit (Fig. 808) may be used for each one. Individual voltages for the various stages can be obtained by appropriate taps on R_2.

Well-regulated bias for several stages may be obtained by the use of gaseous regulator tubes, when the voltage and current ratings of the tubes permit their use. This is shown in Fig. 810. A single tube or two or more in series can be used to give the desired bias-voltage drop; the bias supply voltage must be high enough to provide starting voltage for the tubes in series. R_1 is the protective resistance (§ 8-8); its value should be calculated for minimum stable tube current. The maximum grid current that can be handled is 20 to 25 milliamperes with available regulator tubes.

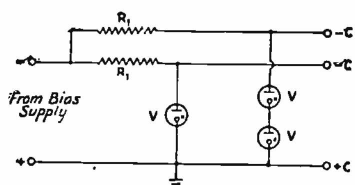


Fig. 810 — Use of VR tubes to stabilize bias voltage.

ℂ 8-10 Miscellaneous Power-Supply Circuits

Voltage dividers — A voltage divider is a resistor connected across a source of voltage and tapped at appropriate points (§ 2-6). Since the voltage at any tap depends upon the current drawn from the tap, the voltage regulation (§ 8-1) is inherently poor. Hence, a voltage divider is best suited to applications where the currents drawn are constant, or where separate voltage-regulating circuits (§ 8-8) are used to compensate for voltage variations at the taps.

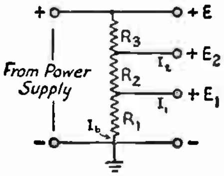
A typical voltage-divider arrangement is shown in Fig. 811. The terminal voltage is E, and two taps are provided to give lower voltages, E_1 and E_2, at currents I_1 and I_2 respectively. The smaller the resistance between taps in proportion to the total resistance, the smaller the voltage between the taps. For convenience, the voltage divider in the figure is considered to be made up of separate resistances, R_1, R_2, R_3, between taps. R_1 carries only the bleeder current, I_b. R_2 carries I_1 in addition to I_b; R_3 carries I_2, I_1 and I_b. To calculate the resistances required, a bleeder current, I_b, must be assumed; generally it is low compared to the total load current (10 per cent or so). Then the required values can be calculated as shown below, I being in amperes.

Fig. 811 — Typical voltage-divider circuit.

$$R_1 = \frac{E_1}{I_b}$$

$$R_2 = \frac{E_2 - E_1}{I_b + I_1}$$

$$R_3 = \frac{E - E_2}{I_b + I_1 + I_2}$$



The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law (§ 2-6) using the voltage drop across it and the total current through it. The power dissipated by each section may be calculated by multiplying I and E.

Transformerless plate supplies — The line voltage is rectified directly, without a step-up power transformer, for certain applications (such as some types of receivers) where the low voltage so obtained is satisfactory. A simple power supply of this variety, often called the "a.c.-d.c." type, is shown in Fig. 812. Rectifier tubes for this purpose have heaters operating at relatively high voltages (12.6, 25, 35, 45, 50, 70 or 115 volts), which can be connected across the a.c. line in series with other tube filaments and/or a resistor, R, of suitable value to limit the current to the rated value for the tubes.

The half-wave circuit shown has a fundamental ripple frequency equal to the line frequency (§ 8-4) and hence requires more inductance and capacity in the filter for a given ripple percentage (§ 8-5) than the full-wave rectifier. A condenser-input filter generally is used. The input condenser should be at least 16 μfd. and preferably 32 or 40 μfd., to keep the output voltage high and to improve voltage regulation. Frequently a second filter section (§ 8-5) is sufficient to provide smoothing.

No ground connection can be used on the

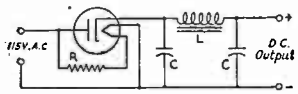


Fig. 812 — Transformerless plate supply with half-wave rectifier. Other filaments are connected in series with R.

power supply unless the grounded side of the power line is connected to the grounded side of the supply. Receivers using an a.c.-d.c. supply usually are grounded through a low capacity (0.05 μfd.) condenser, to avoid short-circuiting the line should the line plug be inserted in the socket the wrong way.

Voltage multiplier circuits — Transformerless voltage multiplier circuits make it possible to obtain d.c. voltages higher than the line voltage without using step-up transformers. By alternately charging two or more condensers to the peak line voltage and allowing them to discharge in series, the total output voltage becomes the sum of the voltages appearing across the individual condensers. The required switching operation is performed automatically by diode rectifier tubes associated with the condensers.

A half-wave voltage doubler is shown in Fig. 813-A. In this circuit when the plate of the lower diode is positive the tube passes current, charging C_1 to a voltage equal to the peak line voltage less the tube drop. When the line polarity reverses at the end of the half cycle the voltage resulting from the charge in C_1 is added to the line voltage, the upper diode meanwhile similarly charging C. C, however, does not receive its full charge because it be-

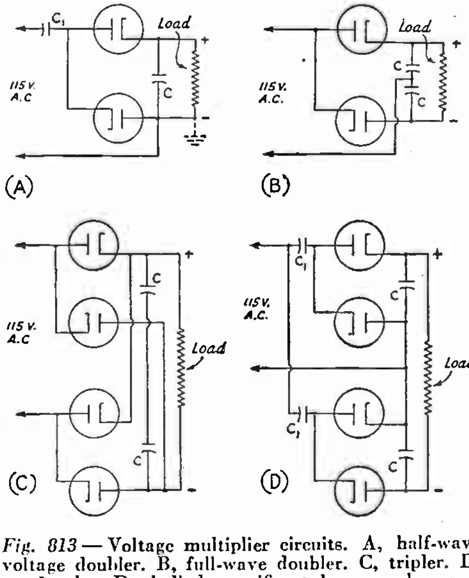


Fig. 813 — Voltage multiplier circuits. A, half-wave voltage doubler. B, full-wave doubler. C, tripler. D, quadrupler. Dual diode rectifier tubes may be used.

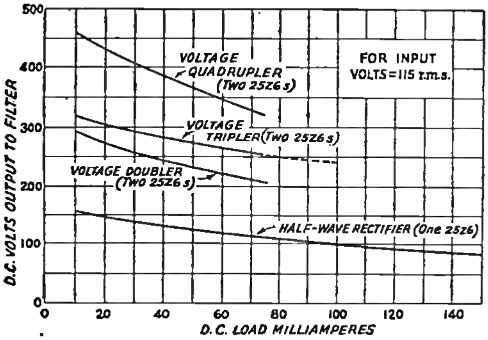


Fig. 814 — Curves showing the d.c. output voltage and the regulation under load for voltage-multiplier circuits.

gins discharging into the load resistance as soon as the upper diode becomes conductive. For this reason, the output is somewhat less than twice the line peak voltage. As with any half-wave rectifier, the ripple frequency corresponds to the line frequency.

The full-wave voltage doubler at B is more popular than the half-wave type. One diode charges C_1 when the polarity between its plate and cathode is positive while the other section charges C when the line polarity reverses. Thus each condenser is charged separately to the same d.c. voltage, and the two discharge in series into the load circuit. The ripple frequency with the full-wave doubler is twice the line frequency (§ 8-4). The voltage regulation is inherently poor and depends critically upon the capacities of C_1 and C_2, being better as these capacities are made larger. A typical supply with 16 μfd. at C_1 and C_2 will have an output voltage of approximately 300 at light loads, as shown in Fig. 814.

The voltage tripler in Fig. 813-C comprises four diodes in a full-wave doubler and full-wave rectifier combination. The ripple frequency is that of the line as in a half-wave circuit, because of the unbalanced arrangement, but the output to the first filter condenser is very nearly three times the line voltage, and the regulation is better than in other voltage multiplier arrangements, as shown in Fig. 814.

Fig. 813-D is a voltage quadrupler with two half-wave doublers connected in series, discharging the sum of the accumulated voltages in the associated condensers into the filter input. The quadrupler is by no means the ultimate limit in voltage multiplication. Practical power supplies have been built using up to twelve doubler stages in series.

In the circuits of Fig. 813 C should have a working voltage rating of 350 volts and C_1 of 250 volts for a 115-volt line. Their capacities should be at least 16 μfd. each. Subsequent filter condensers must, however, withstand the *peak* total output voltage — 450 volts in the case of the tripler and 600 for the quadrupler.

No direct ground can be used on any of these supplies or on associated equipment. If an r.f. ground is made through a condenser the capacity should be small (0.05 μfd.), since it is in shunt from plate to cathode of one rectifier.

Duplex plate supplies — In some cases it may be advantageous economically to obtain two plate-supply voltages from a single power supply, making one or more of the components serve a double purpose. Circuits of this type are shown in Figs. 815 and 816.

In Fig. 815, a bridge rectifier is used to obtain the full transformer voltage, while a connection is also brought out from the center-tap to obtain a second voltage corresponding to half the total transformer secondary voltage. The sum of the currents drawn from the two taps should not exceed the d.c. ratings of the rectifier tubes and transformer. Filter values for each tap are computed separately (§ 8-6).

Fig. 816 shows how a transformer with multiple secondary taps may be used to obtain both high and low voltages simultaneously. A separate full-wave rectifier is used at each tap. The filter chokes are placed in the common negative lead, but separate filter condensers are required. The sum of the currents drawn from each tap must not exceed the transformer rating, and the chokes must be rated to carry the total load current. Each bleeder resistance

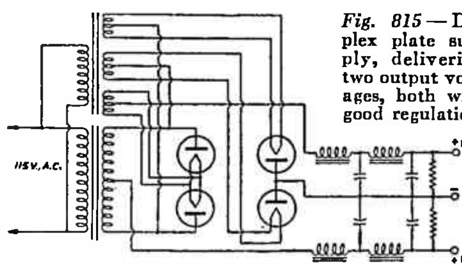


Fig. 815 — Duplex plate supply, delivering two output voltages, both with good regulation.

should have a value in ohms 1000 times the maximum rated inductance in henrys of the swinging choke, L_1, for best regulation (§ 8-6).

Rectifiers in parallel — Vacuum-type rectifiers may be connected in parallel (plate to plate and cathode to cathode) for higher current-carrying capacity with no circuit changes.

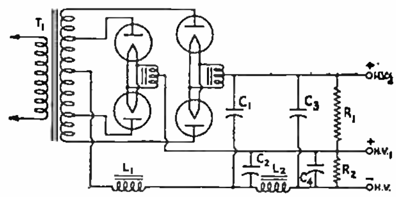


Fig. 816 — Power supply in which a single transformer and set of chokes serve for two different output voltages.

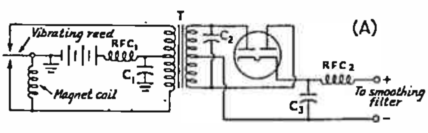
When mercury-vapor rectifiers are connected in parallel, slight differences in tube characteristics may make one ionize at a slightly lower voltage than the other. Since the ignition voltage is higher than the operating voltage the first tube to ionize carries the whole load, since the voltage drop is then too low to ignite the second tube. This can be prevented by connecting 50- to 100-ohm resistors in series with each plate, thereby insuring that a high-enough voltage for ignition will be available.

Vibrator power supplies—The vibrator type of power supply consists of a special step-up transformer combined with a vibrating interrupter (*vibrator*). When the unit is connected to a storage battery, plate power is obtained by passing current from the battery through the primary of the transformer. The circuit is made and reversed rapidly by the vibrator contacts, interrupting the current at regular intervals to give a changing magnetic field which induces a voltage in the secondary (§ 2-5). The resulting square-wave d.c. pulses in the primary of the transformer cause an alternating voltage to be developed in the secondary. This high-voltage a.c. in turn is rectified, either by a vacuum-tube rectifier or by an additional synchronized pair of vibrator contacts. The rectified output is pulsating d.c., which may be filtered by ordinary means (§ 8-5). The smoothing filter can be a single-section affair, but the filter output capacity should be fairly large—16 to 32 μfd.

Fig. 817 shows the two types of circuits. At A is shown the *nonsynchronous* type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet coil is short-circuited, deënergizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

The synchronous circuit of Fig. 817-B is provided with an extra pair of contacts which rectify the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center-tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment.

The buffer condenser, C_2, across the transformer secondary absorbs the surges which



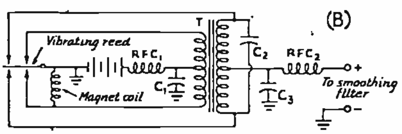


Fig. 817—Basic types of vibrator power-supply circuits.

occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes very high voltages to be induced in the secondary (§ 2-5). Without this condenser excessive sparking occurs at the vibrator contacts, shortening the vibrator life. Correct values usually lie between 0.005 and 0.03 μfd. and for 250–300-volt supplies the condenser should be rated at 1500 to 2000 volts d.c. The exact capacity is critical, and should be determined experimentally. The optimum value is that which results in least battery current for a given rectified d.c. output from the supply. In practice the value can be determined by observing the degree of vibrator sparking as the capacity is changed. When the system is operating properly there should be practically no sparking at the vibrator contacts. A 5000-ohm resistor in series with C_2 will limit the secondary current to a safe value should the condenser fail.

A more exact check on the operation can be secured with an oscilloscope having a linear sweep circuit which can be synchronized with the vibrator. The vertical plates should be connected across the outside ends of the transformer primary winding to show the input voltage waveshape. Fig. 818-B shows an idealized trace of the optimum waveform when the buffer capacity is adjusted to give proper operation throughout the life of the vibrator. The horizontal lines in the trace represent the voltage during the time the vibrator contacts are closed, which should be approximately 90 per cent of the total time. When the contacts are open the trace should be partly tilted and partly vertical, the tilted part being 60 per cent of the total connecting trace. The oscilloscope will show readily the effect of the buffer capacity on the percentage of tilt. In actual patterns the horizontal sections are likely to droop somewhat because of the resistance drop in the battery leads as the current builds up through the primary inductance (Fig. 818-C).

Sparking at the vibrator contacts causes r.f. interference ("hash," which can be distinguished from hum by its harsh, sharper pitch) when used with a receiver. To minimize this, r.f. filters are incorporated, consisting of RFC_1 and C_1, in the battery circuit and RFC_2 with C_3 in the d.c. output circuit. C_1 is usually from 0.5 to 1 μfd., a 50-volt rating being adequate. RFC_1 consists of about 50 turns of No. 12 or No. 14 wound to about half-inch diameter, large wire being required to carry the rather heavy battery current without undue loss of voltage. A choke of these specifications should be adequate, but if there is persistent trouble with hash it may be beneficial to experiment with other sizes. Bank-wound chokes are more compact and give higher inductance for a given resistance. In the secondary filter, C_3 may be of the order of 0.01 to 0.1 μfd., and RFC_2 a 2.5-millihenry r.f. choke of ordinary design.

A 100-μμfd. mica condenser, connected from the positive output lead to the "hot" side

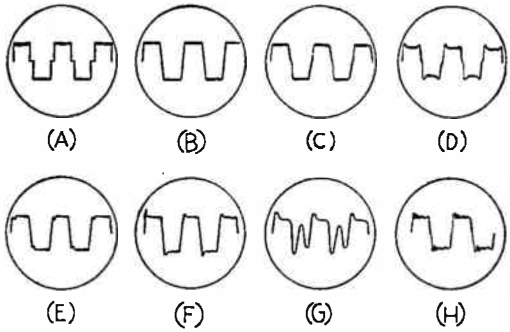


Fig. 818 — Characteristic vibrator waveforms as viewed on the oscilloscope. A, ideal theoretical trace for resistive load; current flow stops instantly when vibrator contacts open and resumes approximately 1 micro-second later (with standard 115-cycle vibration frequency) after interrupter arm moves across for the next half-cycle. B, ideal practical waveform for inductive load (transformer primary) with correct buffer capacity. C, practical approximation of B for loaded nonsynchronous vibrator. D, satisfactory practical trace for synchronous (self-rectifying) vibrator under load; the peaks result from the voltage drop in the primary when the secondary load is connected and not faulty operation.

Faulty operation is indicated in traces E through H: E, effect of insufficient buffering capacity (not to be mistaken for "bouncing" of contacts). The opposite condition — excessive buffering capacity — is indicated by slow build-up with rounded corners, especially on "open." F, overclosure caused by too-small buffer condenser (same condition as in E) with vibrator unloaded. G, "skipping" of worn-out or misadjusted vibrator, with interrupter making poor contact on one side. H, "bouncing" resulting from worn-out contacts or sluggish reed. G and H usually call for replacement of the vibrator.

of the "A" battery, may be helpful in reducing hash in certain power supplies. A trial is necessary to see whether or not it is required. It should be mounted right at the output socket.

Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough r.f. to cause interference in a sensitive receiver.

Testing in connection with hash elimination should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiving antenna leads by radiation from the supply itself and from the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the supply and leads.

The power supply should be built on a metal chassis, with all unshielded parts underneath. A bottom plate to complete the shielding is advisable. The transformer case, vibrator cover and the metal shell of the tube all should be grounded to the chassis. If a glass tube is used it should be enclosed in a tube shield. The battery leads should be evenly twisted, since these leads are more likely to radiate hash than any other part of a well-shielded supply. Experimenting with different values in the hash filters should come *after* radiation from the battery leads has been reduced to a minimum. Shielding the leads is not particularly helpful.

Line-voltage adjustment — In some localities the line voltage may vary considerably from the nominal 115 volts as the load on the power system changes. Since it is desirable to operate tube equipment, particularly filaments and heaters, at constant voltage for maximum life, a means of adjusting the line voltage to the rated value is desirable. This can be accomplished by the circuit shown in Fig. 820, utilizing a step-down transformer with a tapped secondary connected as an autotransformer (§ 2-9). The secondary preferably should be tapped in steps of two or three volts, and should have sufficient total voltage to compensate for the widest variations encountered. Depending upon the end of the secondary to which the line is connected, the voltage to the load can be made either higher or lower than the line voltage. A secondary winding capable of carrying five amperes will serve for loads up to 500 volt-amperes on a 115-volt line.

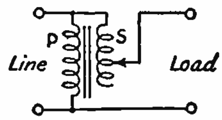


Fig. 819 — Line-voltage compensation by a tapped step-down autotransformer.

8-11 — Emergency Power Supply

Dry batteries — Dry-cell batteries are ideal for emergency receiver and low-power transmitter supplies because they provide steady, pure, direct current. Their disadvantages are weight, high cost, and limited current capability. In addition, they will lose their power even when not in use if allowed to stand idle for periods of a year or more. This makes them uneconomical if not used more or less continuously.

Table I shows the life to be expected from representative types of batteries under various current drains, based on intermittent service simulating typical operation. The continuous-service life will be somewhat greater at very low current drains and from one-half to two-thirds the intermittent life at higher drains.

The secret of long battery life at normal current drains lies in intermittent operation. The duration of "on" periods should be reduced to a minimum. The more frequent the rests given a dry-cell battery, the longer it will last. As an example, one standard type will last 50 per cent longer if it is operated for periods of one minute, with five-minute rest intervals, in 24-hour intermittent operation than if it is operated continuously for four hours per day, although the actual energy consumption in the 24-hour period is the same in both cases.

Storage batteries — The most universally acceptable self-contained power source is the storage battery. It has high initial capacity and can be recharged, so that its effective life is practically indefinite. It can be used to provide filament or heater power directly, and plate power through associated devices such as vibrator-transformers, dynamotors and genemotors, and a.c. converters. For emergency work a storage battery is a particularly convenient power source, since such batteries are universally available. In a serious emergency it is possible to obtain 6-volt storage batteries so long as there are automobiles to borrow them from, and for this reason the 6-volt storage battery makes an excellent unit around which to design a low-powered emergency station.

For maximum efficiency and usefulness the power drain on the storage battery should not exceed 15 or 20 amperes from the ordinary 100- or 120-ampere-hour 6-volt battery. Heavy connecting leads should be used to minimize the voltage drop; similarly, heavy-duty low-resistance switches are required.

Vibrator power supplies — For portable or mobile work, the most common source of power for both filaments and plates is the 6-volt automobile-type storage battery. Filaments may be heated directly from the battery, while plate power is obtained by passing current from the battery through the primary of a suitable transformer, interrupting it at regular intervals and rectifying the secondary output (§ 2-5) providing outputs as high as 400 volts at 200 ma. The high-voltage filter circuit usually is identical with that of an equivalent power source operating from the a.c. line (§ 8-5). Noise suppression filters, serving to minimize r.f. interference caused by the vibrator, are incorporated in manufactured units.

Although vibrator supplies are ordinarily used with 6-volt tubes, their use with 2-volt tubes is quite possible provided additional filament filtration is incorporated. This filter may consist of a small low-resistance iron-core filter choke or the voice-coil winding of a speaker transformer. The field coil of a loudspeaker designed to operate on 4 volts at the total filament current of the receiver may be used. The filaments are then connected in parallel, as usual, and placed in series with this winding across the 6-volt battery. In both 6- and 2-volt receivers, "hash" can be reduced by heavily by-passing the battery at the vibrator supply terminals, using fixed condensers of 0.25 to 1 μfd. capacity or more, and by including an r.f. choke of heavy wire in the battery lead near the condenser. Noise will be minimized if a single ground, consisting of a short, heavy copper strap, is used. Thorough shielding of the vibrator also will contribute to the noise reduction.

Table II lists standard commercial vibrator supplies suitable for use as emergency or portable power sources. Those units which include a hum filter are indicated. The vibrator supplies used with automobile receivers are satisfactory for receiver applications and for use with transmitters where the power requirements are small.

The efficiency of vibrator packs runs between about 60 to 75 per cent.

Dynamotors and genemotors — A dynamotor is a double-armature high-voltage generator, the additional winding serving as a driving motor. Dynamotors usually are operated from 6-, 12- or 32-volt storage batteries, and deliver output voltages from 300 to 1000 or more.

The genemotor is a refinement of the dynamotor, designed especially for automobile receiver, sound truck and similar applications. It has good regulation and efficiency, combined with economy of operation. Standard models of genemotors have ratings ranging from 135 volts at 30 ma. to 300 volts at 200 ma. or 500 volts at 200 ma., as can be seen from Table III. The normal efficiency averages around 50 per cent, increasing to better than 60 per cent in the higher-power units. The voltage regulation of a genemotor is comparable to that of well-designed a.c. supplies.

Successful operation of dynamotors and genemotors requires heavy, direct leads, mechanical isolation to reduce vibration, and thorough r.f. and ripple filtration. The shafts and bearings should be thoroughly "run in" before regular operation is attempted, and thereafter the tension of the bearings should be checked occasionally.

In mounting the genemotor, the support should be in the form of rubber mounting blocks, or equivalent, to prevent the transmission of vibration mechanically. The frame of the genemotor should be grounded through a heavy flexible connector. The brushes on the high-voltage end of the shaft should be by-passed with 0.002-μfd. mica condensers to a common point on the genemotor frame, preferably to a point inside the end cover close to the brush holders. Short leads are essential. It may prove desirable to shield the entire unit, or even to remove the unit to a distance of three or four feet from the receiver.

When the genemotor is used for receiving, a filter should be used similar to that described for vibrator supplies. A 0.01-μfd. 600-volt (d.c.) paper condenser should be connected in shunt across the output of the genemotor, followed by a 2.5-mh. r.f. choke in the positive high-voltage lead. From this point the output should be run through a "brute force" smoothing filter using 4- to 8-μfd. electrolytic condensers with a 15- or 30-henry choke having low d.c. resistance.

A.c.-d.c. converters — In some instances it is desirable to utilize existing equipment built for 115-volt a.c. operation. To operate such equipment with any of the power sources outlined above would require a considerable amount of rebuilding. This can be obviated by using a rotary converter capable of changing the d.c. from 6-, 12- or 32-volt batteries to 110-volt 60-cycle a.c. Such converter units are built to deliver output ranging from 40 to 300 watts.

The conversion efficiency of these units averages about 50 per cent. In appearance and operation they are similar to genemotors of equivalent rating. The over-all efficiency of the converter will be lower, however, because of losses in the a.c. rectifier-filter circuits and the necessity for converting heater as well as plate power.

TABLE I—BATTERY SERVICE HOURS

Estimated to 34-volt end-point per nominal 45-volt section.
Based on intermittent use of 3 to 4 hours daily.
(For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.	Weight		Current Drain in Ma.											
Eveready	Lb.	Oz.	5	10	15	20	25	30	40	50	60	75	100	150
386	14	—	2000	1100	690	510	400	320	200	170	130	100	50	30
486	13	5	1700	880	550	395	300	240	165	125	100	70	45	20
586	12	2	1400	800	530	380	260	185	130	85	60	40	30	14
585	8	13	900	450	290	210	130	100	60	45	25	20	11	5
762	3	3	320	140	81	54	37	27	—	—	—	—	—	—
482	2	—	320	140	81	54	37	27	—	—	—	—	—	—
738	1	2	160	70	30	20	10	7	—	—	—	—	—	—
733	—	10	50	20	11	7	5.2	—	—	—	—	—	—	—
455[1]	—	8.6	70	20	11	7	5.2	—	—	—	—	—	—	—

[1] Same life figures apply to 467, 67½-volt, 10.5 oz.

Estimated to 1-volt end-point per 1.5-volt unit.
Based on intermittent use of 3 to 4 hours daily.
(For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.	Weight		Voltage	Current Drain in Ma.										
Eveready	Lb.	Oz.		50	60	120	150	175	180	200	240	250	300	350
A-1300	8	4	1.25				2000	1715	1500	1333	1250	1200	1000	854
740	6	12	1.5				1400	1200		1050		775	625	
741[1]	2	14	1.5		1100	750				375	300	275	215	175
743	2	1	1.5		750	325				245		180	135	110
7111	2	2	1.5		700	320			200		120		90	
742	1	6	1.5		500	325			155	135	100	95	85	50
A-2300	15	8	2.5				2000	1715	1500	1333	1250	1200	1000	854
723	1		3.0		240	100			70		40		30	
746	1	3	4.5	200										
718[2]	3		6.0	375										

[1] Same life figures apply to 745, wt. 3 lbs.

[2] Same life figures apply to 747, wt. 3 lbs.

TABLE II—GASOLINE-ENGINE-DRIVEN GENERATORS, AIR-COOLED

Manufacturer				Output		Weight	Starter
Eicor	Kato	Onan	Pioneer	Volts	Watts	Lbs.	
3AP6[1]			BD-6[1]	110 a.c. or 6 d.c.	300 200	100	Push-button
	JR-35[2]			110 a.c.	300	65	Push-button
	JRA-3[2]			110 a.c.	350	65	Rope crank
	19-A			110 a.c. or 6 d.c.	350 200	95	Push-button
		358[1,3]		115 a.c.	350	91	Push-button
	JR-10[2]			110 a.c.	400	—	Rope crank
		5L[3]		110 a.c.	500	165	Push-button
	23A			110 a.c. or 6 d.c.	500 200	105	Push-button
6AP1	14A		BA-6[1]	110 a.c.	600	135	Push-button
		7L[3]		115 a.c.	750	195	Push-button
10AP1		10L[3,4]	BA-10[1]	110 a.c.	1000	170	Push-button
	26A			110 a.c.	1000	265	Manual
		OTC		110 a.c.	1500	135	Manual
			BA-15	110 a.c.	1500	365	Push-button

[1] Also available in remote-control models.
[2] Intermittent-duty model.

[3] Also available in manual-started type.
[4] 115-volt output, weight 200 lbs.

TABLE III—VIBRATOR SUPPLIES

Manufacturer's Type Number				Output		Rectifier	Output Filter
American Television and Radio Co.	Electronic Labs	Mallory	Radiart	Volts	Ma.		
VPM-F-7				90	10	Syn.	Yes
		VP-551[1]		125-150-175-200	100 max.	Syn.	No
			4201B[2]	250	50	Syn.	Yes
		VP-540		250	60	Syn.	Yes
			4204F[3]	100-150-250	35-40-60	Syn.	Yes
	605			150-200-250-275	35-40-50-65	Syn.	No
	604[4]	VP-552[5]		225-250-275-300	50-65-80-100	Syn.	No
			4201[6]	150-200-250-275-300	35-40-50-70-100	Syn.	No
	251[7]			300	100	Tube	Yes
		VP-555		300	200	Tube	Yes
VPM-6[8]	311[9]			250-275-300-325	50-75-100-125	Tube	Yes
		VP-557		400	150	Tube	Input cond.
			4202D	300-400	200-150	Tube	Yes
	606[10]			325-350-375-400 and 110 a.c. 60 cycle	125-150-175-200 20 watts	Tube	Input condenser

All inputs 6.3 volts d.c. unless otherwise noted.

[1] VP-553 same with tube rectifier.
[2] In weatherproof case. 4201B2 same with tube rectifier.
[3] 180-cycle vibrator, lightweight. 4204 same without filter.
[4] 601 same with tube rectifier; 602 same except 12 v. d.c. input and tube rectifier; 603 same except 32 v. d.c. input and tube rectifier.
[5] VP-554 same with tube rectifier; VP-G556 same except 12 v. d.c. input; VP-F558 same except 32 v. d.c. input.
[6] 4200D same with tube rectifier; 4200DF same with tube rectifier and output filter.
[7] 551 same with 12 v. d.c. input.
[8] Also available without filter.
[9] 511 same except 12 v. d.c. input.
[10] Input 6 v. d.c. or 110 v. a.c., 607 same except 12 v. d.c. or 110 v. a.c. input; 608 same except 32 v. d.c. or 110 v. a.c. input; 609 same except 110 v. d.c. or 110 v. a.c. input.

TABLE IV—DYNAMOTORS

Manufacturer's Type No.			Input		Output		Weight
Carter.	Eicor	Pioneer	Volts	Amps.	Volts	Ma.	Lbs.
135A			6	1.8	135	30	6½
180A			6	2.2	180	30	6½
240A			6	3.3	200	40	6½
210A			6	6.3	200	100	6½
220A			6	13	200	200	6½
250A	102[1]	E1W272[2]	6	5	250	50	6½
251A		E1W339[3]	6	9	250	100	6½
277A			6	6	275	75	6½
301A	106[4]	E2W351[5]	6	9.7	300	100	6½
315A	158[6]	E2W243[5]	6	15	300	150	7⅞
320A		RAOW158[7]	6	19	300	200	9½
351A			6	10	350	100	6½
355A	108	E2W256[5]	6	15	350	150	7⅞
352A			6	22	350	200	9½
401A			6	13	400	100	7⅞
		E2W438	6	14.2	400	125	9¼
415A	109[8]		6	20	400	150	7⅞
420A			6	25	400	200	9½
425A		RA1W201[9]	6	30	400	225	9½
450A	110[10]		6	33	400	250	9½
		E3W413	6	15	500	100	11
515A	111[11]		6	24	500	150	9½
520AR		RA1W189[12]	6	33	500	200	—

[1] Input current 4.6 amp.; wt. 4⅝ lbs.
[2] Wt. 7½ lbs.
[3] Input current 7.5 amp.; wt. 7½ lbs.
[4] Wt. 5 lbs.
[5] Wt. 9¼ lbs.
[6] Input current 14 amp.; wt. 5¾ lbs.
[7] Wt. 16 lbs.; input current 18 amp.
[8] Input current 17 amp.
[9] Wt. 17½ lbs.; input current 25 amp.
[10] Input current 27.5 amp.; wt. 7⅝ lbs.
[11] Input current 21.5 amp.; wt. 7⅝ lbs.
[12] Input current 27 amp.; wt. 17½ lbs.

Chapter Nine

Wave Propagation

9-1 Characteristics of Radio Waves

Relation to other forms of radiation — Radio waves differ from other forms of electromagnetic radiation principally in the order of their wavelength, which ranges from approximately 30,000 meters to a small fraction of a centimeter; i.e., their frequency ranges between about 10 kc. and 1,000,000 Mc. They travel at the same velocity as light waves (about 300,000,000 meters per second in free space) and can be similarly reflected, refracted and diffracted.

The total energy in a radio wave is evenly divided between traveling electrostatic and electromagnetic fields. The lines of force of these fields are at right angles to each other in a plane perpendicular to the direction of travel, as shown in Fig. 901.

Polarization — The polarization of a radio wave is taken as the direction of the lines of force in the electrostatic field. If the plane of this field is perpendicular to the earth, the wave is said to be *vertically polarized;* if it is parallel to the earth, the wave is *horizontally polarized.* The longer waves, when traveling along the ground, usually maintain their polarization in the same plane as was generated at the antenna. The polarization of shorter waves may be altered during travel, however, and sometimes will vary quite rapidly.

Reflection — Radio waves may be reflected from any sharply defined discontinuity of suitable characteristics and dimensions encountered in the medium in which they are traveling. Any conductor differing in dielectric constant from that medium offers such a discontinuity, if its dimensions are at least comparable to the wavelength. The surface of the earth, the boundaries between ionospheric layers, and boundaries between dissimilar air masses in the lower atmosphere are other examples of such discontinuities. Objects as small as an airplane, a tree or even a man's body will readily reflect the shorter waves.

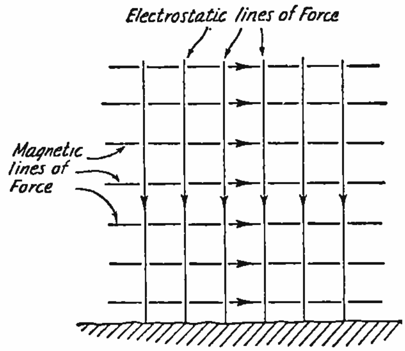


Fig. 901 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

Refraction — As in the case of light, a radio wave is bent when it moves obliquely into any medium having a different refractive index from that of the medium which it leaves. Since the velocity of propagation or travel differs in the two mediums, that part of the wave front which enters first travels faster or slower than the part which enters last, and so the wave front is turned or refracted (usually downward in the vertical plane). Refraction may take place in either the ionosphere (upper atmosphere) or the troposphere (lower atmosphere).

Diffraction — When a wave grazes the edge of an object in passing, it tends to be bent around that edge. This effect, called *diffraction,* results in a diversion of part of the energy of those waves which normally follow a straight or line-of-sight path, so that they may be received at some distance below the summit of an obstruction, or around its edges.

Types of waves — According to the altitude of the paths along which they are propagated, radio waves may be classified as *ionospheric waves, tropospheric waves* or *ground waves.*

The ionospheric wave (sometimes called the "sky wave,") is that part of the total radiation which is directed toward the ionosphere. Depending upon variable conditions in that region, as well as upon wavelength (or frequency), the ionospheric wave may or may not be returned to earth by the effects of refraction and reflection.

The tropospheric wave is that part of the total radiation which undergoes refraction and reflection in regions of abrupt change of dielectric constant in the troposphere, such as the boundaries between air masses of differing temperature and moisture content.

The ground wave is that part of the total radiation which is directly affected by the presence of the earth and its surface features. The ground wave has two components. One is the *surface wave,* which is an earth-guided wave, and the other is the *space wave* (not to be confused with the ionospheric or "sky wave.") The space wave is itself the resultant of two components — the *direct wave* and the *ground-reflected wave,* as shown in Fig. 902.

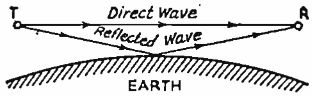


Fig. 902 — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

9-2 Ionospheric Propagation

The ionosphere — Communication between distant points by means of radio waves of frequencies ranging between 3 and 30 Mc. depends principally upon the ionospheric wave. Upon leaving the transmitting antenna, this wave travels upward from the earth's surface at such an angle that it would continue out into space were its path not bent sufficiently to bring it back to earth. The medium which causes such bending is the ionosphere, a region in the upper atmosphere, above a height of about 60 miles, where free ions and electrons exist in sufficient quantity to cause a change in the refractive index. This condition is believed to be the effect of ultraviolet radiation from the sun. The ionosphere is not a single region but is composed of a series of layers of varying densities of ionization occurring at different heights. Each layer consists of a central region of relatively dense ionization which tapers off in intensity both above and below.

Refraction, absorption and reflection — For a given density of ionization, the degree of refraction becomes less as the wavelength becomes shorter (or as the frequency increases). The bending therefore is less at high than at low frequencies, and if the frequency is raised to a sufficiently high value, a point is finally reached where the refractive bending becomes too slight to bring the wave back to earth, even though it may enter the ionized layer along a path which makes a very small angle with the boundary of the ionosphere.

The greater the density of ionization, the greater the bending at any given frequency. Thus, with an increase in ionization, the minimum wavelength which can be bent sufficiently for long-distance communication is lessened and the maximum usable frequency is increased.

The wave necessarily loses some of its energy in traveling through the ionosphere, this absorption loss increasing with wavelength and also with ionization density. Unusually high ionization, especially in the lower strata of the ionosphere, may cause complete absorption of the wave energy.

In addition to refraction, reflection may take place at the lower boundary of an ionized layer if it is sharply defined; i.e., if there is an appreciable change in ionization within a relatively short interval of travel. For waves approaching the layer at or near the perpendicular, the change in ionization must take place within a difference in height comparable to a wavelength; hence, ionospheric reflection is more apt to occur at longer wavelengths (lower frequencies).

Critical frequency — When the frequency is sufficiently low, a wave sent vertically upward to the ionosphere will be bent sharply enough to cause it to return to the transmitting point. The highest frequency at which such reflection can occur, for a given state of the ionosphere, is called the *critical frequency*. Although the critical frequency may serve as an index of transmission conditions, it is not the highest useful frequency, since other waves of the same frequency which enter the ionosphere at angles smaller than 90 degrees (less than vertical) will be bent sufficiently to return to earth. The *maximum usable frequency*, for waves leaving the earth at very small angles to the horizontal, is in the vicinity of three times the critical frequency.

Besides being directly observable, the critical frequency is of more practical interest than the ionization density because it includes the effects of absorption as well as refraction.

Virtual height — Although an ionospheric layer is a region of considerable depth it is convenient to assign to it a definite height, called the *virtual height*. This is the height from which a simple reflection would give the same effect as the gradual refraction which actually takes place, as illustrated in Fig. 903. The wave traveling upward is bent back over a path having an appreciable radius of turning, and a measurable interval of time is consumed in the turning process. The virtual height is the height of a triangle formed as shown, having equal sides of a total length proportional to the time taken for the wave to travel from *T* to *R*.

Normal structure of the ionosphere — The lowest normally useful layer is called the *E* layer. The average height of the region of maximum ionization is about 70 miles. The ionization density is greatest around local noon; the layer is only weakly ionized at night, when it is not exposed to the sun's radiation. The air at this height is sufficiently dense so that free ions and electrons very quickly meet and recombine.

The second principal layer is the *F* layer, which has a height of about 175 miles at night. At this altitude the air is so thin that recombination of ions and electrons takes place very slowly, inasmuch as particles can travel relatively great distances before meeting. The ionization decreases after sundown, reaching a minimum just before sunrise. In the daytime

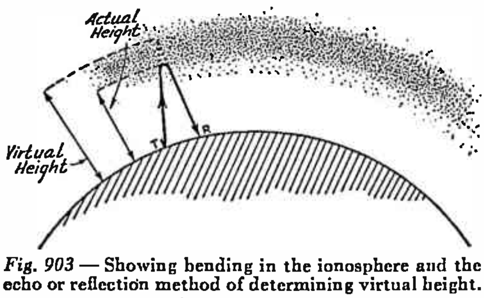


Fig. 903 — Showing bending in the ionosphere and the echo or reflection method of determining virtual height.

the F layer splits into two parts, the F_1 and F_2 layers, with average virtual heights of, respectively, 140 miles and 200 miles. These layers are most highly ionized at about local noon, and merge again at sunset into the F layer.

Cyclic variations in the ionosphere — Since ionization depends upon ultraviolet radiation, conditions in the ionosphere vary with changes in the sun's radiation. In addition to the daily variation depending upon the height of the sun in the observer's sky, seasonal changes result in higher critical frequencies in the E layer in summer (averaging about 4 Mc. above the winter levels), while the F_1 layer, which has a critical frequency near 5 Mc. in summer, usually disappears entirely in winter. The critical frequencies for the F_2 are highest in winter (11 to 12 Mc.) and lowest in summer (around 7 Mc.). The virtual height of the F_2 layer, which is about 185 miles in winter, averages 250 miles in summer.

There are at least three other regular cycles of variation in ionization. One such cyclic period covers 28 days, which corresponds with the period of the sun's rotation. For a short period in each 28-day cycle, favorable transmission conditions reach a peak. Usually this peak is followed by a fairly rapid drop to a lower level, and then a slow building up to the next peak. The 28-day cycle is particularly evident in the 14- and 28-Mc. amateur bands.

The longest cycle yet observed covers about 11 years, corrresponding to a similar cycle of sunspot activity. The effect of this cycle is to shift upward or downward the values of the critical frequencies for F- and F_2-layer transmission. The critical frequencies are highest during sunspot maxima and lowest during sunspot minima. It is during the period of minimum sunspot activity when long-distance transmissions occur on the lower frequencies. At such times the 28-Mc. band is seldom useful for DX work, while the 14-Mc. band performs well in the daytime but is not ordinarily useful at night. The most recent sunspot maximum is considered to have occurred in 1938.

Seasonal transition periods also occur in spring and fall, when ionospheric conditions are found highly variable.

Magnetic storms — During a sunspot maximum, and for some time following, occasonal severe disturbances may develop in the ionosphere. These are accompanied by corresponding abnormal conditions in the earth's magnetic field and by more frequent and spectacular displays of aurora, with the streamers extending further from the polar regions than is the normal case.

During unusual disturbances in the earth's magnetic field (magnetic storms) the daytime signal strength of low-frequency waves is increased, while at night the intensities are subnormal. On the higher frequencies long-distance communication frequently becomes impossible, signal strengths dropping sharply without warning. On the very-high frequencies voice transmissions may be badly distorted, while abnormal distances, of the order of 500 miles, can be covered with c.w. signals or tone-modulated keyed carriers at 56 Mc. At times a characteristic noise, similar to tone modulation, appears on c.w. carriers. Best results in v.h.f. transmission and reception with directive antennas is obtained during these "aurora skip" periods if the array is pointed toward the auroral curtain, without regard to the desired direction of communication. The effect is as though the auroral curtains were serving as global reflectors.

Sporadic E-layer ionization — Occasionally scattered patches or clouds of relatively dense ionization appear at heights approximately the same as that of the E layer. The effect is to raise the critical frequency to a value perhaps twice that which is returned from any of the regular layers by normal refraction. Distances of about 500 to 1400 miles may be covered at 56 Mc. if the ionized cloud is situated midway between transmitter and receiver, or is of any very considerable extent. This effect, while infrequently observed in winter, is prevalent during the late spring and early summer, with no apparent correlation of the condition with the time of day.

The presence of sporadic-E refraction on the 14- and 28-Mc. bands is indicated by an abnormally short distance between the transmitter and the point where the wave first is returned to earth as when, for example, 14-Mc. signals from a transmitter only 100 miles distant may arrive with an intensity usually associated with distances of this order on 7 and 3.5 Mc.

Wave angle — The smaller the angle at which a wave leaves the earth, the less will

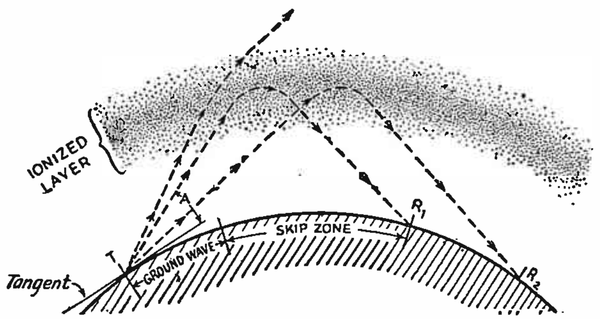


Fig. 904 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than A) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances. Below a certain minimum angle (less than A) the waves do not ever return to earth.

be the bending required in the ionosphere to bring it back and, in general, the greater the distance between the point where it leaves the earth and that at which it returns. This is shown in Fig. 904. The vertical angle which the wave makes with a tangent to the earth is called the *wave angle* or *angle of radiation.*

Skip distance — Since greater bending is required to return the wave to earth when the wave angle is high, at the higher frequencies the refraction frequently is not enough to give the required bending unless the wave angle is smaller than a certain angle called the *critical angle.* This also is illustrated in Fig. 904, where waves at angles of A or less give useful signals while waves sent at higher angles penetrate the layer and are not returned. The distance between T and R_1 is, therefore, the shortest possible distance over which communication by normal ionospheric refraction can be accomplished.

The area between the end of the useful ground wave and the beginning of ionospheric wave reception is called the *skip zone.* The extent of skip zone depends upon the frequency and the state of the ionosphere, and is greater the higher the transmitting frequency and the lower the critical frequency. Skip distance depends also upon the height of the layer in which the refraction takes place, the higher layers giving longer skip distances for the same wave angle. Wave angles at the transmitting and receiving points are usually, although not always, approximately the same for any given wave path.

It is readily possible for the ionospheric wave to pass through the E layer and be refracted back to earth from the F, F_1 or F_2 layers. This is because the critical frequencies are higher in the latter layers, so that a signal too high in frequency to be returned by the E layer can still come back from one of the others, depending upon the time of day and the existing conditions. Depending upon the wave angle and the frequency, it is sometimes possible to carry on communication via either the E or F_1 F_2 layers on the same frequency.

Multi-hop transmission — On returning to the earth the wave can be reflected upward and travel again to the ionosphere. There it may once more be refracted, and again bent back to earth. This process may be repeated several times. Multi-hop propagation of this nature is necessary for transmission over great distances because of the limited heights of the layers and the curvature of the earth, since at the lowest useful wave angles (of the order of a few degrees, waves at lower angles generally being absorbed rapidly at high frequencies by being in contact with the earth) the maximum one-hop distance is about 1250 miles for refraction from the E layer and around 2500 miles for the F_2 layer. However, ground losses absorb some of the energy from the wave on each reflection (the amount of the loss varying with the type of ground and being least for reflection from sea water). Thus, when the distance permits, it is better to have one hop rather than several, since the multiple reflections introduce losses which are higher than those caused by the ionosphere alone.

Fading — Two or more parts of the wave may follow slightly different paths in traveling to the receiving point, in which case the difference in path lengths will cause a phase difference to exist between the wave components at the receiving antenna. The field strength therefore may have any value between the numerical sum of the components (when they are all in phase) and zero (when there are only two components and they are exactly out of phase). Since the paths change from time to time, this causes a variation in signal strength called *fading.* Fading can also result from the combination of single-hop and multi-hop waves, or the combination of a ground wave with an ionospheric or tropospheric wave. Such a condition gives rise to an area of severe fading near the limiting distance of the ground wave, better reception being obtained at both shorter and longer distances where one component or the other is considerably stronger. Fading may be rapid or slow, the former type usually resulting from rapidly changing conditions in the ionosphere, the latter occurring when transmission conditions are relatively stable.

It frequently occurs that transmission conditions are different for waves of slightly different frequencies, so that in the case of voice-modulated transmission, involving side-bands differing slightly from the carrier in frequency, the carrier and various side-band components may not be propagated in the same relative amplitudes and phases they had at the transmitter. This effect, known as *selective fading,* causes severe distortion of the signal, especially in the case of frequency-modulated signals received over other than line-of-sight paths. The distortion results from the fact that the instantaneous frequency of the f.m. signal is subject to continual variation, so that when two waves reach a receiving antenna by different paths they differ in instantaneous frequency. The result is a combined wave in which components of both amplitude and frequency modulation make up a new modulation at a frequency which is not harmonically related to the modulation impressed at the transmitter, but which depends upon the differences in transit time for the different paths traveled by the waves. The resulting distortion is greatest at high modulation frequencies and with high depths of modulation.

9-3 Tropospheric Propagation

Air masses and fronts — In the lower atmosphere wave propagation is affected by the changes in refractive index between differing air masses. A mass of air hundreds of miles in area, miles in depth and millions of tons in weight may remain at rest over one region until it becomes affected by the surface temperature and humidity characteristic of that region. Eventually it is moved on by the forces of atmospheric circulation, often at a tremendous speed. The mass may travel over regions quite different from its origin and retain for some time its original characteristics as warm-dry or warm-moist, cold-dry or cold-moist. When it meets a dissimilar air mass, the lighter, warmer and drier air overruns the heavier cold, moist mass. A boundary is created between the dissimilar masses, called a *front.* This front represents a discontinuity in the dielectric constant of the troposphere, which serves to refract and reflect the higher-frequency radio waves in much the same manner as the ionospheric layers, but at lesser heights and more widely varied angles. As a result 56- and 112-Mc. signals (and possibly those of higher frequencies as well) are returned to earth at distances considerably beyond the range of ground-wave propagation, sometimes up to 400 miles. As the amount of bending is small, the wave angle must be quite low. To achieve this without excessive ground attenuation, it is advantageous to have both receiving and transmitting antennas as high as possible.

Temperature inversions — The condition prevailing when warm air overruns cooler air constitutes one of several types of *temperature inversion,* in this case known as a *dynamic* inversion.

The temperature of the lower atmosphere decreases at a constant rate with increasing height. When for any reason the normal variation or *lapse rate* of approximately 3° F. per 1000 feet of elevation is altered, a temperature inversion is said to take place. The resulting change in the dielectric constants of the air masses affected causes reflection and refraction similar to that in the ionosphere.

Types of inversion other than the dynamic inversion include the *subsidence* inversion, caused by the sinking of an air mass which has been heated by compression; the *nocturnal* inversion, brought about by the rapid cooling of surface air after sunset; and the *cloud-layer* inversion, caused by the heating of air

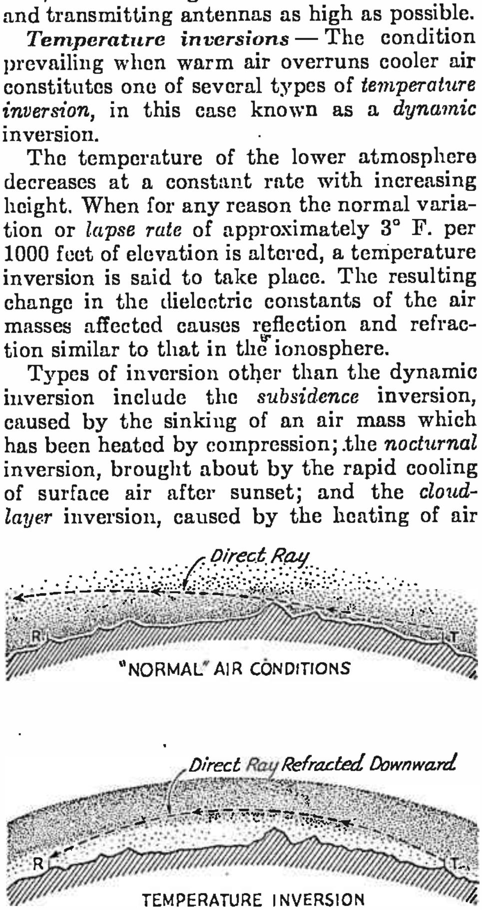


Fig. 905 — Illustrating the effect of a temperature inversion in extending the range of v.h.f. signals.

above a cloud layer by reflection of the sun's rays from the upper surface of the clouds. Refraction and reflection of v.h.f. waves are brought about also, although to a lesser degree, by the presence of sharp transitions in the water-vapor content of the atmosphere. Fig. 905 illustrates the conditions existing when the air is "normal" and when a temperature inversion is present.

Aeroanalysis, the science of the behavior of air masses and fronts, serves both the weather forecaster and the radio experimenter since tropospheric wave propagation is closely correlated with weather conditions.

◖ 9-4 Ground-Wave Propagation

Surface wave — The surface wave is continuously in contact with the surface of the earth and, in cases where the distance of transmission makes the curvature of the earth a factor, extends its range by diffraction. The surface wave is practically independent of seasonal and day and night effects at frequencies above 1500 kc.

The surface wave must be vertically polarized because the electrostatic field of a horizontally polarized wave would be short-circuited by the ground, which acts as a conductor at the frequencies for which the surface wave is of most interest.

The wave induces a current in the ground in traveling along its surface. If the ground were a perfect conductor there would be no loss of energy, but actual ground has appreciable resistance, so that the current flow causes some energy dissipation. This loss must be supplied by the wave which is correspondingly weakened. Hence, the transmitting range depends upon the ground characteristics. Because sea water is a good conductor, the range will be greater over the ocean than over land. The losses increase with frequency, so that the surface wave is rapidly attenuated at high frequencies and above about 2 Mc. is of little importance, except in purely local communication. The range at frequencies in the vicinity of 2 Mc. is of the order of 200 miles over average land and perhaps two or three times as far over sea water, for a medium-power transmitter (500 watts or so) using a good antenna. At higher frequencies the range drops off rapidly.

Space wave — In the v.h.f. portion of the spectrum (above 30 Mc.) the bending of the waves in the normal ionosphere is so slight that the ionospheric wave (§ 9-2) is not ordinarily useful for communication. The range of the surface wave also is extremely limited, as stated above. Hence, normal v.h.f. transmission is by means of the space wave in which the *direct-wave* component travels directly from the transmitter to the receiver through the atmosphere along a line-of-sight path.

Part of the space wave strikes the ground between the transmitter and receiver and is reflected upward at a slight angle, as was shown

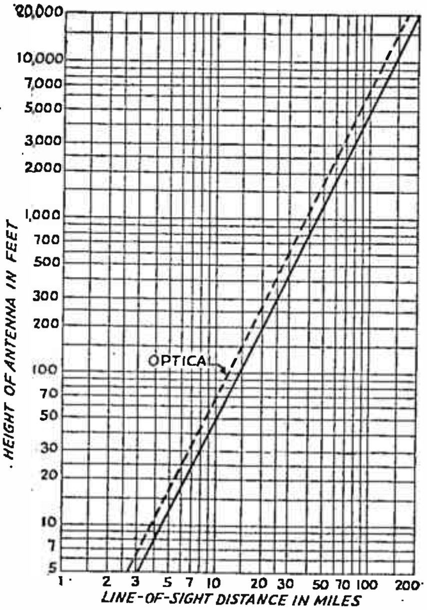


Fig. 906 — Chart for determining line-of-sight distance for v.h.f. transmission. The solid line includes effect of refraction, while the dotted line is the optical distance.

in Fig. 902. The effect of this *ground-reflected wave*, when it is out of phase with the direct wave, is to reduce the net field strength at the receiving point. The degree of cancellation depends upon the heights of the transmitting and receiving antennas above the point of reflection, the ground losses when reflection takes place, and the frequency — the cancellation decreasing with an increase in any of these.

The energy lost in ground absorption by a wave traveling close to the ground decreases very rapidly with its height in terms of wavelengths above the ground. A v.h.f. direct wave, therefore, can be relatively close (in physical height) to the ground without suffering the absorption effects which would occur at the same physical heights with longer wave-lengths.

Normal refraction — There is normally some change in the refractive index of the air with height above ground, its nature being such as to cause the wave to bend slightly towards the ground. Where curvature of the earth must be considered, this has the effect of lengthening the distance over which it is possible to transmit a direct wave. It is convenient to consider the effect of this "normal refraction" as equivalent to an increase in the earth's radius, in determining the antenna heights necessary to provide a clear path for the wave. The equivalent radius, taking refraction into account, is 4/3 the actual radius.

Range vs. height — Since the direct wave travels in practically a straight line, the maximum signal strength can be obtained only when there is an unobstructed atmospheric path between the transmitter and receiver. This means that antennas should be sufficiently elevated to provide such a path. On long paths the curvature of the earth, as well as the intervening terrain, must be taken into account.

The height required to provide a clear line-of-sight path over level terrain from an elevated transmitting point to a receiving point on the surface, not including the effect of refraction, is

$$h = \frac{d_2}{1.51}$$

where h is the height of the transmitting antenna in feet and d the distance in miles. Conversely, the line-of-sight distance in miles for a given height in feet is determined by

$$d = 1.23\sqrt{h}.$$

Taking refraction into account, this equation becomes

$$d = 1.41\sqrt{h}.$$

Fig. 906 gives the answer directly when either value is known.

When transmitter and receiver both are elevated, the maximum direct-wave distance to ground level can be determined separately for each. Adding the two distances thus obtained will give the maximum distance by which they can be separated for direct-wave communication. This is shown in Fig. 907.

Diffraction — At distances beyond the direct-wave path the wave is diffracted around the curvature of the earth. The diffracted wave is attenuated very rapidly, so that beyond the maximum direct-wave distance the signal strength decreases considerably faster with distance than it does within the direct-wave or line-of-sight path.

Antenna design determines useful radiation — It should be borne in mind that the discussion of useful radiation in this chapter under classifications depending upon varying angles of propagation, as "ground waves," "ionospheric waves," etc., is an analytical device. These modes of propagation are not separate and distinct entities. The useful wave for a given frequency and distance is often only a part of the total radiation. In the ideal non-directional antenna the radiation is generated in a hemispherical pattern made up of waves propagated at all angles. By proper antenna design it is possible to concentrate radiation along desired horizontal and vertical angles to suit conditions. Typical patterns of practical antennas will be described in Chapter 10.

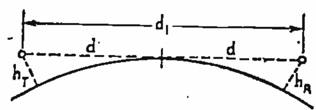


Fig. 907 — Method of determining total line-of-sight distance when both transmitter and receiver are elevated, based on Fig. 906. Since only earth curvature is taken into account in Fig. 906, irregularities in the ground between the transmitting and receiving points must be considered when computing each actual path.

Chapter Ten

Antenna Systems

10-1 Antenna Properties

Wave propagation and antenna design — For most effective transmission, the propagation characteristics of the frequency under consideration must be given due consideration in selecting the type of antenna to use. These have been discussed in Chapter Nine. On some frequencies the angle of radiation and polarization may be of relatively little importance; on others they may be all-important. On a given frequency, the particular type of antenna best suited for long-distance transmission may not be as good for shorter-range work as would a different type.

The important properties of an antenna or antenna system are its polarization, angle of radiation, impedance, and directivity.

Polarization — The polarization of a straight-wire antenna is its position with respect to the earth. That is, a vertical wire transmits vertically polarized waves and a horizontal antenna generates horizontally polarized waves (§ 9-1). The wave from an antenna in a slanting position contains both vertical and horizontal components.

Angle of radiation — The wave angle (§ 9-4) at which an antenna radiates best is determined by its polarization, height above ground, and the nature of the ground. Radiation is not all at one well-defined angle, but rather is dispersed over a more or less large angular region, depending upon the type of antenna. The angle is measured in a vertical plane with respect to a tangent to the earth at the transmitting point.

Impedance — The impedance (§ 2-8) of the antenna at any point is the ratio of voltage to current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load resistance represented by the antenna. At high frequencies the antenna impedance consists chiefly of radiation resistance (§ 2-12). It is understood to be measured at a current loop (§ 2-12), unless otherwise specified.

Directivity — All antennas radiate more power in certain directions than in others. This characteristic, called *directivity*, must be considered in three dimensions, since directivity exists in the vertical plane as well as in the horizontal plane. Thus, the directivity of the antenna will affect the wave angle as well as the actual compass directions in which maximum transmission takes place.

Current — The field strength produced by an antenna is proportional to the current flowing in it. Since standing waves generally are present on an antenna, the parts of the wire carrying the higher current therefore have the greatest radiating effect.

Power gain — The ratio of power required to produce a given field strength, with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the *power gain* of the latter antenna. The term is used in connection with antennas intentionally designed to have directivity, and the field is measured in the optimum direction of the antenna under test. The comparison antenna almost always is a half-wave antenna having the same polarization as the antenna under consideration. Power gain usually is expressed in decibels (§ 3-3).

10-2 The Half-Wave Antenna

Physical and electrical length — The fundamental form of antenna is a single wire whose length is approximately equal to half the transmitting wavelength. It is the unit from which many more complex forms of antennas are constructed. It is sometimes known as a *Hertz* or *doublet* antenna.

The length of a half wave in space is:

$$\textit{Length (feet)} = \frac{492}{\textit{Freq. (Mc.)}} \quad (1)$$

The actual length of a half-wave antenna will not be exactly equal to the half wave in space, but is usually about 5 per cent less because of capacitance at the ends of the wire (*end effect*). The reduction factor increases slightly as the frequency is increased. Under average conditions the following formula will give the length of a half-wave antenna to sufficient accuracy for frequencies up to 30 Mc.:

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times 0.95}{\textit{Freq. (Mc.)}} = \frac{468}{\textit{Freq. (Mc.)}} \quad (2)$$

At 56 Mc. and higher frequencies the somewhat larger end effects cause a slightly greater reduction in length, so that, for these higher frequencies,

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times 0.94}{\textit{Freq. (Mc.)}} = \frac{462}{\textit{Freq. (Mc.)}}, \quad (3)$$

$$\textit{or length (inches)} = \frac{5540}{\textit{Freq. (Mc.)}} \quad (4)$$

Current and voltage distribution — When power is fed to such an antenna the current and voltage vary along its length (§ 2-12). The

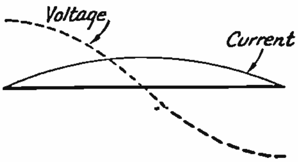


Fig. 1001 — Current and voltage distribution on a half-wave antenna. Current is maximum in center, nearly zero at ends. Voltage distribution is just the opposite.

distribution, which is practically a sine curve, is shown in Fig. 1001. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage. The current does not actually reach zero at the current *nodes* (§ 2-12), because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance of the wire (*ohmic resistance*) and the radiation resistance (§ 2-12). Usually the ohmic resistance of a half-wave antenna is small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

Impedance — The radiation resistance of a half-wave antenna in free space — that is, sufficiently removed from surrounding objects so that they do not affect the antenna's characteristics — is 73 ohms, approximately. The value under practical conditions will vary with the height of the antenna, but is commonly taken to be in the neighborhood of 70 ohms. It is pure resistance, and is measured at the center of the antenna. The impedance is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends (§ 10-1). The actual value at the ends will depend on a number of factors, such as the height, the physical construction, and the position with respect to ground.

Conductor size — The impedance of the antenna also depends upon the diameter of the conductor in relation to its length. The figures above are for wires of practicable sizes. If the diameter of the conductor is made large, of the order of 1 per cent or more of the length, the impedance at the center will be raised and the impedance at the ends decreased. This increase in center impedance (of the order of 50 per cent for a diameter/length ratio of 0.025) is accompanied by a decrease in the Q (§ 2-10, 2-12) of the antenna, so that the resonance curve is less sharp. Hence, the antenna is capable of working over a wider frequency range. This effect is greater as the diameter/length ratio is increased, and is a property of some importance at the very-high frequencies where the wavelength is small.

Radiation characteristics — The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions at right-angles to the wire and zero along the direction of the wire itself, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 1002, which represents the radiation pattern in free space. The relative intensity of radiation is proportional to the length of a line drawn from the center of the figure to the perimeter. If the antenna is vertical, as shown in the figure, then the field strength (§ 9-1) will be uniform in all horizontal directions; if the antenna is horizontal, the relative field strength will depend upon the direction of the receiving point with respect to the direction of the antenna wire.

ℭ 10-3 Ground Effects

Reflection — When the antenna is near the ground the free-space pattern of Fig. 1002 is modified by reflection of radiated waves from the ground, so that the actual pattern is the resultant of the free-space pattern and ground reflections. This resultant is dependent upon the height of the antenna, its position or orientation with respect to the surface of the ground, and the electrical characteristics of the ground. The reflected waves may be in such phase relationship to the directly radiated waves that the two completely reinforce each other, or the phase relationship may be such that complete cancellation takes place. All intermediate values also are possible. Thus, the effect of a perfectly reflecting ground is such that the original free-space field strength may be multiplied by a factor which has a maximum value of 2, for complete reinforcement, and having all intermediate values to zero, for complete cancellation. Since waves are always reflected upward from the ground (assuming that the surface is fairly level), these reflections only affect the radiation pattern in the vertical plane — that is, in directions upward from the earth's surface — and not in the horizontal plane, or the usual geographical directions.

Fig. 1003 shows how the multiplying factor varies with the vertical angle for several representative heights for horizontal antennas. As the height is increased the angle at which complete reinforcement takes place is lowered, until for a height equal to one wavelength it occurs at a vertical angle of 15 degrees. At still greater heights, not shown on the chart, the first maximum will occur at still smaller angles.

When the half-wave antenna is vertical the maximum and minimum points in the curves of Fig. 1003 exchange positions, so that the nulls become maxima, and vice versa. In this

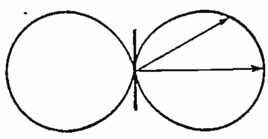


Fig. 1002 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining the drawing glued to a piece of cardboard, with a short length of wire fastened on it to represent the antenna. Twirling the wire will give a visual representation of the solid radiation pattern.

case, the height is taken as the distance from ground to the center of the antenna.

Radiation angle — The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies (§ 9-2, 9-4).

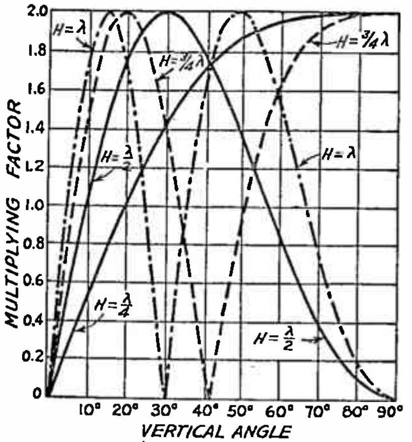


Fig. 1003 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly conducting ground.

It is advantageous, therefore, to erect the antenna at a height which will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low radiation angles usually are desirable, this generally means that the antenna should be high — at least ½ wavelength at 14 Mc., and preferably ¾ or 1 wavelength; at least 1 wavelength, and preferably higher, at 28 Mc. and the very-high frequencies. The physical height decreases as the frequency is increased, so that good heights are not impracticable; a half wavelength at 14 Mc. is only 35 feet, approximately, while the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower frequencies the higher radiation angles are effective, so that again a reasonable antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures generally being preferable where circumstances permit their use.

Imperfect ground — Fig. 1003 is based on ground having perfect conductivity, which is not met with in practice. The principal effect of actual ground is to make the curves inaccurate at the lowest angles; appreciable high-frequency radiation at angles smaller than a few degrees is practically impossible to obtain at heights of less than several wavelengths. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the sort of result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance — Waves which are reflected directly upward from the ground induce a current in the antenna in passing, and, depending on the antenna height, the phase relationship of this induced current to the original current may be such as either to increase or decrease the total current in the antenna. For the same power input to the antenna, an increase in current is equivalent to a decrease in impedance, and vice versa. Hence, the impedance of the antenna varies with height. The theoretical curve of variation of radiation resistance for an antenna above perfectly reflecting ground is shown in Fig. 1004. The impedance approaches the free-space value as the height becomes large, but at low heights may differ considerably from it.

Choice of polarization — Polarization of the transmitting antenna is generally unimportant on frequencies between 3.5 and 30 Mc. However, the question of whether the antenna should be installed in a horizontal or vertical position deserves consideration for other reasons. A vertical half-wave antenna will radiate equally well in all *horizontal* directions, so that it is substantially non-directional, in the usual sense of the word. If installed horizontally, however, the antenna will tend to show directional effects, and will radiate best in the direction at right angles, or broadside, to the wire. The radiation in such a case will be least in the direction toward which the wire points. This can be readily seen by imagining that Fig. 1002 is lying on the ground, and that the pattern is looked at from above.

The vertical angle of radiation also will be affected by the position of the antenna. If it were not for ground losses at high frequencies, the vertical half-wave antenna would be preferred because it would concentrate the radiation horizontally. In practice, however, this theoretical advantage over the horizontal antenna is of little or no consequence.

At 1.75 Mc., vertical polarization will give more low-angle radiation, and hence is better for long-distance transmission; at this fre-

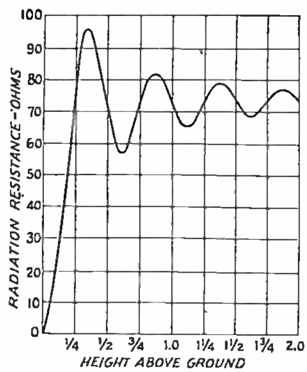


Fig. 1004 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height above perfectly reflecting ground.

quency the ground wave also is useful, and must be vertically polarized. On very-high frequencies, direct-ray and lower troposphere transmission require the same type of polarization at both receiver and transmitter, since the waves suffer no appreciable change in polarization in transmission (§ 9-1). Either vertical or horizontal polarization may be used, the latter being slightly better for longer distances.

Effective radiation patterns — In determining the radiation pattern it is necessary to consider radiation in both the horizontal and vertical planes. When the half-wave antenna is vertical, the vertical angle of radiation chosen does not affect the *shape* of the horizontal pattern, but only its relative amplitude. When the antenna is horizontal, however, both the shape and amplitude are dependent upon the angle of radiation chosen.

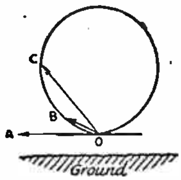


Fig. 1005 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing of the free-space field pattern of a horizontal antenna.

Fig. 1005 illustrates this point. The "free-space" pattern of the horizontal antenna shown is a section cut vertically through the solid pattern. In the direction *OA*, horizontally along the wire axis, the radiation is zero. At some vertical angle, however, represented by the line *OB*, the radiation is appreciable, despite the fact that this line runs *in the same geographical direction* as *OA*. At some higher angle, *OC*, the radiation, *still in the same geographical direction*, is still more intense. The effective radiation pattern therefore depends upon which angle of radiation is most useful, and for long-distance transmission is dependent upon the conditions existing in the ionosphere. These conditions may vary not only from day to day and hour to hour, but even from minute to minute. Obviously, then, the effective directivity of the antenna will change along with transmission conditions.

At very-high frequencies, where only extremely low angles are useful for any but sporadic-*E* transmission (§ 9-2), the effective radiation pattern of the antenna approaches the free-space pattern. A horizontal antenna therefore shows more marked directive effects than it does at lower frequencies, on which high radiation angles are effective.

Theoretical horizontal-directivity patterns for half-wave horizontal antennas at vertical angles of 9, 15, and 30 degrees (representing average useful angles at 28, 14 and 7 Mc. respectively) are given in Fig. 1006. At intermediate angles the values in the affected regions also will be intermediate. Relative field strengths are plotted on a decibel scale (§ 3-3), so that they represent as nearly as possible the actual aural effect at the receiving station.

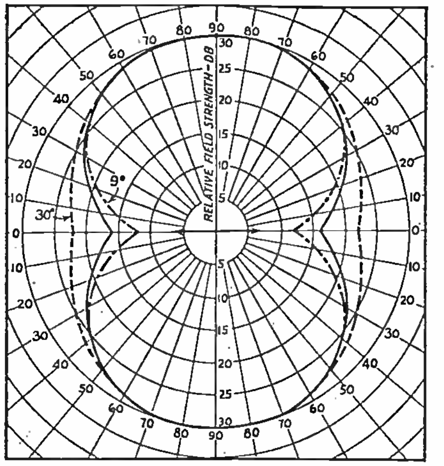


Fig. 1006 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will depend upon the height of the antenna above ground and the vertical angle considered. The patterns for all three angles have been proportioned to the same scale, but this does not mean that the maximum amplitudes necessarily will be the same. The arrow indicates the direction of the horizontal antenna wire.

¶ 10-4 Applying Power to the Antenna

Direct excitation — When power is transferred directly from the source to the radiating antenna, the antenna is said to be directly excited. While almost any coupling method (§ 2-11) may be used, those most commonly employed are shown in Fig. 1007. Power usually is fed to the antenna at either a current or voltage loop (§ 10-2). If power is fed at a current loop, the coupling method is called *current feed;* if at a voltage loop, the method is called *voltage feed.*

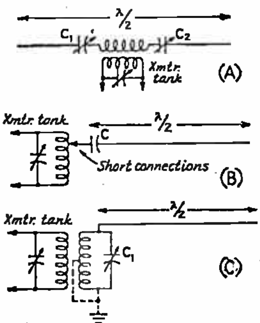


Fig. 1007 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacity coupling; C, voltage feed, with an inductively coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

Current feed — This method is shown in Fig. 1007-A. The antenna is cut at the center and a small coil coupled to the output tank circuit of the transmitter, with adjustable coupling so that the transmitter loading can be controlled. Since the addition of the coil "loads" the antenna, or increases its effective length because of the additional inductance, the series condensers, C_1 and C_2, are used to provide electrical means for reducing the length to its original unloaded value; in other words, their capacitive reactance serves to cancel the effect of the inductive reactance of the coil (§ 2-10).

Voltage feed — In Fig. 1007, at B and C the power is introduced into the antenna at a point of high voltage. In B, the end of the antenna is coupled to the output tank circuit through a small condenser, *C*; in C, a separate tank circuit, connected directly to the antenna, is used. This tank is tuned to the transmitter frequency, and should be grounded at one end or at the center of the coil, as shown.

Adjustment of coupling — Methods of tuning and adjustment of direct-feed systems correspond to those used with transmission lines, which are discussed in § 10-6.

Disadvantages of direct excitation — Direct excitation seldom is used except on the lowest amateur frequencies, because it involves bringing the antenna proper into the operating room and hence into close relationship with the house and electric wiring. This usually means that some of the power is wasted in heating poor conductors in the field of the antenna. Also, it often means that the shape of the antenna must be distorted, so that the expected directional effects are not realized, and likewise that the height will be limited. For these reasons, in high-frequency work practically all amateurs use transmission lines or feeder systems, which permit placing the antenna in a desirable location.

¶ 10-5 Transmission Lines

Requirements — A transmission line (§ 2-12) is used to transfer power, with a minimum of loss, from the transmitter to the antenna from which the power is to be radiated. At radio frequencies, where every wire carrying r.f. current tends to radiate energy in the form of electromagnetic waves, special design is necessary to minimize radiation and thus cause as much of the power as possible to be delivered to the receiving end of the line.

Radiation can be minimized by using a line in which the current is low, and by using two conductors carrying currents of equal magnitudes but opposite phase so that the fields about the conductors cancel each other. For good cancellation of radiation, the two conductors should be kept parallel and quite close to each other.

Types — The most common form of transmission line consists of two parallel wires, maintained at a fixed spacing of two to six inches by insulating spacers or spreaders placed at suitable intervals (*open-wire line*). A second type consists of rubber-insulated wires twisted together to form a flexible line, without spacers (*twisted-pair line*). A third uses a wire inside of and coaxial with a tubing outer conductor, separated from the outer conductor by insulating spacers or "beads" at regular intervals (*coaxial* or *concentric line*). A variation of this type uses solid rubber insulation between the inner and outer conductors, the latter usually being made of metal braid rather than of solid tubing, so that the line will be flexible. Still another type of line uses only a single wire, without a second conductor (*single-wire feeder*); in this type, radiation is minimized by keeping the line current low.

Spacing of two-wire lines — The spacing between the wires of an open-wire line should be small in comparison to the operating wavelength, to prevent appreciable radiation. It is impracticable to make the spacing too small, however, because when the wires swing with respect to each other in a wind the line constants (§ 2-12) will vary, and thus cause a variation in tuning or loading on the transmitter. It is also desirable to use as few insulating spacers as possible, to keep the weight of the line to a minimum. In practice, a spacing of about six inches is used for 14 Mc. and lower frequencies, with four- and two-inch spacings being common on the very-high frequencies.

Balance to ground — For maximum cancellation of the fields about the two wires, it is necessary that the currents be equal in amplitude and opposite in phase. Should the capacity or inductance per unit length in one wire differ from that in the other, this condition cannot be fulfilled. Insofar as the line itself is concerned, the two wires will have identical characteristics only when the two have exactly the same physical relationships to ground and to other objects in the vicinity. Thus, the line should be symmetrically constructed and the two wires should be at the same height. Line unbalance can be minimized by keeping the line as far above the ground and as far from other objects as possible.

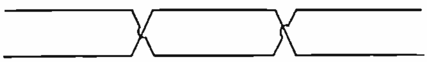


Fig. 1008 — Transposing a two-wire open transmission line preserves balance to ground and to near-by objects.

To overcome unbalance the line sometimes is transposed, which means that the positions of the wires are interchanged at regular intervals (Fig. 1008). This procedure is more helpful on long than on short lines, and usually need not be resorted to for lines less than a wavelength or so long.

Losses — Air-insulated lines operate at quite high efficiency. Parallel-conductor lines average 0.12 to 0.15 db. (§ 3-3) loss per wavelength of line. These figures hold only if the standing-wave ratio is 1. The losses increase with the standing-wave ratio, rather slowly up to a ratio of 15 to 1, but rapidly thereafter. For standing-wave ratios of 10 or 15 to 1, the increase is inconsequential provided the line is well balanced.

Concentric lines with air insulation are excellent when dry, but losses increase if there is moisture in the line. Provision should be therefore made for making such lines airtight, and they should be thoroughly dry when assembled. This type of line has the least radiation loss. The small lines (⅜-inch outer conductor) should not be used at high voltages; hence, it is desirable to keep the standing-wave ratio down.

Good quality rubber-insulated lines, both twisted pair and coaxial, average about 1 db. loss per wavelength of line. At the higher frequencies, therefore, such lines should be used only in short lengths if losses are important. These lines have the advantages of compactness, ease of installation, and flexibility. Ordinary lamp cord has a loss of approximately 1.4 db. per wavelength when it is dry, but its losses become excessive when wet. The parallel moulded-rubber type is best from the standpoint of withstanding wet weather. The characteristic impedance of lamp cord is between 120 and 140 ohms.

The loss in db. is directly proportional to the length of the line. Thus, a line which has a loss of 1 db. per wavelength will have an actual loss of 3 db. if the line is three wavelengths long. In the case of line losses, the length is not expressed in terms of electrical length but in physical length; that is, a wavelength of line, in feet, is equal to 984/frequency (Mc.) for computing loss. This permits a direct comparison of lines having the same physical length. The electrical lengths, of course, may differ considerably.

Resonant and nonresonant lines — Lines are classified as *resonant* or *nonresonant*, depending upon the standing-wave ratio. If the ratio is near 1, the line is said to be nonresonant. Reactive effects will be small, and consequently no special tuning provisions need ordinarily be made for canceling them even when the line length is not an exact multiple of a quarter wavelength. Such a line must be terminated in its characteristic impedance (§ 2-12).

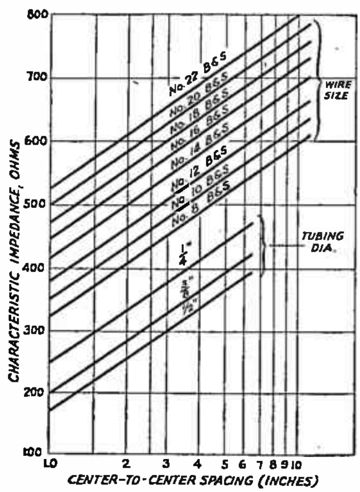


Fig. 1009 — Chart showing the characteristic impedances of typical spaced-conductor transmission lines.

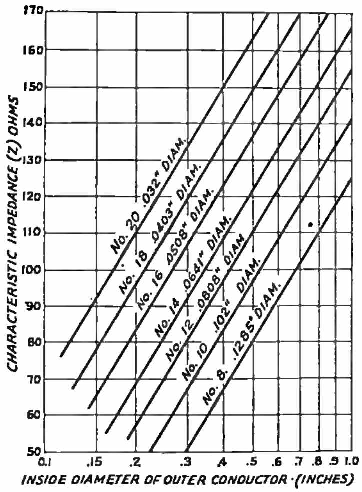


Fig. 1010 — The characteristic impedances of typical concentric lines. Tubing sizes are outside diameters.

If the standing-wave ratio is fairly large, the input reactance must be canceled or "tuned out" unless the line is a multiple of a quarter wavelength, and the line is said to be resonant.

10-6 Coupling to Transmission Lines

Requirements — The coupling system between a transmitter and the input end of a transmission line must provide means for adjusting the load on the transmitter to the proper value (impedance matching), and for tuning out any reactive component that may be present (§ 2-9, 2-10, 2-11). The resistance and reactance considered are those present *at the input end of the line*, and hence have nothing to do with the antenna itself except insofar as the antenna load may affect the operation of the line (§ 10-5).

Untuned coil — One of the simplest systems, shown in Fig. 1011-A, uses a coil of a few turns tightly coupled to the plate tank coil. Since no provision is made for tuning, this system is suitable only for non-resonant lines which show practically no reactance at the input end. Loading on the transmitter may be varied by varying the coupling between the tank inductance and the *pick-up coil*, as it is frequently called, or by changing the number of turns on the pick-up coil. A slight amount of reactance is coupled into the tank circuit by the pick-up coil, since the flux leakage (§ 2-11) is high, so that some slight retuning of the plate tank condenser may be necessary when the load is connected.

Taps on tank circuit — A method suitable for use with open-wire lines is shown in Fig. 1011-B, where the line is tapped on a balanced tank circuit with taps equidistant from the center or ground point. This symmetry is necessary to maintain line balance to ground (§ 10-5). Loading is increased by moving the taps outward from the center. Any reactance present may be tuned out by readjustment of the plate tank condenser, but this method is not suitable for large values of reactance and therefore direct tapping is best confined to use with non-resonant lines.

Adjustment of untuned systems — Adjustment of either of the above systems is quite simple. Starting with loose coupling, apply power to the transmitter, and adjust the plate tank condenser for minimum plate current. If the current is less than the desired load value, increase the coupling and again resonate the plate condenser. Continue until the desired plate current is obtained, always keeping the plate tank condenser at the setting which gives minimum current.

Pi-section coupling — A coupling system which is electrically equivalent to tapping on the tank circuit, but using a capacity voltage divider in the plate tank circuit for the purpose, is shown in Fig. 1011-C. Since one side of the condenser across which the line is connected is grounded, some unbalance will be introduced into the transmission line. This method is used chiefly with low-power portable sets, because it is readily adjustable to meet a fairly wide range of impedance values. A single-ended amplifier, using either a screen-grid tube or a grid-neutralized triode (§ 4-7), is required, since the plate tank circuit is not balanced. Coupling is adjusted by varying C_1, re-resonating the circuit each time by means of C_2 until the desired amplifier plate current is obtained. In general, the coupling will increase as C_1 is made smaller with respect to C_2. Relatively large-capacity condensers are required to give a suitable impedance-matching range while maintaining resonance.

Pi-section filter — The coupling circuit shown in Fig. 1011-D is a low-pass filter capable of coupling between a fairly wide range of impedances. The method of adjustment is as follows: First, with the filter disconnected from the transmitter tank, tune the transmitter tank to resonance, as evidenced by minimum plate current. Then, with trial settings of the clips on L_1 and L_2 (few turns for high frequencies, more for lower), tap the input clips on the final tank coil at points equidistant from the center, so that about half the coil is included between them. A balanced tank circuit must be used. Set C_2 at about half scale, apply power, and rapidly rotate C_1 until the plate current drops to minimum. If this minimum is not the desired full-load plate current, try a new setting of C_2 and repeat. If, for all settings of C_2, the plate current is too high or too low, try new settings of the taps on L_1 and L_2, and also of the taps on the transmitter tank. Do not touch the tank condenser during these adjustments. When, finally, the desired plate current is obtained, set C_1 carefully to the exact minimum plate-current point. *This adjustment is important in minimizing harmonic output.*

With some lengths of resonant lines, particularly those which are not exact multiples of a quarter wavelength, it may be difficult to get proper loading with the pi-section coupler. Usually antennas of these lengths also will be difficult to feed with other systems of coupling, as well. In such cases, the proper output loading often can be obtained by varying the L/C ratio of the filter over a considerably wider range than is necessary for normal loads as specified on page 204.

Series tuning — When the input impedance of the line is low, the coupling method shown in Fig. 1011-E may be used. This system, known as *series tuning*, places the coupling coil, tuning condensers and load all in series, and is particularly suitable for use with resonant lines when a current loop appears at the input end. As shown, two tuning condensers are used, to keep the line balanced to ground. However, one will suffice, the other end of the line being connected directly to the end of L_1.

The tuning procedure with series tuning is as follows: With C_1 and C_2 at minimum capacity, couple the antenna coil, L_1, loosely to the transmitter output tank coil, and observe the plate current. Then increase C_1 and C_2 simultaneously until a setting is reached which gives maximum plate current, indicating that the antenna system is in resonance with the transmitting frequency. Readjust the plate tank condenser to minimum plate current. This is necessary because tuning the antenna circuit will have some effect on the tuning of the plate tank. The new minimum plate current will be higher than with the antenna system detuned, but should still be well below the rated value for the tube or tubes. Increase the coupling between L_1 and L_2 by a small amount, readjust C_1 and C_2 for maximum plate current, and again set the plate tank condenser to minimum. Continue this process until the minimum plate current is equal to the rated plate current for the amplifier. Always use the degree of coupling between L_1 and L_2 which will just bring the amplifier plate current to rated value when C_1 and C_2 pass through resonance. The r.f. ammeters should indicate maximum feeder current at the resonance setting; these meters are not strictly necessary, but are useful in indicating the relative power output.

Parallel tuning — When the line has high input impedance, the use of parallel tuning, as shown in Fig. 1011-F, is required. Here the coupling coil, tuning condenser and line all are in parallel, the load represented by the line being directly across the tuned coupling circuit. If the line is non-reactive, the coupling circuit will be tuned independently to the transmitter frequency; line reactance can be compensated for by tuning of C_1 and, if necessary, adjustment of L_1 by means of taps. Parallel tuning is suited to resonant lines when a voltage loop appears at the input end.

The tuning procedure is quite similar to that with series tuning. Find the value of coupling between L_1 and L_2 which will bring the plate current to the desired value as C_1 is tuned through resonance. Again, a slight readjustment of the amplifier tank condenser may be necessary to compensate for the effect of coupled reactance.

Link coupling — Where tuning of the circuit connected to the line is necessary or desirable, it is possible to separate physically the line-tuning apparatus and the plate tank circuit by means of link coupling (§ 2-11). This is often convenient from a constructional standpoint, and has the advantage that, with proper construction, there will be somewhat less harmonic transfer to the antenna, since stray capacity coupling is lessened with the smaller link coils.

Figs. 1011-G and H show a method which can be considered to be a variation of Fig. 1011-B. The first (G) is suitable for use with a single-ended plate tank, the second (H) for a balanced tank. The auxiliary tank on which the transmission line is tapped may have adjustable inductance as well as capacity, to provide a wide range of reactance variation for compensating for line reactance. The center of the auxiliary tank inductance may be grounded, if desired. The link windings should be placed at the grounded parts of the coils, to reduce capacity coupling and consequent harmonic transfer. With this inductively coupled system, the loading on the auxiliary tank circuit increases as the taps are moved outward from the center, but, since this decreases the Q of the circuit, the coupling to the plate tank simultaneously decreases (§ 2-11). Hence, a compromise adjustment giving proper loading must be found in practice. Loading also may be varied by changing the coupling between one link winding and its associated tank coil; either tank may be used for this purpose. When the auxiliary tank is properly tuned to compensate for line reactance, the plate tank tuning will be practically the same as with no load; hence, the plate tank condenser need be readjusted only slightly to compensate for the small reactance introduced by the link.

Link coupling also may be used with series and parallel tuning, as shown in Figs. 1011-I and J. The coupling between one link and its associated coil may be made variable, to give the same effect as changing the coupling between the plate tank and antenna coils in the ordinary system. The tuning procedure is the same as described above for series and parallel tuning. In the case of single-ended tank circuits the input link is coupled to the grounded end of the tank coil, as in Fig. 1011-G.

Circuit values — The values of inductance and capacity to use in the antenna coupling system will depend upon the transmitting frequency, but are not particularly critical. With series tuning (Fig. 1011-E, I), the coil may consist of a few turns of the same construction as is used in the final tank; average values will run from one or two turns at very-high frequencies to perhaps 10 or 12 at 1.75 Mc. The number of turns preferably should be adjustable so that the inductance can be changed should it not be possible to reach resonance with the condensers used. The series condensers should have a maximum capacity of 250 or 350 μμfd. at the lower frequencies; the same values will serve even at 28 Mc., although 100 μμfd. will be ample for this and the 14-Mc. band. Still smaller condensers can be used at very-high frequencies. Since series tuning is used at a low-voltage point in the feeder system, the plate

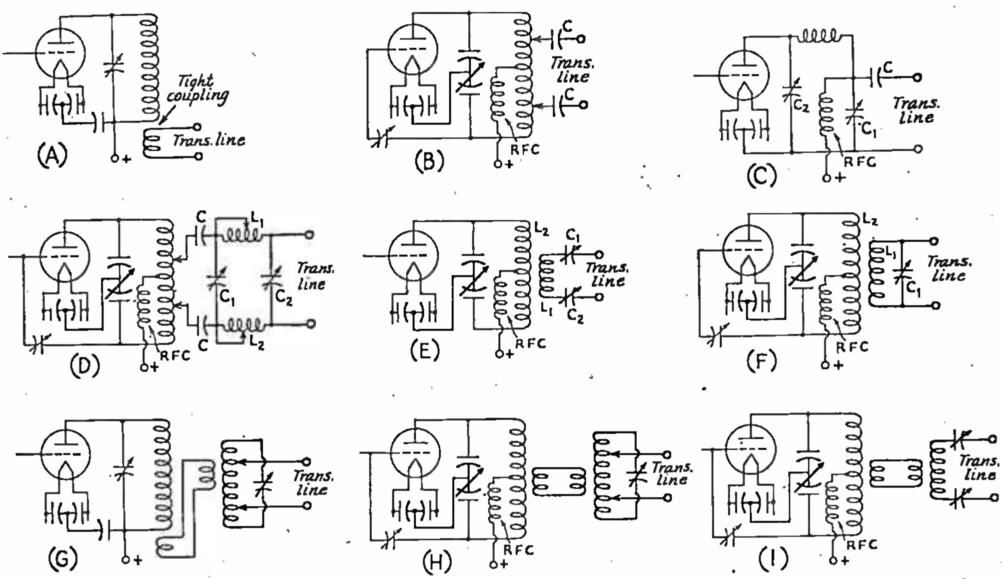


Fig. 1011— Methods of coupling the transmitter output to the transmission line. Application, circuit values and adjustment are discussed in the text. The coupling condensers, *C*, are fixed blocking condensers used to isolate the transmitter plate voltage from the antenna. Their capacity is not critical, 500 μμfd. to 0.002 μfd. being satisfactory values, but their voltage rating should at least equal the plate voltage on the final stage.

spacing of the condensers does not have to be large. Ordinary receiving-type condensers are large enough for plate voltages up to 1000, and the smaller transmitting condensers have high-enough voltage ratings for higher-power applications. In high-power radiotelephone transmitters it may be necessary to use condensers having a plate spacing of approximately 0.15 to 0.2 inch.

In parallel-tuned circuits (F, G, H, J) the antenna coil and condenser should be approximately the same as those used in the final tank circuit. The antenna tank circuit must be capable of being tuned independently to the transmitting frequency, and, if possible, provision should be made for tapping the coil, so that the L/C ratio can be varied to the optimum value (§ 2-11) as determined experimentally.

In Fig. 1011-D, C_1 and C_2 may be 100 to 250 μμfd. each, the higher-capacity values being used for lower-frequency operation (3.5 and 1.75 Mc.). Plate spacing should be, in general, at least half that of the final-amplifier tank condenser. For operation from 1.75 to 14 Mc., L_1 and L_2 each may consist of 15 turns, 2½ inches in diameter, spaced to occupy 3 inches length, and tapped every three turns. Approximate settings are 15 turns for 1.75 Mc., 9 turns for 3.5 Mc., 6 turns for 7 Mc., and 3 turns for 14 Mc. The coils may be wound with No. 14 or No. 12 wire. This method of coupling is very seldom used at very-high frequencies.

Harmonic reduction — It is important to prevent, insofar as possible, harmonics in the output of the transmitter from being transferred to the antenna system. Untuned (Fig. 1011-A) and directly coupled (Figs. 1011-B) systems do not discriminate against harmonics, and hence are more likely to cause harmonic radiation than the inductively coupled tuned systems. Low-pass filter arrangements, such as those at C and D, Fig. 1011, do discriminate against harmonics, but the direct coupling frequently is a source of trouble in this respect.

In inductively coupled systems, care must be taken to prevent capacity coupling between coils. Link coils always should be coupled at a point of ground potential (§ 2-13) on the plate tank coil, as also should series- and parallel-tuned coils (E and F), when possible. Capacity coupling can be practically eliminated by the use of a Faraday shield (§ 4-9) between the plate tank and antenna coils.

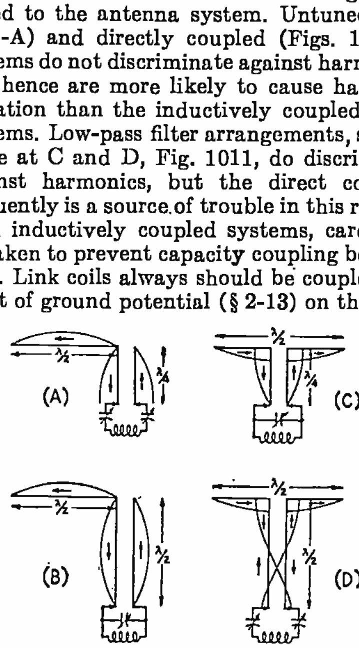


Fig. 1012 — Half-wave antennas fed from resonant lines. A and B are end-feed systems for use with quarter- and half-wave lines; C and D are center-feed systems. The current distribution is shown for all four cases, arrows indicating the instantaneous direction of current flow.

ℂ 10-7 Resonant Lines

Two-wire lines — Because of its simplicity of adjustment and flexibility with respect to the frequency range over which an antenna system will operate, the resonant line is widely used with simple antenna systems. Constructionally, the spaced or "open" two-wire line is best suited to resonant operation; rubber-insulated lines, such as twisted pair, have excessive losses when operated with standing waves.

Connection to antenna — A resonant line is usually — in fact, practically always — connected to the antenna at either a current or voltage loop. This is advantageous, especially when the antenna is to be operated at harmonic frequencies, since it simplifies the problem of determining the coupling system to be used at the input end of the line.

Half-wave antenna with resonant line — It is often helpful to look upon the resonant line simply as an antenna folded back on itself. Such a line may be any whole-number multiple of a quarter wave in length; in other words, any total wire length which will accommodate a whole number of standing waves. (The "length" of a two-wire line is, however, always taken as the length of one of the wires.)

Quarter- and half-wave resonant lines feeding half-wave antennas are shown in Fig. 1012. The current distribution on both antenna and line is indicated. It will be noted that the quarter-wave line has maximum current at one end and minimum current at the other, determined by the point of connection to the antenna. The half-wave line, however, has the same current (and voltage) values at both ends.

If a quarter-wave line is connected to the end of an antenna, as shown in Fig. 1012-A, then at the transmitter end of the line the current is high and the voltage low (low impedance), so that series tuning (§ 10–6) can be used. Should the line be a half-wave long, as at 1012-B, current will be minimum and voltage maximum (high impedance) at the transmitter end of the line, just as it is at the end of the antenna. Parallel tuning therefore is required (§ 10-6). The line could be coupled to a balanced final tank through small condensers, as in Fig. 1011-B, but the inductively coupled circuit is preferable. An end-fed antenna with resonant feeders, as in 1011-A and B, is known as the "Zeppelin" or "Zepp" antenna.

The line also may be inserted at the center of the antenna at the maximum-current point. Quarter- and half-wave lines used in this way are shown at Fig. 1012-C and D. In C, the antenna end of the line is at a high-current low-voltage point (§ 10-2); hence, at the transmitter end the current is low and the voltage high.

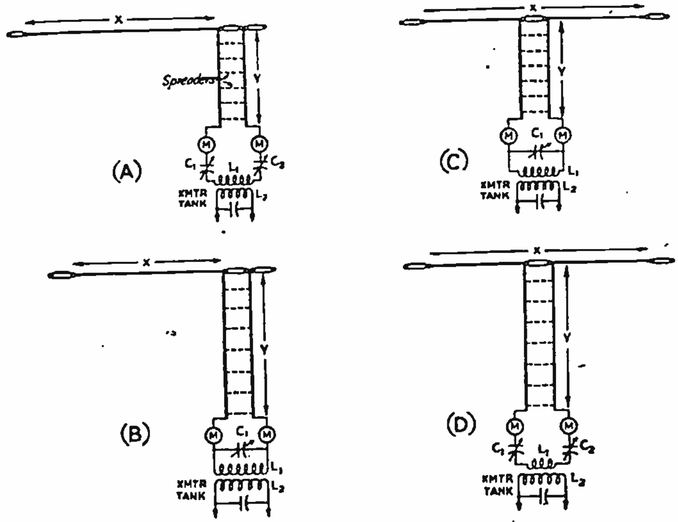


Fig. 1013 — Practical half-wave antenna systems using resonant-line feed. In the center-feed systems, the antenna length, *X*, does not include the length of the insulator at the center. Line length is measured from the antenna to the tuning apparatus; leads in the latter should be kept short enough so their effect can be neglected. The use of two r.f. ammeters, *M*, as shown is helpful for balancing feeder currents; however, one meter is sufficient to enable tuning for maximum output, and may be transferred from one feeder to the other, if desired. The systems at (A) and (C) are for feeders an odd number of quarter waves in length; (B) and (D) are for feeders a multiple of a half wavelength. The detailed drawings shown here correspond electrically to the elementary schematic half-wave antenna systems shown in Fig. 1012.

Parallel tuning therefore is used. The half-wave line at D has high current and low voltage at both ends, so that series tuning is used at the transmitter end.

The four arrangements shown in Fig. 1012 are thoroughly useful antenna systems, and are shown in more practical form in Fig. 1013. In each case the antenna is a half wavelength long, the exact length being calculated from Equations 2, 3 or 4 (§ 10-2) or taken from the charts of Fig. 1016. The line length should be an integral multiple of a quarter wavelength and may be calculated from equation 5 (§ 10-5), the result being multiplied by any whole number which gives a total length convenient for reaching from the antenna to the transmitter. If there is an *odd* number of quarter waves on the line in the case of the end-fed antenna, series tuning should be used at the transmitter end; if an *even* number of quarter waves, then parallel tuning should be used. With the center-fed antenna the reverse is true.

Practical line lengths — In general, it is best to use line lengths that are integral multiples of a quarter wavelength. Intermediate lengths will give intermediate impedance values and will show reactance (§ 10-5) as well. The tuning apparatus is capable of compensating for reactance, but it may be difficult to get suitable transmitter loading because simple series and parallel tuning are suitable for only low and high impedances, respectively, and neither will perform well with impedances of the order of a few hundred ohms. Such values of impedance may reduce the *Q* of the coupling circuit to a point where adequate coupling cannot be obtained (§ 2-11). However, some departure from the ideal length is possible — even as much as 25 per cent of a quarter wave in many cases — without undue difficulty in tuning and coupling. In such cases the type of tuning to use, whether series or parallel, will depend on whether the feeder length is nearer an odd number of quarter waves or nearer an even number, as well as on the point at which the feeder is connected to the antenna — at the end or in the center.

Line current — The feeder current as read by the r.f. ammeters is useful for tuning purposes only; the absolute value is of little importance. When series tuning is used the current will be high, but very little current will be indicated in a parallel-tuned system. This is because of the current distribution on the feeders, as shown by Fig. 1012. With a given antenna and tuning system, of course, the greatest power will be delivered to the antenna when the readings are highest. However, should the feeder length be changed no useful conclusions can be drawn from comparison between the new and old readings. For this reason, any indicator which registers the relative intensity of r.f. current can be used for tuning purposes. Many amateurs, in fact, use flashlight or dial lamps for this purpose instead of meters. Such lamps are inexpensive indicators, and, when shunted by short lengths of wire so that considerable current can be passed without danger of burn-out, will serve very well even with high-power transmitters.

Antenna length and line operation — Insofar as the operation of the antenna itself is concerned, departures of a few per cent from the exact length for resonance are of negligible consequence. Such inaccuracies may influence the behavior of the feeder system, however, and as a result may have an adverse effect on the operation of the system as a whole. This is true particularly of end-fed antennas, such as are shown in Fig. 1013-A and -B.

For example, Fig. 1014-A shows the current distribution on the half-wave antenna and quarter-wave feeder when the antenna length is correct. At the junction of the "live" feeder and the antenna the current is minimum, so that the currents in the two feeder wires are equal at all corresponding points along their

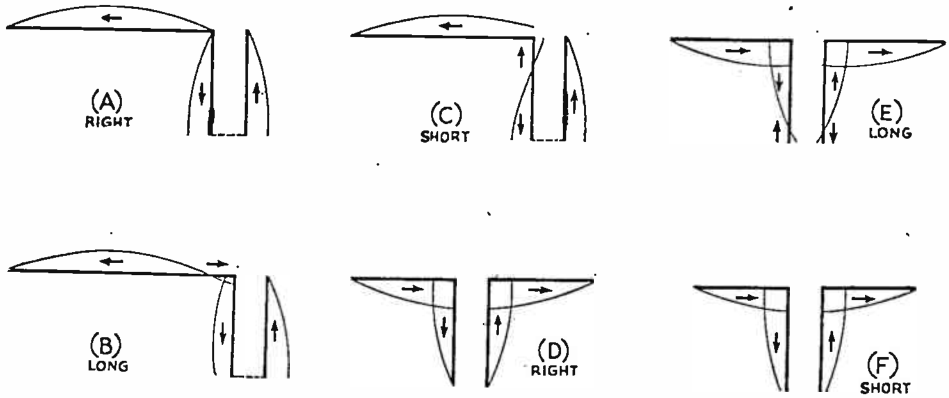


Fig. 1014 — Illustrating the effect on feeder balance of incorrect antenna length for various types of antenna systems. In end-feed systems, the current minimum shifts above or below the feeder junction, unbalancing the line. With center feed, incorrect antenna length does not unbalance the transmission line as it does with end feed.

length. When the antenna is too long, as in B, the current minimum occurs at a point on the antenna proper, so that at the top of the live feeder there is already appreciable current flowing, whereas at the top of the "dead" feeder the current must be zero. As a result the feeder currents are not balanced, and some power will be radiated from the line. In C, the antenna is too short, bringing the current minimum to a point on the live feeder, so that again the currents are unbalanced. The more serious the unbalance, the greater the radiation from the line.

Strictly speaking, a line having an unbalanced connection, such as the one-way termination at the end of an antenna, cannot be truly balanced even though the antenna length is correct. This is because of the difference in loading on the two sides. The effect of this difference is fairly small when the currents are balanced, however.

If the antenna is fed at the center the undesirable effects of incorrect antenna length balance out, so that the line operates properly under all conditions. This is shown in Fig. 1014 at D, E and F. So long as the two halves of the antenna are of equal length the distribution of current on the feeders will be symmetrical, so that no unbalance exists even for antenna lengths considerably removed from the correct value.

ℂ 10-8 Nonresonant Lines

Requirements — The advantages of nonresonant transmission lines — minimum losses, and elimination of the necessity for tuning — make the use of this type of line attractive. The chief disadvantage of the nonresonant line, aside from the necessity for more care in initial adjustment, is that when "matched" to the ordinary antenna the match is perfect only for one frequency, or at most for a small band of frequencies on either side of the frequency for which the matching is done. Except for a few special systems, such an antenna is unsuitable for work on more than one amateur band.

Adjustment of a nonresonant line is simply a process of adjusting the terminating resistance to match the characteristic impedance of the line. To accomplish this the antenna itself must be resonant at the selected frequency, and the line must then be connected to it in such a way that the antenna impedance as looked at by the line is the right value. The matching may be done by connecting the line at the proper spot along the antenna, by inserting an impedance-transforming device between the antenna and line, or by using a line having an impedance equal to the center impedance of the antenna.

In the following discussion of ways in which different types of lines may be matched to the antenna, a half-wave antenna is used as an example. Other types of antennas may be treated by the same methods, making due allowance for the order of impedance that appears at the end of the line when more elaborate systems are used.

Single-wire feed — In the single-wire-feed system, the return circuit is through the ground. There will be no standing waves on the feeder when its characteristic impedance is matched by the impedance of the antenna at the connection point. The principal dimensions (Fig. 1015) are the length of the antenna, *L*, and the distance, *D*, from the exact center of the antenna to the point at which the feeder is attached. Approximate dimensions

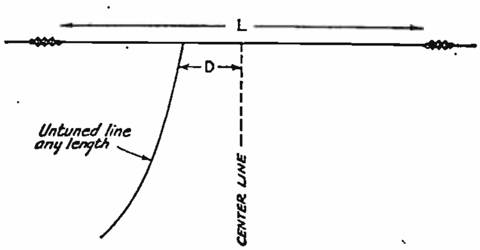


Fig. 1015 — Single-wire-feed system. The length, *L*, (one-half wavelength) and the feeder location, *D*, for various bands are determined from the charts of Fig. 1016.

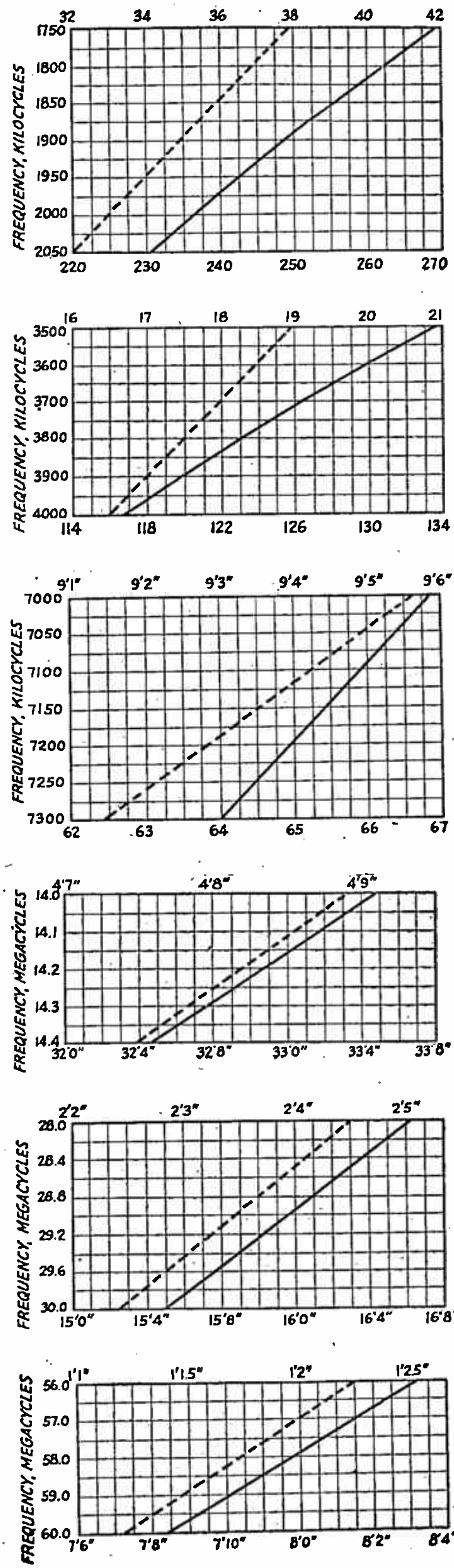


Fig. 1016 — Charts for determining the length of half-wave antennas for use on various amateur frequencies. Solid lines indicate antenna length in feet (lower scale); dotted lines indicate the point of connection for a single-wire feeder (upper scale) measured from center of antenna.

for both antenna length and the feeder connection point can be obtained from Fig. 1016 for an antenna system having a fundamental frequency in any of the most-used amateur bands.

In constructing an antenna system of this type, the feeder must run straight away from the antenna (at a right angle) for a distance of at least one-third the length of the antenna. Otherwise the field of the antenna will affect the feeder and cause faulty operation. There should be no sharp bends in the feeder wire at any point.

With the coupling system shown in Fig. 1017-A, the process of adjustment is as follows: Starting at the ground end of the tank coil, the tap is moved towards the plate end until the amplifier draws the rated plate current. The plate tank condenser should be readjusted each time the tap is changed, to bring the plate current back to minimum. The amplifier is loaded properly when this "minimum" value is the

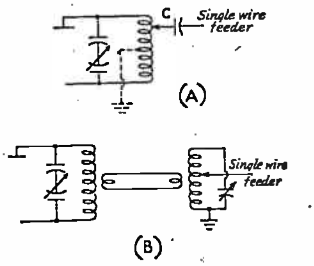


Fig. 1017 — Methods of coupling the feeder to the transmitter in a single-wire-feed system. Circuits are shown for both single-ended and balanced tank circuits.

rated current. The condenser, *C*, in the feeder is for the purpose of insulating the antenna system from the high-voltage plate supply when series plate feed is used. It should have a voltage rating somewhat higher than that of the plate supply. Almost any capacity greater than 500 $\mu\mu$fd. will be satisfactory. The condenser is unnecessary, of course, if parallel plate feed is used.

Inductive coupling to the output circuit is shown in Fig. 1017-B. The antenna tank circuit should tune to resonance at the operating frequency, and the loading is adjusted by varying the coupling between the two tanks, both being kept tuned to resonance.

Regardless of the type of coupling employed, a good ground connection is essential with this system. Single-wire feed works best over moist ground, and comparatively poorly over rock and sand.

Twisted-pair feed — A two-wire line composed of twisted rubber-covered wires can be constructed to have a surge impedance approximately equal to the 70-ohm impedance at the center of the antenna itself, thus permitting connecting the line to the antenna as shown in Fig. 1018. Any discrepancy which may exist between line and antenna impedance can be compensated for by a slight fanning of the line where it connects to the two halves of the antenna, as indicated at B in Fig. 1018.

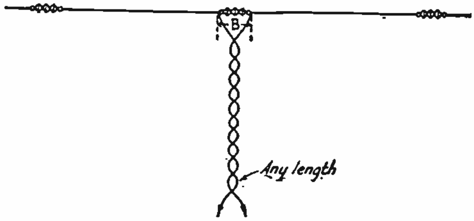


Fig. 1018 — Half-wave antenna center-fed by a twisted-pair line. Fanning (*B*) compensates for line impedance.

The twisted-pair line is a convenient type to use, since it is easy to install and the r.f. voltage on it is low because of the low impedance. This makes the quality of insulation a matter of less importance. Special twisted line for transmitting purposes, having lower losses than ordinary rubber-covered wire, is available.

The antenna should be one-half wavelength long for the frequency of operation, as determined by charts of Fig. 1016 or the formulas (§ 10-2). The amount of "fanning" (dimension *B*) will depend upon the kind of cable used; the required spacing usually will be between 6 and 18 inches. It may be checked by inserting ammeters in each antenna leg at the junction of the feeder and antenna; the value of *B* which gives the largest current is correct. Alternatively, the system may be operated continuously for a time with fairly high r.f. power input, after which the feeder may be inspected (by touch) for hot spots. These indicate the presence of standing waves, and the fanning should be adjusted until they are eliminated or minimized. Each leg of the feeder forming the triangle at the antenna should be equal in length to dimension *B*.

Coupling between the transmitter and the transmission line is ordinarily accomplished by the untuned coil method shown in Fig. 1011-A (§ 10-6).

Concentric-line feed — A concentric transmission line can be constructed to have a surge impedance exactly equal to the 70-ohm impedance at the center of a half-wave antenna.

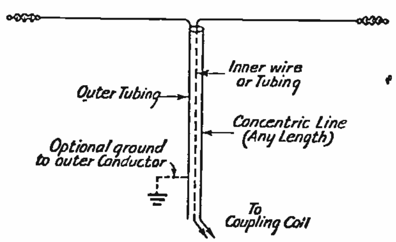


Fig. 1019 — Half-wave antenna center-fed by a concentric transmission line of 70 ohms surge impedance.

Such a line can be connected directly to the center of the antenna, therefore, forming the system shown in Fig. 1019.

Solving Equation 6 (§ 10-5) for an air-insulated concentric line shows that, to obtain a 70-ohm surge impedance, the inside diameter of the outer conductor should be approximately 3.2 times the outside diameter of the inner conductor. This condition can be fulfilled by using standard 5/16-inch (outside diameter) copper tubing for the outer conductor and No. 14 wire for the inner. Ceramic insulating spacers are available commercially for this combination. Rubber-insulated concentric line having the requisite impedance for connection to the center of the antenna also is available.

The operation of such an antenna system is similar to that of the twisted-pair system just described, and the same transmitter coupling arrangements may be used (§ 10-6).

The outer conductor of the line may be grounded, if desired. The feeder system is slightly unbalanced, because the inner and outer conductors do not have the same capacity to ground. There should be no radiation from a line having a correct surge impedance, however.

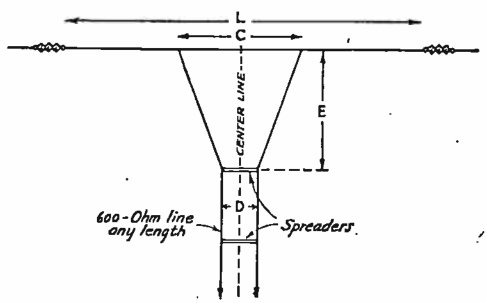


Fig. 1020 — Delta-matched antenna system. The dimensions *C*, *D*, and *E* are found by formulas given in the text. It is important that the matching section, *E*, come straight away from the antenna without any bends.

Delta matching transformer — Because of the extremely close spacing required, it is impracticable to construct an open-wire transmission line which will have a surge impedance low enough to work directly into the center of a half-wave antenna. Such wire lines usually have impedances between 400 and 700 ohms, 600 ohms being a widely used value. It is necessary, therefore, to use other means for matching the line to the antenna.

One method of matching is illustrated by the system shown in Fig. 1020. The matching section, *E*, is "fanned" to have a gradually increasing impedance so that its impedance at the antenna end will be equal to the impedance of the antenna section, *C*, while the impedance at the lower end matches that of a practicable transmission line.

The antenna length, *L*, the feeder clearance, *E*, the spacing between centers of the feeder wires, *D*, and the coupling length, *C*, are the important dimensions of this system. The system must be designed for exact impedance values as well as frequency values, and the dimensions therefore are fairly critical.

The length of the antenna is figured from the formula (§ 10-2) or taken from Fig. 1016.

The length of section C is computed by the formula:

$$C\ (feet) = \frac{148}{Freq.\ (Mc.)}$$

The feeder clearance, E, is found from the equation:

$$E\ (feet) = \frac{123}{Freq.\ (Mc.)}$$

The above equations are for feeders having a characteristic impedance of 600 ohms and will not apply to feeders of any other impedance. The proper feeder spacing for a 600-ohm transmission line is computed to a sufficiently close approximation by the following formula:

$$D = 75 \times d$$

where D is the distance between the centers of the feeder wires and d is the diameter of the wire. If the wire diameter is in inches the spacing also will be in inches, and if the wire diameter is in millimeters the spacing also will be in millimeters.

Methods of coupling to the transmitter are discussed in § 10-6, those shown in Figs. 1011-C, D, G and H being suitable.

"Q"-section transformer — The impedance of a two-wire line of ordinary construction (400 to 600 ohms) can be matched to the impedance of the center of a half-wave antenna by utilizing the impedance-transforming properties of a quarter-wave line (§ 10-5). The matching section must have low surge impedance and therefore is commonly constructed of large-diameter conductors such as aluminum or copper tubing, with fairly close spacing. This system is known as the "Q" antenna. It is shown in Fig. 1021. The important dimensions are the length of the antenna, the length of the matching section, B, the spacing between the two conductors of the matching section, C, and the impedance of the untuned transmission line connected to the lower end of the matching section.

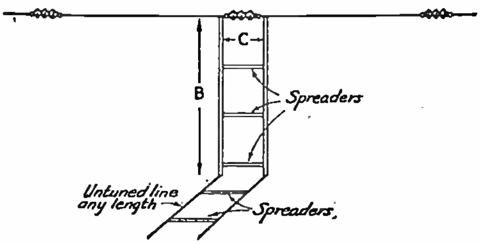


Fig. 1021 — The "Q" antenna, using a quarter-wave impedance-matching section with close-spaced conductors.

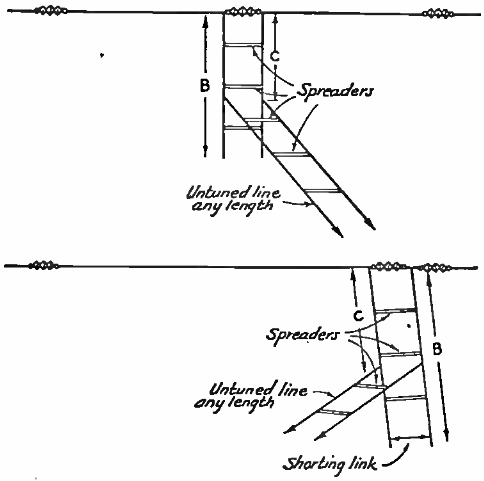


Fig. 1022 — Half-wave antenna systems with quarter-wave open-wire linear impedance-matching transformers.

The required surge impedance for the matching section is

$$Z_s = \sqrt{Z_1 Z_2} \qquad (9)$$

where Z_1 is the input impedance and Z_2 the output impedance. Thus a quarter-wave section matching a 600-ohm line to the center of a half-wave antenna (72 ohms) should have a surge impedance of 208 ohms. The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 1009. With ½-inch tubing, the spacing should be 1.5 inches for an impedance of 208 ohms.

The length of the matching section, B, should be equal to a quarter wavelength, and is given by

$$\text{Length of quarter-wave line (feet)} = \frac{234}{Freq.\ (Mc.)}$$

The length of the antenna can be calculated from the formula (§ 10-2), or taken from the charts of Fig. 1015.

This system has the advantage of the simplicity of adjustment of the twisted-pair feeder system and at the same time the superior insulation of an open-wire system. Figs. 1011-B, D, G and H (§ 10-6) represent suitable methods of coupling to the transmitter.

Linear transformers — Fig. 1022 shows two methods of coupling a non-resonant line to a half-wave antenna through a quarter-wave linear transformer (§ 10-5) or matching section. In the case of the center-fed antenna, the free end of the matching section, B, is open (high impedance), since the other end is connected to a low-impedance point on the antenna. With the end-fed antenna, the free end of the matching section is closed through a shorting bar or link; this end of the section has low impedance, since the other end is connected to a high-impedance point on the antenna (§ 10-7).

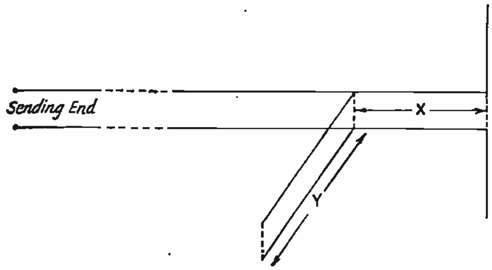


Fig. 1023 — When antenna and transmission line differ in impedance, they may be matched by a short length of transmission line, *Y*, called a stub. Determination of the critical dimensions, *X* and *Y*, for proper matching depends on whether the stub is open or closed at the end.

In contrast to the quarter-wave impedance matching transformer of Fig. 1021, the functioning of linear matching transformer sections or "stubs" is based on the fact that, as is shown in Fig. 1024, the impedance of a short resonant line varies along its length, the distribution depending upon whether the section is shorted or open. To obtain any desired impedance requires only that the section be cut off at the proper point. The design of a matching stub to connect a transmission line and an antenna of differing impedances, may be described by referring to Fig. 1023, which shows a two-wire transmission line feeding a half-wave antenna. At any distance (*X*) from the antenna, the line will have an impedance which may be considered to be made up of reactance (either inductive or capacitive) and resistance, in parallel. The reactive component can be eliminated by shunting the line at distance *X* from the antenna with another reactance equal in value but opposite in sign for the reactance presented by the line at that point. If distance *X* is such that the line presents an inductive reactance, a corresponding shunting capacitive reactance will be required.

The similarity between this linear circuit and the more familiar lumped tank circuit is apparent. In this case, the required compensating reactance may be supplied by shunting the line with a stub cut to proper length, *Y*. With the reactances canceled only a pure resistance remains as a termination for the remainder of the line between the sending end and the stub, and this resistance can be adjusted to match

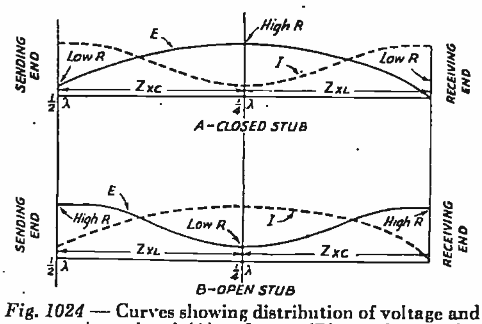


Fig. 1024 — Curves showing distribution of voltage and current along closed (A) and open (B) matching stubs.

the characteristic impedance of the line by adjusting the distance *X*.

A match may be obtained with either a closed or an open stub. If a closed stub is used, it will be found that the match occurs closer to the antenna where the antenna impedance is higher than the characteristic impedance of the line. For an open stub the same is true where the antenna impedance is lower than the line impedance. Since standing waves will exist between the antenna and the matching point, it is desirable to keep this distance as short as possible.

When the connection between the matching section and the antenna is unbalanced, as in the end-fed system, it is important that the antenna be the right length for the operating frequency if a good match is to be obtained (§ 10-7). The balanced center-fed system is less critical in this respect. The shorting-bar method of tuning the center-fed system to resonance may be used if the matching section is extended to a half wavelength, bringing a current loop at the free end.

In the center-fed system, the antenna and matching section should be cut to lengths found from the equations in §10-2 and §10-5. Any necessary on-the-ground adjustment can be made by adding to or clipping off the open ends of the matching section. In the end-fed system the matching section can be adjusted by making the line a little longer than necessary and adjusting the system to resonance by moving the shorting link up and down. Resonance can be determined by exciting the antenna at the proper frequency from a temporary antenna near by and measuring the current in the shorting bar by a low-range r.f. ammeter or galvanometer using one of the devices of this type described in Chapter Twenty. The position of the bar should be adjusted for maximum current reading. This should be done before the transmission line is attached to the matching section.

The position of the line taps will depend upon the impedance of the line as well as on the antenna impedance at the point of connection. The procedure is to take a trial point, apply power to the transmitter, and then check the transmission line for standing waves. This can be done by measuring the current in or voltage along the wires. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a quarter wavelength will indicate whether or not standing waves are present.

While the distance *X* may be determined experimentally, in as much as both *X* and *Y* are interdependent the cut-and-try method of experimental adjustment is not only laborious but of doubtful reliability. With the aid of the graphs, Figs. 1025 and 1026, the process can be reduced to one simple operation in arithmetic.

To make this calculation it is necessary to know the degree of mismatch, which can be determined by measuring the minimum and maximum amplitudes of the standing waves on the line. This measurement may be made either in terms of impedance (voltage) or current ratios: it makes no difference. However, if the readings are taken on a non-linear meter, such as a current-squared galvanometer, an appropriate correction must be made. The ratio of maximum to minimum current or voltage thus measured along the line is equal to the ratio of the antenna impedance to the characteristic impedance, Q.

With the antenna connected to the line and the stub disconnected, the r.f. meter should be moved along the line toward the sending end until the first point of maximum deflection (V_{max}) is reached. This is a reference point for all future measurements, so its location must be carefully determined. Record the magnitude of V_{max} and its distance from the antenna. Now move the meter farther toward the sending end, until a node (V_{min}) is found. Record the magnitude of V_{min}. As a cross-check for wavelength, the distance between V_{max} and V_{min} should be ¼ wavelength.

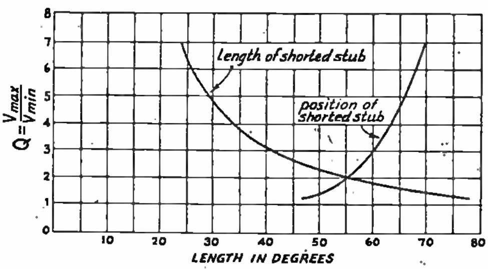


Fig. 1025 — Graphs for determining position and length of a shorted stub. Dimensions may be converted to linear units after values have been taken from the graph.

With the ratio of V_{max} to V_{min} known, the correct location of the stub and its length can be determined directly from the graphs. Fig. 1025 is used to find the position and length for a shorted stub, while Fig. 1026 is for an open stub. The only additional operation is conversion to linear units. This conversion is accomplished by simply dividing the fundamental wavelength in meters by 360 × the measured distance between the two points of maximum indication (also in meters). By first calculating the results in degrees, they may be made to apply to any wavelength where it is physically possible to use this method.

It will not usually be possible to obtain complete elimination of standing waves when the matching stub is exactly resonant, but the line taps should be adjusted for the smallest obtainable standing-wave ratio. Then a further "touching up" of the matching-stub tuning will eliminate the remaining standing wave, provided the adjustments are carefully made. The stub must be readjusted, because when resonant it exhibits some reactance as well as resistance at all points except at the ends, and

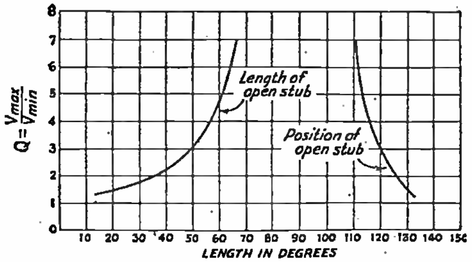


Fig. 1026 — Graphs for determining position and length of an open stub. Dimensions may be converted to linear units after values have been taken from the graph.

a slight lengthening or shortening of the stub is necessary to tune out this reactance.

An impedance mismatch of several per cent is of little consequence so far as power transfer to the antenna is concerned. It is relatively easy to get the standing-wave ratio down to 2 or 3 to 1, a perfectly satisfactory condition in practice. Of considerably greater importance is the necessity for getting the currents in the two wires balanced, both as to amplitude and phase. If the currents are not the same at corresponding points on adjacent wires and the loops and nodes do not also occur at corresponding points, there will be considerable radiation loss. Perfect balance can be brought about only by perfect symmetry in the line, particularly with respect to ground. This symmetry should extend to the coupling apparatus at the transmitter. An electrostatic shield between the line and the transmitter coupling coils often will be of value in preventing capacity unbalance, and at the same time will reduce harmonic radiation.

¶ 10-9 Long-Wire Antennas

Definition — An antenna will be resonant so long as an integral number of standing waves of current and voltage can exist along its length; in other words, so long as its length is some integral multiple of a half wavelength. When the antenna is more than a half-wave long it usually is called a long-wire antenna, or a harmonic antenna.

Current and voltage distribution — Fig. 1027 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is equal to a half wavelength) and at its second, third and fourth harmonics. For example, if the fundamental frequency of the antenna is 7 Mc., the current and voltage distribution will be as shown at A. The same antenna excited at 14 Mc. would have current and voltage distribution as shown at B. At 21 Mc., the third harmonic of 7 Mc., the current and voltage distribution would be as in C; and at 28 Mc., the fourth harmonic, as in D. The number of the harmonic is the number of half waves contained in the antenna at the particular operating frequency.

The polarity of current or voltage in each standing wave is opposite to that in the ad-

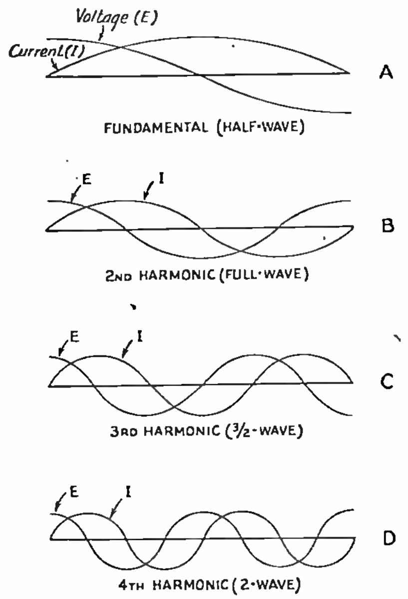


Fig. 1027 — Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.

jacent standing waves. This is shown in the figure by drawing the current and voltage curves successively above and below the antenna (taken as a zero reference line), to indicate that the polarity reverses when the current or voltage goes through zero. Currents flowing in the same direction are *in phase;* in opposite directions, *out of phase.*

It is evident that one antenna may be used for harmonically related frequencies, such as the various amateur bands. The long-wire or harmonic antenna is the basis of multi-band operation with one antenna.

Physical lengths — The length of a long-wire antenna is not an exact multiple of that of a half-wave antenna because the end effects (§ 10-2) operate only on the end sections of the antenna; in other parts of the wire these effects are absent, and the wire length is approximately that of an equivalent portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$\textit{Length (feet)} = \frac{492\ (N\text{-}0.05)}{\textit{Freq. (Mc.)}} \qquad (10)$$

where N is the number of *half* waves on the antenna. From this, it is apparent that an antenna cut as a half wave for a given frequency will be slightly off resonance at exactly twice that frequency (on the second harmonic) because of the different behavior of end effects when there is more than one standing wave on the antenna. For instance, if the antenna is cut to have exact fundamental resonance on the second harmonic (full-wave operation) it should be 2.6 per cent longer, and on the fourth harmonic (two-wave), 4 per cent longer. The effect is not very important except for a possible unbalance in the feeder system (§ 10-7), which may result in some radiation from the feeder in end-fed systems.

Impedance and power gain — The radiation resistance as measured at a current loop becomes larger as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable direction than does a half-wave antenna in its most favorable direction. This power gain is secured at the expense of radiation in other directions. Fig. 1028 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

Directional characteristics — As the wire is made longer in terms of the number of half wavelengths, the directional effects change. Instead of the "doughnut" pattern of the half-wave antenna, the directional characteristic splits up into "lobes" which make various angles with the wire. In general, as the length of the wire is increased the direction in which maximum radiation occurs tends to approach the line of the antenna itself.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 1029, 1030 and 1031, for three vertical angles of radiation. Note that, as the wire length increases, the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to have practically "end-on" directional characteristics, even at the lower radiation angles.

Methods of feeding — In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig.

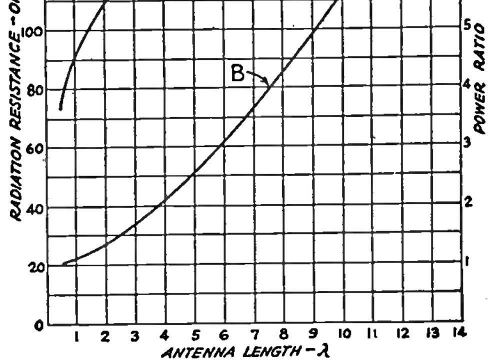


Fig. 1028 — Curve A shows variation in radiation resistance with antenna length. Curve B shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

1032 and Fig. 1027. The feeder system must not upset this phase relationship. This requirement is met by feeding the antenna at either end or at any current *loop*. A two-wire feeder cannot be inserted at a current *node*, however, because this invariably brings the currents in two adjacent half-wave sections in phase; if the phase in one section could be reversed, then the currents in the feeders necessarily would have to be in phase and the feeder radiation would not be canceled out.

Either resonant or non-resonant feeders may be used. With the latter, the systems employing a matching section (§ 10-8) are best. The non-resonant line may be tapped on the matching section, as in Fig. 1022, or a "Q" type section, Fig. 1021, may be employed. In such case, Fig. 1033 gives the required surge impedance for the matching section. It can also be calculated from Equation 9 (§ 10-8) and the radiation resistance data in Fig. 1028.

Methods of coupling the line to the transmitter are the same as described in § 10-6 for the particular type of line used.

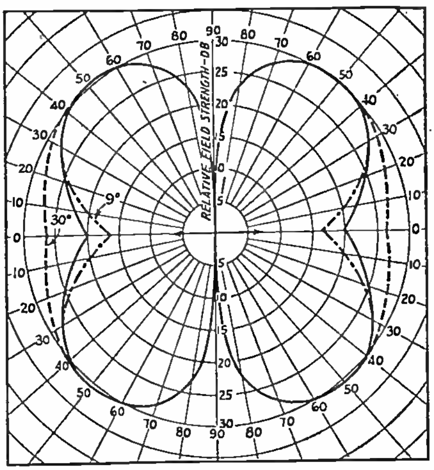


Fig. 1029 — Horizontal patterns of radiation from a full-wave antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale; actual amplitudes will depend upon the height of the antenna.

10-10 Multiband Antennas

Principles — As suggested in the preceding section, the same antenna may be used for several bands by operating it on harmonics. When this is done it is necessary to use resonant feeders, since the impedance matching for non-resonant feeder operation can be accomplished only at one frequency unless means are provided for changing the length of a matching section and shifting the point at which the feeder is attached to it. A matching section which is only a quarter-wavelength long at one frequency will be a half-wavelength long at twice that frequency, and so on; and changing the length of the wires, even by switching, is so inconvenient as to be impracticable.

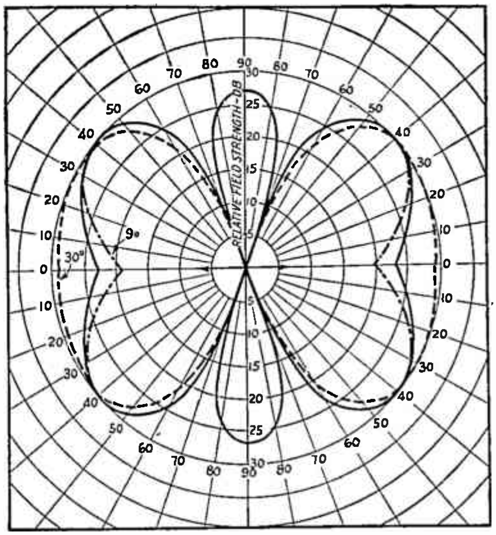


Fig. 1030 — Horizontal patterns of radiation from an antenna three half-waves long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

Furthermore, the current loops shift to a new position on the antenna when it is operated on harmonics, further complicating the feed situation. It is for this reason that a half-wave antenna which is center-fed by a rubber-insulated line is practically useless for harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses arise in the rubber dielectric.

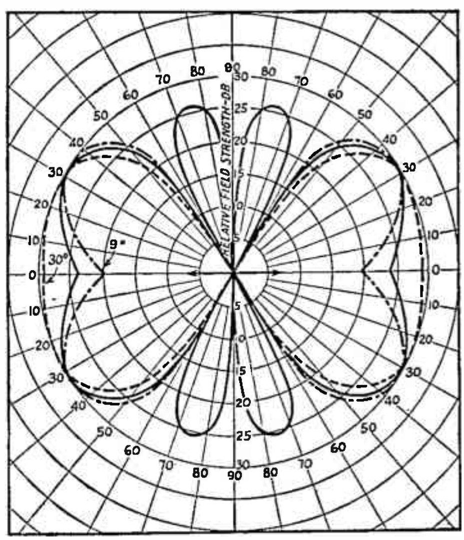


Fig. 1031 — Horizontal patterns of radiation from an antenna two wavelengths long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

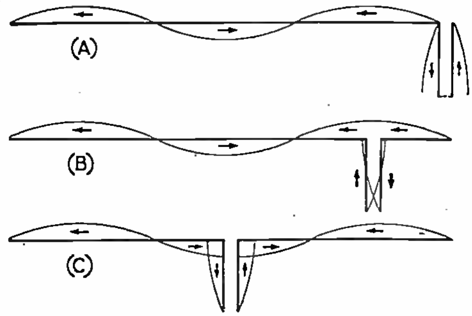


Fig. 1032 — Current distribution and feed points for long-wire antennas. A 3/2-wave antenna is used as an illustration. With two-wire feed, the line may be connected at the end of the antenna or at any current *loop* (but not at a current node) for harmonic operation.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will vary with the band in use.

Simple systems — Any of the antenna arrangements shown in § 10-7 may be used for multi-band operation by making the antenna a half wave long at the lowest frequency to be used. The feeders should be a quarter wave, or some multiple of a quarter wave, long at the same frequency. Typical examples, together with the type of tuning to be used, are given in Table I. The figures given represent a compromise designed to give satisfactory operation on all the bands considered, taking into account the change in required length as the order of the harmonic goes up.

A center-fed half-wave antenna will not operate as a long wire on harmonics, because of the phase reversal at the feeders previously mentioned (§ 10-9). On the second harmonic the two antenna sections are each a half wave long, and, since the currents are in phase, the directional characteristic is different from that of a full-wave antenna even though the over-all length is the same. On the fourth harmonic each section is a full wave long, and, again because of the direction of current flow, the system will not operate as a two-wave-length antenna. It should not be assumed that these systems are not effective radiators; it simply means that the directional characteristic will not be that of a long wire having the same over-all length. Rather, it will resemble the characteristic of one side of the antenna, although not necessarily having the same exact form.

Antennas with a few other types of feed systems may be operated on harmonics for the higher-frequency bands, although their performance is some-

TABLE I

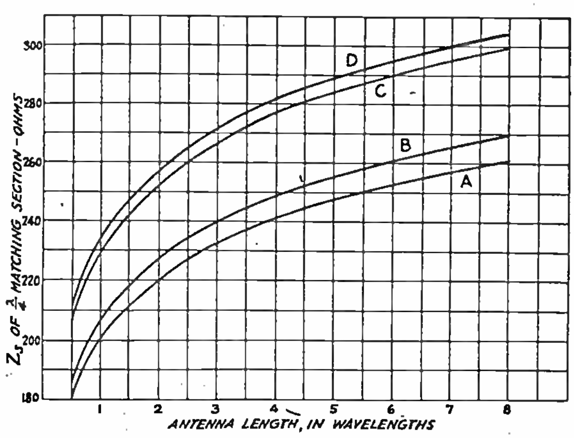
Multi-Band Resonant-Line-Fed Antennas

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed:			
243	120	1.75-Mc. 'phone	series
		4-Mc. 'phone	parallel
		14 Mc.	parallel
		28 Mc.	parallel
136	67	3.5-Mc. c.w.	series
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
134	67	3.5-Mc. c.w.	series
		7 Mc.	parallel
67	33	7 Mc.	series
		14 Mc.	parallel
		28 Mc.	parallel
With center feed:			
272	135	1.75 Mc.	parallel
		3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
137	67	3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
67.5	34	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc., but will work well in the region which quadruples into the 14-Mc. band (3500–3600 kc.). Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length; the 272-foot antenna, for instance, may be used for both c.w. and 'phone on either 1.75 or 4 Mc. without loss of efficiency.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

Fig. 1033 — Required surge impedance of quarter-wave matching sections for radiators of various lengths. Curve A is for a transmission-line impedance of 440 ohms, Curve B is for 470 ohms, Curve C for 580 ohms and Curve D for 600 ohms. Dimensions for matching sections of the required impedance are obtained from Fig. 1009.



what impaired. The singlewire-fed antenna (§ 10-8) may be used in this way; the feeder and antenna will not be matched exactly on harmonics, with the result that standing waves will appear on the feeder, but the system as a whole will radiate. A better match will be obtained if the point of connection of the feeder to the antenna is made exactly one-third the over-all antenna length from one end. While this disagrees slightly with the figures given for a half-wave antenna, it has been found to work better on the harmonic frequencies.

The "Q" antenna system (§ 10-8) also can be operated on harmonics, but the line cannot operate as a non-resonant line except at the fundamental frequency of the antenna. For harmonic operation the line must be tuned, and therefore the feeder length is important. The tuning system will depend upon the number of quarter waves on the line, including the "Q" bars. The concentric-line-fed antenna (§ 10-8) may be used on harmonics, if the concentric line is air-insulated. Its operation on harmonics is similar to that of the "Q." This antenna is not recommended for multi-band operation with a rubber-insulated line, however.

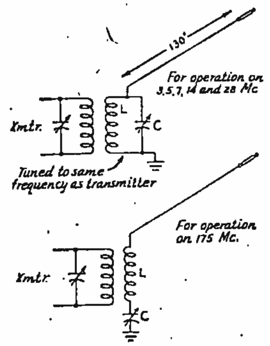


Fig. 1034 — A simple antenna system for five amateur bands. The antenna is voltage fed on 3.5, 7, 14 and 28 Mc., working on the fundamental, second, fourth and eighth harmonics, respectively. For 1.75 Mc. the system is a quarter-wave grounded antenna, in which case series tuning must be used. The antenna wire should be kept well in the clear and should be as high as possible. If the length of the antenna is increased to approximately 260 feet, voltage feed can be used on all five bands.

The delta-match system (§ 10-8) can be used on harmonics, although some standing waves will appear on the line. For that matter, *any* antenna system can be used on harmonic frequencies by tying the feeders together at the transmitter end and feeding the system as a single wire by means of a tuned circuit coupled to the transmitter.

A simple antenna system without feeders, useful for operation on five bands, is shown in Fig. 1034. On all bands from 3.5 Mc. upward it operates as an end-fed antenna — half wave on 3.5 Mc., long wire on the other bands. On 1.75 Mc. it is only a quarter wave in length, and must be worked against ground (§ 10-14). On this band, since it is fed at a high-current point, series tuning (§ 10-6) must be used.

Antennas for restricted space — If the space available for the antenna is not large enough to accommodate the length necessary for a half wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half wave long. Nevertheless, such a system is useful where operation on the desired band otherwise would be impossible.

Resonant feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance. With end feed the feeder currents become badly unbalanced, and, since lengths midway between those requiring series or parallel tuning ordinarily must be used to bring the entire system to resonance, coupling to the transmitter often becomes difficult.

With center feed practically any convenient length of antenna can be used, if the feeder length is adjusted to accommodate at least one half wave around the whole system. Typical cases are shown in Fig. 1035, one for an antenna having a length of one quarter wave (A) and the other for an antenna somewhat longer (C) but still not a half wave long. Current distribution is shown for both fundamental and second harmonic. From the points marked *X*, resonant feeders any convenient number of quarter waves in length may be extended to the operating room. The sum of the distances on each wire from *X* to the antenna end must equal a half wave. It is sufficiently accurate to use Equation 2 (§ 10-2) in calculating this length. Note that *X*-*X* is a high-current point on these shortened antennas, corresponding to the center of a half-wave antenna. It is also apparent that the antenna at A is a half-wave antenna on the next higher-frequency band (B).

A practical antenna of this type can be made as shown in Fig. 1036. Table II gives a few

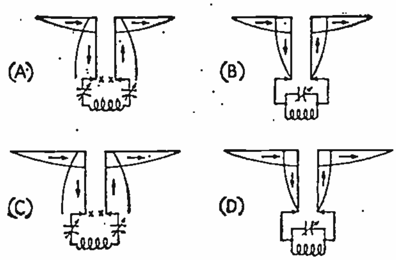


Fig. 1035 — Current distribution on short antennas. Those at the left are too short for fundamental operation, one (A) having an over-all length of one quarter wave; the other (C) being longer but not a half-wave long. These systems may be used wherever space to erect a full half-wave antenna is not available. The current distribution for second harmonic-operation is shown at the right of each figure (B and D). In A and C, the total length around the system is a half-wave at the fundamental. In B and D, the over-all length is a full wave. Arrows show the instantaneous direction of current flow.

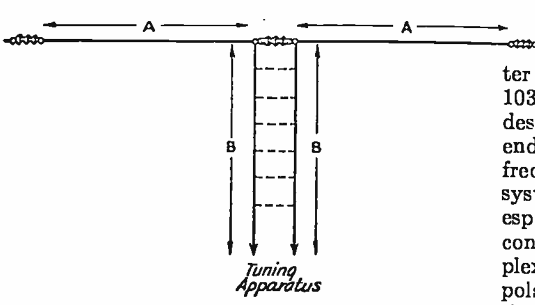


Fig. 1036 — Practical arrangement of a shortened antenna. The total length, A + B + B + A, should be a half wavelength for the lowest-frequency band, usually 3.5. Mc. See Table II for lengths and tuning data.

recommended lengths. Remembering the preceding discussion, however, the antenna can be made any convenient length, provided the feeder is considered to "begin" at *X-X* and the line length is adjusted accordingly.

Bent antennas — Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a half-wave antenna (the center quarter wave, approximately) does most of the radiating (§ 10-1). Advantage can be taken of this fact when the space available does not permit erecting an antenna a half-wave long. In this case the ends may be bent, either horizontally or vertically, so that the total length equals a half wave, even though the straightaway horizontal length may be as short as a quarter wave.

TABLE II

Antenna and Feeder Lengths for Short Multi-Band Antennas, Center-Fed

Antenna length (ft.)	*Feeder length (ft.)*	*Band*	*Type of tuning*
137	68	1.75 Mc.	series
		3.5 Mc.	parallel
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
100	38	3.5 Mc.	parallel
		7 Mc.	series
		14 Mc.	series
		28 Mc.	series or parallel
67.5	34	3.5 Mc.	series
		7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
50	43	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
33	51	7 Mc.	parallel
		14 Mc.	parallel
		28 Mc.	parallel
33	31	7 Mc.	parallel
		14 Mc.	series
		28 Mc.	parallel

The operation is illustrated in Fig. 1037. Such an antenna will be a somewhat better radiator than the arrangement of Fig. 1035-A on the lowest frequency, but is not so desirable for multi-band operation because the ends play an increasingly important part as the frequency is raised. The performance of the system in such a case is difficult to predict, especially if the ends are vertical (the most convenient arrangement), because of the complex combination of horizontal and vertical polarization which results as well as the dissimilar directional characteristics.

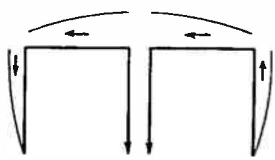


Fig. 1037 — Folded arrangement for shortened antennas. The total length is a half wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter-wave long.

ℂ 10-11 Long-Wire Directive Arrays

The "V" antenna — It has been emphasized that, as the antenna length is increased, the lobe of maximum radiation makes a more acute angle with the wire (§ 10-9). Two such wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bidirectional) along a line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate. It can also be used on harmonics, so that it is suitable for multi-band work. The "V" antenna is shown in Fig. 1038.

Fig. 1039 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multiband operation. Angle α

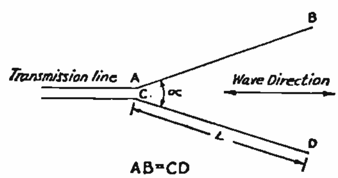


Fig. 1038 — The "V" antenna, made by combining two long wires in such a way that each reinforces the radiation from the other. The important quantities are the length of each leg and the angle between the legs.

is approximately equal to twice the angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 1039 is the vertical angle of maximum radiation (§10-1). Tilting the whole horizontal plane of the "V" will tend to increase the low-angle radiation off the low end and decrease it off the high end.

The gain increases with the length of the wires, but is not exactly twice the gain for a single long wire as given in Fig. 1028. In the longer lengths the gain will be somewhat increased, because of mutual coupling between the wires. A "V" eight wavelengths on a leg, for instance, will have a gain of about 12 db. over a half-wave antenna, whereas twice the gain of a single 8-wavelength wire would be only approximately 9 db.

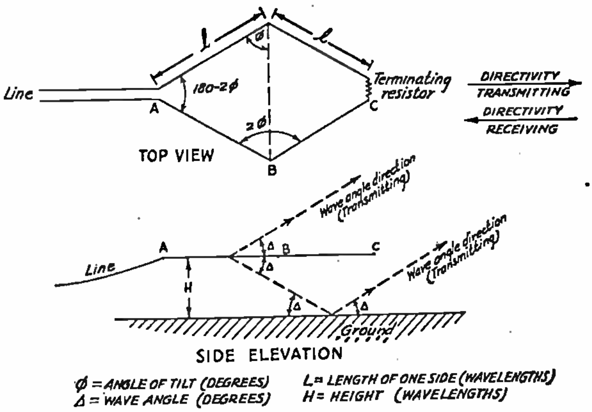


Fig. 1040 — The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

The two wires of the "V" must be fed out of phase, for correct operation. A resonant line may simply be attached to the ends, as shown in Fig. 1038. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a non-resonant line (§ 10-8). If the antenna wires are made multiples of a half wave in length (use Equation 10, § 10-9, for computing the length), the matching section will be closed at the free end.

The rhombic antenna — The horizontal rhombic or "diamond" antenna is shown in Fig. 1040. Like the "V," it requires a good deal of space for erection, but it is capable of giving excellent gain and directivity. It also can be used for multi-band operation. In the terminated form shown in Fig. 1040, it operates like a non-resonant transmission line, without standing waves, and is unidirectional. It may also be used without the terminating resistor, in which case there are standing waves on the wires and the antenna is bidirectional.

The important quantities influencing the design of the rhombic antenna are shown in Fig. 1040. While several design methods may be used, the one most applicable to the conditions existing in amateur work is the so-called "compromise" method. The chart of Fig. 1041 gives design information based on a given length and wave angle to determine the remaining optimum dimensions for best operation. Curves for values of length of 2, 3 and 4 wavelengths are shown, and any intermediate values may be interpolated.

With all other dimensions correct, an increase in length causes an increase in power gain and a slight reduction in wave angle. An increase in height also causes a reduction in wave angle and an increase in power gain, but not to the same extent as a proportionate increase in length. For multiband work, it is satisfactory to design the rhombic antenna on the basis of 14-Mc. operation, which will permit work on the 7- and 28-Mc. bands as well.

A value of 800 ohms is correct for the terminating resistor for any properly constructed rhombic, and the system behaves as a pure resistive load under this condition. The terminating resistor must be capable of safely dissipating one-half the power output (to eliminate the rear pattern), and should be noninductive. Such a resistor may be made up from a carbon or graphite rod or from a long 800-ohm transmission line using resistance wire. If the carbon rod or a similar form of lumped resistance is used, the device should be suitably protected from weather effects, i.e., it should be covered with a good asphaltic compound and sealed in a small, light-weight box or fibre tube. Suitable nonreactive terminating resistors are also available commercially.

For feeding the antenna, the antenna impedance will be matched by an 800-ohm line, which may be constructed

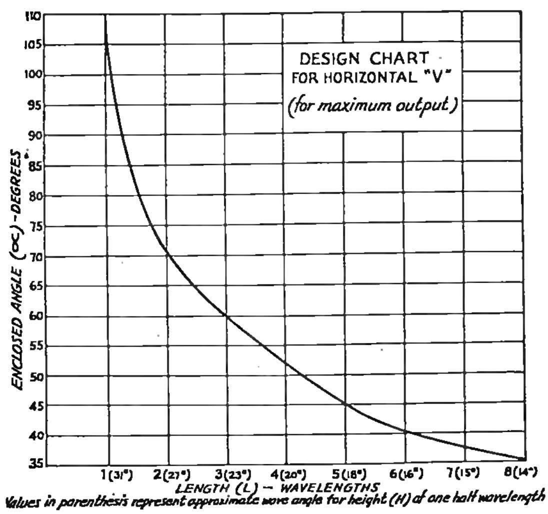


Fig. 1039 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides vs. the length of the wires.

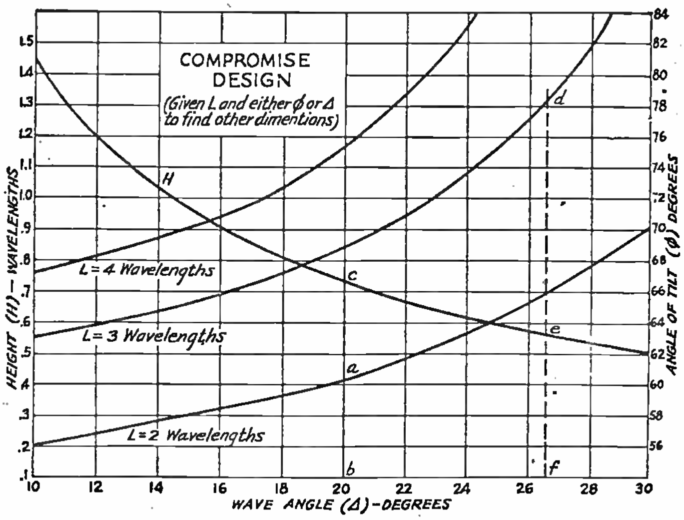


Fig. 1041 — Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:

(1) Given:
Length (L) = 2 wavelengths.
Desired wave angle (Δ) = 20°.
To Find: H, Φ.
Method:
Draw vertical line through point *a* (L = 2 wavelengths) and point *b* on abscissa (Δ = 20°.) Read angle of tilt (Φ) for point *a* and height (H) from intersection of line *ab* at point *c* on curve H.
Result:
Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:
Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.
To Find: H, Δ.
Method:
Draw a vertical line from point *d* on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point *e*) for height, and intersection at point *f* on the abscissa for Δ.
Result:
H = 0.56 wavelength.
Δ = 26.6°.

from No. 16 wire spaced 20 inches or from No. 18 wire spaced 16 inches. The 800-ohm line is somewhat ungainly to install, however, and may be replaced by an ordinary 600-ohm line with only a negligible mismatch. Alternatively, a matching section may be installed between the antenna terminals and a low-impedance line. However, when such an arrangement is used, it will be necessary to change the matching-section constants for each different band on which operation is contemplated.

The same design details apply to the unterminated rhombic as to the terminated type. When used without a terminating resistor, the system is bidirectional. Resonant feeders are preferable for the unterminated rhombic. A non resonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to multiband work.

Rhombic antennas will give a power gain of 8 to 12 db. or more for leg lengths of two to four wavelengths, when constructed according to the charts given. In general, the larger the antenna, the greater the power gain.

◖ 10-12 Directive Arrays with Driven Elements

Principles — By combining individual half-wave antennas into an *array* with suitable spacing between the antennas (called *elements*) and feeding power to them simultaneously, it is possible to make the radiated fields from the individual elements add in a favored direction, thus increasing the field strength in that direction as compared to that produced by one antenna element alone. In other directions the fields will more or less oppose each other, giving a reduction in field strength. Thus a power gain in the desired direction is secured at the expense of a power reduction in other directions.

Besides the spacing between elements, the instantaneous direction of current flow (*phase*) in individual elements determines the directivity and power gain. There are several methods of arranging the elements. If they are strung end to end, so that all lie on the same straight line, the elements are said to be *collinear*. If they are parallel and all lying in the same plane, the elements are said to be *broadside* when the phase of the current is the same in all, and *end-fire* when the currents are not in phase. Elements which receive power from the transmitter through the transmission line are called *driven elements.*

The power gain of a directive system increases with the number of elements. The proportionality between gain and number of elements is not simple, however. The gain depends upon the effect which the spacing and phasing has upon the radiation resistance of the elements, as well as upon their number.

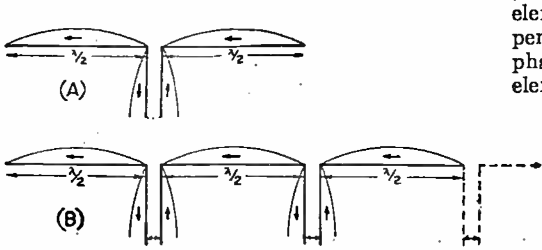


Fig. 1042 — Collinear half-wave antennas in phase. The system at A is generally known as "two half waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

Collinear arrays — Simple forms of collinear arrays, with the current distribution, are shown in Fig. 1042. The two-element array at A is popularly known as "two half waves in phase." It will be recognized as simply a center-fed antenna operated at its second harmonic. The way in which the number of elements may be extended for increased directivity and gain is shown in Fig. 1042-B. Note that quarter-wave transmission lines are used between each element; these give the reversal in phase necessary to make the currents in

TABLE III

THEORETICAL GAIN OF COLLINEAR HALF-WAVE ANTENNAS

Spacing between centers of adjacent half waves	*Number of half waves in array vs. gain in db.*				
	2	3	3	5	6
½ Wave	1.8	3.3	4.5	5.3	6.2
¾ Wave	3.2	4.8	6.0	7.0	7.8

individual antenna elements all flow in the same direction at the same instant. Another way of looking at it is to consider that the whole system is a long wire, with alternate half-wave sections folded so that they do not radiate. Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder (§ 10-8), or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end-fed by any of the systems previously described (§ 10-7, 10-8), or any element may be center-fed. It is best to feed at the center of the array, so that the energy will be distributed as uniformly as possible among the elements.

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table III. Although ¾-wave spacing gives greater gain, it is difficult to construct a suitable phase-reversing system when the ends of the antenna elements are widely separated. For this reason, the half-wave spacing is most generally used in actual practice.

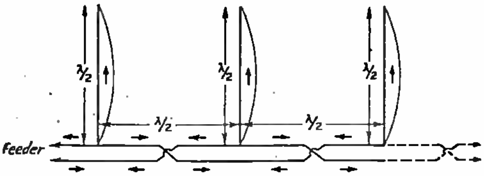


Fig. 1043 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the plane of the antennas (perpendicularly through this page).

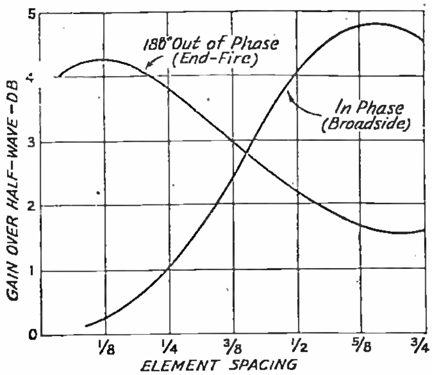


Fig. 1044 — Gain vs. spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mounting gives increased horizontal directivity, while the vertical directivity remains the same as for a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles. It is seldom practicable to use more than two elements vertically at frequencies below 14 Mc. because of the excessive height required.

TABLE IV

THEORETICAL GAIN VS. NUMBER OF BROADSIDE ELEMENTS (HALF-WAVE SPACING)

No. of elements	*Gain*
2	4 db.
3	5.5 db.
4	7 db.
5	8 db.
6	9 db.

Broadside arrays — Parallel antenna elements with currents in phase may be combined as shown in Fig. 1043 to form a *broadside* array, so named because the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the number of elements and the spacing, the gain for different spacings being shown in Fig. 1044. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table IV gives theoretical gain as a function of the number of elements with half-wave spacing.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other (*stacked*). In the former case the horizontal pattern becomes quite sharp, while the vertical pattern is the same as that of one element alone. If the array is suspended horizontally, the horizontal pattern is equivalent to that of one element

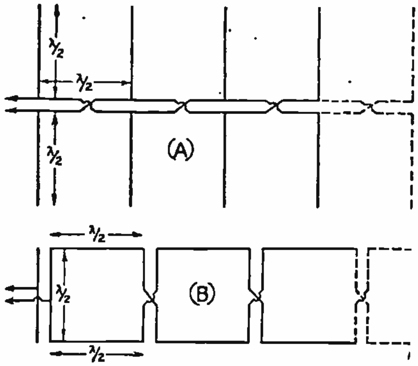


Fig. 1045 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table IV) plus the gain of one set of collinear elements (Table III). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.), giving a total gain of 8.8 db. In B, each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.), making the total gain 7.3 db. The result is not strictly accurate, because of mutual coupling between the elements, but is good enough for practical purposes.

while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines (§ 10-7) or through quarter-wave matching sections and nonresonant lines (§ 10-8). In Fig. 1043, note the "crossing over" of the feeders, which is necessary to bring the elements in proper phase relationship.

Combined broadside and collinear arrays — Broadside and collinear arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. The general plan of constructing such antennas is shown in Fig. 1045. The lower angle of radiation resulting from stacking elements in the vertical plane is desirable at the higher frequencies. In general, doubling the number of elements in an array by stacking will raise the gain from 2 to 4 db., depending upon whether vertical or horizontal elements are used — that is, whether the stacked elements are of the broadside or collinear type.

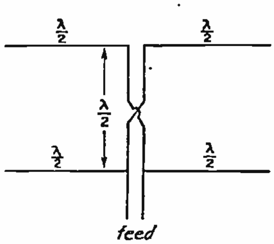


Fig. 1046 — A four-element combination broadside-collinear array, popularly known as the "lazy H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

The arrays in Fig. 1045 are shown fed from one end, but this is not especially desirable in the case of large arrays. Better distribution of energy between elements, and hence better all-around performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the 8-element array at A, the feeders could be introduced at the middle of the transmission line between the second and third set of elements, in which case the connecting line would not be transposed. Alternatively, the antenna could be constructed with the transpositions as shown and the feeder connected between the adjacent ends of either the second or third pair of collinear elements.

A four-element array of the general type shown in Fig. 1045-B, known as the "lazy H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 1046.

End-fire arrays — Fig. 1047 shows a pair of parallel half-wave elements with currents out of phase. This is known as an *end-fire* array, because it radiates best along the line of the antennas, as shown.

The end-fire array may be used either vertically or horizontally (elements at the same height), and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 1044 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and

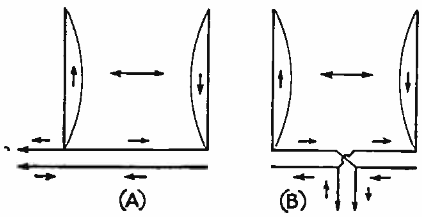


Fig. 1047 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 1044. Direction of maximum radiation is shown by the large arrows.

broadside elements to give a further increase in gain and directivity.

Either resonant or nonresonant lines may be used with this type of array. Nonresonant lines preferably are matched to the antenna through a quarter-wave matching section (§ 10-8).

Checking phasing — Figs. 1045 and 1047 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 1047, when the transmission line is connected as at A there is no crossover in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the crossover becomes necessary (B). This is because in B the two halves of the connecting line are simply *branches* of the same line. In other

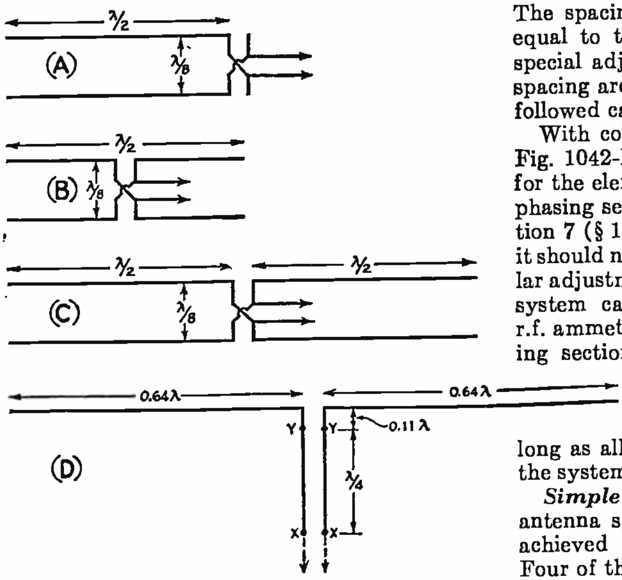


Fig. 1048 — Simple directive antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is a simple two-element broadside array using extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5 db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db. In A, B and C, the phasing line contributes about 1/16th wavelength to the transmission line; when B is used on the second harmonic, this contribution is ⅛ wavelength. Alternatively, the antenna ends may be bent to meet the transmission line, in which case each feeder is simply connected to one antenna. In D, points *Y-Y* indicate a quarter-wave point (high current) and *X-X* a half-wave point (high voltage). The line may be extended in multiples of quarter waves, if resonant feeders are to be used. A, B, and C may be suspended on wooden spreaders. The plane containing the wires should be parallel to the ground.

words, even though the connecting line in B is a half wave in length, it is not actually a half-wave line but *two quarter-wave lines in parallel*. The same thing is true of the untransposed line of Fig. 1045. Note that, under these conditions, the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two antenna elements and does not have the feed line connected to its center, as in Fig. 1043.

Adjustment of arrays — With arrays of the types just described, using half-wave spacing between elements, it will usually suffice to make the length of each element that given by the equation for a half-wave antenna in § 10-2, while the half-wave phasing lines between the parallel elements can be calculated from the formula:

$$\textit{Length of half-wave line (feet)} = \frac{492 \times 0.975}{\textit{Freq. (Mc.)}} = \frac{480}{\textit{Freq. (Mc.)}}$$

The spacing between elements can be made equal to the length of the phasing line. No special adjustments line or element length or spacing are needed, provided the formulas are followed carefully.

With collinear arrays of the type shown in Fig. 1042-B, the same formula may be used for the element length while the quarter-wave phasing section can be calculated from Equation 7 (§ 10-5). If the array is fed at its center it should not be necessary to make any particular adjustments, although, if desired, the whole system can be resonated by connecting an r.f. ammeter in the shorting link on each phasing section and moving the link back and forth to find the maximum current position. This refinement is hardly necessary in practice, however, so long as all elements are the same length and the system is symmetrical.

Simple arrays — Several simple directive antenna systems using driven elements have achieved rather wide use among amateurs. Four of these systems are shown in Fig. 1048. Tuned feeders are assumed in all cases; however, a matching section (§ 10-8) readily can be substituted if a nonresonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths can be calculated from the equations in § 10-2 for the antenna and from Equation 7 (§ 10-5) for the resonant transmission line or matching section. In cases where the transmission-line proper connects to the mid-point of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 1044) for such operation.

A close-spaced four-element array is shown at C. It will give about 2 db. more gain than the two-element array.

The antenna at D, commonly known as the "extended double Zepp," is designed to take advantage of the greater gain possible with collinear antennas having greater than half-wave center-to-center spacing, but without introducing feed complications. The elements are made longer than a half wave in order to bring this about. The gain is 3 db. over a single half-wave antenna, and the broadside directivity is quite sharp.

The antennas of A and B may be mounted either horizontally or vertically; horizontal suspension (with the elements in a plane parallel to the ground) is recommended, since this tends to give low-angle radiation without an unduly sharp horizontal pattern. Thus these systems are useful for coverage over a wide horizontal angle. The system at C, when mounted horizontally, will have a sharper horizontal pattern than the two-element arrays.

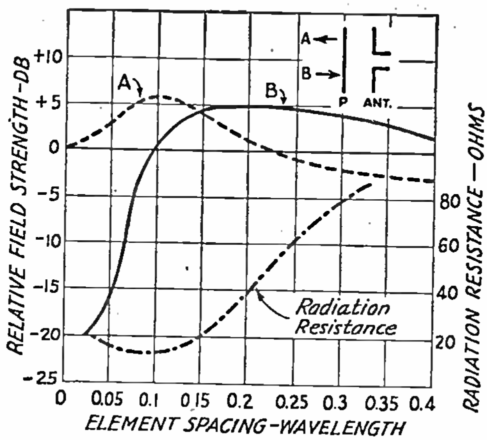


Fig. 1049—Gain vs. element spacing for an antenna and one parasitic element. The reference point, 0 db. is the field strength from a half-wave antenna alone. The greatest gain is in direction A at spacings of less than 0.14 wavelength, and in direction B at greater spacings. The front-to-back ratio is the difference in db. between curves A and B. Variation in radiation resistance of the driven element also is shown. These curves are for a self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; the gain as a director can be increased by shortening. This also improves the front-to-back ratio.

ℂ 10-13 Directive Arrays with Parasitic Elements

Parasitic excitation — The antenna arrays described in § 10-12 are bidirectional; that is, they will radiate in directions both to the "front" and to the "back" of the antenna system. If radiation is wanted in only one direction (for instance, north only, instead of north-south), it is necessary to use different element arrangements. In most of these arrangements the additional elements receive power by induction or radiation from the driven element, generally called the "antenna," and re-radiate it in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line.

The parasitic element is called a *director* when it reinforces radiation on a line pointing to it from the antenna, and a *reflector* when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic element tuning (which usually is adjusted by changing its length), and, particularly when the element is self-resonant, upon the spacing between it and the antenna.

Gain vs. spacing — The gain of an antenna-reflector or an antenna-director combination varies chiefly with the spacing between the elements. The way in which gain varies with spacing is shown in Fig. 1049, for the special case of self-resonant parasitic elements. This chart also shows how the attenuation to the "rear" varies with spacing. The same spacing does not necessarily give both maximum forward gain and maximum backward attenuation. Backward attenuation is desirable when the antenna is used for receiving, since it greatly reduces interference coming from the opposite direction to the desired signal.

Element lengths — The antenna length is given by the formulas in § 10-2. The director and reflector lengths must be determined experimentally for maximum performance. The preferable method is to aim the antenna at a receiver a mile or more distant and have an observer check the signal strength (on the receiver "S" meter) while the reflector or director is adjusted a few inches at a time, until the length which gives maximum signal is found. The attenuation may be similarly checked, the length being adjusted for minimum signal. In general, for best front-to-back ratio the length of a director will be about 4 per cent less than that of the antenna. The reflector will be about 5 per cent longer than the antenna.

Simple systems; the rotary beam — Four practical combinations of antenna, reflector and director elements are shown in Fig. 1050. Spacings which give maximum gain or maximum front-to-back ratio (ratio of power radiated in the desired direction to power radiated in the opposite direction) may be taken from Fig. 1049. In the chart, the front-to-back ratio in db. will be the sum of gain and attenuation at the same spacing.

Systems of this type are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directiv-

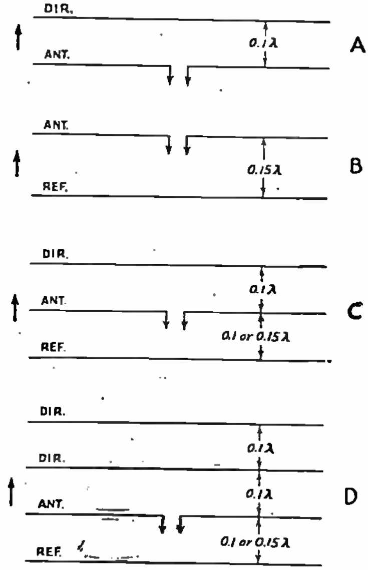


Fig. 1050—Half-wave antennas with parasitic elements. A, with reflector; B, with director; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 1049, in the first two cases, and depends upon the spacing and length of the parasitic element. In the three- and four-element arrays a reflector spacing of 0.15 wavelength will give slightly more gain than 0.1-wavelength spacing. Arrows show direction of maximum radiation. The array should be mounted horizontally (top views are shown).

ity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

Arrays using more than one parasitic element, such as those shown at C and D in Fig. 1050, will give more gain and directivity than is indicated for a single reflector and director by the curves of Fig. 1049. The gain with a properly adjusted three-element array (antenna, director and reflector) will be 5 to 7 db. over a half-wave antenna. Somewhat higher gain still can be secured by adding a second director to the system, making a four-element array. The front-to-back ratio is correspondingly improved as the number of elements is increased.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half- to one-inch diameter, both to reduce the ohmic resistance (§ 10-2) of the conductors and to secure mechanical rigidity. If the elements are free to move with respect to each other, the array will tend to show detuning effects under windy conditions.

Feeding close-spaced arrays — While any of the usual methods of feed may be applied to the driven element of a parasitic array, the fact that, with close spacing, the radiation resistance as measured at the center of the driven element drops to a very low value makes some systems more desirable than others. The preferred methods are shown in Fig. 1045. Resonant feeders are not recommended for lengths greater than a half wavelength.

The quarter- or half-wave matching stubs shown at A and B in Fig. 1045 preferably should be constructed of tubing with rather close spacing, in the manner of the "Q" section. This lowers the impedance of the matching section and makes the position of the line taps somewhat less difficult to determine accurately. The line adjustment should be made only with the parasitic elements in place, and after the correct element lengths have been determined, it should be checked to compensate for changes likely to occur because of element tuning. The procedure is the same as that described in § 10-8.

The concentric-line matching section at C will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as an impedance-inverting transformer, and, if its characteristic impedance is 70 ohms, it will give an exact match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2 to 1. The length of the quarter-wave section may be calculated from Equation 7 (§ 10-5).

The delta matching transformer shown at D is an excellent arrangement for parasitic arrays, and is probably easier to install, mechanically, than any of the others. The positions of the taps (dimension a) must be determined experimentally, along with the length, b, by checking the standing-wave ratio on the line as adjustments are made. Dimension b should be about 15 per cent longer than a.

Sharpness of resonance — Peak performance of a multi-element directive array depends upon proper phasing or tuning of the elements, which in all but the simplest systems can be exact for one frequency only. However, there is some latitude, and most arrays will work well over a relatively narrow region such as the 14 Mc. band. If frequencies in all parts of the band are to be used, the antenna system should be designed for the mid-frequency; on the other hand, if only one frequency in the band will be used for the greater portion of the time, the antenna might be designed for that frequency and some degree of misadjustment tolerated on the occasionally used spare frequencies.

When reflectors or directors are used the tolerance is usually less than in the case of driven elements, partly because the parasitic-element lengths are fixed and the operation may change appreciably as the frequency passes from one side of resonance to the other, and partly be-

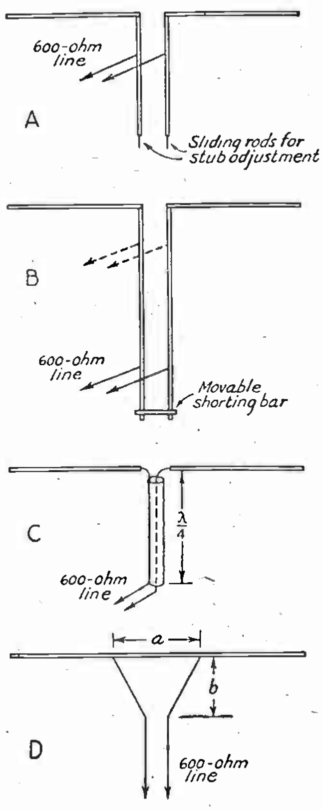


Fig. 1051 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer. Adjustment details are discussed in the text.

cause the close spacing ordinarily used results in a sharp-tuning system. With parasitic elements, operation should be confined to a small region about the frequency for which the antenna is adjusted if peak performance is to be secured.

Combination arrays — It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broadside-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array could be treated in the same fashion.

When combination arrays are built up, a rough approximation of the gain to be expected may be obtained by adding the gains for each type of combination. Thus the gain of two broadside sets of four collinear arrays with a set of reflectors, one behind each element, at quarter-wave spacing for the parasitic elements, would be estimated as follows: From Table III, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 1044 or Table IV, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 1049, the gain of a parasitic reflector at quarter-wave spacing is 4.5 db. The total gain is then the sum, or 13 db. for the sixteen elements. Note that using two *sets* of elements in broadside is equivalent to using two elements, so far as gain is concerned; similarly with sets of reflectors, as against one antenna and one reflector. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements and upon the effect which mutual coupling between elements has upon the radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive antenna combinations can be worked out by combining elements according to these principles.

10-14 Miscellaneous Antenna Systems

Grounded antenna — The grounded antenna is used almost exclusively for 1.75-Mc. work, where the length required for a half-wave antenna would be excessive for most locations. An antenna worked "against ground" need be only a quarter-wave long, approximately, because the earth acts as an electrical "mirror" which supplies the missing quarter wave. The current is maximum at the ground connection with a quarter-wave antenna, just as it is at the center of a half-wave antenna.

On 1.75 Mc. the most useful radiation is from the vertical part of the antenna, since vertically polarized waves are characteristic of ground-wave transmission. It is therefore desirable to make the down-lead as nearly vertical as possible, and also as high as possible. This gives low-angle sky-wave transmission, which is most useful for long-distance work at night, in addition to a good ground wave for local work. The horizontal portion contributes to high-angle sky-wave transmission, which is useful for covering short distances on this band at night.

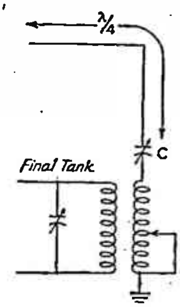


Fig. 1052 — Typical grounded antenna for 1.75 Mc., consisting of a vertical section and a horizontal section having a total length (including the ground lead, if the latter is more than a few feet long) of one-quarter wavelength. Coil *L* should have about 20 turns of No. 12 wire on a 3-inch diameter form, tapped every two or three turns for adjustment. *C* is a 250 to 500 $\mu\mu$fd. variable. The inductive coupling between *L* and the final tank coil should be variable.

Fig. 1052 shows a grounded antenna with the top folded to make the length equal to a quarter wave. The antenna coupling apparatus consists of the coil, *L*, tuned by the series condenser, *C*, with *L* inductively coupled to the transmitter tank circuit (§ 10-4, 10-6).

For computation purposes, the *over-all* length of a grounded system is given by

$$\text{Length (feet)} = \frac{236}{f\ (\text{Mc.})}$$

This is the *total* length from the far end of the antenna to the ground connection. The length is not critical, since departures of the order of 10 to 20 per cent can be compensated by the tuning apparatus.

The ground should preferably be one with conductors buried deep enough to reach natural moisture. In urban locations, good grounds can be made by connecting to the water mains where they enter the house; the pipe should be scraped clean and a low-resistance connection made with a tightly fastened ground clamp. If no water supply pipes are available, several rods or pipes six to eight feet long may be driven into the ground at intervals of six or eight feet, all being connected together. The transmitter should be located so as to make the ground lead as short as possible.

In locations where it is impossible to secure a good ground connection, because of sandy soil or other considerations, it is preferable to use a counterpoise or capacity ground instead of an actual ground connection. The counterpoise consists of a system of wires, insulated from ground and running horizontally above the earth beneath the antenna. The counterpoise should have a sufficient number of wires of sufficient length to cover well the area immediately under the antenna. The wires may be formed into any convenient shape; i.e., they may be spread out fan-shape, in a radial pattern, or as three or more parallel wires separated a few feet and running beneath the antenna. The counterpoise may be elevated six feet or so above the ground, so that it will not interfere with persons walking under it. A low-resistance connection should be made between the usual ground terminal of the transmitter and each of the wires in the counterpoise.

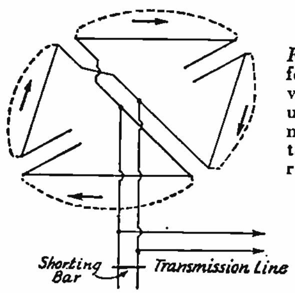


Fig. 1053—The Alford loop antenna for v.h.f. and u.h.f. is made up of resonant elements fed in phase rotation, and has high radiation efficiency.

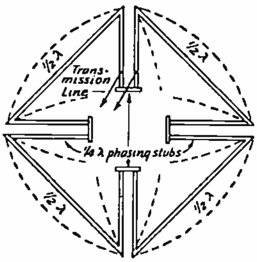


Fig. 1054—Various feed and phasing arrangements may be used with v.h.f. loops. The shorted ends of the closed quarter-wave matching stubs may be grounded to a metal mast or other support.

Loaded antennas—Methods of securing maximum usable radiation from a grounded vertical antenna of limited height utilize loading coils and capacity tops. The latter may be in the form of a ring or spider or a top-mounted outrigger. Capacity effect raises the maximum current point nearer the top of the antenna.

Another form of top loading which involves the insertion of an inductance coil near the top, enclosed within a shield can for protection and to increase the top capacity, is particularly suited to mobile installations.

The advantage of top loading in short vertical antennas is that it forces the upper portion of the antenna to carry a more substantial current, making the effective height approach more closely to the actual physical height.

V.h.f. loop antennas—Although the radiation resistance of an ordinary loop transmitting antenna is very low, at the very-high frequencies, the Alford loop shown in Fig. 1053 permits the use of resonant dimensions of the order of ⅛ to ¼ wavelength on each side, resulting in relatively high radiation efficiency as compared with ordinary loop antennas for the lower frequencies.

Various configurations and feed methods are possible, following this general pattern. In the form shown in Fig. 1054, the sides of the loop are half-wave resonant sections linked by half-wave transmission-line matching stubs so arranged that there is a current loop at the center of each side, with the currents in the various sections all in phase rotation. Since the shorted ends of the quarter-wave stubs are at a current node, the system may be directly attached at these points to a grounded metal tower or similar structure.

Center-fed dipoles with low impedance coaxial lines or delta-matched lines may be used, the correct phasing for each line being arranged at the feed-line terminals.

"J" antenna—This type of antenna, frequently used on the very-high frequencies when vertical polarization is desired, is simply a half-wave radiator fed through a quarter-wave matching section (§ 10-8), the whole being mounted vertically as shown in Fig. 1055. Adjustment and tuning are as described in § 10-8. The bottom of the matching section, being at practically zero r.f. potential, can be grounded for lightning protection.

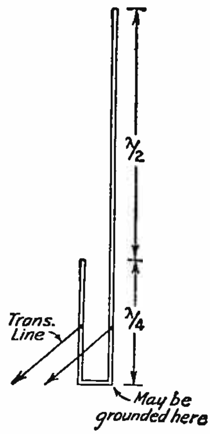


Fig. 1055—The "J" antenna, usually constructed of hard-drawn metal tubing. The ¾-wave vertical section may be mounted as an extension of a grounded metal mast. The matching stub may be adjusted by a sliding shorting bar.

Coaxial antenna—With the "J" antenna radiation from the matching section and the transmission line tends to combine with the radiation from the antenna in such a way as to raise the angle of radiation. At v.h.f. the lowest possible radiation angle is essential, and the coaxial antenna shown in Fig. 1056 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric transmission line is extended one quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is non-resonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna (§ 10-2).

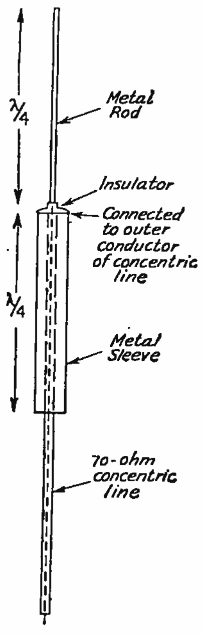


Fig. 1056—Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems in its performance at low radiation angles.

Even with coaxial antennas radiation by the line cannot always be entirely eliminated, particularly in the ultrahigh-frequency region. With other varieties of coaxial-line-fed systems that even slight asymmetry in systems of such dimensions as to be capable of resonant operation is exceedingly difficult to avoid. The

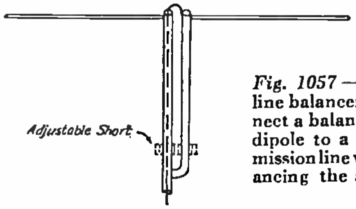


Fig. 1057 — A "bazooka" line balancer, used to connect a balanced center-fed dipole to a coaxial transmission line without unbalancing the antenna load.

only sure solution is to detune the line completely. This can best be achieved by the use of a quarter-wave detuning sleeve, similar to the phase reversing or coupling sleeve of the coaxial antenna except that it is located one-half wavelength down the line and is connected with the open end at the top. Such a sleeve corresponds to a matching stub (§ 10-8) as used with parallel-wire lines, and the length must be critically adjusted for maximum efficiency.

Turnstile antenna — The turnstile antenna consists of two half-wave radiators crossing each other at right angles and excited 90 degrees out of phase. A number of these are sometimes arranged in an array in which the individual turnstiles in a horizontal plane are spaced one above another at half-wave intervals. Such an array gives nearly uniform radiation in all horizontal directions together with directivity in a vertical plane.

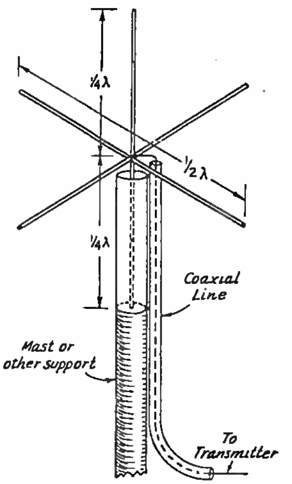


Fig. 1058 — The "ground-plane" antenna gives low vertical-angle radiation with a circular horizontal pattern. The quarter-wave mounting section of large-diameter tubing may be mounted on a metal mast or other support.

Wide-band Antennas

Cylindrical antennas—Low-impedance radiators such as are used by television and broadband f.m. stations are of interest in amateur v.h.f. operation because they function at high efficiency without adjustment throughout the width of an amateur band.

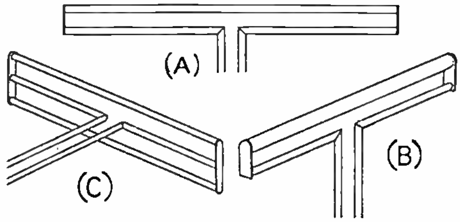


Fig. 1059 — Folded dipoles are an elementary form of broad-band antenna, simply constructed and easily fed.

At the very-high frequencies an ordinary dipole or equivalent antenna made of small wire is purely resistive only over a very small frequency range. Its Q, and therefore its selectivity, is sufficient to limit its optimum performance to a narrow frequency range and readjustment of the length or tuning is required for each narrow slice of the spectrum. With tuned transmission lines the effective length of the antenna can be shifted by retuning the system. In the case of antennas fed by matched-

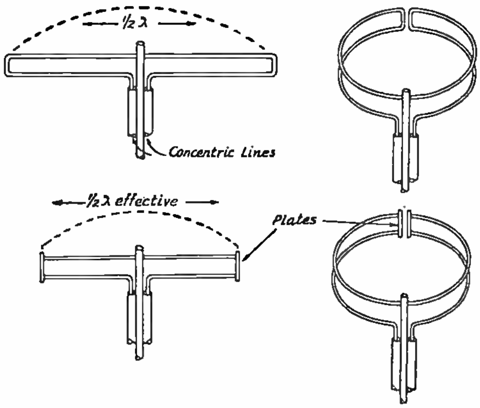


Fig. 1060 — The compact circular loop antenna is derived from a folded dipole by bending it into a circle. To reduce length and concentrate current distribution, capacity-loading end plates are added as in lower view. Circular loops may be stacked at one-half wavelength intervals for increased vertical directivity. Circular arrays of three or four folded dipoles, bent into 120° or 90° arcs and fed in phase rotation, are extensively used for u.h.f.

impedance lines, any appreciable frequency change requires an actual mechanical adjustment of the system. Otherwise the resulting mismatch with the line will be sufficient to cause significant loss.

A properly designed and constructed wide-band antenna, on the other hand, will exhibit very nearly constant input impedance over a range of several megacycles.

The simplest method of obtaining a broad-band characteristic is the use of what is termed a "cylindrical" antenna. This is no more than a conventional doublet in which large-diameter tubing is used for the elements. The ratio of diameter to length being large, in comparison to the length. This raises the input impedance (§ 10-2) and simultaneously lowers the Q, thus broadening the resonance characteristic.

The broad-band characteristic is achieved by increasing the C/L ratio. This lowers the effective Q of the antenna by reducing the characteristic impedance, the radiation resistance remaining approximately the same.

Folded dipoles — A system combining the radiation characteristics of a half-wave antenna with the impedance-transforming properties of a quarter-wave line (§ 10-5) is shown in Fig. 1059-A. Essentially, it is a center-fed half-wave antenna with another half-wave element connected directly between its ends. The spacing between the two sections should be quite close — not more than a few per cent of the wavelength. As used at very-high frequencies, the spacing is of the order of an inch to two when the elements are constructed of metal tubing.

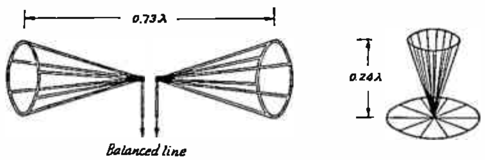


Fig. 1061 — Conical broad-band antennas have relatively constant impedance over a wide frequency range. The three-quarter wavelength dipole at left and the quarter-wave vertical with ground plane at right have the same input impedance — approximately 65 ohms. Sheet-metal or spine-type construction may be used.

The total required length around the loop may be calculated by Equation 10 (§ 10-9) for a total length of one wavelength.

The impedance at the terminals of the dipole is four times that of a half-wave antenna, or nearly 300 ohms, when the antenna conductors are both the same diameter. A 300-ohm line will therefore be nonresonant when the antenna is connected to its output end (§ 10-5), while the standing-wave ratio with a 600-ohm line will be only of the order of 2 to 1.

An exact match with a 600-ohm line can be obtained by either of two modifications. One is to double the size of one of the elements, as shown in Fig. 1059-B; the other is to add an additional element in parallel, as in Fig. 1059-C.

Cone antennas — From the cylindrical antenna various specialized forms of broadly resonant radiators have been evolved, including the ellipsoid, spheroid, cone, diamond and double diamond. Of these, the conical antenna is perhaps the most interesting. With large angles of revolution, ψ, the characteristic impedance can be reduced to a very low value suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines, as in Fig. 1061.

Plane Reflector Antennas

Plane-sheet reflectors — The small physical size in terms of linear dimensions of v.h.f. antennas makes practical many methods not feasible on lower frequencies. For example, a plane flat-sheet reflector may be used with a half-wave dipole, obtaining gains of 5 to 7 db. Much higher gains are attainable with a number of stacked dipoles, spaced ¼ or ¾ wavelength apart, and a larger reflecting sheet; such an arrangement is called a "billboard" array.

Plane reflectors need not be constructed of solid sheets. Wire mesh or a grid of a closely spaced parallel wire spines are not only more easily erected but offer less wind resistance.

Parabolic reflectors — Sheets formed into the shape of a section of a parabolic cylinder are used with a driven radiator situated at the focus as highly directive antenna systems. If the parabolic reflector is sufficiently large so that the distance to the focal point is a number of wavelengths, optical conditions are approached and the wave across the mouth is a plane wave. However, if the reflector is of the same order of dimensions as the operating wavelength, or less, the driven radiator is appreciably coupled to the reflecting sheet and minor lobes occur in the pattern.

Plane sheets shaped to a parabolic curve are used to obtain high directivity in a single plane. With apertures of the order of 10 or 20 wavelengths, a beam width of 5° may be achieved.

A reflecting paraboloid must be carefully designed and constructed to obtain ideal performance. The antenna must be located at the focal point. The most desirable focal distance is that which places the radiator along the plane of the mouth of the parabola, corresponding to one-half its radius. At other focal distances interference fields may deform the pattern or cancel a portion of the radiation.

Corner reflector antenna — The "corner" reflector consisting of two flat conducting sheets which intersect at a designated angle.

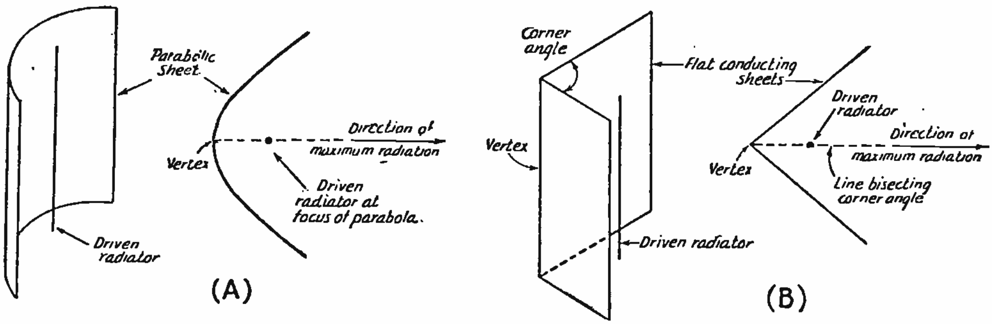


Fig. 1062 — Plane-sheet reflectors for v.h.f. and u.h.f. A shows a parabolic sheet and B a square-corner reflector.

TABLE V

Frequency Band	Length of Side	Length of Reflector Elements	Number of Reflector Elements	Spacing of Reflector Elements	Spacing of Driven Dipole to Vertex
224–230 Mc. (1¼ meter)	4′2″	4′7″	20	5″	2′ 2″
112–116 Mc. (2½ meter)	8′4″	5′2″	20	10″	4′ 4″
112–116 Mc.* (2½ meter)	6′8″	5′2″	16	10″	3′ 6″
56–60 Mc. (5 meter)	16′8″	10′4″	20	1′ 8″	8′ 8″
56–60 Mc.* (5 meter)	13′4″	10′4″	16	1′ 8″	6′11″

Dimensions of square-corner reflector for the 224-, 112-, and 56-Mc. bands. Alternative designs are listed for the 112- and 56-Mc. bands. These designs, marked (*), have fewer reflector elements and shorter sides, but the effectiveness is only slightly reduced. There is no reflector element at the vertex in any of the designs.

The corner reflector antenna is particularly useful at v.h.f. where structures one or two wavelengths in maximum dimensions are more practical to build than larger systems.

The plane surfaces are set at an angle of 90°, with the antenna set on a line bisecting this angle. For maximum performance, the distance of the antenna from the vertex should be 0.5 wavelength, but compromise designs can be built with closer spacings (see Table V). The plane surfaces need not be solid sheets; spines spaced about 0.1 wavelength apart will serve as well. The spines do not have to be connected together electrically.

If the driven radiator is situated on a line bisecting the corner angle, as shown in Fig. 1062, maximum radiation is in the direction of this line. There is no focus point for the driven radiator, as with a parabolic reflector, and the radiator can be placed at a variety of positions along the bisecting line.

Corner angles larger than 90° can be used, with some decrease in gain. A 180° "corner" is equivalent to a single flat-sheet reflector. With angles smaller than 90°, the gain theoretically increases as the corner angle is decreased. However, to realize this gain the size of the reflecting sheets must also be increased.

At a spacing of 0.5 wavelength from the driven dipole to the vertex, the radiation resistance of the driven dipole is approximately twice the radiation resistance of the same dipole in free space. Smaller spacings of driven dipole and vertex are practical, but at a slight sacrifice in efficiency. The alternative design for the 112- and 56-Mc. square-corner reflector in Table I has a dipole-to-vertex spacing of 0.4 wavelength. At this spacing the driven dipole radiation resistance is still somewhat higher than its free space value, but is considerably less than when the spacing is 0.5 wavelength.

Horn radiators — On the ultrahigh frequencies a metal horn can be used to guide and concentrate the wave in a sharp beam. Highest directivity is secured when the mouth of the horn has a dimension large compared with the wavelength. Factors governing the gain include flare angle, length and mouth diameter.

Various types of horn radiators include the simple *sectoral* horn, flared linearly in only one dimension; the *pyramidal* horn, flared in two dimensions; the *conical* horn, a section of a cylindrical cone whose apex is terminated in a cylindrical wave guide or cylindrical coupling section; and the *biconical* horn, consisting of two cones joined back to back at the apex.

Receiving antennas — Nearly all of the properties possessed by an antenna as a radiator also apply when it is used for reception. Current and voltage distribution, impedance, resistance and directional characteristics are the same in a receiving antenna as if it were used as a transmitting antenna. This reciprocal behavior makes possible the design of a receiving antenna of optimum performance based on the same considerations as have been discussed for transmitting antennas.

The simplest receiving antenna is a wire of random length. The longer the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, a large antenna is not necessary for picking up signals at good strength. An indoor wire only 15 to 20 feet long will serve; although a longer wire outdoors is better.

The use of a tuned antenna improves the operation of the receiver, however, because the signal strength is raised more in proportion to the stray noises picked up than is the case with wires of random length. Since the transmitting antenna usually is given the best location, it can also be expected to serve best for receiving. This is especially true when a directive antenna is used, since the directional effects and power gain of directive transmitting antennas are the same for receiving as for transmitting. A change-over switch or relay, connected in the antenna leads, can be used to transfer the connections from the receiver to the transmitter.

In selecting a directional receiving antenna it is preferable to choose a type which gives very little response in all but the desired direction (small minor lobes). This is even more important than high gain in the desired direction, because the cumulative response to noise and unwanted-signal interference in the smaller lobes may offset the advantage of increased desired-signal gain.

The auxiliary elements in the antenna system — transmission line and receiver coupling — should be arranged to avoid direct pick-up of undesired-signal or noise energy, which would impair the directivity. This requires a balanced transmission line without standing waves, a carefully shielded and balanced receiver input circuit, and a good ground on the receiver.

Chapter Eleven

Construction Practice

In contrast to the earlier days of amateur radio, when many components were obtainable only at prohibitive prices or not at all, the construction of a piece of equipment these days resolves itself chiefly into the proper assembly and wiring of manufactured components from the wide assortment available.

Tools

While an easier, and perhaps a better, job can be done with a greater variety of tools available, by taking a little thought and care it is possible to turn out a fine piece of equipment with only a few of the common hand tools. A list of tools which will be found indispensable in the construction of radio equipment will be found on this page. With these tools it should be possible to perform any of the required operations in preparing panels and metal chassis for assembly and wiring. A few additional tools will make certain operations easier, so it is a good idea for the amateur who does constructional work at intervals to add to his supply of tools from time to time. The following list will be found helpful in making a selection:

- Bench vise, 4-inch jaws.
- Tin shears, 10-inch, for cutting thin sheet metal.
- Taper reamer, ½-inch, for enlarging small holes.
- Taper reamer, 1-inch, for enlarging holes.
- Countersink for brace.
- Carpenter's plane, 8- to 12-inch, for woodworking.
- Carpenter's saw, cross-cut.
- Motor-driven emery wheel for grinding.
- Long-shank screwdriver with screw-holding clip for tight places.
- Set of "Spintite" socket wrenches for hex nuts.
- Set of small, flat, open-end wrenches for hex nuts.
- Wood chisel, ½-inch.
- Cold chisel, ½-inch.
- Wing dividers, 8-inch, for scribing circles.
- Set of machine-screw taps and dies.
- Folding rule, 6-foot.
- Dusting brush.

Several of the pieces of light woodworking machinery, often sold in hardware stores and mail-order retail stores, are ideal for amateur radio work, especially the drill press, grinding head, band and circular saws, and joiner. Although not essential, they are desirable for anyone in a position to acquire them.

Care of Tools

The proper care of tools is not alone a matter of pride to a good workman. He also realizes the energy which may be saved and the annoyance which may be avoided by the possession of well-kept, sharp-edged tools. A few minutes spent with the oil stone or emery wheel now and then will maintain the fine cutting edges of knives, drills, chisels, etc.

Drills should be sharpened at frequent intervals so that grinding is kept at a minimum each time. This makes it easier to maintain the rather critical surface angles required for best cutting with least wear. Occasional oil-stoning of the cutting edges of a drill or reamer will extend the time between grindings. Stoned cutting edges also will stand more feed and speed.

The soldering iron can be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale which may have accumulated. An oxidized tip may be cleaned by dipping it in sal ammoniac while hot and then wiping it clean with a rag. If the tip becomes pitted, it should be filed until smooth and bright, and then tinned by dipping it in solder.

All tools should be wiped occasionally with an oily cloth to prevent rust.

INDISPENSABLE TOOLS

- Long-nose pliers, 6-inch.
- Diagonal cutting pliers, 6-inch.
- Screwdriver, 6- to 7-inch, ¼-inch blade.
- Screwdriver, 4- to 5-inch, ⅛-inch blade.
- Scratch awl or scriber for marking lines.
- Combination square, 12-inch, for laying out work.
- Hand drill, ¼-inch chuck or larger, 2-speed type preferable.
- Electric soldering iron, 100 watts.
- Hacksaw, 12-inch blades.
- Center punch for marking hole centers.
- Hammer, ball peen, 1-lb. head.
- Heavy knife.
- Yardstick or other straight-edge.
- Carpenter's brace with adjustable hole cutter or socket-hole punches (see text).
- Pair of small C-clamps for holding work.
- Large, coarse, flat file.
- Large round or rat-tail file, ½-inch diameter.
- Three or four small and medium files—flat, round, half-round, triangular.
- Drills, particularly ¼-inch and Nos. 18, 28, 33, 42 and 50.
- Combination oil stone for sharpening tools.
- Solder and soldering paste (non-corroding).
- Medium-weight machine oil.

Useful Materials

Small stocks of various miscellaneous materials will be required in constructing radio apparatus, most of which are available from hardware or radio supply stores. A representative list follows:

- ½ × 1/16-inch brass strip for brackets, etc. (half-hard for bending).
- ¼-inch square brass rod or ½ × ½ × 1/16-inch angle brass for corner joints.
- ¼-inch diameter round brass rod for shaft extensions.
- Machine screws: Round-head and flat-head, with nuts to fit. Most useful sizes: 4-36, 6-32 and 8-32, in lengths from ¼-inch to 1½-inch. (Nickel-plated iron will be found satisfactory except in strong r.f. fields, where brass should be used.)
- Bakelite and hard rubber scraps.
- Soldering lugs, panel bearings, rubber grommets, terminal-lug wiring strips, varnished-cambric insulating tubing.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

Construction Planning

The construction of any piece of radio equipment requires careful planning, proper coördination of parts, circuit and layout to achieve the desired result.

Equipment can be divided into three main classifications — experimental, temporary and permanent. Each class has its own peculiarities and limitations affecting design and constructional details.

Experimental equipment, such as the gear thrown together to investigate the possibilities of some newly published circuit, or an original idea, requires a simpler approach and less work than a unit to be used in the regular station. Experimental equipment may be built "breadboard" style on a board faced with a thin sheet of metal for grounding purposes, or even on an old chassis from the junk box. If the chassis has been previously used the old socket and screw holes may save time and effort. Random parts and a semi-makeshift arrangement can be used. Plenty of space for changes in wiring and components must be available. While temporary equipment such as a power supply built in an emergency, to replace a defective transmitter bias supply, does not require the same amount of planning and care in assembly as did the original bias supply, it should be attached firmly to the chassis and wired securely to prevent breakdowns. Connections should be soldered and safety precautions taken since it is difficult to anticipate the exact use or required length of service life of this type of equipment.

Permanent equipment requires the most careful planning and assembly since it must necessarily fit in with other units. Permanent equipment consists of three main classes — fixed station, mobile and portable.

In fixed station usage, several types of construction are available. For example, take the case of a proposed exciter power supply. Will this unit be made a permanent part of the exciter but not located adjacent to it; will it be removed and used as a source of power for some other equipment such as an experimental amplifier; or will it be constructed as an integral part of the exciter. The type of construction chosen for any given unit must depend on the foreseeable uses that will be made of it. Thus, in the case of the exciter power supply, if it is to be used with but not attached to the exciter, it should be packaged so it can be moved and connected to other equipment. For maximum utility, both screw-type terminals and plug- and socket-type connections should be available. If it is desirable to use the supply in the field such as on Field Day, it must be more sturdily built and should be provided with a protective cabinet or box.

If the exciter supply is made a permanent part of the exciter, its design must be coördinated with the exciter unit as a whole, a chassis of suitable size and form must be selected and a layout made to fit all the components into the available space.

In fixed station applications, assemblies of small units built to conform with the available space may prove to have more convenience and utility than large masses of assembled parts, such as a cabinet type, in which it is extremely difficult to replace defective parts or to make changes readily. This type of equipment includes power supplies, volt-ohmmeter units, audio amplifiers and any type equipment which may have more than one use in or around the station. For example, if the speech amplifier were designed to be removed readily it could perform double duty as a public address amplifier and used for a club "jam" session. This would be feasible, however, only if provisions were made for quick and easy removal, and connection to the normal gear. Such an amplifier should be built self-contained, with power supply, and terminated with plug-in type connectors to fit both the phonograph pick-up and the phone rig connections. All multi-purpose equipment must be built solidly, readily demountable and with some system of universal connections.

The desirable features of portable equipment combine those of fixed and mobile station apparatus plus lightness and compactness. Portables are usually packaged in at least two units, one containing the transmitter-receiver (or transceiver) and the other, the power supply or source.

Specialized Construction Technique

Mobile equipment must be laid out and assembled to prevent damage due to vibration and shock. In addition to the standard good practices of construction, mobile equipment requires additional care in the mounting of components, the placement of parts to prevent detrimental heating effects and in the arrangement of the wiring. Heavy leads should be pre-formed to fit between the connecting points in order to prevent mechanical strain on the components. Fixed resistors and condensers should be fastened at both ends, clipping the wire leads short and attaching them to terminal strips or blocks, and large units should also be fastened at the center. Transformers should be securely bolted to the chassis, using bolts that fill the mounting holes in the transformer. Chassis should be solidly constructed of heavy metal. Ordinary chassis spot-welded in the corners will not be satisfactory for mobile sets. The chassis should be of the type in which the corners are bent over and securely riveted, then welded, and should have a lip at the bottom for rigidity. Cast chassis are usually excellent for mobile units. Cross-bracing of a chassis will strengthen it. Coils should be wound on rigid low-loss forms and securely mounted. For example, the output tank coil of a 56 Mc. transmitter can be wound on a solid grooved dielectric rod, which is then mounted vertically on the chassis, adjacent to the plate tank condenser, using one large size brass machine screw. The antenna coil can be also wound on the same rod.

Detuning and loss of efficiency would result if, for example, the same coils were mounted directly on the terminals of their respective variable condensers without having a solid support and mounting.

Ground connections must not be spot-soldered to the chassis. Instead, they should be made to ground lugs or straps provided for that purpose. Lockwashers or locknuts must be used on all screws. Stranded hook-up wire, laced into cables and securely fastened down, has been found highly desirable in mobile sets.

The use of tube locks is almost imperative on any tubes except the smaller metal type (such as 6C5, 6H6, etc.). Certain common types of ceramic sockets require tube locks for all tubes. Fiber wafer sockets should be avoided because of their lack of mechanical strength and holding ability.

Mobile installations are affected by shock and vibration and every effort must be taken to prevent mechanical or electrical damage and to prevent parts from shaking loose. Special components such as variable condensers, coils and transformers are available for such use and should be included in this type of equipment.

In the construction of v.h.f. equipment many familiar practices must be discarded or modified. Actual physical relationships between components becomes extremely important since every inch of wire constitutes a tuned circuit and every condenser is also an inductance. Stray capacitance and inductance may lead to a loss of gain or sensitivity and may cause detuning and instability. Grounds must be grouped and connected to definite points rather than indiscriminately to the chassis. Special by-pass technique is required since the condensers ordinarily employed for that purpose will not function in the same manner as on the lower frequencies. For example, the following table shows the approximate value of the usual postage-stamp condenser capacity which, together with the inductance of the leads, will be approximately self-resonant in the amateur bands shown. At signal frequencies, no greater by-pass capacity should be used (for an indicated lead length) than the one shown for the highest frequency to be covered.

Maximum Frequency	*Lead Length (total)* ¼″	½″	1″	2″
28–30 mc. ..	3000μμfd.	2000μμfd.	1000μμfd.	500μμfd.
56–60.....	750	500	350	150
112–116.....	200	150	75	40
224–230.....	80	40	20	10
448–........	15	8	4	—

Symmetry of push-pull circuits is essential in v.h.f., both from an electrical and a mechanical viewpoint.

Copper straps may be utilized for connections in place of straight copper wire, which has considerable inductance at the higher frequencies.

More effective by-pass condensers for v.h.f. may be made by attaching a square inch or so of flat copper or brass strip, insulated by mica or polystyrene, to the chassis, immediately adjacent to the connection to be by-passed.

Allowance must be made for the capacity and inductance of components, such as tube elements, leads, chassis, and metal shielding.

All joints must be soldered, using plenty of heat and care to ensure a good sweated joint.

Particular care must be taken to provide adequate conductor size where large r.f. currents are present such as in tuned lines.

All vibration and movement of components must be completely eliminated since the slightest change in capacity will affect the critical circuits. The elements in resonant cavities must be rigidly fastened.

Wire-wound resistors must be avoided in any circuit where r.f. is present. Carbon resistor values will not be reliable as the frequency is increased and the metallized filament type resistor must be used where critical values are required.

Since the successful performance of v.h.f. equipment largely depends of the absence of stray and undesired capacities and inductance, extreme care must be taken in the mechanical as well as the electrical organization of the chassis. For example, the correct rotation of a socket may shorten an important tube lead as much as an inch — which may have enough inductance to resonate with the tube element capacity at 500 Mc.

Chassis Construction

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is no more of a chore than building with wood, and a more satisfactory job results.

The placing of components on the chassis is shown quite clearly in the photographs in this *Handbook.* Aside from certain essential dimensions, which usually are given in the text, exact duplication is not necessary.

Much trouble and energy can be saved by spending sufficient time in planning the job. When all details are worked out beforehand the actual construction is greatly simplified.

Cover the top of the chassis with a piece of wrapping paper or, preferably, cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Then assemble the parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath, so that interferences in mounting may be avoided. Place condensers and other parts with shafts extending through the panel first, and arrange them so that the controls will form the desired pattern on the panel. Be sure to line up the shafts squarely with the chassis front. Locate any partition shields and panel brackets next, and then the tube sockets and any other parts, marking the mounting-hole centers of each accurately on the paper. Watch out for condensers whose shafts do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i.f. transformers, etc., as well as holes for wiring leads.

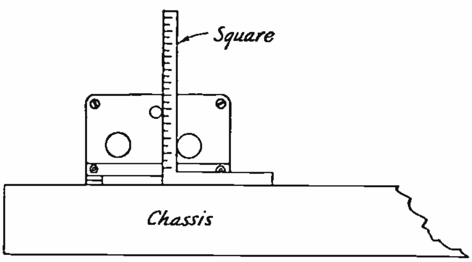


Fig. 1101 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line, the panel being fastened on temporarily. The hole centers may then be punched in the chassis with the center punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, making sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

Next, mount on the chassis the condensers and any other parts with shafts extending to the panel, and measure accurately the height of the center of each shaft above the chassis, as illustrated in Fig. 1101. The horizontal displacement of shafts having already been marked on the chassis line on the panel, the vertical displacement can be measured from this line. The shaft centers may now be marked on the back of the panel, and the holes drilled. Holes for any other panel equipment coming above the chassis line may then be marked and drilled, and the remainder of the apparatus mounted.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hacksaw, it may be marked with scratches as deep as possible along the line of the cut on both sides of the sheet and then clamped in a vise and worked back and forth until the sheet breaks at the line. Do not carry the bending so far that the break begins to weaken; otherwise the edge of the sheet may become bent. A pair of iron bars or pieces of heavy angle stock, as long or longer than the width of the sheet, to hold it in the vise will make the job easier. C-clamps may be used to keep the bars from spreading at the ends. The rough edges may be smoothed up with a file or by placing a large piece of emery cloth or sandpaper on a flat surface and running the edge of the metal back and forth over the sheet.

Bends may be made similarly. The sheet should be scratched on both sides, but not so deeply as to cause it to break.

Drilling and Cutting Holes

When drilling holes in metal with a hand drill it is important that the centers first be located with a center punch, so that the drill point will not "walk" away from the center when starting the hole. Care should be taken not to use too much pressure with small drills, which bend or break easily. Whenever the drill starts to break through, special care must be used. Often it is an advantage to shift a two-speed drill to low gear at this point. Holes more than ¼-inch in diameter may be started with a smaller drill and reamed out with the larger drill.

The chuck on the usual type of hand drill is limited to ¼-inch drills. Although it is rather tedious, the ¼-inch hole may be filed out to larger diameters with round files. Another method possible with limited tools is to drill a series of small holes with the hand drill along the inside of the diameter of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit into the carpenter's brace will make the job easier. A large rat-tail file clamped in the brace makes a very good reamer for holes up to the diameter of the file, if the file is revolved counter-clockwise.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. The cutter should be kept well-sharpened. Occasional application of machine oil in the cutting groove will help. The cutter first should be tried out on a block of wood, to make sure that it is set for the correct diameter. Probably the most convenient device for cutting socket holes is the socket-hole punch. The best type is that which works by turning a take-up screw with a wrench.

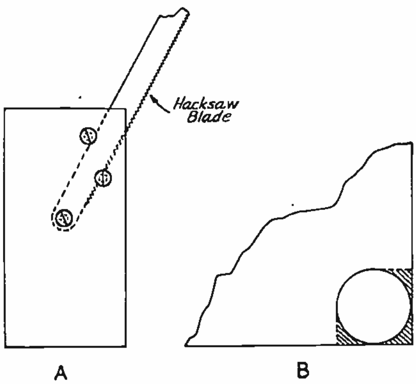


Fig. 1102 — To cut rectangular holes in a chassis, corner holes may be filed out as shown in the shaded portion of B, making it possible to start the hacksaw blade along the cutting line. A shows how a single-ended handle may be constructed for a hacksaw blade.

Square or rectangular holes may be cut out by making a row of small holes as previously described, but is more easily done by drilling a ½-inch hole inside each corner, as illustrated in Fig. 1102, and using these holes for starting and turning the hacksaw. The socket-hole punch also may be of considerable assistance in cutting out large rectangular openings.

The burrs or rough edges which usually result after drilling or cutting holes may be removed with a file, or sometimes more conveniently with a sharp knife or chisel. It is a good idea to keep an old wood chisel sharpened and available for this purpose. A burr reamer will also be useful.

ℂ Crackle Finish

Wood or metal parts can be given a crackle finish by applying one coat of clear Duco or Tri-Seal and allowing it to dry over night. A coat of Kem-Art Metal Finish is then sprayed or applied thickly with a brush, taking care that the brush marks do not show. This should be allowed to dry for two or three hours and the part should then be baked in the kitchen oven at 225° for one and one-half hours. This will produce a regular commercial job. This finish, which comes in several different colors, is made by Sherwin-Williams Paint Co.

NUMBERED DRILL SIZES

Number	*Diameter (mils)*	*Will Clear Screw*	*Drilled for Tapping Iron, Steel or Brass**
1	228.0	—	—
2	221.0	12–24	—
3	213.0	—	14–24
4	209.0	12–20	—
5	205.0	—	—
6	204.0	—	—
7	201.0	—	—
8	199.0	—	—
9	196.0	—	—
10	193.5	10–32	—
11	191.0	10–24	—
12	189.0	—	—
13	185.0	—	—
14	182.0	—	—
15	180.0	—	—
16	177.0	—	12–24
17	173.0	—	—
18	**169.5**	**8–32**	—
19	166.0	—	12–20
20	161.0	—	—
21	159.0	—	10–32
22	157.0	—	—
23	154.0	—	—
24	152.0	—	—
25	149.5	—	10–24
26	147.0	—	—
27	144.0	—	—
28	**140.0**	**6–32**	—
29	**136.0**	—	**8–32**
30	128.5	—	—
31	120.0	—	—
32	116.0	—	—
33	**113.0**	**4–36 4–40**	—
34	111.0	—	—
35	**110.0**	—	**6–32**
36	106.5	—	—
37	104.0	—	—
38	101.5	—	—
39	099.5	3–48	—
40	098.0	—	—
41	096.0	—	—
42	**093.5**	—	**4–36 4–40**
43	089.0	2–56	—
44	086.0	—	—
45	082.0	—	3–48
46	081.0	—	—
47	078.5	—	—
48	076.0	—	—
49	073.0	—	2–46
50	**070.0**	—	—
51	067.0	—	—
52	063.5	—	—
53	059.5	—	—
54	055.0	—	—

*Use one size larger for tapping bakelite and hard rubber.

ℂ Twist Drills

Twist drills are made of either high-speed steel or carbon steel. The latter type are more common and will usually be supplied unless specific request is made for high speed drills. The carbon drill will suffice for most ordinary equipment construction work and costs less than the high speed type.

While twist drills are available in a number of sizes those listed in bold-faced type above will be the drills most commonly used in construction of amateur radio equipment. It is usually desirable to purchase several of each of the commonly used sizes rather than a quantity of odd sizes, most of which will be used infrequently, if at all.

Cutting Threads

Brass rod may be threaded, or the damaged threads of a screw repaired, by the use of *dies.* Holes of suitable size (see drill chart) may be threaded for screws by means of *taps.* Taps and dies are obtainable in all standard machine-screw sizes. A set usually consists of taps and dies for 4–36, 6–32, 8–32, 10–32 and 14–20 sizes, with a holder suitable for use with either tap or die. The die may be started easily by first filing a sharp taper or bevel on the end of the rod. In tapping a hole, extreme care should be used to prevent breaking the tap. The tap should be kept at right angles to the surface of the material, and rotation should be reversed a revolution or two whenever the tap begins to turn hard. With care, holes can be tapped rapidly by clamping the tap in the chuck of the hand drill and using slow speed. Machine oil applied to the tap usually makes cutting easier and sticking less troublesome.

Cleaning and Finishing Metal

Parts made of aluminum can be cleaned up and given a satin finish, after all holes have been drilled, by placing them in a solution of lye for one-half to three-quarters of an hour. Three or four tablespoonfuls of lye should be used to each gallon of water. If more than one piece is treated in the same bath, each piece should be separated from the others so as to expose all surfaces to the solution. Overlapping of pieces may result in spots or stains.

Hook-Up Wire

A popular type of wire for receivers and low-power transmitters is that known as "push-back" wire. It comes in sizes No. 16, 18, 20, etc., which is sufficiently large for all power circuits except filament. The insulating covering, which is sufficient for circuits where voltages do not exceed 400 or 500, can be pushed back a few inches at the end, making cutting of the insulation unnecessary when making a connection. Filament wiring should be done with sufficiently large conductors to carry the required current without appreciable voltage drop (see Copper Wire Table in the Appendix). Rubber-covered house-wire sizes No. 14 to No. 10 are suitable for heavy-current transmitting tubes, while No. 18 to No. 14 flexible wire is satisfactory for receivers and low-drain transmitting tubes where the total length of the leads is not excessive.

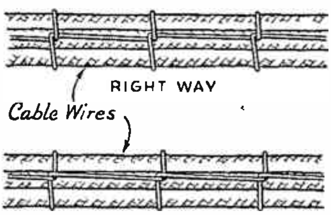


Fig. 1103 — Right and wrong methods of lacing cable. With the right way the leading line is pinched under each turn and will not loosen if a break occurs in the lacing.

Stiff bare wire, sometimes called *bus wire* or *bus bar,* is most favored for the high r.f.-potential wiring of transmitters and, where practicable, in receivers. It comes in sizes No. 14 and No. 12 and is usually tin-dipped. Soft-drawn antenna wire also may be used. Kinks or bends can be removed by stretching 10 or 15 feet of the wire and then cutting it into small usable lengths.

The insulation covering power wiring which is to carry high transmitter voltages should be appropriate for the voltage involved. Wire with rubber and varnished cambric covering, similar to ignition cable, is available from radio parts dealers. The smaller sizes have sufficient insulation to be safe at 1000 to 1500 volts, while the more heavily insulated types should be used for voltages above 1500.

Wiring Transmitters and Receivers

It is usually advisable to do the power-supply wiring first. The leads should be bunched together as much as possible and kept down close to the surface of the chassis. The lacing of power wiring in cable form not only improves its appearance but also strengthens the wiring. Fig. 1103 shows the correct way of lacing cabled wires. When done correctly the leading line is held tightly pinched in place after tension has been removed, and therefore does not loosen readily. When the wrong method is used the turns will loosen up as soon as tension is removed.

Chassis holes for wires should be lined with *rubber grommets* which fit the hole, to prevent chafing of the insulation. In cases where power-supply leads have several branches, it is often convenient to use fibre *terminal strips* as anchorages. These strips also form handy mountings for wire-terminal resistors, etc. When any particular unit is provided with a nut or thumb-screw terminal, soldering-lug wire terminals to fit are useful.

High-potential r.f. wiring should be well spaced from the chassis or other grounded metal surfaces and should be run as directly as possible between the points to be connected, without fancy bends. When wiring balanced or push-pull circuits, care should be taken to make the r.f. wiring on each side of the circuit as symmetrical as possible. Where it is necessary to pass r.f. wiring through the chassis, either a *feed-through* insulator of low-loss material should be used or the hole in the chassis should be of sufficient size to provide plenty of air space around the wire. Large-diameter rubber grommets can be used to prevent accidental short-circuits to the chassis.

By-pass condensers should be connected directly to the point to be by-passed and grounded immediately at the nearest available mounting screw, making certain that the screw makes good electrical contact with the chassis. Care should be taken to connect the marked side of tubular paper by-pass condensers to ground. Blocking and coupling condensers should be well spaced from the chassis.

High-voltage wiring should have exposed points kept at a minimum and those which cannot be avoided rendered as inaccessible as possible to accidental contact.

Soldering

The secret of good soldering is in allowing time for the *joint*, as well as the solder, to attain sufficient temperature. Enough heat should be applied so that the solder will melt when it comes in contact with the wires being joined, without touching the solder to the iron.

Wartime solder, which has a much smaller ratio of tin to lead, requires considerably more heat, and it becomes especially important to keep the iron clean at all times. More care must be exercised in making the joint because the new solder does not flow as readily, and also has a tendency to crystallize.

Soldering paste, if of the non-corroding type, is extremely helpful when used correctly. In general, it should not be used for radio work except when necessary. The joint should first be warmed slightly and the soldering paste applied with a piece of wire. Only the bit of paste which melts from the warmth of the joint should be used. If the soldering iron is clean it will be possible with one hand to pick up a drop of solder on the tip of the iron which can be applied to the joint, while the other hand is used to hold the connecting wires together. The use of excessive soldering paste causes the paste to spread over the surface of adjacent insulation, causing leakage or breakdown of the insulation. Except where absolutely necessary, solder should never be depended upon for the mechanical strength of the joint; the wire should be wrapped around the terminals or clamped with soldering terminals.

Do not attempt to make ground connections to a cadmium-plated chassis by soldering to the surface of the chassis, since the plating may be loosened by the heat and later fall off, breaking the connection. Drill a hole in the chassis and solder the wire in the hole.

Construction Notes

Lockwashers should be used under nuts to prevent loosening with use, particularly when mounting tube sockets or plug-in coil receptacles subject to frequent strain.

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation should be used. Satisfactory support for the shaft extension can be provided by means of a *metal* panel bearing made for the purpose. Never use panel bearings of the non-metal type unless the condenser shaft is grounded. *The metal bearing should be connected to the chassis with a wire or grounding strip.* This prevents any possible danger of shock.

The standard way of mounting toggle switches is with the switch "On" when the lever is in the upward position.

Variable condensers and resistors, having one-hole mountings, should be firmly fastened using the special lockwashers provided for shaft nuts.

The use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

Coil Winding

Dimensions for coils for the various units described in the constructional chapters are given under the circuit diagrams. Where no wire size is given, the power is sufficiently low to permit use of any available size within reason.

Unless a close-wound winding is definitely specified, the number of turns indicated should be spaced out to fill the specified length on the form. The length should be marked on the form and holes drilled opposite the pins to which the ends of the winding are to connect. Scrape one end of the wire and pass it through the lower hole in the form to the pin to which the bottom end of the winding is to connect, and solder this end fast. Unroll a length of wire approximately sufficient for the winding, and clamp the spool in a vise so it will not turn. The wire should be pulled out straight and the winding started by turning the form in the hands and walking toward the vise. A fair tension should be kept on the wire at all times. The spacing can be judged by eye. If, as the winding progresses, it becomes evident that the spacing is going to be incorrect to fill the required length, the winding can be started over again with a different spacing. If the spacing is only slightly off, the winding may be finished, the top end fastened, and the spacing corrected by pushing each turn. When complete, the turns should be fastened in place with coil cement. After a little practice, the job of determining the correct spacing will not be difficult.

Sometimes it is necessary to adjust the number of turns on a coil experimentally. The easiest way to do this is to bring a wire up from one of the pins, extending it through a hole in the form for a half-inch or so. The end of the winding may then be soldered to this extension rather than to the pin itself, and the nuisance of repeatedly fishing the wire through the pin avoided until the correct size of the winding has been determined.

Coil Cement

Duco cement, obtainable universally at hardware, stationery or 5-and-10-cent stores, is satisfactory for fastening coil turns. For small coils, a better-looking job will result if it is thinned out with acetone (amyl acetate), sometimes referred to as banana oil. If desired, the solution may be made thin enough to permit application with a brush.

Special low-loss coil "dopes" are available, including some with a polystyrene base.

Chapter Twelve

RECEIVER CONSTRUCTION

A Two-Tube Regenerative Receiver

The regenerative-detector receiver has long been a favorite with beginners, since it is comparatively easy to construct and adjust. A receiver of this type is shown in Figs. 1201, 1202 and 1204. It is designed to operate with tubes of either battery or a.c. type, and covers a total frequency range of 550 kc. to 32 Mc. with a series of plug-in coils. Sufficient audio output is available to operate a small permanent-magnet loudspeaker.

The circuit diagram appears in Fig. 1203. The antenna is coupled to the input circuit of the detector by means of the adjustable mica condenser, C_1. The input circuit is tuned to the frequency of the incoming signal by means of the variable condensers, C_2 and C_3. C_3 is used for general-coverage tuning, while C_2 is provided for bandspread tuning over a narrow band of frequencies within the total receiver range. C_3 serves also as the band-set condenser to locate the starting point of the range covered by C_2. C_3 has sufficient capacity range to cover the entire broadcast band with a single coil when used for general-coverage tuning.

Feed-back for regeneration is supplied by L_2. The amount of regeneration may be adjusted by means of R_4 which varies the screen voltage.

The output of the detector is coupled to the input of the audio amplifier by means of the audio reactor L_3, and the coupling condenser, C_8. While it is preferable to use a high-inductance (up to 1000 henries) reactance designed for this purpose, an ordinary filter choke of 15 to 30 henries will make an acceptable substitute. Volume may be adjusted for speaker or headphones by R_2. T is the output transformer coupling the plate of the audio amplifier to the speaker. When headphones are used they are plugged into the closed-circuit jack, J, which automatically short-circuits and thereby silences the speaker.

The components for the receiver shown in the photographs were assembled on an 8 × 6½ × 2-inch chassis bent up from sheet metal, although a standard steel chassis 7 × 7 × 2 inches will accommodate the parts equally well without any change in the relative positions shown in the photographs. The two tuning condensers are mounted on ceramic

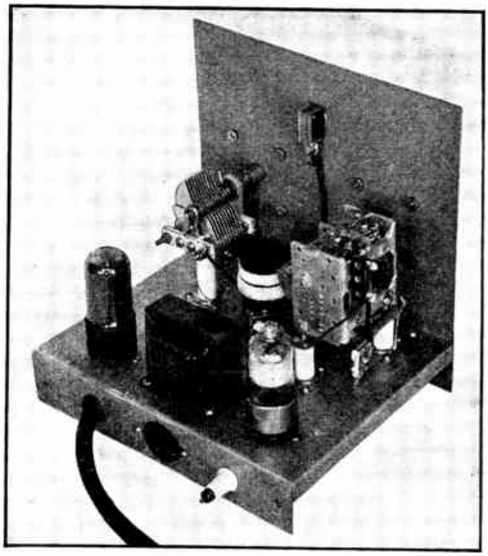


Fig. 1202 — Rear view of the two-tube regenerative receiver. Near the panel, from left to right are the bandspread tuning condenser, C_2, the plug-in coil and the general-coverage tuning condenser, C_3. The audio-amplifier tube, the output transformer, T, and the detector tube are lined up along the rear. The 4-prong socket set in the rear edge of the chassis is for connecting a small p.m. loudspeaker. The antenna terminal is at the right. The heater switch is mounted on the panel.

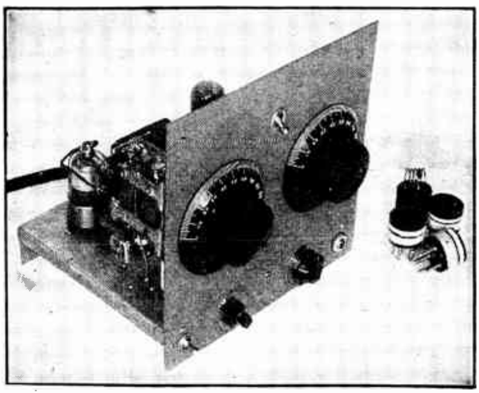


Fig. 1201 — Panel view of the two-tube regenerative receiver. The dial to the left controls the general-coverage or band-set tuning condenser, C_3, while the one to the right is the bandspread tuning dial which controls C_2. At the bottom the regeneration control is on the right and the audio volume control, R_2, at the left. The stand-by switch, S_1, is in the lower left-hand corner and the headphone jack occupies the opposite corner. The toggle switch at the top of the panel is the heater switch, S_2. Tube-base coils are shown at the right. This view shows also the method of mounting the antenna coupling condenser, C_1, the grid condenser, C_5, and the grid-leak resistor, R_1, on the frame of C_3.

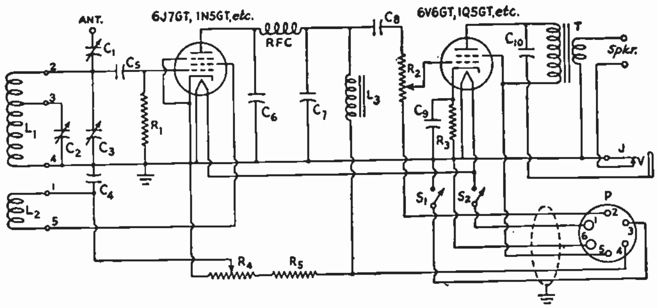


Fig. 1203 — Circuit diagram of the two-tube regenerative receiver.
C_1 — 3-30-$\mu\mu$fd. mica trimmer.
C_2 — 100-$\mu\mu$fd. midget variable condenser (National EX-100).
C_3 — 365-$\mu\mu$fd. variable (in the unit pictured, one section of a dual b.c. replacement variable, Meissner 21-5214).
C_4 — 0.001-μfd. mica.
C_5 — 250-$\mu\mu$fd. mica.
C_6, C_7 — 100-$\mu\mu$fd. mica.
C_8 — 0.05-μfd. paper.
C_9 — 10-μfd., 25-volt electrolytic.
C_{10} — 0.1-μfd. paper.
R_1 — 2 megohms, ½ watt.
R_2 — 500,000-ohm volume control.
R_3 — Cathode resistor, see text.
R_4 — 25,000-ohm potentiometer.
R_5 — 15,000 ohms, 1 watt.
RFC — 15-mh. r.f. choke.
L_1, L_2 — See text and coil table.
L_3 — Audio coupling reacter, see text.
T — Pentode output-to-speaker transformer, universal type.
S_1, S_2 — Single-throw, single-pole toggle switch.
J — Closed-circuit jack.
P — 6-prong tube socket.

pillars or metal spacers so that the two shafts are elevated to the same height above the chassis level while their centers are separated by a distance of 4¼ inches.

The coil socket is located midway between the two variable condensers and also is elevated above the chassis on small metal pillars so that its terminals will be accessible for connections to the condensers. One side of the antenna coupling condenser, C_1, is fastened to a stator terminal of C_3, while a wire running through a small hole in the chassis directly underneath connects the opposite side of C_1 to the feed-through antenna terminal set in the rear edge of the chassis.

The grid condenser, C_5, is supported on a small fibre lug strip fastened near the top of the frame of C_3 to bring it up to the level of the grid terminal of the detector tube, while the grid leak, R_1 is supported by its leads between one terminal of C_5 and one of the mounting screws in the chassis.

The two tubes and the audio reactor, L_3 are mounted in line along the rear edge of the chassis. The tube sockets are submounted by cutting holes to fit in the chassis so that their terminals will come below the surface of the chassis. Connections between the coil and detector-tube sockets are made with insulated wire running through a ¼-inch hole drilled directly underneath the coil socket.

Most of the small components, such as resistors, by-pass condensers and the r.f. choke, as well as the speaker coupling transformer are placed underneath the chassis as shown in the bottom-view photograph of Fig. 1204. One side of each by-pass condenser is connected as close as possible to the point to be by-passed and the other terminal grounded at the nearest point on the chassis.

The regeneration control, volume control, stand-by switch, S_1, and the headphone jack are mounted along the front edge of the chassis. The latter must be insulated by means of fibre washers when it is mounted.

The wires of the power-supply cable are anchored at an insulated lug strip located underneath the chassis at a point where the cable enters the chassis through a grommetted hole in the rear edge. The 4-prong socket for speaker connections also is mounted in the rear edge.

The panel is 8 × 8 inches and is fastened to the chassis by means of three 6-32 machine screws. Holes must be drilled along the bottom edge of the panel to pass the shafts of the controls and the shanks of the toggle switch and headphone jack. The heater switch, S_2, is mounted in a hole drilled near the top of the panel. Each dial requires four mounting holes and a half-inch hole to clear the shaft. These holes require careful lining up with the shafts of the tuning condensers.

Coils — Coil dimensions are given in the accompanying table both for standard 1½-

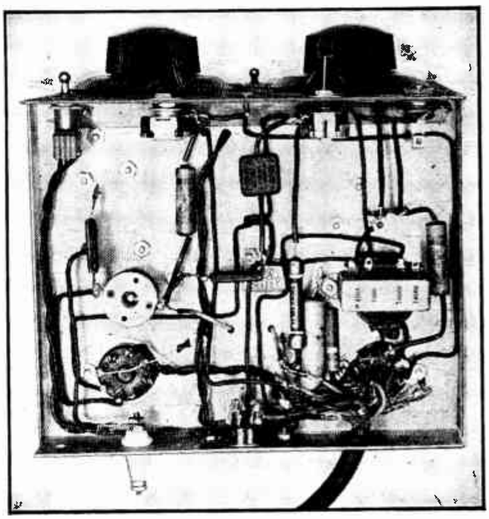


Fig. 1204 — Bottom view of the regenerative receiver. At the left above the detector-tube socket are the r.f. choke and the two by-pass condensers, C_6 (right) and C_7 (left). Above C_6 are C_8 to the left and C_4 to the right. Of the two resistors slightly to the right of center, R_5 is to the left and R_3 to the right with C_9 in between. To the right are the output transformer and C_{10}. The audio-tube socket is in the lower right-hand corner.

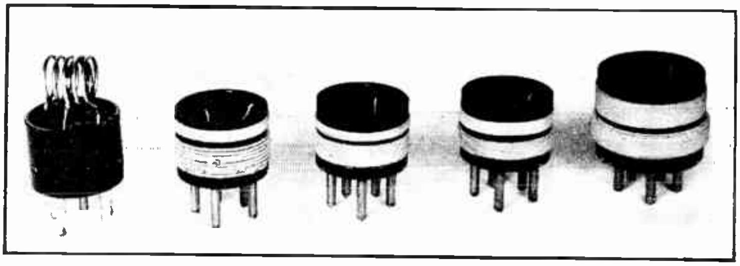


Fig. 1205 — A close-up view of the tube-base coils for the regenerative receiver. Band-spread taps are required on only the two highest-frequency coils. A special air-wound coil is necessary to cover the highest-frequency range of 12.8 to 32 Mc.

inch-diameter coil forms and for old bakelite tube bases, as shown in Fig. 1205, which may be used in case standard coil forms are not available. If tube bases are used, it will be necessary to wind the two largest coils in layers, since the base will not accommodate a sufficiently large single-layer winding. After winding the turns of the first layer in the conventional manner, the second layer is wound over the first in zig-zag fashion, taking one-half turn to cross back to the starting edge of the first layer and another half turn to cross back over the first layer to the finishing edge. Each successive turn of the zig-zag winding overlaps preceding turns at each reversal of the cross over. The finished coil actually has more than two layers at some points which explains the larger number of turns given for the "second layer" in the coil table. While an ordinary "scramble-wound" coil is not as neat in appearance, it will work almost as well if the constructor does not wish to bother with the more complicated winding. Connections to the coil-form pins and the coil-socket prongs are shown in Fig. 1206.

The self-supporting air-core coil No. 5 is made by winding 5 turns of No. 18 wire around a half-inch form, such as the shank of a drill, and spreading the turns out to the required coil length of ⅝ inch by running a screwdriver or knife between the turns after they have been removed from the form. The tickler winding also should be wound on a half-inch form and the turns should be fastened together with Duco waterproof cement or low-loss coil "dope" before removing from the form. After the cement has dried, the coil is inserted inside the "ground" end of L_1. Its position should be varied until smooth regeneration is obtained and then it may be fastened in place with cement. Care should be taken to make all tickler windings in the same direction as the turns of L_1, otherwise the circuit will not regenerate with the coil connections shown.

Choice of tubes — A wide variety of tubes will work satisfactorily in this receiver. Any of the 6.3- or 2.5-volt a.c., or 2- or 1.4-volt battery r.f. amplifier pentodes listed in the tube tables of the appendix may be used as the detector, while any of the audio power pentodes or beam tubes listed with a similar filament- or heater-voltage rating may be used in the audio amplifier, providing its rated plate current does not exceed 50 ma. Among the more widely available types are 57, 58 and 78 for detector in the old 2.5-volt a.c. series, with the 2A5 or 59 as the corresponding audio-amplifier type. In the 6.3-volt series with old-style bases are the 6C6 and 6D6 as detector and the 41, 42 or 89 for the audio stage. In the more modern octal series are the 6J7 or 6K7 for detector and the 6F6 or 6V6 for output amplifier. Their glass equivalents are equally suitable, of course. In the loktal series, the 7A7, 7B7 or 7W7 may be used in the detector stage, while the 7A5 or 7B5 will serve as audio amplifier.

For battery operation, the 1D5GP or 1E5GP in the 2-volt-filament class are suitable for

COIL TABLE FOR THE 2-TUBE REGENERATIVE RECEIVER

No	*Range Mc*	*Amateur Band Mc.*	*Dimensions (inches)* *A*	*B*	*Turns* L_1	L_2	*Bandspread Tap*
1	0.55– 1.6		1⅜	⅛	65½[1] (No. 32 d.c.c.)	16¾ (No. 32 d.c.c.)	
			1½	⅛	90½ "	20¾ "	
2	1.2 – 3.35	1.75	1⅛	⅛	29½[2] "	9¾ "	
			1½	⅛	32½ "	10¾ "	
3	2.7 – 7.7	3.5	1⅛	⅛	14½ "	5¾ "	
			1½	⅛	10½[3] "	5¾ "	
4	5.35–14.6	7.0	1⅛	⅛	7½[4] (No. 24 d.c.c.)	5¾ "	4¼
			1½	⅛	6½[4] "	5¾ "	3½
5	12.8 –32	14–28	⅝	½	5[5] (No. 18 d.c.c.)	6 (No. 28 d.c.c.)	2½

Coil dimensions *A* and *B* and socket connections are shown in Fig. 1206. Specifications for standard 1½" diameter coil forms are also shown. Direction of winding is the same for all coils. All windings are close-wound unless otherwise indicated. Taps are counted from the ground end of the coil.

[1] First layer, 18 turns, close-wound; second layer 47½ turns (see text).

[2] First layer, 12 turns, close-wound; second layer 17½ turns (see text).

[3] Spaced to cover ¼ inch.

[4] Spaced to cover ⅜ inch.

[5] Self-supporting (see text).

detector, while the 1F5GP or 1G5G of the same class may be used in the audio stage. In the 1.5-volt octal series there are the 1N5G and 1P5G for detector and the 1A5G and 1C5G for audio amplifier. The loktal types 1LC5 and 1LN5 will make satisfactory detectors with a 1LA4 or 1LB4 amplifier.

When the receiver is to be used for portable as well as home-station operation the selection of a.c. and battery tubes whose sockets and connections make them interchangeable is desirable. In the detector circuit the 1D5GP, 1D5GT and the 1E5GP in the 2-volt battery series, or the 1N5G and 1P5G in the 1.5-volt battery series are directly interchangeable with the 6M7G, 6T6GM, 6U7G, 6W7G, 6J7GT/G, 6K7GT/G and the 6S7 in the 6.3-volt octal-base series. If loktal-base tubes are preferred, the 1LC5 and 1LN5 battery tubes are directly interchangeable with the 7V7, 7A7, 7B7, 7C7, 7G7, 7H7, 7L7 or 7T7 types.

In the audio amplifier the 1G5G and 1J5G of the 2-volt battery series or the 1A5G, 1C5G, 1Q5G and 1T5G in the 1.5-volt battery series are directly interchangeable with the 6AG6G, 6K6G, 6M6G or 6V6 with octal bases. The loktal types 1LA4 and 1LB4 also are interchangeable with the indirectly-heated types 7A5 and 7B5 of the loktal series.

It may pay to try different values of grid-leak resistance if detector-tube types other than those mentioned in Fig. 1203 are used and some slight alteration in the number of turns in L_2 may be necessary for best performance. The audio tube selected will require a certain biasing voltage which may be taken from the tube tables. With battery tubes, this means simply selecting the proper "C"-battery voltage. With a.c. tubes, however, bias is obtained from the voltage drop across the cathode resistor, R_3. A satisfactory resistance value for any particular audio tube may be easily calculated by adding the rated plate and screen currents given in the tube tables and then dividing the required biasing voltage by this sum. If the current value is in terms of milliamperes, the answer from the above operation must be multiplied by 1000 to obtain the required resistance value in ohms. For example, the tube table shows that a 7C5 requires a biasing voltage of 12.5 and that the screen current averages about 6 ma., while the plate current averages 46 ma. Adding the two currents gives a total of 52 ma. Dividing 12.5 by 52 gives a result of 0.24. When this is multiplied by 1000, the answer of 240 ohms is obtained.

Power Supply— Suitable power-supply diagrams are shown in Fig. 1207. The VR75 in the a.c. circuit in Fig. 1207-A provides a regulated voltage of 75 for the detector which will help materially in obtaining smooth regeneration. However, if one is not available, it may be omitted, if R is increased to 150,000 ohms, at some sacrifice in smoothness of regeneration control. The small condensers, C_1 and C_2 are to help reduce "tunable hum" which may occur at certain points in the frequency range of the receiver. Components for the a.c. power supply may be mounted on a 7 × 7 × 2-inch steel chassis or a baseboard made of wood. The placement of parts is not important. If the steel chassis is used, the smaller components may be mounted underneath. The voltage of the filament winding should, of course, correspond to the rated heater voltage of the tubes used (6.3 or 2.5 volts). In Fig. 1207-B is shown the proper connections for a battery supply. The voltage of the "A" battery will depend upon the rated filament voltage of the tubes selected (1.5 or 2 volts), while the "B" battery should consist of two 45-volt "B" batteries connected in series. The voltage of the "C" battery will depend upon the biasing voltage required for the tube in the audio

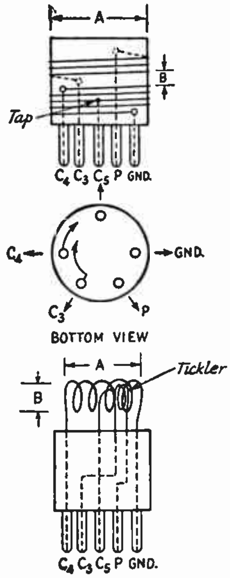


Fig. 1206 — Sketch showing coil-form and coil-socket connections for the plug-in coils for the two-tube regenerative receiver. The bottom sketch shows the special construction for Coil No. 5 listed in the accompanying coil table.

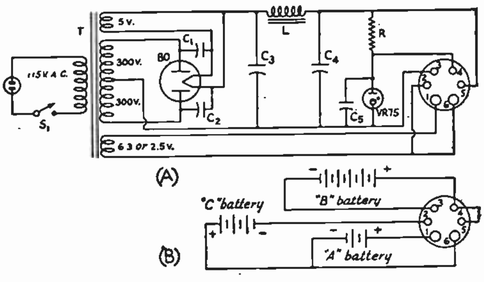


Fig. 1207 — Circuit diagram of a power supplies suitable for small receivers. The a.c. supply shown at A is suggested for the two-tube regenerative and the three-tube superheterodyne receivers described in this chapter. The battery supply shown at B is arranged especially to fit the plug connections of the regenerative receiver of Fig. 1201.

C_1, C_2 — 0.001-μfd., 1,000-volt mica.
C_3, C_4, C_5 — 8-μfd., 450-volt electrolytic.
R — 10,000-ohm, 10-watt wire-wound for the 2-tube regenerative receiver, 5000 ohms, 10 watts for the 3-tube superheterodyne receiver of Fig. 1213. If the VR-75 is omitted, R, should be increased to 100,000 and 50,000 ohms respectively.
T — Standard replacement-type power transformer with 6.3-volt, 5-volt, and 600-volt center-tapped windings, 70 ma. d.c. output rating.
L — Standard replacement-type filter choke, 15–30 henries at 70 ma.
S_1 — S.p.s.t. toggle switch.

amplifier as mentioned previously. The 6-prong socket for connections to the receiver may be mounted on a board or connected to one end of a short cable, the wires at the other end of the cable being connected to the batteries as shown.

Adjustment — The most difficult part of adjusting the receiver after it has been constructed and the power supply connected is that of adjusting the coils for proper regeneration. It is perhaps best to start out with the lowest-frequency coil (b.c. band) and work up through the higher frequencies. One end of a single-wire antenna should be connected to the antenna terminal and the chassis connected to the nearest water pipe or other ground connection. The antenna preferably should be 50 to 100 feet in length. With the antenna connected, R_4 should be set for maximum screen voltage (to the right toward R_5 in Fig. 1203), C_3 should be turned to maximum capacity (plates completely meshed) and C_2 at minimum capacity. With these adjustments, a click should be heard in the headphones each time the grid terminal of the detector tube is touched with the finger. If no click is heard, the capacity of C_1 should be reduced a bit at a time, by loosening up on the adjusting screw, until the click is heard. If no click is heard with C_1 at minimum capacity, it will be necessary to add a turn or two to L_2.

After the click has been obtained, it should be possible to turn R_4 back and forth with a "plop" in the headphones each time R_4 passes through a certain point in its range. On the high-voltage side of the "plop" point, the click should be heard when the grid terminal is touched with the finger, while it will not be heard when R_4 is turned to the low-voltage side of this point. The detector is oscillating under the condition where the click is heard and is not oscillating when the click cannot be obtained. With correct circuit adjustments it should be possible to bring the detector into oscillation with a soft rushing noise and not a loud "plop." For c.w. code reception R_4 is adjusted so that the detector is oscillating, but very close to the point where oscillation ceases. On the other hand for the reception of modulated signals ('phone or music) the detector is adjusted to a non-oscillating condition, but very close to the point where it goes into oscillation.

In listening over any band for signals, it is common practice to set the detector into oscillation for both 'phone and c.w. signals. When the steady whistle of a 'phone station is tuned in, the regeneration control is then backed off to stop oscillation when the whistle will disappear, and only the modulation will be heard.

C_3 is used for general-coverage tuning, while C_2 is used for bandspread tuning. If the coil dimensions given in the table have been followed closely, C_3 should be set at approximately the following points on the dial for each of the amateur bands: 1.75-Mc. band, 52; 3.5-Mc. band, 39; 7-Mc. band, 34; 14- and 28-Mc. bands, 93. These dial settings assume the use of a dial which reads 100 at *minimum* capacity. C_2 then should be used for tuning over the band. Coil No. 5 is used for both the 14- and 28-Mc. bands. At the higher frequencies especially, it may be necessary to retune each time the regeneration control is adjusted to keep the signal tuned in.

A Two-Tube Superheterodyne Receiver

Although all the advantages of the superheterodyne-type receiver cannot be secured without going to rather elaborate multi-tube circuits, it is possible to use the superhet principle to overcome most of the disadvantages of the simple regenerative receiver. These are chiefly the necessity for critical adjustment of the regeneration control with tuning, antenna "dead spots," lack of stability (both in the detector circuit itself and because of slight changes in frequency when the antenna swings with the wind), and blocking, or the tendency for strong signals to pull the detector into zero beat. These effects can be largely eliminated by making the regenerative detector operate on a fixed low frequency and designing it for maximum stability. The incoming signal is then converted to the fixed detector frequency before being detected.

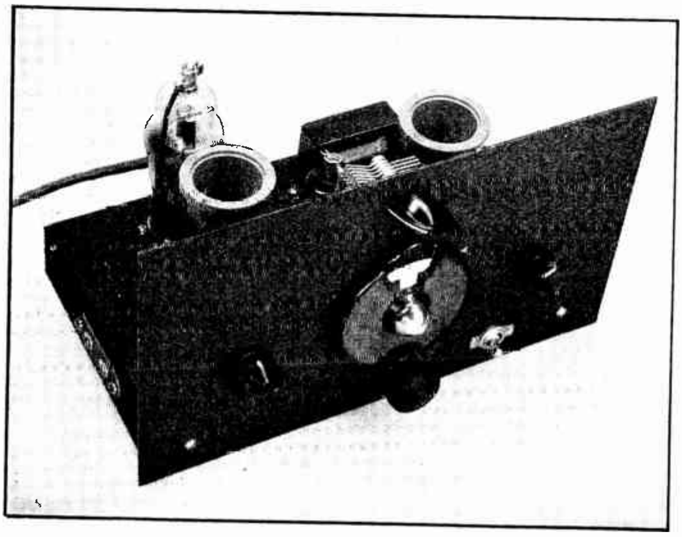


Fig. 1208 — The two-tube superheterodyne shown here has one more operating control than the ordinary regenerative-detector receiver, but it is more stable.

A two-tube receiver operating on this principle is shown in Figs. 1208 to 1212.

The circuit diagram is given in Fig. 1210. A 6K8 is used to convert the frequency of the incoming signal to the fixed or intermediate frequency, and the two triode sections of a 6C8G serve as the regenerative detector and audio amplifier respectively. L_1C_1 is the r.f. circuit, tuned to the signal, and L_2 is the antenna coupling coil. C_7 is a by-pass condenser across the 1.5-volt battery used to bias the signal grid of the 6K8. The high-frequency oscillator tank circuit is $L_3C_3C_4$, with C_3 for band-setting and C_4 for bandspread.

The i.f. tuned circuit (or regenerative detector circuit) is L_5C_5. This must be a high-C circuit if stability better than that of an ordinary regenerative detector is to be secured. The frequency to which it is tuned should be in the vicinity of 1600 kc.; the exact frequency does not matter so long as it falls on the low-frequency side of the 1750-kc. band. L_5 and its tickler coil, L_6, are wound on a small form, and L_5 is tuned by a fixed mica condenser of the low-drift type. Since these condensers are rated with a capacity tolerance of 5 per cent, it is sufficient to wind L_5 as specified under Fig. 1210. The resulting resonant frequency will be in the correct region. No manual tuning is necessary, and therefore the frequency of this circuit need not be adjusted. C_2 is the regeneration-control condenser, isolated from the d.c. supply by the choke, *RFC*. Only enough turns need be used on L_6 to make the detector oscillate readily when C_2 is at half capacity or more.

The second section of the 6C8G is transformer-coupled to the detector. The grid is biased by the same battery which furnishes bias for the 6K8.

Looking at the top of the chassis from in front, the r.f. or input circuit is at the left, with C_1 on the panel and L_1L_2 just behind it. The 6C8G is directly to the rear of the coil. The 6K8 converter tube is centered on the chassis, with C_3 and C_4 on the panel directly in front of it. C_4 is driven by the vernier dial and C_3 is toward the top of the panel. The coil at the right is L_3L_4, in the oscillator tuned circuit. The regeneration-control condenser, C_2, is at the right on the panel. The audio transformer, T_1, is behind the oscillator coil.

Looking at the bottom of the chassis, the antenna-ground terminals are at the left, with a lead going directly to L_2 on the coil socket. The bias battery is fastened to a two-lug insulating strip by means of wires soldered to the battery. The zinc can is the negative end and the small cap the positive terminal. By-pass condenser C_7 is mounted on the coil socket.

The i.f. coil is mounted on the chassis midway between the socket for the 6C8G and that for the 6K8. In winding the coil the ends of the wires are left long enough to reach to the various tie-in points. The grid condenser, C_9, is

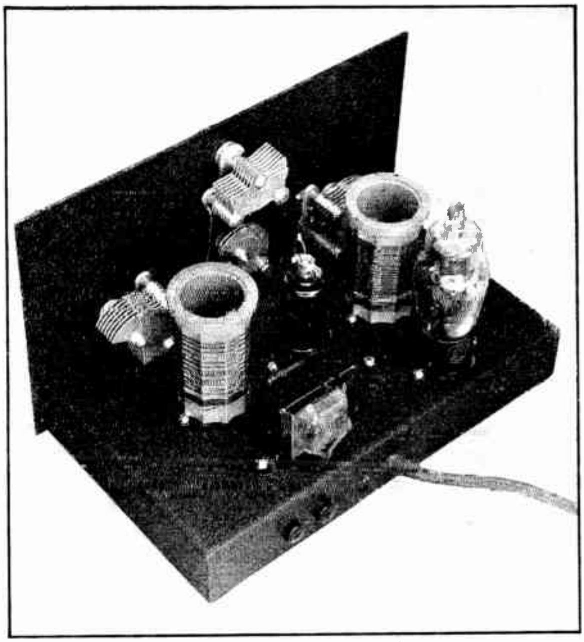


Fig. 1209 — A back-of-panel view of the two-tube superhet, showing the arrangement of parts on top of the $5\frac{1}{2} \times 9\frac{1}{2} \times 1\frac{1}{2}$-inch chassis.

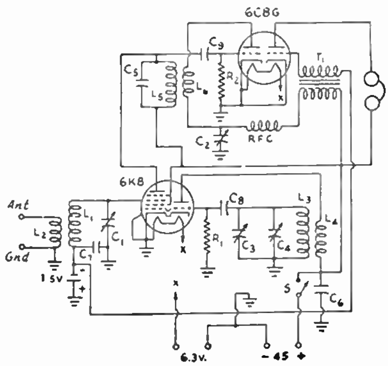


Fig. 1210 — Circuit diagram of the two-tube superheterodyne.

C_1, C_2, C_3 — 100-$\mu\mu$fd. variable (Hammarlund SM-100).
C_4 — 15-$\mu\mu$fd. variable (Hammarlund SM-15).
C_5 — 250-$\mu\mu$fd. silvered mica (Dubilier Type 5-R).
C_6 — 0.01-μfd. paper.
C_7 — 0.005-μfd. mica.
C_8, C_9 — 100-$\mu\mu$fd. mica.
R_1 — 50,000 ohms, ½ watt.
R_2 — 1 megohm, ½ watt.
RFC — 2.5-mh. r.f. choke.
T_1 — Audio transformer, interstage type, 3:1 ratio (Thordarson T-13A34).
L_1–L_4, inc. — See coil table.
L_5 — 55 turns No. 30 d.s.c., close-wound on ¾-inch diameter form (National PRF-2); inductance 40 microhenrys.
L_6 — 18 turns No. 30 d.s.c., close-wound, on same form as L_5; see Fig. 1212.
S — S.p.s.t. toggle switch.

TWO-TUBE SUPERHET COIL DATA

Coil	Grid Winding (L_1 and L_3)	Antenna (L_2) or Tickler (L_4)
A	56 turns No. 22 enameled	10 turns No. 24 enameled
B	32 " " " "	8 " " " "
C	18 " " " "	7 " " " "
D	12 " " " "	7 " " " "
E	10 " " " "	7 " " " "

All coils wound on 1½-inch diameter forms (Hammarlund SWF-4). Grid windings on coils B-E, inclusive, are spaced to occupy a length of 1½ inches; grid winding on coil A is close-wound. Antenna-tickler coils are all close-wound, spaced ⅛-inch from bottom of grid winding. See Fig. 1212.

Frequency Range	*Coil at L_1-L_2*	*Coil at L_3-L_4*
1700 to 3200 kc.	A	B
3000 to 5700 kc.	B	C
5400 to 10,000 kc.	C	D
9500 to 14,500 kc.	E	D

supported by the grid terminal on the tube socket and the end of the grid winding, L_5. R_2 is mounted over the 6C8G socket. The i.f. tuning condenser, C_5, is mounted by its terminals between the plate and screen prongs on the 6K8 socket, the ends of L_5 being brought to the same two points.

The oscillator grid condenser, C_8, is connected between the coil-socket prong and the oscillator grid prong on the 6K8 socket. By-pass condenser C_6 is mounted alongside the oscillator coil socket, as shown. The connections to the rotors of the tuning condensers for both coils go through holes in the chassis near the front edge. Grounds are made directly to the chassis in all cases; make sure that there is an actual connection to the metal.

The "B" switch is a single-pole single-throw toggle. 'Phone-tip jacks on the rear chassis edge provide means for connecting the audio output to the headphones.

The method of winding coils is indicated in Fig. 1212; if the connections to the circuit are made as shown, there will be no trouble in obtaining the necessary oscillation. Both coils on each form should be wound in the same direction.

Adjustment — To test the receiver, first try out the i.f. circuit. Connect the filament and "B" supply and place both tubes in their sockets. Put a high-frequency coil in the r.f. socket, but do not insert a coil in the oscillator socket. The only test which need be made is to see if the detector oscillates properly. Advance C_2 from minimum capacity until the detector goes into oscillation, which will be indicated by a soft hiss. This should occur at around half scale on the condenser. If it does not occur, check the coil (L_5L_6) connections

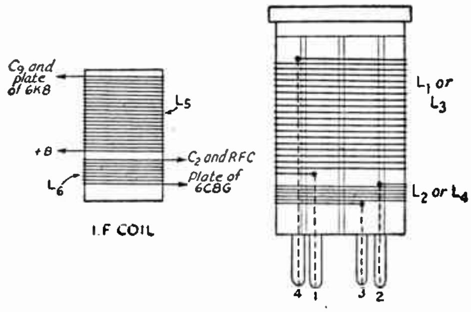


Fig. 1212 — How the coils for the two-tube superheterodyne are wound. The bottom end of the i.f. coil in this drawing is the end mounted adjacent to the chassis. L_5 and L_6 are wound in the same direction. On the r.f. socket, pin 4 connects to the No. 3 grid (top cap) of the 6K8 and stator of C_1, pin 1 to C_7, pin 2 to ground and pin 3 to the antenna post. On the oscillator socket, pin 4 goes to C_8 and the stators of C_3 and C_4, pin 1 to ground, pin 2 to "B" + and pin 3 to the 6K8 oscillator plate. Both windings are in the same direction on each coil.

and winding direction and, if these seem right, add a few turns to the tickler, L_6. If the detector oscillates with very low capacity at C_2, it will be advisable to take a few turns off L_6 until oscillation starts at about midscale.

After the i.f. has been checked, plug in an oscillator coil for a range on which signals are likely to be heard at the time. The 5400–10,000-kc. range is usually a good one. The coils are arranged so that a minimum number is needed, even though two are used at a time. With coil C in the r.f. socket and D in the oscillator circuit, set C_1 at about half scale and turn C_3 slowly around midscale until a signal is heard. Then tune C_1 for maximum volume. Should no signals be heard, the probability is that the oscillator section of the 6K8 converter tube is not

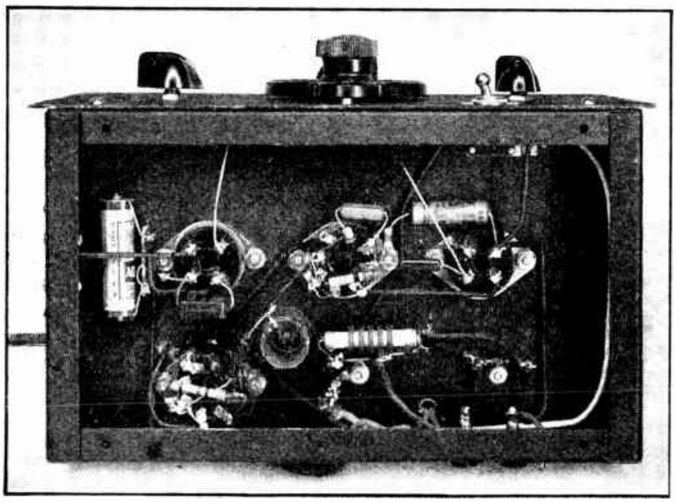


Fig. 1211 — Below-chassis view of the two-tube superhet. The i.f. circuit is underneath the chassis; no adjustment of its frequency is necessary. Since few parts are required, the construction, assembly and wiring are quite simple.

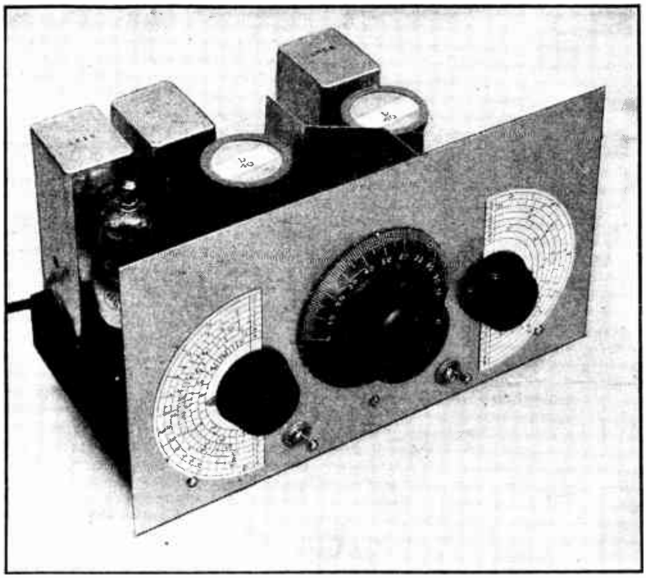


Fig. 1213 — A three-tube superheterodyne receiver, designed for either a.c. or d.c. heater operation and for 90-volt "B" battery plate supply.

working, in which case the same method of testing is used as described above for the i.f. detector — check wiring, direction of windings of coils, and finally, add turns to the tickler, L_4, if necessary.

The same oscillator coil, D, is used for two frequency ranges. This is possible because the oscillator frequency is placed on the low-frequency side of the signal on the higher range. This gives somewhat greater stability at the highest-frequency range. Some pulling — a change in beat-note as the r.f. tuning is varied by means of C_1 — will be observed on the highest-frequency range, but it is not serious in the region of resonance with the incoming signal frequency.

The receiver will respond to signals either 1600 kc. lower or 1600 kc. higher than the oscillator frequency. The unwanted response is discriminated against by the selectivity of the r.f. circuit. On the three lower-frequency ranges, when it is possible to find two tuning spots on C_1 at which incoming noise peaks up, the lower-frequency peak is the right one. The oscillator frequency is 1600 kc. higher than that of the incoming signal on these three ranges and 1600 kc. lower on the fourth range. Bandspread is not needed in the r.f. circuit.

The regeneration control may be set to give desired sensitivity and left alone while tuning; only when an exceptionally strong signal is encountered is it necessary to advance it more to keep the detector in oscillation. It should be set just on the edge of oscillation for 'phone reception.

The heater requirements of the set are 0.6 amperes at 6.3 volts, approximately. Either a.c. or d.c. may be used. The "B" battery current is between 4 and 5 ma., so that a standard 45-volt block will last hundreds of hours

A Three-Tube General Coverage and Bandspread Superheterodyne

A superhet receiver of simple construction, having a wide frequency range for general listening-in as well as full bandspread for amateur-band reception, is shown in Figs. 1213 to 1217. The circuit uses only three tubes and gives continuous frequency coverage from about 75 kc. (4000 meters) to 60 Mc. (5 meters). The receiver is intended for operation from either a 6.3-volt transformer or 6-volt storage battery for filament supply, and a 90-volt "B" battery for plate supply.

The circuit diagram is given in Fig. 1214. A 6K8 is used as a combined oscillator-mixer followed by a 6SK7 i.f. amplifier. The intermediate frequency is 1600 kc., a frequency which reduces image response on the higher frequencies and simplifies the design for low-frequency operation in the region below the broadcast band. One section of the 6C8G dou-

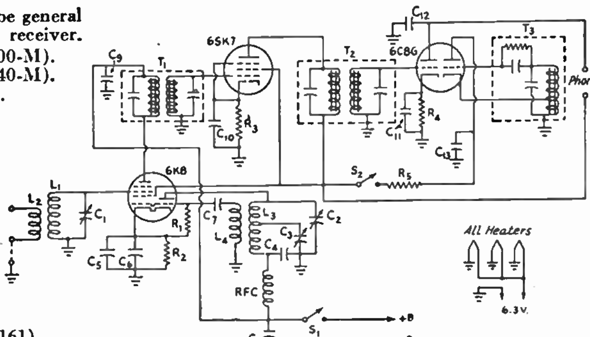


Fig. 1214 — Wiring diagram for the three-tube general coverage and bandspread superheterodyne receiver.

C_1 — 100-µµfd. variable (Hammarlund MC-100-M).
C_2 — 140-µµfd. variable (Hammarlund MC-140-M).
C_3 — 35-µµfd. variable (Hammarlund HF-35).
C_4 — Oscillator padder; see coil table.
C_5 — 0.1-µfd. paper.
C_6 — 0.002-µfd. mica.
C_7 — 250-µµfd. mica.
C_8 — 0.002-µfd. mica.
C_9, C_{10} — 0.01-µfd. paper.
C_{11} — 5-µfd. electrolytic, 50 volts.
C_{12}, C_{13} — 0.002-µfd. mica.
R_1 — 50,000 ohms, ½ watt.
R_2, R_3 — 250 ohms, ½ watt.
R_4 — 12,000 ohms, ½ watt.
R_5 — 50,000 ohms, ½ watt.
T_1, T_2 — 1600-kc. i.f. transformer (Millen 64161).
T_3 — 1600-kc. oscillator transformer (Millen 65163).
L_1, L_2, L_3, L_4 — See coil table.
S_1, S_2 — S.p.s.t. toggle switch.
RFC — 2.5-mh. r.f. choke.

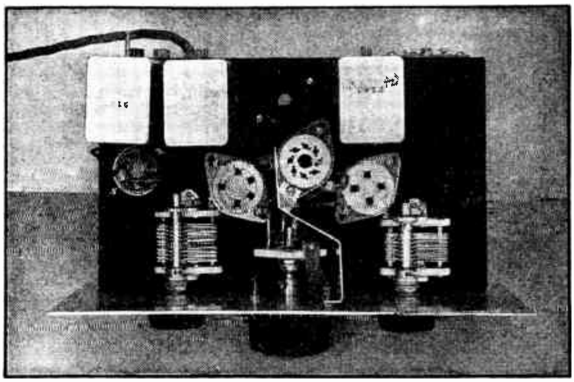


Fig. 1215 — A plan view of the three-tube superheterodyne with the coils and tubes removed. The chassis measures 5½ × 9½ × 1½ inches and the panel size is 10½ × 6 inches.

ble triode is used as a second detector and the other section as a beat-frequency oscillator. Headphone output is taken from the plate circuit of the second detector.

To simplify construction, the antenna and oscillator circuits are separately tuned. The antenna tuning control, C_1, may be used as a volume control by detuning from resonance. The oscillator circuit, $L_3C_2C_3$, is tuned 1600 kc. higher than the signal on frequencies up to 5 Mc.; above 5 Mc. the oscillator is 1600 kc. lower than the signal. C_2 is the general coverage or band-setting condenser, C_3 the bandspread or tuning condenser. C_4 is a tracking condenser which sets the oscillator tuning range on each band so that it coincides with the tuning range in the mixer grid circuit.

The i.f. stage uses permeability-tuned transformers with silvered-mica fixed padding condensers. The second detector is cathode-biased by R_4, by-passed by C_{11} for audio frequencies.

The second 6C8G section is the beat oscillator, using a permeability-tuned transformer. The grid condenser and leak are built into the transformer. The plate is fed through the b.o. on-off switch and a dropping resistor, R_5, the latter serving both to reduce the "B" current drain and to cut down the output of the oscillator to a value suitable for good heterodyning.

No special coupling is needed between the beat oscillator and the second detector.

The plates and screens of all tubes except the beat oscillator are operated at the same voltage — 90 volts. The "B" current drain is approximately 15 milliamperes, which is about the normal drain for medium-size "B" batteries. The receiver will operate satisfactorily, although with somewhat reduced volume, using a single 45-volt battery for "B" supply.

The parts arrangement is shown in the photographs of Figs. 1215 and 1216. The mixer tuning condenser, C_1, is at the right. The bandspread oscillator tuning condenser, C_3, is in the center, controlled by the National Type-A 3½-inch dial, and the bandset condenser, C_2, is at the left.

Referring to the top view, Fig. 1215, the i.f. section is along the rear edge, with T_1 at the right. Next is the socket for the 6SK7, then T_2, and finally T_3 at the extreme left. The socket for the 6C8G is just in front of T_3. The triode section in which the grid is brought out to the top cap is the one which is used for the beat oscillator.

The r.f. section has been arranged for short leads to favor high-frequency operation. The three sockets grouped closely together in the center are, from left to right, the oscillator-coil socket, socket for the 6K8, and the mixer-coil socket. All are mounted above the chassis by means of mounting pillars, so that practically all r.f. leads are above deck. The oscillator grid leak, R_1, and the high-frequency cathode by-pass condenser, C_6, should be mounted directly on the socket before it is installed. So also should the oscillator grid condenser, C_7, which can be seen extending to the left toward the oscillator-coil socket in Fig. 1215. Power-supply connections should be soldered to the 6K8 socket prongs before the socket is mounted, and these leads brought down through a hole in the chassis.

The general-coverage condensers, C_1 and C_2, are mounted directly on the chassis. C_3 is held from the panel by means of a small bracket made from metal strip, bent so that the condenser shaft lines up with

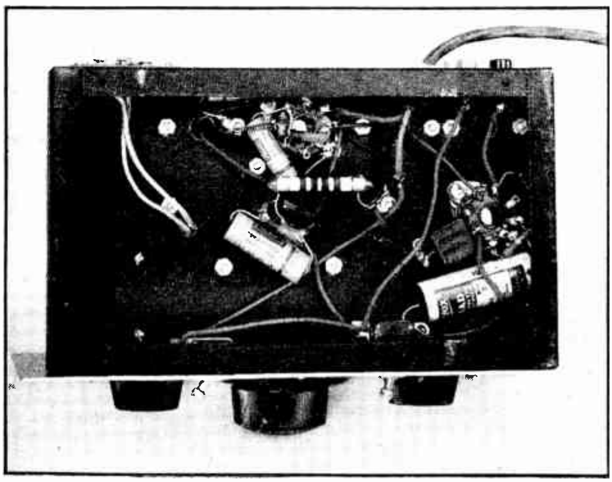


Fig. 1216 — Below the chassis of the three-tube receiver. The r.f. choke is mounted near the oscillator coil socket to keep the r.f. leads short. In the i.f. stage, care should be taken to keep the plate and grid leads from the i.f. transformer short and well separated. A four-wire cable is used for power-supply connections. The headphone-tip jacks may be seen near the upper right-hand corner.

the dial coupling. A baffle shield made of aluminum separates the oscillator and mixer sections; this shield is essential to prevent coupling between the two circuits which might otherwise cause interaction and poor performance.

The first step in putting the receiver into operation is to align the i.f. amplifier. This should preferably be done with the aid of a test oscillator, but if one is not available the circuits may be aligned on hiss or noise. The beat oscillator can also be used to furnish a signal for alignment. Further information on alignment may be found in Chapter Seven.

The coils are wound as shown in Fig. 1217. A complete set of specifications is given in the coil table. Ordinary windings are used for all oscillator coils, and for all mixer coils for frequencies above 1600 kc. Below 1600 kc., readily available r.f. chokes are used for the tuned circuits. For the broadcast band and the 600–750-meter ship-to-shore channels, the mixer coil is a Hammarlund 2.5-mh. r.f. choke, with the pies tapped as shown in Fig. 1217. The grid end and the intermediate tap are connected to machine screws mounted near the top of the coil form, and a flexible lead is brought out from the grid pin in the coil form to be fastened to either lead as desired. Mixer coils for the two lowest-frequency ranges are constructed as shown. The antenna winding in each case is a coil taken from an old 465-kc. i.f. transformer, having an inductance of about 1 millihenry. The inductance is not particularly critical, and a pie from a 2.5-mh. choke may be used instead.

With the i.f. aligned, the mixer grid and oscillator coils for a band can be plugged in. C_3 should be set near minimum capacity and C_2 tuned from minimum capacity until a signal is heard. Then C_1 is adjusted for maximum signal strength. If C_2 is set at the high-frequency end of an amateur band, further tuning should be done with C_3, and the band should be found to cover about seventy-five per cent of the dial. C_3 can of course be used for bandspread tuning outside as well as inside the amateur bands. It is convenient to calibrate the receiver, using homemade paper scales for the purpose as shown in Fig. 1213. Calibration points may be taken from incoming signals whose frequencies are known, from a calibrated test oscillator, or from the harmonics of a 100-kc. oscillator as described in Chapter Twenty. The mixer calibration need only be approximate, since tuning of the mixer circuit has little effect on the oscillator frequency. It is sufficient to make a

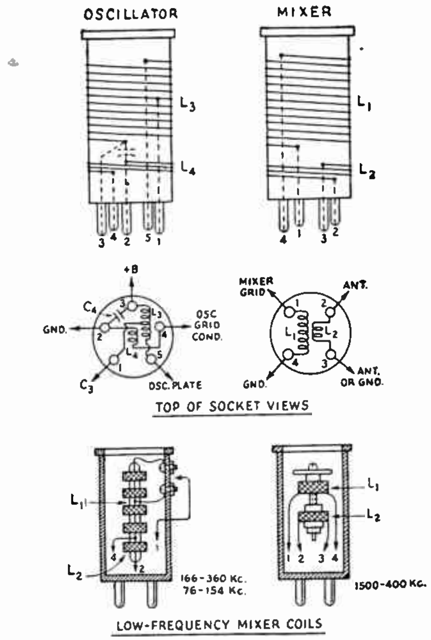


Fig. 1217 — How the coils for the three-tube superheterodyne are constructed. On the hand-wound oscillator and mixer coils, all windings are in the same direction.

COIL DATA FOR THE THREE-TUBE SUPERHETERODYNE

Range	*Turns* L_1	L_2	L_3	L_4	L_3 *Tap*	C_4
A — 76–154 kc.	30 mh.	1 mh.				
166–360 kc.	8 mh.	1 mh.	65	12	Top	300 µµfd.
400–1500 kc.	2.5 mh.*	*				
B — 1.6 to 3.2 Mc. (160 meters)	56	10	42	11	Top	75 µµfd.
C — 3.0 to 5.7 Mc. (80 meters)	32	8	27	9	Top	100 µµfd.
D — 5.4 to 10.0 Mc. (40 meters)	18	8	22	9	12	0.002 µfd.
E — 9.5 to 18.0 Mc. (20 meters)	10	8	12	3½	6	400 µµfd.
F — 15.0 to 30 Mc. (10 meters)	6	4	6	2½	2½	400 µµfd.
G — 30 to 60 Mc. (5 meters)	3	3	3½	1	1	300 µµfd.

* See Fig. 1217 and text for details. C_4 is mounted inside oscillator coil form; see Fig. 1217. Bandspread taps on L_3 measured from bottom ("B" + end) of coil. L_3-A and L_1-B coils close-wound with No. 22 enameled wire; L_3-B close-wound with No. 20 enameled; all other L_1 and L_3 coils wound with No. 18 enameled, spaced to give a length of 1½ inches on a 1½-inch diameter form (Hammarlund SWF) except the G coils, which are spaced to a length of 1 inch on 1-inch diameter forms (Millen 45004 and 45005). Antenna and tickler coils, L_2 and L_4, are close-wound with No. 24 enameled, spaced about ⅛th-inch from bottoms of grid coils, except for L_4-G, which is interwound with L_3.

calibration which ensures that the mixer is tuned to the desired signal rather than to the image.

On the broadcast band, the tuning range is such that, with C_2 set at 1500 kc., the entire band will be covered on C_3. It is necessary, however, to change the tap on the mixer coil to make the antenna circuit cover the entire band. Only one oscillator coil is needed for the range from 75 to 1500 kc., but a series of coils is needed to cover the same range in the mixer circuit.

Adding an audio stage to the three-tube superhet — The three-tube receiver just described is designed for headphone operation, but readily can be converted to a four-tube set for use with a speaker. For this purpose a 6F6 pentode can be added to the circuit diagram, as shown in Fig. 1219. Figs. 1218 and 1220 show the receiver when completed.

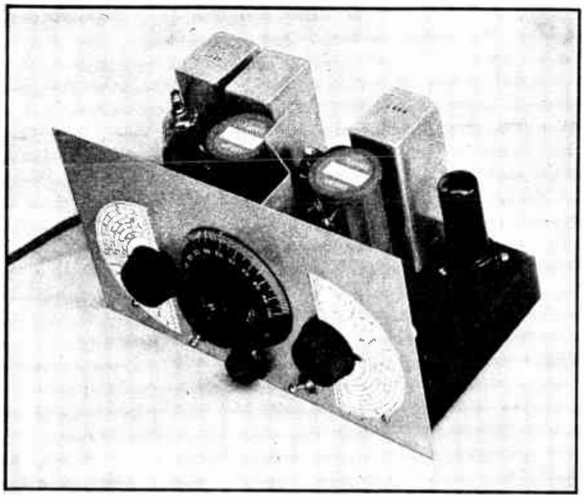


Fig. 1218 — The modified three-tube superheterodyne receiver with the audio amplifier stage added for loudspeaker operation.

For the purpose of driving the audio stage, resistance coupling is used from the plate of the second detector to the grid of the 6F6. A volume control is used for the grid resistor of the 6F6, and a jack is installed in the second-detector plate circuit so that a headphone plug may be inserted. The volume control, R_7, should be of the midget type so that it will fit in the chassis; it is installed with its shaft projecting under the tuning dial. In the bottom view, Fig. 1220, the 6F6 socket is in the upper left corner, along with the cathode resistor and by-pass condenser, R_8 and C_{15}. The coupling condenser, C_{14}, and the plate resistor, R_6, are mounted on an insulated lug strip near the volume control.

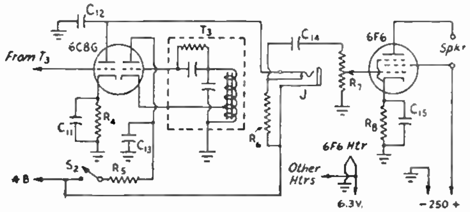


Fig. 1219 — Circuit diagram of the pentode audio-amplifier stage for loudspeaker operation of the three-tube superheterodyne. Except as noted below, values for components correspond to those bearing the same numbers in Fig. 1214.

C_{14} — 0.1-μfd. paper.
C_{15} — 25-μfd. electrolytic, 50 volts.
R_6 — 120,000 ohms, ½ watt.
R_7 — 500,000-ohm volume control.
R_8 — 400 ohms, 1 watt.
J — Closed-circuit jack.

The 6F6 will require a plate supply of 250 volts at about 40 milliamperes. This may be taken from a regular power pack, and a five-wire connection cable is used to provide an extra lead for the purpose. The first three tubes may be operated from a "B" battery, as before. Alternatively, the power supply may be constructed with a tap giving 90 or 100 volts for these tubes, the tap being connected to the proper wire in the connection cable. For best performance, the output voltage should be regulated by a VR105-30 regulator tube. A suitable power-supply circuit is shown in Fig. 1207-A. In this case the value of R should be 5000 ohms, 10 watts.

The primary winding of the speaker output transformer always should be connected in the plate circuit of the 6F6. Operation without the plate circuit closed is likely to damage the screen-grid. Any speaker having a transformer with a primary impedance of 7000 ohms will be satisfactory; a permanent-magnet dynamic is convenient, since no field supply is necessary.

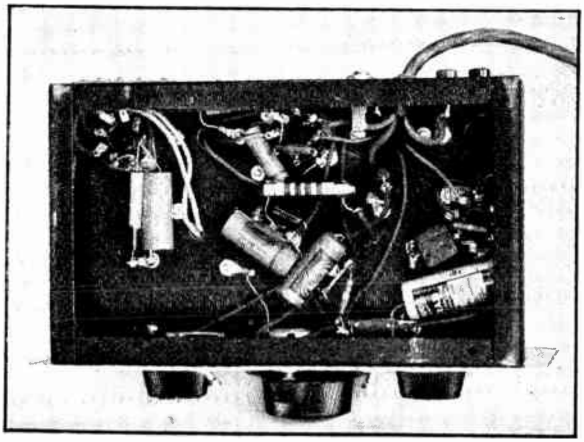


Fig. 1220 — The additional parts for the audio output stage can be identified in this sub-chassis view of the three-tube receiver.

A Regenerative Single-Signal Receiver

An inexpensive amateur-band receiver using i.f. regeneration for single-signal reception is shown in Fig. 1221. Fig. 1223 gives the circuit diagram. Regeneration also is used in the mixer circuit to improve the signal-to-image ratio and to give added gain. This receiver is designed to give the maximum of performance, in the hands of a capable operator, at minimum cost; selectivity, stability and sensitivity are the primary considerations.

The mixer, a 6SA7, is coupled to the antenna and is separately excited by a 6J5 oscillator. There is a single 460-kc. i.f. stage, using a 6SK7 and permeability-tuned transformers. The second detector and first audio amplifier is a 6SQ7, and the audio output tube for loudspeaker operation is a 6F6. The separate beat-oscillator circuit uses a 6C5. A VR105-30 voltage-regulator tube is used to stabilize the plate voltage on the oscillators and the screen voltage on the mixer and i.f. tubes.

To make construction easy and to avoid the necessity for additional trimmer condensers on each coil, the mixer and high-frequency oscillator circuits are separately tuned. Main tuning is by the oscillator bandspread condenser, C_3, which is operated by the calibrated dial. C_2 is the oscillator band-setting condenser. The mixer circuit is tuned by C_1. Regeneration in this circuit is controlled by R_{15}, connected across the mixer tickler coil, L_3.

R_{16} is the i.f.-amplifier gain control, which also serves as an i.f. regeneration control when this stage is made regenerative. C_{15} is the regeneration condenser; it is adjusted to feed back a small amount of i.f. energy from the plate to the grid of the 6SK7, and thus produce regeneration. If the high selectivity afforded by i.f. regeneration is not wanted, C_{15} may be omitted.

Diode rectification is used in the second-detector circuit. One of the two diode plates in the 6SQ7 is used for developing a.v.c. voltage, being coupled through C_{22} to the detector diode. The detector load resistor consists of R_5 and R_7 in series, the tap being used for r.f. filtering of the audio output to the triode section of the tube. R_{18} is the a.v.c. load resistor; R_9, C_{14} and C_{12} constitute the a.v.c. out of circuit by grounding the rectifier output. The headphones are connected in the plate circuit of the triode section of the 6SQ7. R_{17} is the audio volume control potentiometer.

The top and bottom views, Figs. 1222 and 1224, show the layout clearly. The bandspread tuning condenser, C_3, is at the front center; at the left is C_1, the mixer tuning condenser; and at the right, C_2, the oscillator band-set condenser. The oscillator tube is directly behind C_3, with the mixer tube to the left on the other side of a baffle shield which separates the two r.f. sections. This shield, measuring $4\frac{1}{4} \times 4\frac{1}{2}$ inches, is used to prevent coupling between oscillator and mixer. The mixer coil socket is at the left edge of the chassis behind C_1; the oscillator coil socket is between C_2 and C_3.

The i.f. and audio sections are along the rear edge of the chassis. The transformer in the rear left corner is T_1; next to it is the i.f. tube, then T_2. Next in line is the 6SQ7, followed by the 6C5 beat oscillator, the b.o. transformer, T_3, and finally the 6F6. The VR105-30 is just in front of T_3. The i.f. transformers should be mounted with their adjusting screws projecting to the rear where they are easily accessible.

The beat oscillator is coupled to the second detector by the small capacity formed by running an insulated wire from the grid of the 6C5 close to the detector diode plate prong on the 6SQ7 socket. Very little coupling is needed for satisfactory operation.

In wiring the i.f. amplifier, keep the grid and plate leads from the i.f. transformers fairly close to the chassis and well separated. Without C_{15}, the i.f. stage should be perfectly stable and should show no tendency to oscillate at full gain.

The method of winding the plug-in coils is shown in Fig. 1225, and complete specifications are given in the coil table. Ticklers (L_3) for the mixer circuit are scramble-wound to a diameter which will fit readily inside the coil form and mounted on stiff leads going directly

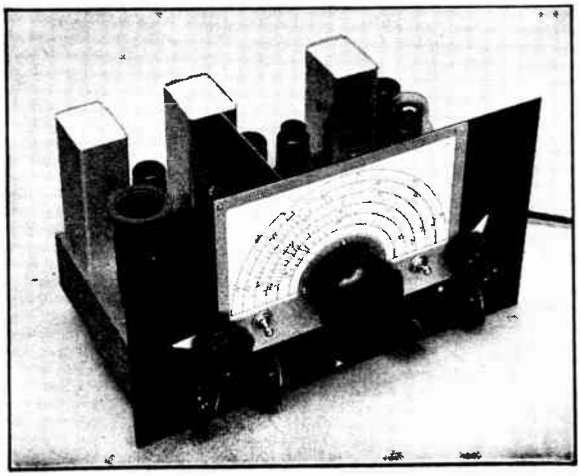


Fig. 1221 — A 7-tube superheterodyne using regeneration in the i.f. amplifier to give single-signal reception and improved image ratio. The dial (National ACN) may be directly calibrated for each amateur band. The chassis is $11 \times 7 \times 2$ inches and the panel 7×12 inches. The controls along the bottom edge of the panel are, from left to right, the mixer regeneration control, R_{15}, the i.f. gain control, R_{16}, the audio volume control, R_{17}, and the beat-oscillator vernier condenser, C_{21}. The latter has the corner of one rotor plate bent over so that when the condenser plates are fully meshed the tuned circuit is short-circuited, thus stopping the h.f.o. oscillation.

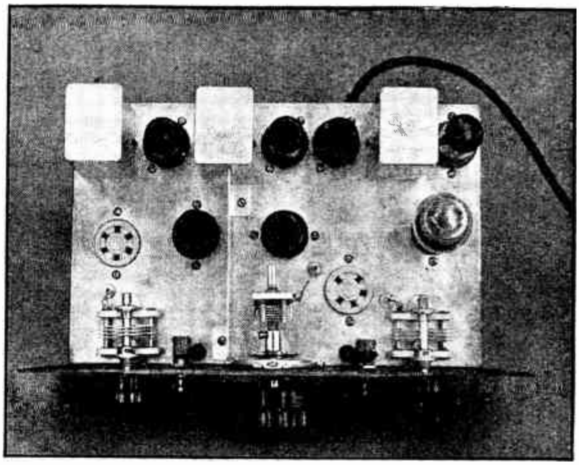


Fig. 1222 — Top view of the 7-tube superheterodyne with plug-in coils removed. Placement of the parts is discussed in the text.

to the proper pins in the form. The leads should be long enough to bring the coils inside the grid winding at the bottom. The amount of feed-back is regulated by bending the tickler coil with respect to the grid coil. Maximum feed-back is secured with the two coils coaxial, minimum when the tickler axis is at right angles to the axis of L_1. The position of L_3 should be adjusted so that the mixer goes into oscillation with R_{15} set at one-half to three-fourths of its maximum resistance.

Alignment — The oscillator circuit has been adjusted to make the proper value of rectified grid current flow in the 6SA7 injection-grid (No. 1) circuit on each amateur band. This calls for a fairly strong value of feed-back, with the result that when the band-set condenser is set toward the high-frequency end of its range the oscillator may "squeg." This is of no consequence unless the receiver is to be used for listening outside the amateur bands, in which case it may be corrected by taking a few turns off the tickler coil, L_6. This can be done only at some sacrifice of conversion efficiency in the amateur band for which the coil was designed, however.

The i.f. amplifier can be aligned most conveniently with the aid of a modulated test oscillator. The initial alignment should be made with C_{15} disconnected so that the performance of the amplifier in a non-regenerative condition can be checked. Headphones or a loudspeaker may be used as an output indicator. The mixer and oscillator coils should be out of their sockets, and R_{15} should be set at zero resistance.

Connect the test oscillator output across C_1, which should be set at minimum capacity. Adjust the test-oscillator frequency to 460 kc. Then, using a modulated signal, adjust the trimmers on T_1 and T_2 for maximum volume.

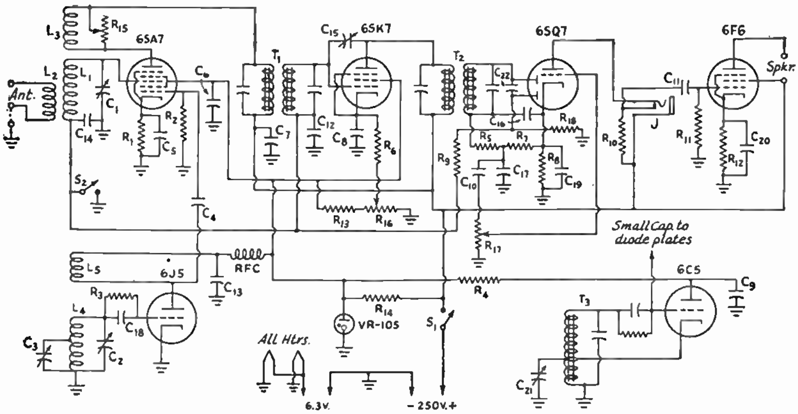


Fig. 1223 — Circuit diagram of the single-signal superheterodyne receiver with regenerative i.f. and mixer stages.

C_1, C_2 — 50-$\mu\mu$fd. variable (Hammarlund MC-50-S).
C_3 — 35-$\mu\mu$fd. variable (National UM-35).
C_4 — 50-$\mu\mu$fd. mica.
C_5, C_6, C_7, C_8 — 0.1-μfd. paper, 600 volts.
C_9, C_{10}, C_{11}, C_{12} — 0.01-μfd. paper, 600 volts.
C_{13}, C_{14} — 0.005-μfd. mica.
C_{15} — 3-30-$\mu\mu$fd. trimmer (National M-30); see text.
C_{16} — 250-$\mu\mu$fd. mica.
C_{17}, C_{18}, C_{22} — 100-$\mu\mu$fd. mica.
C_{19}, C_{20} — 25-μfd. electrolytic, 50 volts.
C_{21} — 25-$\mu\mu$fd. variable (Hammarlund SM-25).
R_1 — 200 ohms, ½ watt.
R_2 — 20,000 ohms, ½ watt.
R_3, R_4, R_5 — 50,000 ohms, ½ watt.
R_6 — 300 ohms, ½ watt.
R_7 — 0.2 megohm, ½ watt.
R_8 — 2000 ohms, ½ watt.
R_9 — 1 megohm, ½ watt.
R_{10} — 0.1 megohm, ½ watt.
R_{11} — 0.5 megohm, ½ watt.
R_{12} — 450 ohms, 1 watt.
R_{13} — 75,000 ohms, 1 watt.
R_{14} — 5000 ohms, 10 watts.
R_{15} — 10,000-ohm volume control (mixer regeneration).
R_{16} — 25,000-ohm volume control (i.f. regeneration).
R_{17} — 2-megohm volume control.
R_{18} — 2 megohms, ½ watt.
T_1 — 460-kc. permeability-tuned i.f. transformer, interstage type (Millen 64456).
T_2 — 460-kc. permeability-tuned i.f. transformer, diode type (Millen 64454).
T_3 — 460-kc. beat-oscillator transformer (Millen 65456).
RFC — 2.5-mh. r.f. choke.
J — Closed-circuit jack.
S_1, S_2 — S.p.s.t. toggle.
L_1–L_6, inc. — See coil table.

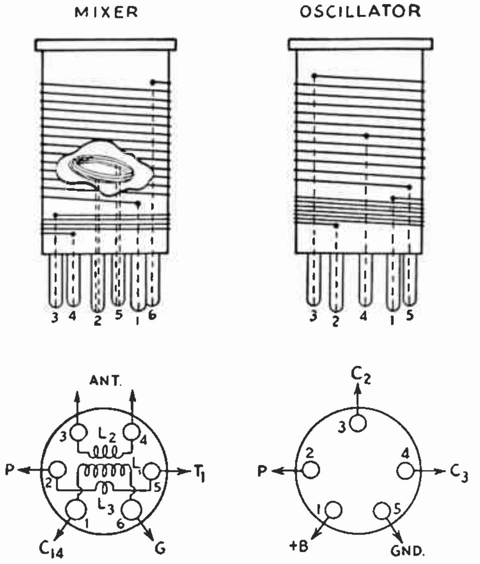


Fig. 1225 — **Mixer and oscillator coil and socket connections for the seven-tube superheterodyne receiver.**

R_{16} should be set for maximum gain, and the beat oscillator should be off. As the successive circuits are brought into line, reduce the oscillator output to keep from overloading any of the amplifiers, since overloading might cause a false indication.

After the i.f. is aligned, plug in a set of coils for some band on which there is a good deal of activity. Set the oscillator padding condenser, C_2, at approximately the right capacity; with the coil specifications given, the proportion of the total capacity of C_2 in use on each band will be about as follows: 1.75 Mc., 90 per cent; 3.5 Mc., 75 per cent; 7 Mc., 95 per cent; 14 Mc., 90 per cent; 28 Mc., 45 per cent. Set the mixer regeneration control, R_{15}, for minimum regeneration — i.e., with no resistance left in the circuit.

Now connect an antenna to the input terminals for L_2. Switch the beat oscillator on by turning C_{21} out of the maximum position, and adjust the trimmer screw on T_3 until the characteristic beat-oscillator hiss is heard.

Next tune C_1 slowly over its scale, starting from maximum capacity. Using the 7-Mc. coils as an example, when C_1 is at about half scale there should be a definite increase in the noise level as well as in the strength of the signals which may be heard. Continue on past this point toward minimum capacity until a second peak is reached on C_1; at this peak the input circuit is tuned to the frequency which represents an image in normal reception. The oscillator in the receiver is designed to work on the high-frequency side of the incoming signal, so that C_1 always should be tuned to the peak which occurs with most capacity.

After the signal peak on C_1 has been identified, tune C_3 over its whole range, following with C_1 to keep the mixer circuit in tune, to see how the band fits the dial. With C_2 properly set, the band edges should fall the same number of main dial divisions from 0 and 100; if the band runs off the low-frequency edge, less capacity is needed at C_2, while the converse is true if the band runs off the high edge. Once the band is properly centered on the dial, the panel may be marked at the appropriate point so that C_2 may be reset readily when changing bands.

To check the operation of the mixer regeneration, tune in a signal on C_3, adjust C_1 for maximum volume, and slowly advance the regeneration control, R_{15}. As the resistance is increased, retune C_1 to maximum volume, since the regeneration control will have some effect on the mixer tuning. As regeneration is increased signals and noise both will become louder, and C_1 will tune more sharply. Finally the mixer circuit will break into oscillation and, when C_1 is right at resonance, a loud carrier will be heard, since the oscillations generated will go through the receiver in exactly the same way as an incoming signal. As stated before, oscillation should occur with R_{15} set at from one-half to three-quarters full scale. In practice, it is best always to work with the mixer somewhat below the critical regeneration point and never permit it actually to oscillate. On the lower frequencies, where images are less serious, the tuning is less critical if the mixer is made non-regenerative. In this case, always set the regeneration control at zero, since there will be a range on the resistor where, without definite regeneration, the signal strength will be less than it is with zero resistance.

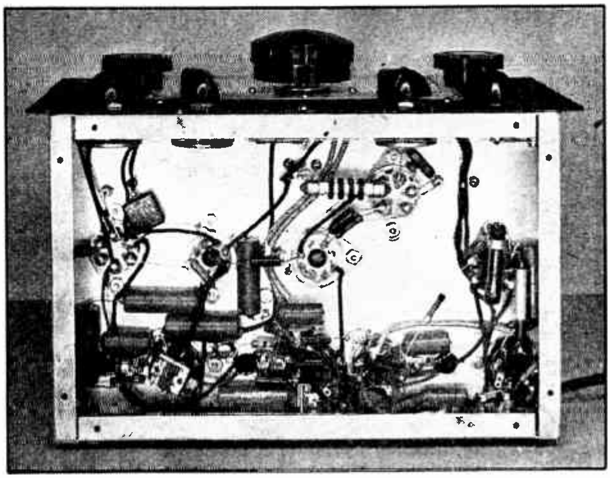


Fig. 1224 — **The below-chassis wiring and location of parts is shown in this bottom view of the seven-tube regenerative single-signal receiver.**

Should the mixer fail to oscillate, adjust the coupling by changing the position of L_3 with respect to L_1. If the two coils happen to be "poled" incorrectly, the circuit will not oscillate. This condition can be cured by rotating L_3 through 180 degrees. It is recommended that the mixer regeneration be tested first with the antenna disconnected, since antenna loading effects may give misleading results until it is known that L_3 is properly adjusted to produce oscillation.

After the preceding adjustments have been completed the i.f. regeneration may be added. Install C_{15}, taking out the adjusting screw and bending the movable plate to make an angle of about 45 degrees with the fixed plate. Realign the i.f. As the circuits are tuned to resonance the amplifier will oscillate, and each time this happens the gain control, R_{16}, should be backed off until oscillations cease. Adjust the trimmers to give maximum output with the lowest setting of R_{16}. At peak regeneration the signal strength should be about the same with this setting, despite reduced gain in the amplifier, as it is without regeneration at full gain. Too much gain with regeneration will have an adverse effect on the selectivity.

For single-signal c.w. reception, set the beat oscillator so that, when R_{16} is advanced to make the i.f. stage just go into oscillation, the resulting tone is the desired beat-note frequency. Then back off on R_{16} to obtain the desired degree of selectivity. Maximum selectivity will be secured with the i.f. amplifier just below the oscillating point. The "other side of zero beat" will be much weaker than the desired side.

COIL DATA FOR 7-TUBE SUPERHET

Band	Coil	Wire Size	Turns	Length	Tap
1.75 Mc.	L_1	24	70	Close-wound	—
	L_2	24	15	" "	—
	L_3	22	15	——	—
	L_4	22	42	Close-wound	Top
	L_5	24	15	" "	—
3.5 Mc.	L_1	22	35	" "	—
	L_2	22	9	" "	—
	L_3	22	12	——	—
	L_4	22	25	1 inch	18
	L_5	22	10	Close-wound	—
7 Mc.	L_1	18	20	1 inch	—
	L_2	22	5	Close-wound	—
	L_3	22	9	——	—
	L_4	18	14	1 inch	6
	L_5	22	6	Close-wound	—
14 Mc.	L_1	18	10	1 inch	—
	L_2	22	5	Close-wound	—
	L_3	22	7	——	—
	L_4	18	7	1 inch	2.4
	L_5	22	4	Close-wound	—
28 Mc.	L_1	18	4	1 inch	—
	L_2	22	4	Close-wound	—
	L_3	22	1.5	——	—
	L_4	18	3.6	1 inch	1.4
	L_5	22	2.4	Close-wound	—

All coils except L_3 are 1½ inches in diameter, wound with enameled wire on Hammarlund SWF forms. Spacing between L_1 and L_2, and between L_4 and L_5, is approximately ⅛ inch. Bandspread taps are counted from bottom (ground) end of L_4.

L_3 for 28 Mc. is interwound with L_1 at the bottom end. L_3 for all other coils is self-supporting, scramble-wound to a diameter of ¾ inch, mounted inside the coil form near the bottom of L_1.

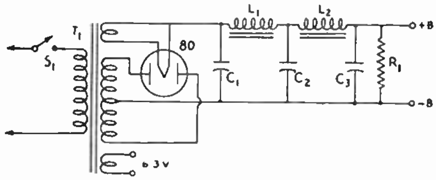


Fig. 1226 — Power-supply for the regenerative superhet.

C_1, C_2 — 8-μfd. electrolytic, 450 volts.
C_3 — 16-μfd. electrolytic, 450 volts.
R_1 — 25,000 ohms, 10 watts.
L_1, L_2 — 12 henrys, 80 ma., 400 ohms.
T_1 — 350 volts each side of center-tap, 80–90 ma.; 6.3 volts at 2.5 amperes or more; 5-volt 2-ampere rectifier-filament winding.
S_1 — S.p.s.t. toggle switch.

Dual-unit electrolytic condensers may be used. This supply will give 275 to 300 volts with full receiver load.

A useful feature of the bandspread dial is that it can be directly calibrated in frequency for each band. These calibrations may be made with the aid of a 100-kc. oscillator, such as is described in Chapter Twenty. Ten-kilocycle points can be plotted if a 10-kc. multivibrator is available, but, since the tuning is almost linear in each band, a fairly accurate plot will result if each 100-kc. interval is simply divided off into ten equal parts when the dial calibrations are marked.

The power-supply requirements for the receiver are 2.2 amperes at 6.3 volts for the heaters and 80 ma. at 250 volts for the plates. Without the 6F6 pentode output stage, a supply giving 6.3 volts at 1.5 amperes and 250 volts at 40 ma. would be sufficient. The circuit of a suitable power supply is given in Fig. 1226.

A 12-Tube Crystal-Filter Receiver

The 12-tube crystal-filter, single-signal super-heterodyne receiver shown in the photographs of Figs. 1227, 1228 and 1230 was built by W4CBD. It is representative of the more elaborate amateur-constructed receivers. The circuit diagram is shown in Fig. 1229.

The r.f. section consists of two stages of tuned r.f. amplification using 6SK7s, a 6SA7 mixer and a 6J5 h.f. oscillator. The tuning condensers of these stages, C_1, C_2, C_3 and C_4 are ganged together and operated by the main tuning dial. The two r.f. stages are similar except that the first stage is not tied into the a.v.c. circuit. While the first tube runs at maximum gain all the time, a grid resistor inserted in the ground return protects the tube against strong r.f. fields. A.v.c. is applied only to the second r.f. tube and the mixer. This provides sufficient a.v.c. action while it also produces a greater deflection of the signal meter than would be obtained with more stages tied to the a.v.c. line. The manual r.f. gain control, R_{11}, controls all stages except the second r.f. stage. C_{18} is used to neutralize the space-charge coupling between No. 1 grid and the signal grid of the mixer tube.

For general coverage the tuning condensers in the r.f. and mixer stages are connected across the entire coil. C_5, C_6 and C_7 are air trimmers ganged to one of the small controls along the lower part of the panel. Since the stray capacities in the mixer stage are slightly higher than in the r.f. stages, C_8 and C_9 are added to permit compensation. Two sets of coils are required to cover the frequencies between one amateur band and the next. Although this means quite a few coils, it provides a good degree of bandspread for even the general-coverage ranges.

When bandspread tuning is desired, the main tuning condensers are tapped down on the coils of the r.f. and mixer stages by a switching system in the bottom of the coil form, as shown in the detail photograph and the sketch of Fig. 1231. This connection would cause considerable non-linearity in calibration, with crowding at one end of the scale, were it not for the fixed padder condensers C_{15}, C_{16}, C_{17}, and C_{10}. C_{10} is an air-insulated condenser which is mounted inside the oscillator shield compartment. After it is set initially to about 25 μμfd., no further adjustment is required. The other condensers are mica units, especially selected for equal capacities.

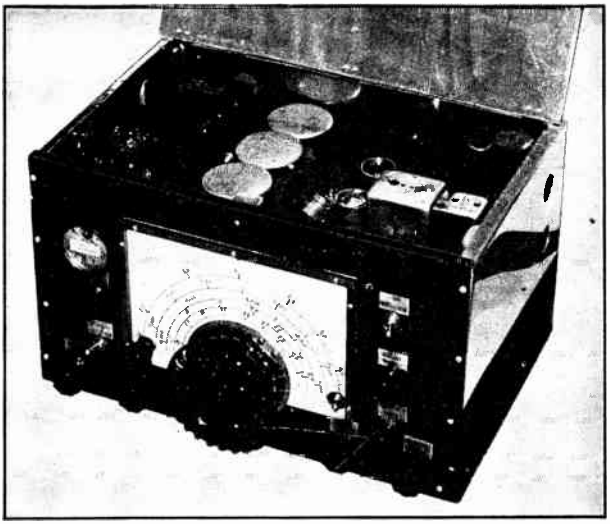


Fig. 1227 — Panel view of the completed 12-tube receiver. Below the "S" meter are the b.f.o. tuning control, the power switch, the stand-by switch and the r.f. gain control. The ganged-trimmer control is in the lower left-hand corner of the dial chart. To the right are the b.f.o. switch, the audio gain control and the crystal-filter and noise-silencer adjusting knobs. The a.v.c. switch is in the lower right-hand corner of the dial chart. This view shows the shield cans in place over the coils.

The mixer output transformer feeds the grids of the first i.f. stage and the 6J7 noise-amplifier stage in parallel while the crystal filter is coupled to the output of the 6L7. The 6J7 amplifies the noise and the 6H6 rectifies the noise and applies the d.c. impulse to the injector grid of the 6L7, cutting it off for the duration of the noise impulse. R_{20} provides the threshold adjustment. The output of the crystal filter is coupled to the input of the second i.f. amplifier which feeds the diode second detector.

The 6H6 second detector is connected so that one section handles the audio signal while the other section supplies a.v.c. voltage. In this arrangement a bias of several volts is placed on the a.v.c. side, since the cathode of the 6H6 is returned to the 6J5 audio cathode rather than

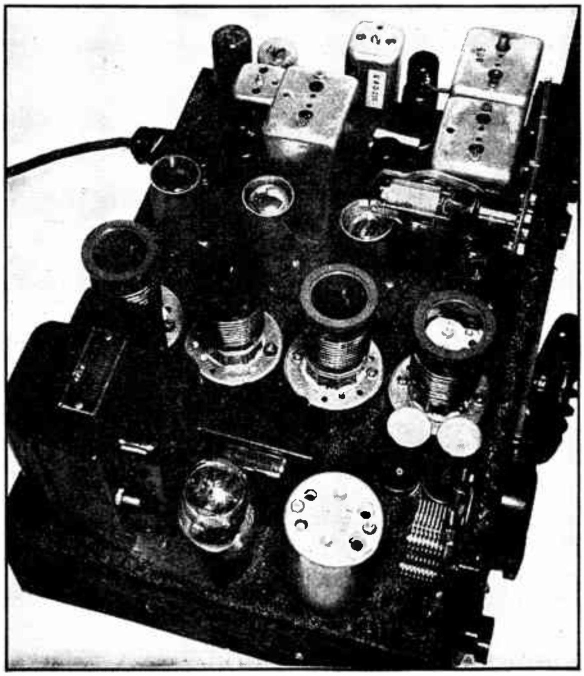


Fig. 1228 — Plan view of the crystal-filter, single-signal superhet receiver. To the left, from front to rear are the "S" meter, filter condensers, b.f.o. tube and tank circuit (T_8), filter choke, rectifier tube and power transformer. The line of empty sockets are for the plug-in coils of the first and second r.f. stages, the mixer and the h.f. oscillator in order from front to rear. The parallel line of corresponding tubes is to the right. The h.f. oscillator tube is hidden by the voltage-regulator tube mounted on the panel. The two shielded transformers to the right near the panel are the crystal-filter input and output transformers, T_4 and T_5. In line in back of T_4 are the 6L7, the first i.f. transformer, T_3, the 6J7 and the 6H6 noise rectifier. Along the right-hand edge of the chassis, from front to back, are the 6K7, the diode coupling transformer, T_6, the 6H6 second detector and the two audio tubes. Between the two lines are the crystal and noise-silencer transformer, T_7.

to ground. Because the 6J5 cathode is above ground for d.c., no a.v.c. action is obtained until the signal level exceeds the bias. Thus a.v.c. action causes no reduction in sensitivity for weak signals. The delayed a.v.c. effect can be further manipulated by adjustment of the r.f. and audio gain controls.

The beat-oscillator circuit is similar to that used in the h.f. oscillator. It is operated at a fairly low level and the output to the diode detector is taken from the cathode. Thorough shielding of the lead to the 6H6 is important, since it is about 24 inches long. The tuning condenser, C_{14}, is connected from cathode to

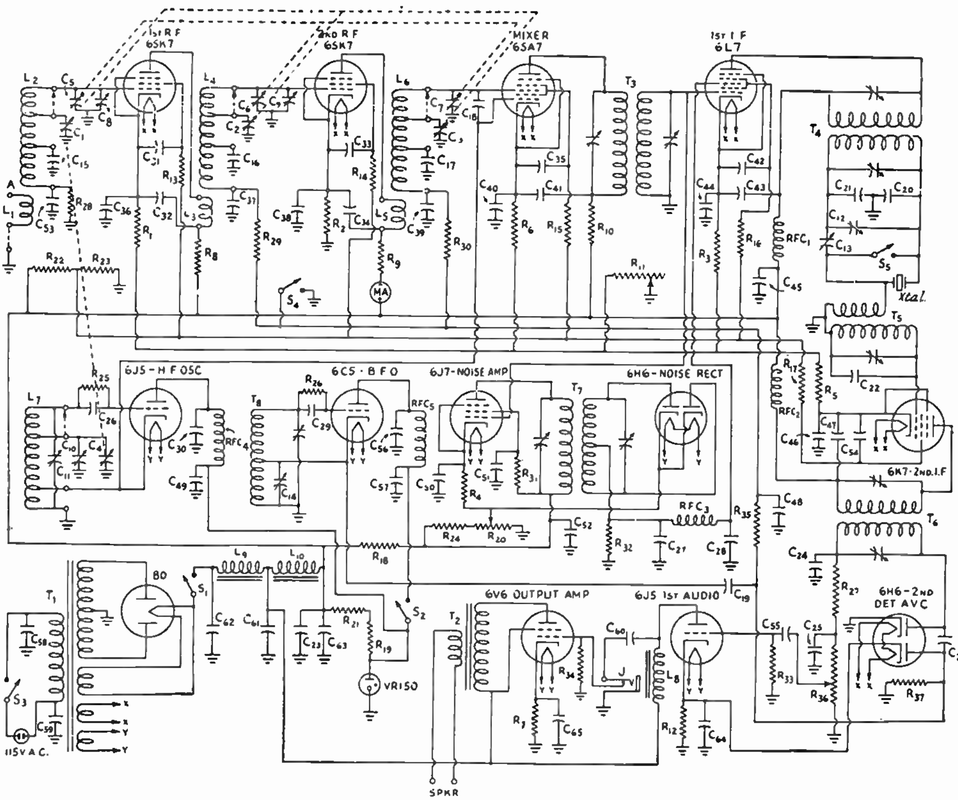


Fig. 1229 — Circuit diagram of the ham-band receiver.

C_1, C_2, C_3, C_4 — 50-μμfd. ganged tuning condensers.
C_5, C_6, C_7 — 15-μμfd. ganged r.f. and mixer trimmers.
C_8, C_9 — 15-μμfd. variable (stray-capacity equalizer).
C_{10} — 50-μμfd. variable air padder (see text).
C_{11} — Oscillator padder inside L_7 (see coil table).
C_{12} — 50-μμfd. variable (crystal selectivity control).
C_{13} — 15-μμfd. variable (rejection control).
C_{14} — 140-μμfd. variable (b.o. tuning control).
C_{15}, C_{16}, C_{17} — 25-μμfd. fixed mica padder.
C_{18} — Approximately 1 μμfd. (twisted insulated leads).
C_{19} — 10-μμfd. mica.
C_{20}, C_{21}, C_{22} — 50-μμfd. mica.
C_{23}, C_{24}, C_{25}, C_{26}, C_{27}, C_{28} — 100-μμfd. mica.
C_{29} — 0.001-μfd. mica.
C_{30}, C_{31}, C_{32}, C_{33}, C_{34} — 0.002-μfd. paper, 600 volts.
C_{35}, C_{36} C_{37}, C_{38}, C_{39}, C_{40}, C_{41}, C_{42}, C_{43}, C_{44}, C_{45}, C_{46} C_{47}, C_{48}, C_{49}, C_{50}, C_{51} — 0.01-μfd. paper.
C_{52}, C_{53}, C_{54}, C_{55}, C_{56}, C_{57}, C_{58}, C_{59}, C_{60} — 0.1-μfd. paper.
C_{61}, C_{62}, C_{63} — 8-μfd. electrolytic, 450 volts.
C_{64}, C_{65} — 40-μfd. electrolytic, 25 volts.
R_1, R_2, R_3 — 250 ohms, 1 watt.
R_4, R_5 — 400 ohms, 1 watt.
R_6 — 500 ohms, 1 watt.
R_7 — 500 ohms, 10 watts, wire-wound.
R_8, R_9, R_{10} — 1000 ohms, 1 watt.
R_{11} — 1000-ohm r.f. gain control, wire-wound.
R_{12} — 1500 ohms, 1 watt.
R_{13}, R_{14}, R_{15}, R_{16}, R_{17}, R_{18} — 2000 ohms, 1 watt.
R_{19} — 5000 ohms, 10 watts, wire-wound.
R_{20} — 5000-ohm potentiometer (silencer gain control).
R_{21} — 7000 ohms, 10 watts, wire-wound.
R_{22} — 10,000 ohms, 10 watts, wire-wound.
R_{23} — 15,000 ohms, 10 watts, wire-wound.
R_{24} — 20,000 ohms, 1 watt.
R_{25}, R_{26}, R_{27} — 50,000 ohms, ½ watt.
R_{28}, R_{29}, R_{30}, R_{31}, R_{32} — 100,000 ohms, 1 watt.
R_{33}, R_{34} — 500,000 ohms, ½ watt.
R_{35} — 1 megohm, ½ watt.
R_{36} — 1-megohm audio gain control.
R_{37} — 2 megohms, ½ watt.
L_1, L_2, L_3, L_4, L_5, L_6, L_7 — See coil table.
L_8 — Audio coupling impedance (primary winding of audio transformer).
L_9 — Filter choke (Thordarson T-849C41).
T_1 — Power transformer (Thordarson T-87R85).
T_2 — Speaker output transformer, universal type.
T_3 — 465-kc. air-tuned i.f. transformer.
T_4 — 465-kc. i.f. transformer altered (see text).
T_5 — 465-kc. b.f.o. unit (see text).
T_6 — 465-kc. diode input transformer.
T_7 — 465-kc. diode input transformer (see text).
T_8 — 465-kc. b.f.o. unit.
RFC_1, RFC_2, RFC_3 — Approximately 11 mh. (replacement 175-kc. i.f. coil).
RFC_4, RFC_5 — 2.5-mh. r.f. choke.
S_1 — S.p.d.t. switch.
S_2, S_3, S_4 — S.p.s.t. switch.
S_5 — Crystal-filter switch (see text).
J — Double-circuit jack.
M — Signal-strength meter (7-ma. movement).

ground to keep the r.f. voltage across it low and thus minimize pickup in neighboring r.f. circuits. This connection makes it necessary to use the unusually large capacity of 140 μμfd. to cover the desired frequency range. The amount of oscillator voltage fed into the detector is low enough so that good limiting of volume on c.w. signals is obtained, and the hiss level is low.

A power-supply unit is built into the receiver. High voltage for the audio section is taken from a tap at the output of the first filter section. The field of the external speaker is used as the inductance in an additional filtering section for the voltage which supplies the other stages. The VR150 voltage-regulator tube holds the plate voltage for the two oscillators constant.

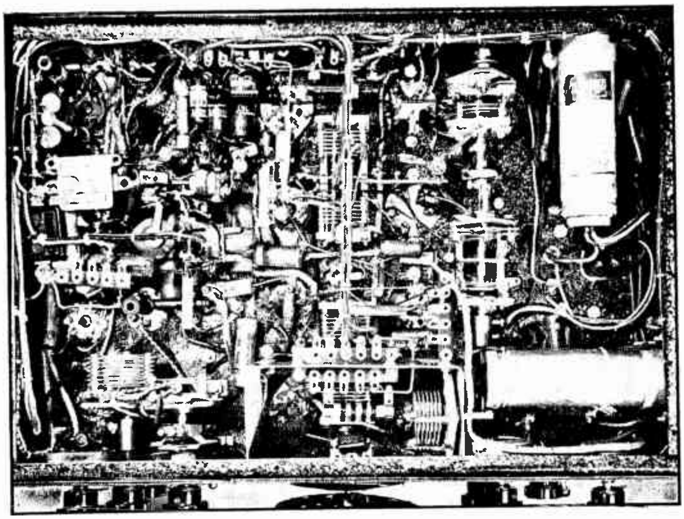


Fig. 1230 — Bottom view of the twelve-tube superheterodyne. The four-gang tuning-condenser is at the center. The ganged trimmers, C_5, C_6 and C_7, are at the right, while the crystal-filter selectivity control, C_{12}, and phasing condenser, C_{13}, are near the front at the left. C_8 and C_9 are fastened to C_5 and C_6, respectively. A separate shielded compartment at the center contains the h.f. oscillator components, with C_{10} at the right. By-pass condensers are placed close to the points to be by-passed and other small parts are placed in the nearest available space. The electrolytic units at the right are filter condensers.

Construction — The chassis is formed from 0.050-inch sheet steel. Reinforcing braces were spot welded in the corners and L-shaped strips are added along the bottom edges of the chassis for reinforcement and to form a shelf to which the bottom cover could be attached with sheet-metal screws. The cover plate is equipped with rubber mounting feet, one at each corner. The panel is formed from 0.062-inch sheet steel. A ½-inch edge with a slight radial bend was formed along the top. All holes are drilled first, and later everything is given a thick coat of baked-on crackle enamel.

The tuning control, built around a National Velvet Vernier dial mechanism, is similar to the ACN model except that it is larger. The ACN may be substituted if available. The handwheel is a 4-inch valve-control knob.

The general lay-out plan of the receiver is shown quite clearly in the photographs. The main essential is, of course, the close grouping of components in the high-frequency stages. All parts, especially those forming the various tuned circuits, should be mounted with good mechanical anchoring to prevent any slight movement which might cause a noticeable change in frequency. Care should be exercised in lining up the units of the ganged condensers so that they will not spring when the shaft is turned. All r.f. wiring should be made as short as possible and kept well spaced from the chassis. Power wiring may be cabled and laid flat against the chassis wherever it is convenient to do so.

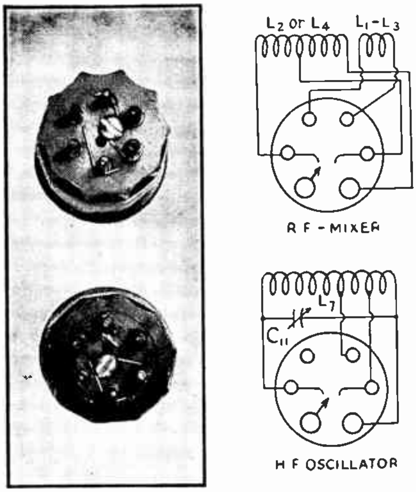


Fig. 1231 — Pin connections for the r.f. and h.f.o. plug-in coils for the 12-tube superhet.

The bottom views of the coil forms in the photograph show the bandspread switching arrangement. The screw head completes the connection between either pair of pins, depending upon its position (see sketch at right).

Coils — All of the coils in each set shown in the table are wound to be as nearly identical as possible. The r.f. and mixer coils are then adjusted to exactly the same inductance by spacing the turns and heavily doping all but one or two turns at one end with clear nail polish. When the dope has set, a further adjustment may be made by moving the free turns on the end and then cementing them firmly in place. The inductance of the coils can be checked by interchanging two coils at a time in the r.f. stages. If there is a difference in inductance, the stray-capacity equalizers, C_8 and C_9, will have to be readjusted when the coils are interchanged. When the inductance of the three coils is adjusted correctly, it should be possible to place the coils in the three positions in any sequence without necessity for readjustment of any trimmer to restore resonance.

The i.f. coupling transformers are modified to fit the circuits. About one-fourth of the turns should be removed from the secondary of T_4, and C_{20} and C_{21} are mounted inside the shield can. The primary and secondary windings should be pushed a little closer together. T_5 is a b.f.o. unit. The tickler winding should be replaced with a 100-turn coil of No. 34 enameled wire, which becomes the new primary. The two windings are placed close together for tight coupling. A 50-$\mu\mu$fd. fixed condenser, C_{22}, must be added to the secondary to hit resonance at 465 kc. An auxiliary brass contact is added to C_{13}, so that the crystal may be shorted out for straight operation. T_6 is tuned by a mica trimmer, but may be replaced by an air-insulated trimmer if drift is excessive.

Adjustment — With the minimum and stray capacities in each stage set at the same value, it is easy to secure good tracking of the r.f. circuits. It is necessary for them to track accurately, since the over-all selectivity of the three resonant r.f. circuits is high. If one of the circuits is detuned by moving a trimmer 2 or 3 $\mu\mu$fd. away from resonance, the signal meter will indicate a drop of several db.

When the adjustment of C_{18} is correct, there is no observable interaction between the oscillator and mixer tuning. Should there be any, the bias on the signal grid should be checked. It should be at least 5 volts.

If during the adjustment of the crystal filter it is found that the rejection control allows rejection of interference on one side of the desired signal but not on the other it may be necessary to add a little capacity, consisting of a pair of twisted wires across the crystal holder, to get the rejection slot to move to the other side of the signal.

COIL TABLE FOR THE 12-TUBE SUPERHETERODYNE

Band	*Coil*	*Turns*	*Wire Size*	*Cathode Tap*	*B.S. Tap*
1.7–2.4 Mc.	L_2, L_4, L_6	60	26 d.c.c.	x	x
	L_1, L_3, L_5	8	28 enam.	x	x
	L_7	51	26 d.c.c.	6	47
2.7–4 Mc. or	L_2, L_4, L_6	42	22 d.c.c.	x	24
3.5–4 Mc.	L_1, L_3, L_5	8	28 enam.	x	x
	L_7	37	22 d.c.c.	5	20
3.4–4.8 Mc.	L_2, L_4, L_6	30	22 d.c.c.	x	x
	L_1, L_3, L_5	4	22 d.c.c.	x	x
	L_7	25	22 d.c.c.	4	x
4.8–7.2 Mc. or	L_2, L_4, L_6	19	22 d.c.c.	x	6¾
7.0–7.3 Mc.	L_1, L_3, L_5	4	22 d.c.c.	x	x
	L_7	15	22 d.c.c.	3	5¾
7.0–10 Mc.	L_2, L_4, L_6	14	22 d.c.c.	x	x
	L_1, L_3, L_5	4	22 d.c.c.	x	x
	L_7	12½	22 d.c.c.	3	x
10–14.2 Mc. or	L_2, L_4, L_6	10⅓	16 bare	x	4
14.0–14.4 Mc.	L_1, L_3, L_5	4	22 d.c.c.	x	x
	L_7	9¾	16 bare	2¾	3
22–30 Mc.	L_2, L_4, L_6	5	16 bare	x	x
	L_1, L_3, L_5	4	22 d.c.c.	x	x
	L_7	4½	16 bare	2	x

NOTE: All coils are close-wound on 1½-inch diameter forms except L_2, L_4, L_6 and L_7 for the 10- to 14.2-Mc. range, where the turns are spaced the diameter of the wire, and the same coils for the 22- to 30-Mc. range, where the turns are spaced to make the coil length 1¼ inches. Taps are made the specified number of turns from the bottom or ground ends of the windings.

The Panoramic Receiver

The panoramic receiver incorporates two signal channels, the channel for audio-output signals normal to a communications-type receiver and an additional channel for reproducing the received signal in visual form on the screen of a cathode-ray tube. The effective acceptance bandwidth of the channel for visual signals usually is made much wider than the channel for audio output — 50 to 100 kc. or more, so that it is possible to observe simultaneously signals over a wide frequency range without destroying the high selectivity for signals delivered to the audio amplifier.

The circuit diagram of an adapter which may be applied to any existing communications-type receiver is shown in Fig. 1234. The signal input to the adapter is taken from the output of the mixer in the receiver. It then passes through a broadly-tuned i.f. stage (6SJ7) at the receiver's i.f. and thence into a second mixer (6SA7) whose output is tuned to 100 kc. The tuning of the oscillator section of this stage is varied over a range of 50 kc. either side of 356 kc. (456 kc., the usual receiver i.f. minus 100 kc.) at a supervisible rate — 25 to 30 times per second — by the 6AC7 reactance modulator. The reactance tube and the horizontal sweep of the 902 cathode-ray tube are driven in synchronism by the 7F7 saw-tooth oscillator. C_{18} is for the purpose of adjusting the phasing between the r.f. plate current and tank voltage of the 6AC7 to the desired value of 90 degrees.

Since the tuning of the r.f. circuits at the front end of the receiver is not swept in this system, the first i.f. stage in the adapter is designed to give a rising characteristic either side of the center frequency to compensate as much as possible for the decreasing characteristic introduced by the selectivity of the r.f. stages. This is done by overcoupling in the input and output transformers of the first 6SJ7 stage.

The 6SA7 mixer is followed by a 6SJ7 i.f. stage tuned to 100 kc. This stage is tuned

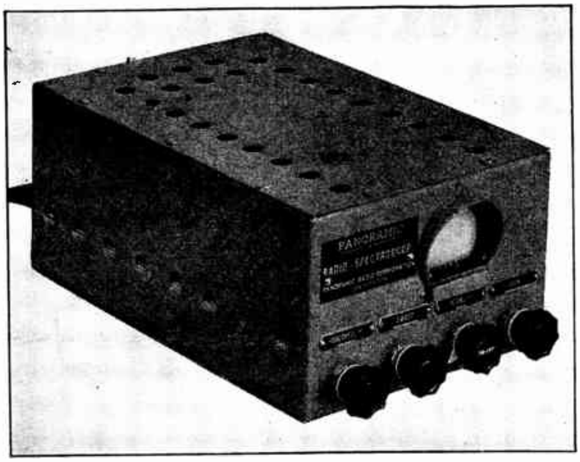


Fig. 1232 — A commercial panoramic adapter. The cathode-ray tube is provided with a scale calibrated in kc. The four controls are for horizontal positioning (R_{19}), sweep (R_{35}), intensity (R_{47}) and gain (R_2). The controls for vertical positioning (R_{17}) and focusing (R_{22}) are mounted in the rear and adjustable by screwdriver.

quite sharply. Thus, while the preceding stages are broadly tuned to cover the "sweep" range, the "instantaneous" selectivity is controlled by this 6SJ7 stage. This selectivity controls the definition or sharpness of the individual signal patterns on the screen.

The signal is rectified and amplified in the 6SQ7 and then applied to the vertical deflection plates of the cathode-ray tube. A typical pattern showing several signals of different amplitudes is shown in Fig. 1235.

Plate and screen voltages for all of the tubes in the adapter as well as anode voltages for the cathode-ray tube are obtained from a voltage-doubling circuit using a 117Z6GT rectifier. The screen voltage for the 6AC7 is held constant by the neon voltage-regulator tube. The various adjusting controls are indicated in the diagram.

The unit shown in Figs. 1232, 1233 and 1236 is a commercial model, but the amateur may build one following the same general lines.

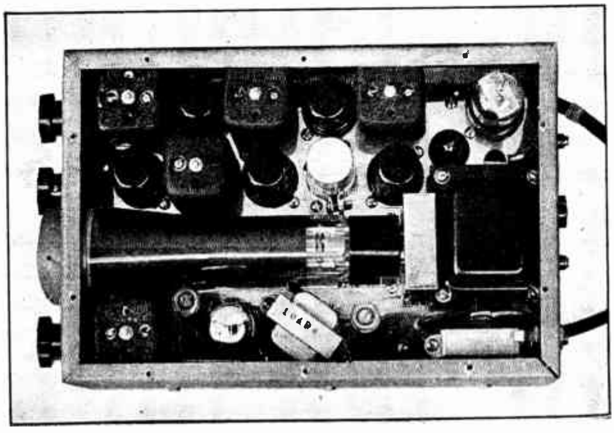
Testing and Alignment — At the positive terminal of C_{22} the voltage to ground should measure 300 volts, approximately, and the same voltage should appear between the negative terminal of C_{25} and chassis. The screen voltage on the two 6SJ7s should be approximately 100 (at full gain).

The cathode-ray tube makes a convenient indicating device in alignment of the r.f. and i.f. stages. The sweep generator should give no difficulty, although it will be helpful to check the shape of the saw-tooth. A 'scope having the regular complement of amplifiers and a linear sweep is necessary for this. Connect the grounded side of the vertical input of the 'scope to chassis and the high side through a condenser (0.1 μfd. or so) to the ungrounded side of C_{10}, when a saw-tooth should appear on the oscilloscope screen, if the oscilloscope sweep frequency is of the order of 30 cycles. With the vertical amplifier connected to the grid of the saw-tooth oscillator a sharp pulse should appear on the screen. Synchronization can be checked by connecting the 'scope across R_{34} and adjusting the oscilloscope sweep to include three or more cycles of the 60-cycle voltage which appears across R_{34}. At each oscillator grid pulse a small transient will appear in the pattern (it may be only a small gap in the 60-cycle trace) and when R_{33} is adjusted so that one of these appears at the same point on every other cycle the saw-tooth oscillator is synchronized at 30 cycles.

The saw-tooth should be reasonably straight (make allowance for possible poor linearity of the sweep in the oscilloscope at this low frequency) and the fly-back time, or horizontal duration of the vertical part of the saw-tooth, should be very short. Should the oscillator not operate at all, reverse the leads of the plate winding of T_6.

With the saw-tooth oscillator in operation, apply voltages to the 902. The saw-tooth applied to the horizontal deflection plates should give a horizontal line on the screen, focusing and intensity being adjustable by means of R_{22} and R_{47}, respectively. The width of the line can be adjusted by the horizontal size control,

Fig. 1233 — Top view of the panoramic adapter. The socket of the 902 is mounted on a metal plate provided with slots so that the tube may be rotated to place the screen in proper position. The power transformer is immediately behind this plate. Along the top edge, from left to right, are the filter choke, horizontal size control (R_{31}), sweep-oscillator transformer (T_6), the 7F7, sweep-frequency control (R_{33}), and the input i.f. transformer (T_1). In the line of components below the 902, from left to right, are the neon voltage-regulator tube, the 6AC7, a triple condenser unit (C_{21}, C_{22}, C_{23}), the 6SQ7, the mixer-oscillator transformer (T_5) and the first 6SJ7 i.f. tube. In the bottom row, from left to right, are the power rectifier tube, the i.f. output transformer (T_4), the second 6SJ7 i.f. tube, the third i.f. transformer (T_3), the 6SA7 and the second i.f. transformer (T_2).



R_{37}, and its position on the screen by R_{19}, the horizontal positioning control, and R_{17}, the vertical positioning control.

A test oscillator is practically a necessity for the preliminary alignment of the r.f. and i.f. amplifiers, if only to get them on the right frequency. The i.f. should be aligned first, tuning the trimmers in T_3 and T_4 for maximum response throughout. As a trimmer is tuned through resonance the line on the cathode-ray tube screen will move upward, the extent of the movement indicating the amplitude of the output voltage from the 6SQ7.

The r.f. circuits (T_1 and T_2) can be aligned with the help of a test oscillator tuned to the intermediate frequency in the receiver. Connect the oscillator output between the plate of the 6SJ7 amplifier and ground, using a blocking condenser in the hot lead to isolate the plate voltage. Then adjust C_{29} in the oscillator transformer, T_5, to give a beat of 100 kc., which will be amplified and give maximum deflection on the cathode-ray tube screen. The sweep control, R_{35}, should be set at zero so that the oscillator will not be frequency modulated. Adjust the secondary trimmer in T_2 for maximum response. Then move the test oscillator output to the grid of the 6SJ7 and adjust the primary trimmer of T_2 to resonance. Align T_1 similarly with the test oscillator connected between ground and the clip which goes to the receiver mixer plate prong.

The next step is to adjust the oscillator sweep, and for the sake of illustration we will assume that the receiver i.f. is 456 kc. With the test oscillator at 456 kc., and with the sweep padder, R_{36}, at about half scale, increase R_{35} slowly from zero. As the amplitude

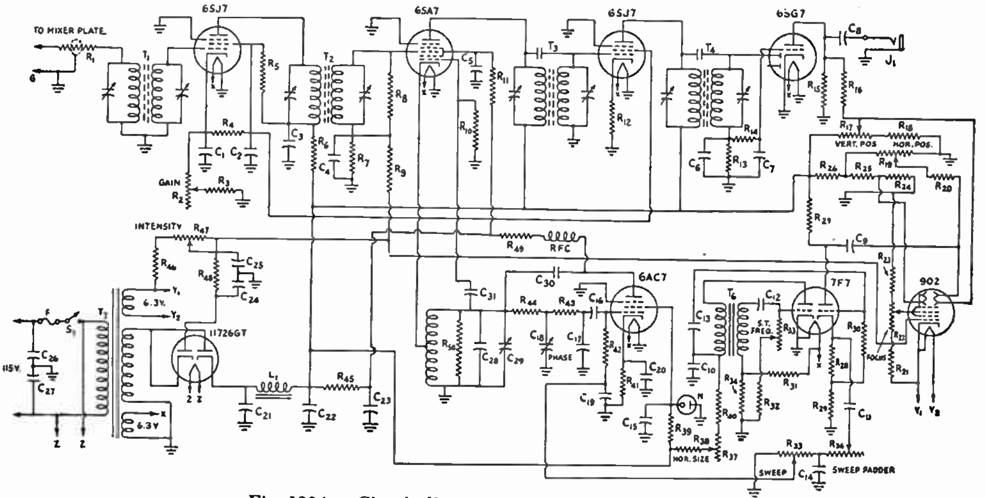


Fig. 1234 — Circuit diagram of the panoramic adapter.

C_1, C_2, C_3, C_4, C_5, C_8, C_{15}, C_{20}, C_{26}, C_{27} — 0.01-μfd. paper, 600 volts.
C_6, C_7, C_{14} — 500-μμfd. mica.
C_9, C_{13} — 0.05-μfd. 400-volt paper.
C_{10} — 0.1-μfd. 400-volt paper.
C_{11} — 0.25-μfd. 400-volt paper.
C_{12} — 0.01-μfd. mica.
C_{16} — 100-μμfd. mica.
C_{17} — 30-μμfd. mica.
C_{18} — 1-10-μμfd. mica padder.
C_{19} — 250-μμfd. mica.
C_{21}, C_{22}, C_{23} — 10-μfd. 450-volt electrolytic.
C_{24}, C_{25} — 4-μfd. 450-volt electrolytic.
C_{28}, C_{31} — 100-μμfd. mica (in oscillator unit, T_5).
C_{29} — 30-240-μμfd. mica padder (in oscillator unit, T_5).
C_{30} — 500-μμfd. mica (in oscillator unit, T_5).
R_1, R_{16}, R_{27} — 0.25 megohm, ½ watt.
R_2 — 10,000-ohm potentiometer.
R_3, R_{12}, R_{34} — 200 ohms, ½ watt.
R_4, R_{43}, R_{44} — 50,000 ohms, ½ watt.
R_5, R_{29} — 25,000 ohms, ½ watt.
R_6, R_7, R_{28}, R_{45} — 5000 ohms, ½ watt.
R_8, R_{18}, R_{21}, R_{23} — 0.1 megohm, ½ watt.
R_9, R_{13}, R_{14}, R_{38}, R_{40} — 1 megohm, ½ watt.
R_{10} — 0.11 megohm, ½ watt.
R_{11} — 45,000 ohms, ½ watt.
R_{15}, R_{32} — 0.5 megohm, ½ watt.
R_{17}, R_{35}, R_{47} — 0.1-megohm potentiometer.
R_{19}, R_{22} — 0.25-megohm potentiometer.
R_{20}, R_{30} — 2 megohms, ½ watt.
R_{24} — 25,000 ohms, 1 watt.
R_{25} — 33,000 ohms, ½ watt.
R_{26} — See note.
R_{31} — 500 ohms, ½ watt.
R_{33}, R_{36}, R_{37} — 1-megohm potentiometer.
R_{39} — 75,000 ohms, ½ watt.
R_{41} — 1000 ohms, ½ watt.
R_{42} — 0.2 megohm, ½ watt.
R_{46}, R_{48} — 10,000 ohms, ½ watt.
R_{49} — 3000 ohms, ½ watt (in oscillator unit, T_5).
R_{50} — 25,000 ohms, ½ watt (in oscillator unit, T_5).
T_1 — R.f. input transformer, 456 kc.
T_2 — R.f. interstage transformer, 456 kc.
T_3 — I.f. input transformer, 100 kc.
T_4 — I.f. output transformer, 100 kc.
T_5 — Oscillator transformer, 356 kc.
T_6 — Saw-tooth oscillator transformer (2:1 or 3:1 midget audio).
T_7 — Power transformer; two 6.3-v. windings, h.v. winding, 300-v. a.c., 40 ma.
L_1 — Filter choke, 40 ma., 350 ohms (app. 5-10 henrys).
F — 2-amp. fuse.
S_1 — Toggle switch (on R_{47}).
J_1 — Open circuit jack.
N — ½-watt neon bulb without base resistor.
RFC — 30-mh. r.f. choke (in oscillator unit, T_5).

Note: R_{26} needed only in case horizontal positioning control (R_{19}) is critical in adjustment or total plate voltage exceeds 300, approximately. It may be omitted in this circuit, the junction of R_{25} and R_{19} being connected directly to + B.

Transformers T_1-T_4, inclusive, are available from Panoramic Radio Corp., New York City.

of the sweep voltage applied to the grid of the 6AC7 reactance modulator increases, the pattern on the cathode-ray tube screen should change, showing the signal as a hump on the horizontal base line, which should move downward to the position it had originally when no signal was applied to the vertical plates. A suitable height for the signal trace can be obtained by adjustment of the gain control, R_2, or the output of the test oscillator.

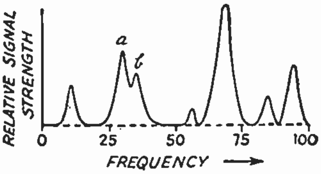


Fig. 1235 — Representation of panoramic reception over a 100-kc. band. The cathode-ray tube beam traces the pattern which would result from plotting instantaneous response of the receiver to signals of differing amplitude and frequency, assuming that such a plot could be made instantaneously. The signals represented by peaks *a* and *b* are so close that the i.f. selectivity is not sufficient to make them appear as isolated peaks.

Should the signal trace not be in the center of the screen, or should it move horizontally as the sweep amplitude is increased (either or both probably will be the case at first trial), adjust C_{29} while varying R_{35} until the signal remains fixed in position on the horizontal base line, regardless of the setting of R_{35}. When the proper adjustment is found the signal will not necessarily appear in the center of the screen, but it can be brought to center by readjusting the horizontal positioning control, R_{19}. The phasing control (C_{18}) adjustment is not critical, and this control may be set simply near but not quite at maximum capacity.

With a 456-kc. signal centered on the screen, tune the test oscillator slowly toward 506 kc., watching the signal trace move horizontally on the screen as the oscillator frequency is changed. R_{35} should be set at maximum. With the oscillator frequency at 506 kc. the signal trace should be just at the edge of the screen; if it is not, it can be brought there by adjustment of the sweep padder, R_{36}. Tuning in the opposite direction to 406 kc. then should move the trace to the opposite end of the screen. When this adjustment is made the maximum sweep will be 100 kc. It may be set at any desired figure between 100 and zero kc. by adjustment of R_{35}.

The next and final step in adjustment is to align T_1 and T_2 to compensate for the r.f. selectivity of the receiver. Set the receiver at about 3 Mc., set the test oscillator to the same frequency and tune the signal to the center of the screen, using the regular receiver tuning control. Then move the test oscillator frequency 50 kc. higher or lower, putting the signal at one edge of the cathode-ray tube screen. Note the amplitude as compared to the amplitude at the center, and adjust the i.f. transformer trimmers to make the amplitude approximately equal to that at the center. Then move the test oscillator 50 kc. on the other side of the center frequency and readjust the trimmers to make the amplitude equal to that at the center. This will upset the first adjustment, so it will be necessary to go back and forth, making compromise adjustments which finally result in making the gain as uniform as possible over the whole 100-kc. band. The desirable condition, of course, is one in which the height of the test signal does not change as the frequency is varied over the 100-kc. range. Probably it will not be possible to get perfect compensation, but there should be no difficulty in coming reasonably close to it. At frequencies higher than 3 Mc. it is to be expected that the signal amplitude will increase toward the edges of the pattern, and that it will decrease at frequencies lower than 3 Mc.

The frequency-modulated oscillator in the unit provides an excellent means for final alignment of the 100-kc. amplifier. Tune in a test signal to the center of the screen and adjust the trimmers in T_3 and T_4 to give the sharpest and most symmetrical pattern. The signal on the screen is actually a trace of the selectivity curve of the 100-kc. amplifier, and corresponds exactly to the similar type of trace obtained when aligning an ordinary superhet with the aid of a frequency-modulated test oscillator and oscilloscope.

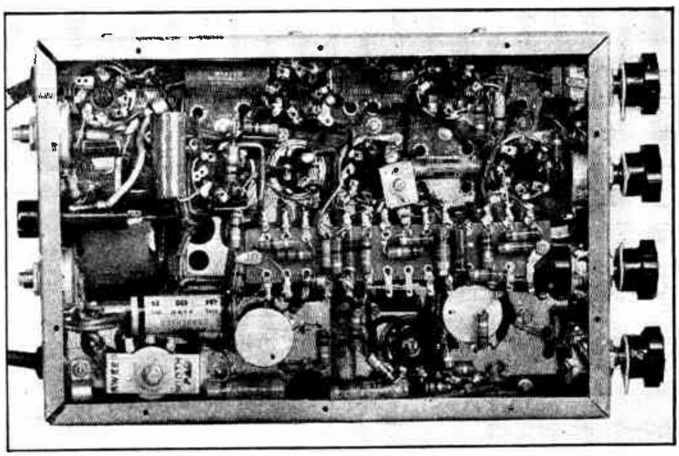


Fig. 1236 — Bottom view of the panoramic adapter. Condensers and resistors in the r.f. circuits are placed close to associated circuits following usual practice. C_{24} and C_{25} are in the upper left-hand corner. The sweep padder (R_{36}) is mounted on a bracket in the lower left-hand corner. R_{33} and R_{37} are the two volume controls near the bottom edge. Near the center, over the 6AC7 socket, is the phasing trimmer condenser, C_{18}. R_{17} and R_{22} may be seen mounted at the left-hand edge with their shafts protruding to the rear of the cabinet.

Chapter Thirteen

Transmitter Construction

In the descriptions of apparatus to follow, not only the electrical specifications but also the manufacturer's name and type number have been given for most components. This is for the convenience of the builder who may wish to make an exact copy of some piece of equipment. However, it should be understood that a component of different manufacture, provided it is of equivalent quality and has the same electrical specifications, may be substituted in most cases.

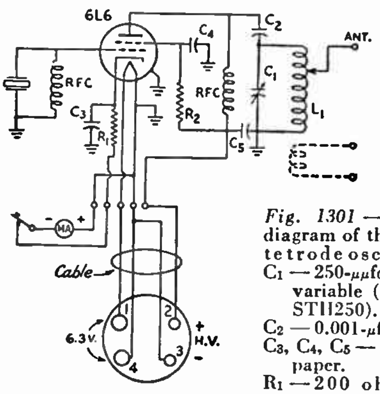


Fig. 1301 — Circuit diagram of the simple tetrode oscillator.
C_1 — 250-μμfd. midget variable (National STH250).
C_2 — 0.001-μfd. mica.
C_3, C_4, C_5 — 0.01-μfd. paper.
R_1 — 200 ohms, 2-watt.
R_2 — 15,000 ohms, 2-watt. RFC — 2.5-mh. r.f. choke.
L_1 — 1.75 Mc. — 42 turns No. 22 e., 2 inches long.
3.5 Mc. — 21 turns No. 18 e., 2 inches long.
7 Mc. — 15 turns No. 18 e., 2 inches long.
All coils wound on 4-prong, 1½-inch diameter forms.

Any unusual characteristics in tuning or operation are explained in the text material in these pages describing the construction of each unit. For information concerning straightforward transmitter adjustments, such as the tuning and neutralizing of standard circuits, the reader should consult Chapter Four. Chapter Ten contains information on the adjustment of antenna tuners for the various types of antennas. Keying systems are treated in Chapter Six. The construction of meter shunts is covered in Chapter Twenty, while operating data on transmitting tubes not specifically included in this chapter will be found in the vacuum-tube tables in the Appendix.

To reduce repetition and make possible a treatment of wider scope, liberal reference will be made to material appearing in other chapters in this *Handbook*.

A Simple Tetrode Oscillator

The unit shown in Figs. 1301–1302 represents one of the simplest types of amateur transmitters. The various parts are assembled on a breadboard purchased already finished at a "dime" store. Rubber feet at the corners elevate the base to clear mounting screws. A "ground" wire is run from one side of the crystal socket to one side of the coil socket, to which all ground connections shown in the diagram are made.

Since parallel plate feed is used, the only exposed high-voltage points are the plate-circuit r.f. choke and the high-voltage power terminal. Grid bias is obtained entirely from the cathode resistance. Either simple voltage feed to a half-wave antenna or an antenna a multiple of one-half wavelength long, or link coupling to an antenna tuner by adding a link winding at the bottom of the form as indicated in the schematic diagram (Fig. 1301), may be used.

Although a 6L6 tube is shown in the photograph, a 6V6 might be used at lower plate voltage without circuit alteration. Any available power supply delivering up to 450 volts or so may be used, the power output obtainable increasing with the voltage applied. The unit shown in Fig. 1304 is suitable. The two units are connected by a four-wire battery cable with

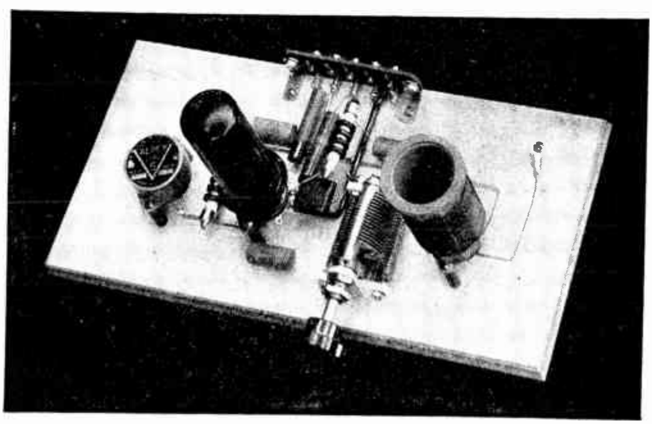


Fig. 1302 — A simple breadboard crystal-controlled transmitter. The grid r.f. choke is located between the crystal and 6L6 and the plate choke is to the right of the tube. The cathode and screen resistors are to the rear of the 6L6. The blocking condenser, C_2, is between the tube and tank condenser.

a four-prong plug at the power-supply end to fit the outlet in the power supply.

Since the circuit is not designed for frequency doubling, a separate crystal will be required for each frequency to be used.

Tuning — A milliammeter with a scale of 100 or 200 ma. should be connected in series with the key, as shown in Fig. 1301, as an aid in tuning. With a suitable coil and crystal in place and the high voltage turned on, a rise in plate current should occur when the key is closed. The plate tank condenser, C_1, should then be rotated until there is a pronounced dip in plate current at resonance. If the voltage-fed antenna is used, it may now be connected to the antenna terminal and a temporary wire run from the antenna terminal to reach the coil, L_1. Starting at a point one-third or half-way up from the bottom of the coil, scrape the wire at a spot, being careful not to short-circuit turns, and let the antenna wire rest against the bare spot. Tuning the transmitter as before, the plate-current dip should again be found, although less pronounced this time. The tap should be moved gradually toward the top of the coil until only a slight dip in plate current is observed as the plate tank circuit is tuned through resonance. At each adjustment of the antenna tap, the transmitter should be tested to make sure that the circuit keys well. Should a point be reached where it is difficult to get the crystal to start, the tap should be backed off somewhat. It will be found possible to load up the circuit more with certain crystals than others, while still maintaining good starting and keying characteristics. When a satisfactory point has been found, the tap may be soldered in place permanently and a connection made through one of the unused pins on the coil form.

With a 6L6 tube and a plate supply delivering 400 volts, the screen voltage will be about 250 volts. The tube will draw about 75 ma. non-oscillating, dipping to about 50 ma. at resonance with the antenna disconnected. It should be possible to load up the circuit until the tube draws about 80 ma. at resonance. Under these conditions, the power output on each band should be 15 to 20 watts.

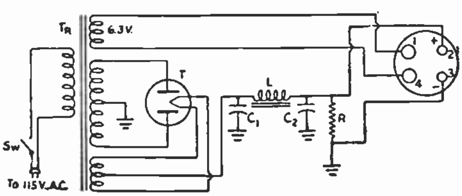


Fig. 1303 — Circuit of the 450-volt power supply.
C_1 — 4-μfd. 600-volt electrolytic (Mallory HS691).
C_2 — 8-μfd. 600-volt electrolytic (Mallory HS693).
L — Filter choke, 10 henrys, 175 ma., 100 ohms (Utah 4667).
R — 15,000 ohms, 25-watt.
T — Type 80 rectifier tube.
T_R — Power transformer, 400 volts each side of center-tap; rectifier filament winding, 5 volts, 3 amperes; r.f. filament winding, 6.3 volts, 6 amperes (Utah Y616).

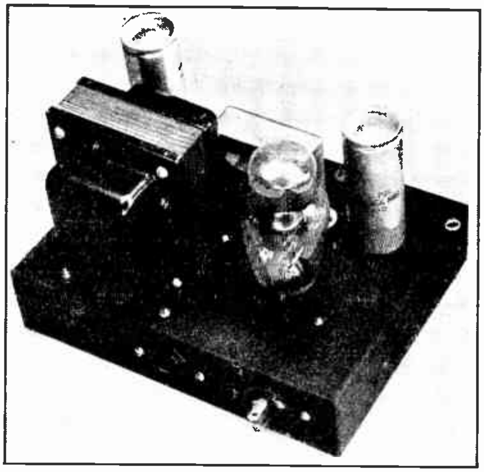


Fig. 1304 — This power supply delivers 450 volts at a full-load current of 130 ma., with 0.3 per cent ripple and measured regulation of 17 per cent. If converted to a choke-input filter by inserting a similar choke between the rectifier and present filter, the output voltage is reduced to about 300 volts. The chassis measures 7 × 9 × 2 inches. Filament and plate voltages are brought out to a four-prong socket. The circuit is given in Fig. 1303.

A Low-Power Antenna Tuner

If an antenna with tuned feeders is used, the antenna tuner shown in Figs. 1305–1306 may be used to couple the 6L6 oscillator-transmitter to the feeders. The link winding of the transmitter and that of the antenna tuner should be connected with a pair of closely spaced wires.

The circuit, shown in Fig. 1306, is arranged so that different tuning combinations may be obtained by shifting the clips F, G and H. When F is connected to A, H is connected to D, and B and C are connected together, the two sections of C_1 in series are connected across L_1, forming a low-capacity parallel-tuned circuit. When H is connected to E and G to D, the other connections remaining the same, a high-capacity parallel circuit is obtained. For series tuning, H is connected to E, F to B and G to C. A low-capacity series-tuned circuit is provided by connecting F to B and H to C.

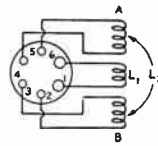
Dimensions are given for antenna coils of four different sizes, which are approximately correct for the band indicated when parallel tuning is required. For series tuning, the coil for the next-higher frequency band usually will be satisfactory. In some cases, where the feeders are not close to exact multiples of one-quarter wavelength for the frequency in use, slight alterations in coil dimensions may be required to permit tuning the system to resonance. The high-capacity circuits usually will be required for the lower frequencies, while the low-capacity connections will serve for the higher frequencies. Coupling may be adjusted by altering the number of turns in the windings at each end of the link line.

Construction — The two uprights and the strip supporting the indicating lamps are pieces of "1 by 2" stock. The uprights are each 13 inches long and the cross-strip 12 inches long; these dimensions may be changed to suit the constructor. The shelf for the condenser and coil is made of a piece of crate wood 4½ inches wide. The panel is made of plywood and is 7 inches high.

The dial lamps are soldered to a pair of parallel wires supported at each end on small stand-off insulators. The bottom of the neon bulb is soldered to a short piece of wire between a third pair of stand-offs. The piece of grounded metal next to the neon bulb is about 1¾ inches square. This provides a capacity to ground which enables the neon bulb to operate without touching the hand to it.

The socket for the plug-in coil is mounted on the shelf with spacers and wood screws. The shield between the two sections of the variable condenser is removed to permit mounting it by a screw through the hole in the ceramic to the shelf. The shaft of the condenser is cut off and an insulating coupling inserted between the shaft and the control knob. The contacts for shifting connections for the different tuning combinations consist of machine screws set in a small strip of bakelite.

The neon bulb and the dial lamps can be used to indicate resonance in the antenna circuit and relative (not actual) power output. The lamps will be useful whenever the length from the far end of the antenna to the feeder terminals is near an odd multiple of one-quarter wavelength for the frequency of operation, while the neon bulb will be useful where the length is near an even multiple. In tuning with the lamps, all sockets should be filled at the start. If, as an indication of resonance is obtained by an increase in plate current, the lamps show no indication, they should be removed, one at a time, until the remaining lamps start to glow. Sufficient lamps should be kept in the circuit to prevent danger of burn-out. After the antenna has been tuned for maximum power output, the lamps should be short-circuited with the switch or clip, *S*.



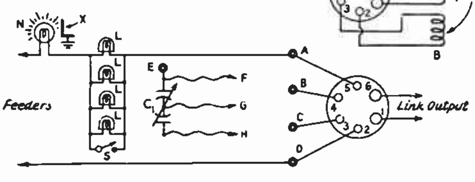


Fig. 1306 — Wiring diagram of the low-power antenna tuner.

At the top are the connections to the coil socket. C_1 has a capacity of 140 μμfd. per section (Hammarlund MCD-140). *L* is a 250-ma. dial light, No. 46. *N* is a ¼-watt neon bulb. *X* is a grounded piece of metal which provides capacitive coupling for igniting the neon bulb. *S* is used for short-circuiting the lamps after tuning.

Below are the connections to the 6-prong coil form. L_1, whose approximate dimensions are given below, is wound in two sections, with the link winding, L_2, in between them.

L_1 — 1.75 Mc. — 20 turns No. 22 e., ¾-inch long each section, ½-inch between sections, 40 turns total.
3.5 Mc. — 11 turns No. 20 e., ¾-inch long each section, ½-inch between sections, 22 turns total.
7 Mc. — 6 turns No. 20 e., ¾-inch long each section, ½-inch between sections, 12 turns total.
14 Mc. — 3 turns No. 20 e., ½-inch long each section, ¼-inch between sections, 6 turns total.

The number of turns for the link winding, L_2, will vary from 2 to 8 turns, depending upon coupling required for proper loading. All coils are wound on Hammarlund 6-prong 1½-inch diameter forms.

When using the neon bulb, the grounded metal plate should be bent near it until the bulb lights (assuming the transmitter has been tuned to approximate resonance by the plate-current meter). The grounded plate should be placed no closer to the bulb than is necessary to obtain a satisfactory indication.

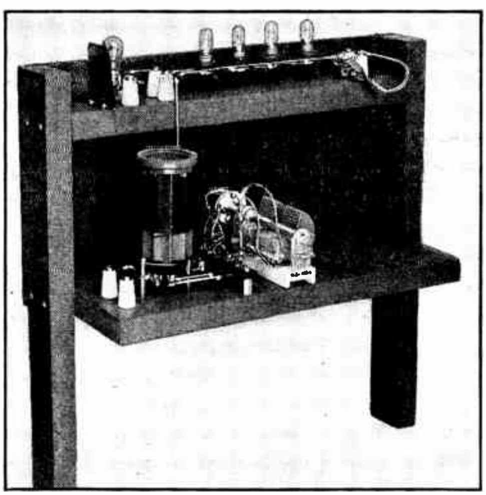


Fig. 1305 — Rear view of an antenna tuner for low-power transmitters. Dial lamps and a neon bulb are used as r.f. indicators. The unit is made to fit over the transmitter shown in Fig. 1302. Circuit is given in Fig. 1306.

Complete 15- to 25-Watt Oscillator Transmitter

The three units of Figs. 1302, 1304 and 1305 may be combined to form a simple, inexpensive low-power transmitter, complete from power supply to antenna tuner.

For convenience and economy of space, the units may be assembled on a vertical relay rack. The plate milliammeter may be mounted in the antenna-tuner unit, if desired.

A Two-Tube Plug-In Coil Exciter

In the two-tube exciter or low-power transmitter pictured in Figs. 1307, 1309 and 1310, a 6L6 oscillator is used to drive an 807 as an amplifier-doubler. As shown in Fig. 1308, a Tri-tet circuit, used to obtain harmonic output, is reduced to a simple tetrode circuit for oscillator output at the crystal fundamental by short-circuiting the cathode tank circuit. Sufficient oscillator output at the fourth harmonic

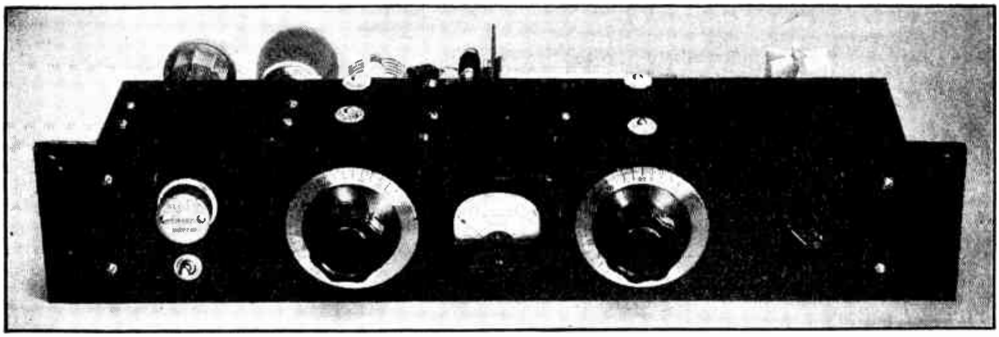


Fig. 1307 — The two-tube plug-in coil exciter is built to conserve space in the relay rack. The panel is 3½ × 19 inches. A clearance hole is cut in the left end of the panel for the crystal socket, which is mounted in the chassis directly above the cathode-circuit switch. The left-hand dial controls the tuning of the oscillator plate-tank circuit; the dial to the right tunes the output tank circuit. The switch at the right-hand end is for the 200-ma. milliammeter.

of the crystal frequency is obtainable to drive the 807, which may be operated as either a straight amplifier or frequency doubler, providing output of 25 to 50 watts or more in four bands from a single crystal.

The entire unit is designed to operate from a single 250-ma. power supply delivering up to 750 volts (see Fig. 1318), the maximum rating for the 807. Fixed bias of 45 volts, which may be obtained from a dry battery, is required for the 807. In the system shown, both oscillator and amplifier are keyed simultaneously in the common cathode lead. A single 200-ma. milliammeter may be switched to read the plate current of either stage.

Tuning — Because it is possible to double or quadruple frequency in the plate circuit of the oscillator and to double in the plate circuit of the 807, as well, there are several possible combinations of coils and crystals which will produce the same output frequency. Since much better efficiencies are obtainable, it is advisable to operate the 807 as a straight amplifier rather than as a doubler. This is possible in all cases except where it is necessary to obtain output at the eighth harmonic of the crystal frequency — 14-Mc. output from a 1.75-Mc. crystal, or 28-Mc. output from a 3.5-Mc. crystal. The accompanying chart shows the combination required for the desired output from any given crystal. This chart also indicates the position for SW_1. Be sure that the

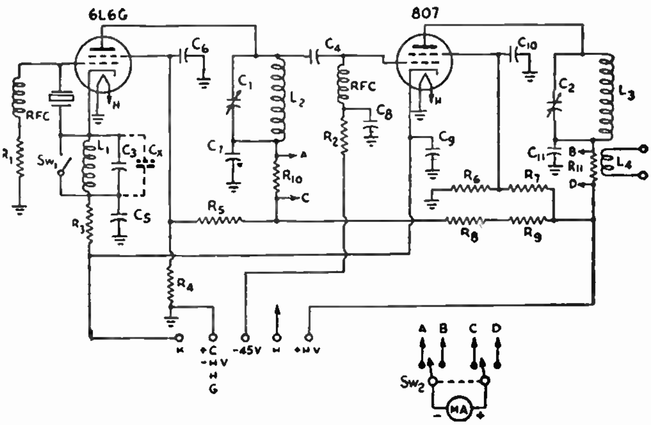


Fig. 1308 — Circuit diagram of the two-tube plug-in coil exciter unit.

C_1 — 140-μμfd. variable (Hammarlund MC-14M).
C_2 — 150-μμfd. variable (Cardwell MR150BS).
C_3 — 100-μμfd. mica.
C_4 — 20-μμfd. mica.
C_5, C_6, C_7, C_8, C_9, C_{10} — 0.01-μfd., 600-volt paper.
C_{11} — 0.01-μfd. 1000-volt paper.
C_x — 100-μμfd. mica (used only on 3.5 Mc.).
MA — Milliammeter, 0–200-ma.
R_1 — 20,000 ohms, 1-watt.
R_2 — 25,000 ohms, 2-watt.
R_3 — 200 ohms, 2-watt.
R_4 — 10,000 ohms, 25-watt.
R_5 — 3500 ohms, 25-watt.
R_6, R_7 — 15,000 ohms, 25-watt.
R_8, R_9 — 1250 ohms, 50-watt.
R_{10}, R_{11} — 10 ohms, 1-watt.
RFC — 2.5-mh. r.f. choke.
SW_1 — S.p.s.t. toggle switch.
SW_2 — D.p.d.t. rotary switch (Mallory 3222J).
L_1 — 1.75-Mc. crystals — 32 turns No. 22 d.s.c., close-wound.
3.5-Mc. crystals — 10 turns No. 22 d.s.c., 1 inch long. Note: *Cx* mounted in form.
7-Mc. crystals — 6½ turns No. 22 d.s.c., ¾-inch long.
All wound on Hammarlund 1½-inch diam. 4-pin forms.
L_2 — 1.75 Mc. — 56 turns, 1¼-inch diameter, 1¾ inches long, 54 μhys. (National AR80 — no link).
3.5 Mc. — 28 turns, 1¼-inch diameter, 1½ inches long, 15 μhys. (National AR40 — no link).
7-Mc. — 14 turns, 1¼-inch diameter, 1¼ inches long, 4.2 μhys. (National AR20 — no link).
14 Mc. — 8 turns, 1¼-inch diameter, 1½ inches long, 1.25 μhys. (National AR10 — no link).
28 Mc. — 4 turns, 1¼-inch diameter, ¾-inch long, 0.5 μhy. (National AR10, 4 turns removed — no link).
L_3 — 1.75 Mc. — 50 turns, 1½-inch diameter, 2⅛ inches long. 52 μhys. (Coto CS6160E).
3.5 Mc. — 25 turns, 1½-inch diameter, 1⅝ inches long, 16 μhys. (Coto CS680E).
7 Mc. — 16 turns, 1½-inch diameter, 1⅞ inches long, 5.7 μhys. (Coto CS640E).
14 Mc. — 8 turns, 1½-inch diameter, 1⅝ inches long, 1.5 μhys. (Coto CS620E).
28 Mc. — 4 turns, 1½-inch diameter, 1½ inches long, 0.7 μhy. (Coto CS610E).

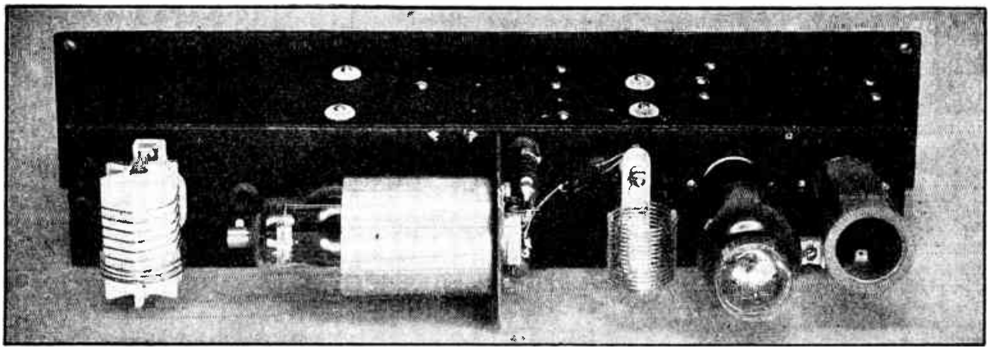


Fig. 1309 — The four-prong socket for the cathode coil, the octal socket for the 6L6 oscillator and the five-prong socket for the Coto coils used in the output tank circuit are sub-mounted along the rear of the chassis. The mounting for the National AR coils used in the oscillator plate circuit is fastened on short cone insulators, while the socket for the 807 is sub-mounted in the small steel partition. The grid r.f. choke and screen and cathode by-pass condensers are fastened directly to the socket. Large clearance holes lined with grommets are provided for passing the connections through the chassis from the oscillator plate coil to the tank condenser and for the 807 plate lead. A pair of pin jacks serves as the link output terminals. Power-supply connections are made to a terminal strip at the right.

harmonics of the crystal frequency fall in the band in which operation is to occur.

With the proper coils and crystal in place, SW_1 in the correct position and both condensers set at minimum capacity (100 on the dial), the plate voltage should be applied with the meter reading plate current to the 807. If all resistances are correct and the plate voltage is 750, the plate current should run approximately 25 ma. With the key closed, tune the oscillator tank condenser for maximum amplifier plate current. (Do not hold the key closed for long periods under this high-current condition.) As soon as the peak has been obtained, tune the amplifier plate tank condenser for resonance as indicated by a pronounced dip in plate current. Should the points of response on either condenser be found at points on the scale differing appreciably from those given in the accompanying table, each circuit should be checked with an absorption frequency meter to make sure that it is tuned to the correct frequency, since the ranges covered by some of the coils include odd harmonics falling outside the amateur bands. Once checked, the dial settings can be logged for quick resetting.

When the amplifier has been tuned, the meter switch may be set to read oscillator plate current and the oscillator tank circuit tuned for minimum plate current consistent with satisfactory keying. Active crystals usually will oscillate continuously in the Tri-tet circuit, regardless of the setting of the tank condenser. With the tetrode circuit, however, the circuit will oscillate only within relatively narrow limits. SW_1 must be closed when the oscillator plate circuit is tuned to the crystal frequency. The oscillator plate current will vary widely, depending upon whether output is taken at the fundamental, second harmonic or fourth harmonic. At the specified plate voltage, it should run between 40 and 50 ma. with the plate circuit tuned to the crystal fundamental or second harmonic. When tuned to the fourth harmonic, the plate current will normally run between 85 and 95 ma.

Because the plate and screen of the 6L6 are operated from a voltage divider, their voltages vary with tuning. Plate voltage varies between 400 and 450, except at the fourth harmonic when it falls to 340 volts or so. The screen voltage varies from 280 to 210 volts.

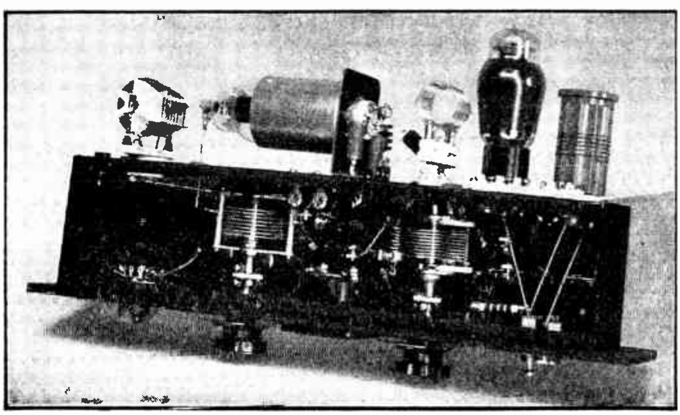


Fig. 1310 — Bottom view of the plug-in exciter. Space inside the 4 × 17 × 3-inch chassis has been utilized to the greatest extent possible while preserving accessibility. Voltage divider resistors R_8 and R_9 are to the right of the oscillator tank condenser, while R_4, R_5, R_6 and R_7 are mounted to the rear of the meter. The oscillator r.f. choke and grid leak are fastened to the crystal socket. Connections between the crystal socket and cathode switch are made directly and kept well spaced. The oscillator circuit may be arranged for v.f.o. input as shown in Fig. 1372. Meter-shunting resistances are fastened to the meter switch. Both tank-condenser shafts must be fitted with insulated couplings and panel bearings.

COIL AND TUNING TABLE FOR TWO-TUBE PLUG-IN COIL EXCITER

Xtal Band Mc.	*Output Band Mc.*	SW_1	L_1 *Band Mc.*	C_1L_2 *Band Mc.*	C_2L_3 *Band Mc.*	C_1*	C_1*
1.75	1.75	Closed	1.75	1.75	1.75	10	10
1.75	3.5	Open	1.75	3.5	3.5	10	30
3.5	3.5	Closed	3.5	3.5	3.5	10	30
1.75	7	Open	1.75	7	7	20	50
3.5	7	Open	3.5	7	7	20	50
7	7	Closed	7	7	7	20	50
1.75	14	Open	1.75	7	14	20	70
3.5	14	Open	3.5	14	14	35	70
7	14	Open	7	14	14	35	70
3.5	28	Open	3.5	14	28	35	80
7	28	Open	7	28	28	75	80

* Approximate settings for low-frequency ends of bands with dial reading zero at full capacity of condenser.

The plate current should be limited to 70 ma. at 28 Mc. and 80 ma. at 14 Mc. when doubling frequency in the output stage, and to 90 ma. when operating the 807 as a straight amplifier at 28 Mc. Power output under these conditions should average 40 to 55 watts on all bands. When doubling frequency in the output circuit to 14 and 28 Mc., the output will be reduced to about 27 and 18 watts respectively.

If the exciter is operated from a power supply of lower voltage, the values of resistance specified for the voltage dividers may be altered to increase the voltages on the oscillator plate and screen and also the screen of the 807. With a 600-volt supply, R_8 and R_9 should be 1000 ohms each, R_4, 20,000 ohms, and R_5, 10,000 ohms. Power output will average 30 to 35 watts from the 807 as a straight amplifier.

ℂ Complete 75-Watt All-Band Transmitter with Plug-in Coils

If it is desired to feed the unit of Fig. 1307 into an antenna as a complete transmitter, it may be combined with the power-supply unit of Fig. 1318, which will furnish heater and plate voltages, and the antenna-tuning unit of Fig. 1320 using the large condensers. A 45-volt dry battery will be required for bias. The three units may be placed in a small table rack with a total height of only 17½ inches.

ℂ A Band-Switching Exciter with 807 Output Stage

The exciter or low-power transmitter pictured in Figs. 1311, 1312 and 1314 is designed for flexibility, being adaptable for use on all bands from 1.75 to 28 Mc., with crystals cut for different bands, and also for quick band changing over three bands. It consists of a 6C5 triode oscillator followed by two triode doubler stages in one tube, a 6N7; by means of a switch, S_2, the output of any of the three stages can be connected to the grid of the final tube, which is an 807 beam tetrode. The circuit diagram is given in Fig. 1313.

The oscillator coil and the first and second doubler plate coils, L_1, L_3 and L_2 respectively, need not be changed for crystals ground for a given band. The switching circuit is so arranged that the grids of unused stages are automatically disconnected from the preceding stage and grounded, so that excitation is not applied to the idle doubler tubes.

Capacity coupling between stages is used throughout. The plates of the first three stages are parallel-fed so that the plate tuning condensers can be mounted directly on the metal chassis. The 6C5, 6N7, and the 807 screen all operate from a 250-volt supply. Series feed is used in the 807 plate circuit, the tank con-

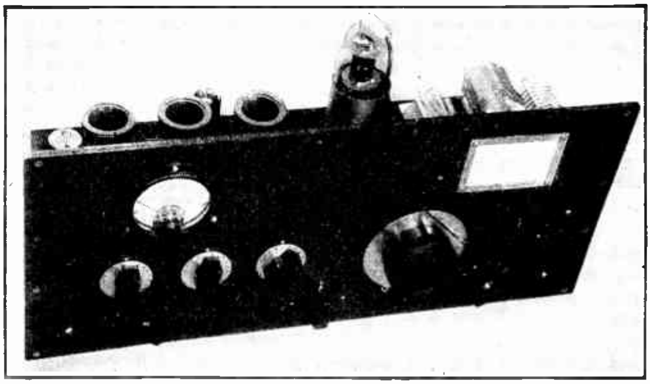


Fig. 1311 — An 807 exciter or low-power transmitter combining the flexibility of plug-in coils with the convenience of band-switching. A band-switching plug-in coil assembly changes tank coils in the 807 plate circuit. Crystal switching and meter switching also are provided. Plate currents for all tubes and screen current for the 807 are read on a 200-ma. meter which can be switched to any circuit. Keying is in the oscillator cathode circuit, for break-in operation. The panel is 8¾ inches high and of standard rack width. The chassis measures 8 x 17 x 2 inches. The unit requires two power supplies, one delivering 250 volts at approximately 75 ma. and the other 750 volts at 100 ma.

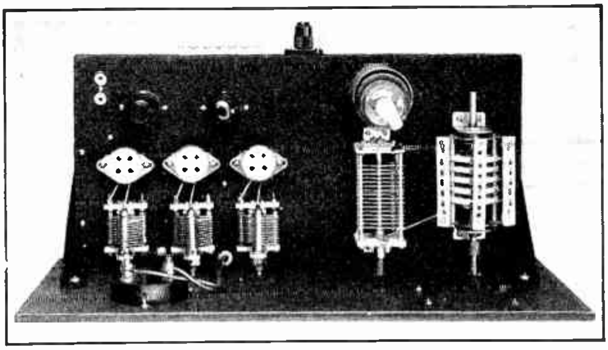


Fig. 1312 — Top view of the band-switching exciter with coils removed. At the left rear are the spare crystal socket, the 6C5 and the 6N7. Directly in front of these are the tuning condensers (mounted directly on the chassis) and the coil sockets (mounted on pillars) for the oscillator and doubler stages. Grouped to the right are the 807, the amplifier tank condenser (which must be insulated from the chassis) and the switch assembly. The "hot" leads from the coils are brought through grommeted holes in the chassis. The amplifier switch assembly should be mounted far enough back from the panel so that the coils will clear the side of the relay rack or cabinet. Leads between the switch and C_4 should be kept as short as possible.

denser being of the type which is insulated from the chassis. Fixed bias of about 75 volts is used on the 807 grid.

Plate currents for all tubes are read by a 200-ma. meter which can be switched to any circuit by means of S_4. Keying is done in the oscillator cathode circuit providing break-in operation.

Since in normal operation the crystal tank circuit, C_1L_1, is tuned well on the high-frequency side of resonance, there is a tendency for the first doubler section to break into a "tuned-grid tuned-plate" type of oscillation when the key is up; this is prevented by a small amount of inductive neutralization provided by the single-turn coils, L_5 and L_6, wound as closely as possible to the ground end of each tank coil. The 28-Mc. coil does not need such a neutralizing winding, since it is used only in the second doubler stage. L_5 and L_6 should be so connected as to prevent self-oscillation of the first 6N7 section when the key is open; the proper connections should be found by trial.

In the bottom view, Fig. 1314, the meter switch with its shunting resistors is at the left, with the 807 plate by-pass condenser, C_{11}, just above it. The stage switch, S_2, is in the center. R.f. leads to this switch should be kept separated as much as the layout will permit. R.f. junction points are insulated by small ceramic pillars. In this view, the right-hand section of the 6N7 is the first doubler. The rotor contact of the section of S_2 nearest the panel goes to the grid of the first doubler, the middle section to the second-doubler grid, and the third section to the 807 grid.

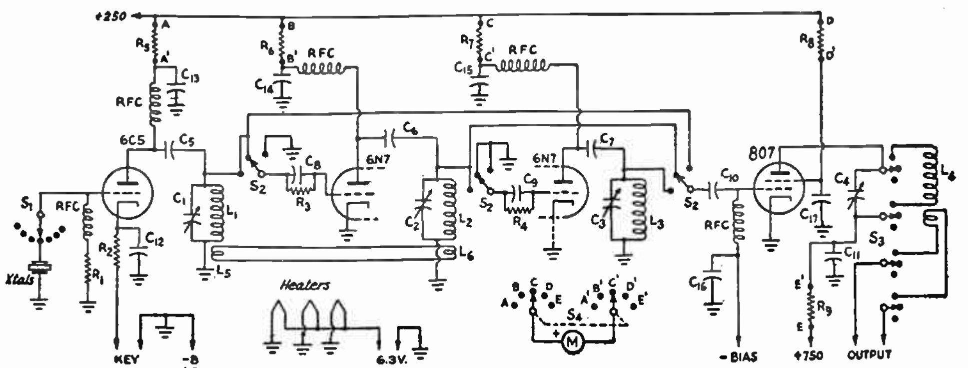
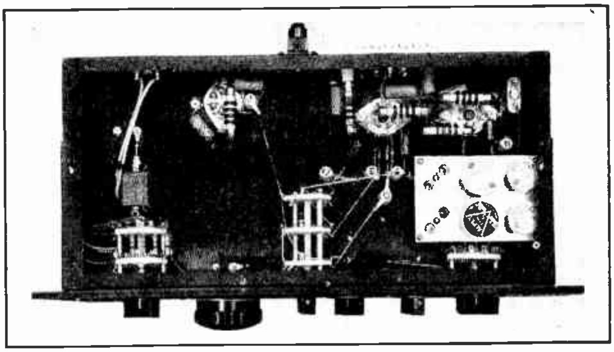


Fig. 1313 — Circuit diagram of the crystal-controlled 807 band-switching exciter or low-power transmitter.

C_1, C_2, C_3 — 100-μμfd. variable (National ST-100).
C_4 — 150-μμfd. variable, 0.05-inch plate spacing (Hammarlund HFB-150-C).
C_5, C_6, C_7 — 0.002-μfd. 500-volt mica.
C_8, C_9, C_{10} — 100-μμfd. 500-volt mica.
C_{11} — 0.002-μfd. 2500-volt mica.
C_{12}–C_{17}, inc. — 0.01-μfd. 600-volt paper.
R_1 — 10,000 ohms, ½-watt.
R_2 — 300 ohms, 1-watt.
R_3, R_4 — 25,000 ohms, ½-watt.
R_5–R_9, inc. — 25 ohms, ½-watt.
RFC — 2.5-mh. r.f. choke.
S_1 — Ceramic wafer switch, 6 or more positions.
S_2 — Three-gang three-position ceramic wafer switch (Yaxley 163C).
S_3 — Band-switch in coil assembly (Coto type 700).
S_4 — Two-gang 6-position (5 used) ceramic wafer switch.
M — 0-200 d.c. milliammeter, bakelite case.
L_1, L_2, L_3 — 1.75 Mc.: 50 turns No. 22 d.s.c. close-wound.
3.5 Mc.: 26 turns No. 18, length 1½ inches.
7 Mc.: 17 turns No. 18, length 1½ inches.
14 Mc.: 8 turns No. 18, length 1½ inches.
28 Mc.: 3 turns No. 18, length 1 inch.
All wound on 1½-inch diameter forms (Hammarlund SWF-4); turns spaced evenly to fill specified winding length.
L_4 — 1.75 Mc. — 50 turns, 1½-inch diameter, 2⅜ inches long, 52 μhys. (Coto CI6160E).
3.5 Mc. — 25 turns, 1½-inch diameter, 1⅝ inches long, 16 μhys. (Coto CI680E).
7 Mc. — 16 turns, 1½-inch diameter, 1⅛ inches long, 5.7 μhys. (Coto CI640E).
14 Mc. — 8 turns, 1½-inch diameter, 1⅝ inches long, 1.5 μhys. (Coto CI620E).
28 Mc. — 4 turns, 1½-inch diameter, 1½ inches long, 0.7 μhys. (Coto CI610E).
L_5, L_6 — One turn at bottom of L_1 and L_2. See text.

Fig. 1314 — Bottom view of the band-switching exciter, showing the meter switch at the left, the hand-switch in the center and the crystal switch at the right. The multiple crystal mounting, which holds six crystals, is made of a 3 × 4½-inch aluminum plate fitted with Amphenol crystal sockets, the assembly being elevated from the chassis by metal pillars. A seventh socket is provided on top of the chassis for a spare crystal or for e.c.o. input. The 750-volt lead is brought through a Millen safety terminal, and all other power connections come to a terminal strip at the rear which has barriers between the terminals to prevent accidental contact. All grounds are made directly to the 8 × 17 × 2-inch chassis.



Figs. 1315 and 1318 show suitable 250- and 750-volt power-supply units for this transmitter. Heater voltage and grid bias are obtained from the 250-volt supply. If desired, both these power units may be assembled on one large chassis.

Tuning — To operate the exciter, coils for consecutively higher-frequency bands are plugged in at L_1, L_2 and L_3; only five are necessary for operation with any crystal from 1.75 to 7 Mc. and for output from 1.75 to 28 Mc. For example, with 3.5-Mc. crystals, the 3.5-, 7- and 14-Mc. coils would be plugged in at L_1, L_2 and L_3 respectively. For 1.75-Mc. crystals, the 1.75-, 3.5- and 7-Mc. coils would be used, and so on. The plate coils for the 807 should cover the same bands as the low-level coils.

Preliminary tuning should be done with the plate voltage for the 807 disconnected. Set S_2 so that all tubes are in use. Switch the milliammeter to the oscillator circuit and close the key. Rotate C_1 for the dip in plate current which indicates oscillation. The non-oscillating plate current should be between 20 and 25 ma., dropping to 15 or 20 when oscillating. Switch the meter to the doubler plate and adjust C_2 to minimum plate current, or resonance. The off-resonance plate current should be about 30 ma. or more and the reading should be between 10 and 15 at resonance. Check the second-doubler plate current and tuning similarly; the off-resonance plate current should again be around 30 ma., dropping to 15 or 20 at resonance. At this point the 807 screen current should be measured; with too much excitation it will be considerably higher than the rated value (about 12 ma.) and the excitation should not be kept on for more than a second or two.

Next, the plate voltage may be applied to the 807. The amplifier should not be operated without load for more than a few moments at a time, because under these conditions the screen dissipation is excessive. Use a 70-ohm dummy antenna or a 60-watt lamp connected to the output link. The three bands may be checked in order by appropriate switching of S_2 and S_3. With the 807 fully loaded, check the screen current to make sure it does not exceed 10 or 12 ma. If it is too high, reduce the excitation by detuning the crystal oscillator until it reaches the proper value. The 807 grid current may be measured with a lower-range milliammeter connected in series with the bias source, if desired. Maximum output will be secured with a grid current of about 3 or 4 milliamperes, a value which also will give about rated screen current. The screen current is, in fact, a very good indicator of excitation. The 807 should show no tendency to oscillate by itself when the key is open.

The current to each section of the 6N7 should be 20 ma. with the key open (no excitation). If the two currents are not the same or show changes when C_2 and C_1 are tuned with key open, the first doubler may be acting as a t.p.t.g. oscillator, as previously mentioned, and the neutralizing circuit should be checked. Do not use more than 250 volts for the low-voltage supply, as higher values will cause excessive 807 screen dissipation. Care also should be taken to avoid excessive excitation. In normal operation, with C_1 detuned to reduce excitation to the proper value, the doubler plate currents will show little change between resonance and off-resonance tuning.

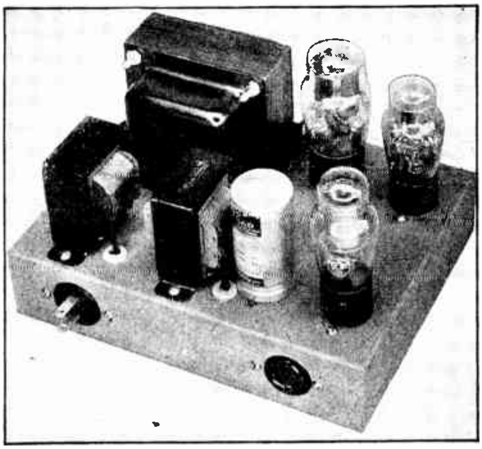


Fig. 1315 — A combination power-supply unit delivering 250 or 300 volts for exciter plate-supply and 75 volts of fixed bias. The unit is designed especially to work with the band-switching exciter of Fig. 1311 or the transmitter of Fig. 1330. If desired, the components may be combined with the components for a high-voltage plate-supply on a single chassis. The circuit diagram of the combination unit is shown in Fig. 1316.

With maximum input to the 807 plate (75 watts) the output is approximately 50 watts on all bands except 28 Mc., where greater circuit losses decrease it to about 40 watts. The excitation provided by the 6N7 doubler is more than ample on all bands.

The oscillator circuit may be arranged for v.f.o. input as in Fig. 1372.

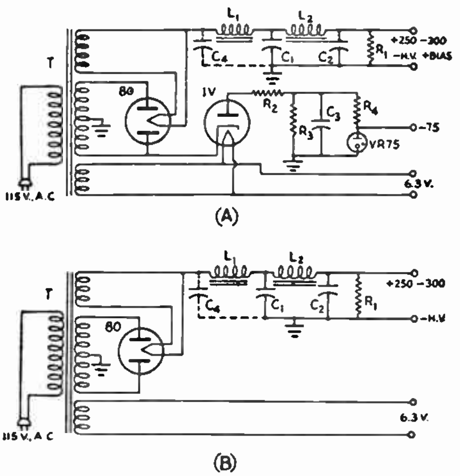


Fig. 1316 — Circuit diagram of the combination plate, screen and grid-bias power supply shown in Fig. 1315.

C_1, C_2 — Sections of 8-μfd. 450-volt dual electrolytic.
C_3 — 8-μfd. 450-volt paper.
C_4 — Same as C_3 (used only for 300-volt output).
L_1, L_2 — 6 henry, 80 ma., 138 ohms (Thordarson T-57C51).
R_1 — 20,000 ohms, 10-watt.
R_2 — 20,000 ohms, 2-watt.
R_3 — 25,000 ohms, 2-watt.
R_4 — 15,000 ohms, 2-watt.
T — 300 volts r.m.s., each side of centertap, 90 ma.; 5 volts, 3 amperes; 6.3 volts, 3.5 amperes (Thordarson T-13R13).

If desired, the bias branch may be omitted, as shown in the alternative diagram at B. All values remain as above.

A Combination Low-Voltage Plate or Screen Supply and Fixed-Bias Pack

Fig. 1315 illustrates a combination pack which will deliver 250 or 300 volts, 75 ma., for supplying plate voltage for receiving-tube exciter stages as well as screen and fixed-bias voltage for a beam-tube driver stage.

The circuit diagram is shown in Fig. 1316-A. In addition to the usual full-wave rectifier circuit employing a type 80 tube, a 1V half-wave rectifier is also connected across one-half of the transformer secondary in reverse direction to provide a negative biasing voltage which is held constant at 75 volts by the VR75-30 regulator tube. With the dropping resistor shown, the regulator tube will pass a grid current of 25 ma. without overload. The 1V rectifier is indirectly heated, so that it may be operated from the same 6.3-volt winding provided to supply the r.f. tubes in the transmitter.

The output voltage at a normal load current of about 75 ma. can be increased from 250 to about 300 by the addition of an input filter condenser, C_4, the connections for which are shown in dotted lines.

If the bias section is not needed, plate or screen voltage may be obtained with the simplified circuit shown in Fig. 1316-B, eliminating the bias section.

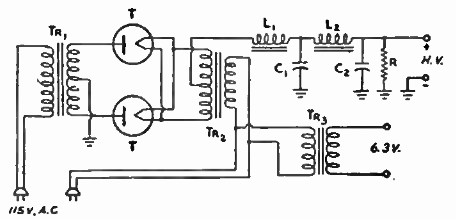


Fig. 1317 — Circuit of the power supply in Fig. 1318.

C_1 — 2-μfd. 1000-volt paper (Sprague OT21).
C_2 — 4-μfd. 1000-volt paper (Sprague OT41).
L_1 — Input choke, 6–19 henrys, 300 ma., 125 ohms (Kenyon T-510).
L_2 — Smoothing choke, 11 henrys, 300 ma., 125 ohms (Kenyon T-166).
R — 20,000 ohms, 50-watt.
T — Type 866 Jr. rectifier.
Tr_1 — 925 or 740 volts r.m.s. each side of center-tap, 300 ma. d.c. (Kenyon T-656).
Tr_2 — 2.5 volts, 10 amperes, 2000-volt insulation (Kenyon T-352).
Tr_3 — 6.3-volt 3-ampere filament transformer.

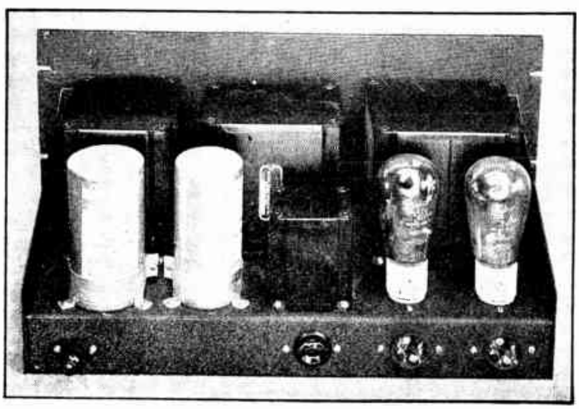


Fig. 1318 — This power-supply unit delivers either 620 or 780 volts at a full-load current of 260 ma. with 0.4 per cent ripple and regulation of 22 per cent. Voltage is changed by a tap on the plate-transformer primary winding. The filter chokes are at the left and the plate power transformer at the right on the panel side of the chassis. The can-type 1000-volt filter condensers are at the left in front and the rectifier tubes at the right, with the rectifier filament transformer in between. All exposed component terminals are underneath the chassis. The panel is 8¾×19×3 inches. The 2.5-volt 10-ampere rectifier filament transformer should have 10,000-volt insulation. A 6.3-volt filament transformer is included for heating the filaments of r.f. tubes. This transformer is mounted underneath the chassis; its output terminals are brought out to a standard a.c. receptacle in the rear. The circuit diagram is shown in Fig. 1317.

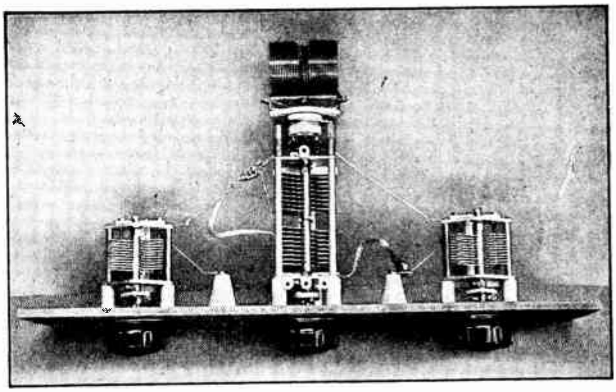


Fig. 1319 — A rack-mounting antenna tuner for low-power transmitters. C_1 is in the center, with C_2 and C_3 on either side. All of the components are mounted directly on the 5¼-inch panel. The variable condensers are mounted on the assembly rods on National type GS-1 insulating pillars which are fastened to the condenser end plates with machine screws from which the heads have been removed. Small Isolantite shaft couplings are used to insulate the controls. The coil socket is fastened to the rear end plate of the parallel condenser, C_1, with spacers to clear the prongs. Clips with flexible leads are provided for the split-stator parallel condenser, C_1, so that its sections may be connected either in parallel or in series to form either a high- or low-capacity tank circuit as required.

ℂ A Low-Power Antenna Tuner for Rack Mounting

In the rack-mounted low-power antenna tuner shown in Fig. 1319, separate series and parallel condensers are used. This arrangement, while requiring three variable condensers, has the advantage that no switching is necessary when changing over from series to parallel tuning. It also makes possible the use of the tuner to cover a considerably wider range of antenna and transmission-line conditions, because the series condensers can be adjusted in conjunction with the parallel condenser to shorten the electrical length of the feeders whenever this is required to make parallel tuning effective. In addition, the series condensers are also useful in that they provide a measure of control over the amplifier loading when parallel tuning is used.

Clips with flexible leads attached are provided for the parallel condenser, C_1, so that the sections may be connected either in parallel or in series to form either a high- or low-capacity tank circuit, as required. When the high-C parallel tank is desired, the two stators are clipped together, as shown by the dotted lines in the circuit diagram of Fig. 1320, and the rotor is connected to the opposite feeder. When the two sections are connected in series, for low-*C* operation, the break-down voltage is increased.

Under the circuit diagram, Fig. 1320, two sets of variable condensers are suggested. The smaller receiving-type condensers with 0.03-inch air gap should be satisfactory for low-power transmitters operating at plate voltages of 400 to 450 volts, while the larger condensers with 0.045-inch spacing will be required for transmitters using plate voltages up to about 750 or 1000 volts.

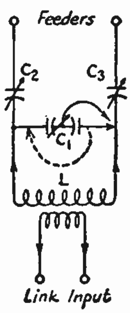


Fig. 1320 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers which are operated at less than 1000 volts on the plate.

C_1 — 100 μμfd. per section, 0.045-inch spacing (National TMK-100-D) for higher voltages; receiving-type for lower voltages (Hammarlund MCD-100).

C_2, C_3 — 250 μμfd., 0.026-inch spacing (National TMS-250) for higher voltages; receiving-type for lower voltages (Hammarlund MC250).

L — B&W JVL series coils. Approximate dimensions for parallel tuning for each band are as follows:
- 1.75-Mc. band — 56 turns No. 24.
- 3.5-Mc. band — 40 turns No. 20.
- 7-Mc. band — 24 turns No. 16.
- 14-Mc. band — 14 turns No. 16.
- 28-Mc. band — 8 turns No. 16.

All coils are 1⅞ inches in diameter and 2¼ inches long, with the variable link located at the center. For series tuning, use the coil specified for the next-higher frequency band, which will be approximately correct.

ℂ Complete 75-Watt Multi-Band Transmitter

If it is desired to use the band-switching 807 exciter unit shown in Fig. 1311 as a complete transmitter feeding the antenna, it may be combined with the power-supply units of Figs. 1315 and 1318 and the antenna tuner of Fig. 1319 (using the large 0.045-inch spacing condensers) to make a complete 75-watt transmitter unit.

The combination 250-volt power supply of Fig. 1315 will supply plate voltage for the oscillator and doubler stages, as well as screen and bias voltages for the 807. Filament supply also is obtainable from this unit. Plate voltage for the 807 is furnished by the power supply unit of Fig. 1318.

The combined height of all units (assuming the power-supply unit of Fig. 1315 to be mounted on a 7-inch panel) will be 29¾ inches. The separate filament transfromer, Tr_3, shown in the diagram of Fig. 1317 will not be required since the necessary heater power for the transmitter can be obtained from the 250-volt supply.

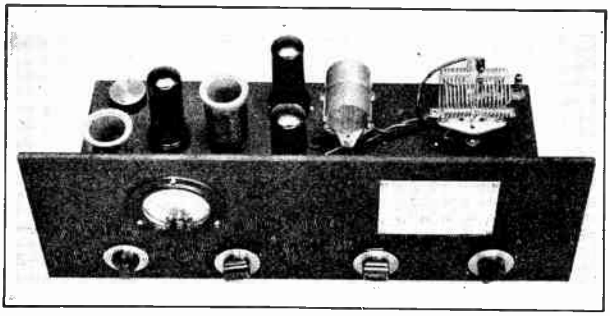


Fig. 1321 — A 90-watt c.w. transmitter using a 6L6 Tri-tet oscillator and a push-pull 6L6 amplifier. The rack-width panel of the transmitter is 7 inches high. The single milliammeter is switched from the oscillator to the amplifier by the rotary switch at the lower left. The three remaining controls are for tuning the oscillator plate, amplifier plate and antenna tank circuits. All sockets, except those for the amplifier- and antenna-tank coils are submounted. The three insulated terminals just visible at the right rear behind the antenna coil, L_4, are the binding-post output connections for the antenna tuner.

A 90-Watt C.W. Transmitter Using Push-Pull 6L6s

In the 90-watt c.w. transmitter shown in Figs. 1321 and 1322, a 6L6 Tri-tet oscillator drives a pair of 6L6s in a push-pull inverted amplifier circuit (also known as cathode coupling — § 3-3 and 4-7). The circuit diagram appears in Fig. 1323.

The sockets for the crystal and the cathode coil are wired as shown in Fig. 1372, to permit feeding with a v.f.o. unit if desired. The plate circuit of the oscillator is parallel-fed to permit grounding of the rotor of C_2 in mounting. A high-capacity tank condenser is used so that two bands may be covered with one coil, reducing coil-changing when shifting from one band to another. The cathode coil, L_5, by which the oscillator and amplifier are coupled, is center-tapped to provide push-pull input to the amplifier stage.

While neutralization is not required, a certain amount is introduced through the fixed condensers C_9 and C_{10} from plates to cathodes partially to nullify the effects of degeneration inherent in this type of circuit and thereby reduce excitation requirements. Neutralization is not carried to the point where there is danger of instability. All r.f.-wiring leads in the amplifier should be made as short and direct as possible. The individual grid condensers, C_7 and C_8 should be connected directly to the grid terminals at each socket.

The output of the amplifier is link-coupled to an antenna tuner. The lower stator of C_4 is fitted with a flexible lead terminated in an insulated banana plug which may be plugged into any one of the antenna terminals, which are jack-top binding posts. These posts are insulated from the chassis by mounting them in National polystyrene button-type insulators which have been drilled out. Series tuning with high capacity is obtained by placing the plug in terminal No. 1 and connecting the feeders to terminals Nos. 2 and 3, and series tuning with low capacity by leaving the plug free and connecting the feeders to terminals Nos. 2 and 3. High-capacity parallel tuning is obtained by placing the plug in terminal 1, shorting terminals 2 and 3, and connecting the feeders between 1 and 3, while parallel tuning with low capacity is obtained by placing the plug in terminal 3 and connecting to 1 and 3.

Both stages are keyed simultaneously in the cathode return leads. The milliammeter, *MA*, can be switched from the oscillator-cathode circuit to that of the amplifier. Switching of the meter is simplified by inclusion of the shunting resistances, R_6 and R_7, which are sufficiently high in value to have negligible effect upon the reading of the meter.

The transmitter can be operated at maximum input from the 450-volt power supply shown in Fig. 1304, provided a 200-ma. power transformer (Utah Y620E) and filter choke (Utah 4668) are substituted for those specified.

Tuning — Tuning of the transmitter is quite simple. It should be borne in mind that output from the oscillator may be obtained at

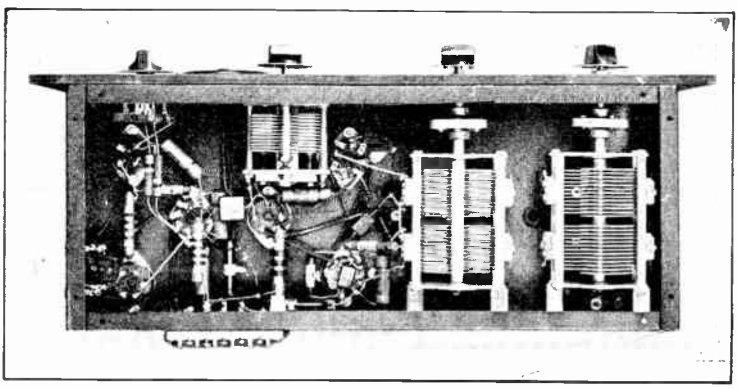


Fig. 1322 — The three tank condensers are mounted underneath the chassis of the 90-watt transmitter. The two split-stator condensers are mounted from the rear edge with insulating pillars, and their shafts are fitted with insulating couplings and panel bearings. They must be mounted so their shafts come level with that of C_2 to the left, which is mounted directly on the chassis. Heavy bare-wire leads through grommeted holes connect the amplifier and antenna tank condensers and coils.

either the fundamental frequency of the crystal or at the second harmonic of that frequency, and that the selection of the proper coil for L_1 depends upon the crystal frequency and not the output frequency of the oscillator. Using the oscillator plate coils listed under Fig. 1323, the lowest-frequency band will be found near the maximum-capacity end on the dial of C_2, while the higher-frequency bands will be found near the minimum capacity end of its tuning range.

With the milliammeter switched to the oscillator circuit, the plate-current reading should be about 60 ma. when the key is closed if a full 350 volts is used on the plate. As C_2 is tuned through resonance, the oscillator plate current will dip to about 25 ma. at the lower frequencies and to about 50 ma. at the higher frequencies.

When the meter is switched to the amplifier stage, a plate-current reading of about 260 ma. should be obtained with the key closed. A plate-current dip to 50 ma. or less should be obtained when C_3 is tuned to resonance.

Once these adjustments have been completed, the antenna may be coupled and tuned. When the plate current of the amplifier under load increases to 200 ma. as C_4 is tuned to resonance, this represents about the optimum loading condition. Using a plate voltage of 450 and with proper adjustment of the amplifier, it should be possible to obtain a power output of 50 to 60 watts on all bands.

Because of the oscillator reaction caused by modulation, resulting from use of the inverted amplifier circuit, this transmitter is recommended for c.w. work only.

Complete 90-Watt C.W. Transmitter

The 90-watt 6L6 r.f. unit of Fig. 1321 may be combined with the power-supply unit showing Fig. 1318 (with the separate 6.3-volt filament transformer, Tr_3, included to supply the greater heater power requirements of the 6L6s) to form a complete c.w. transmitter. The two units will have a combined height of 15¾ inches when they are mounted in a standard relay rack or cabinet.

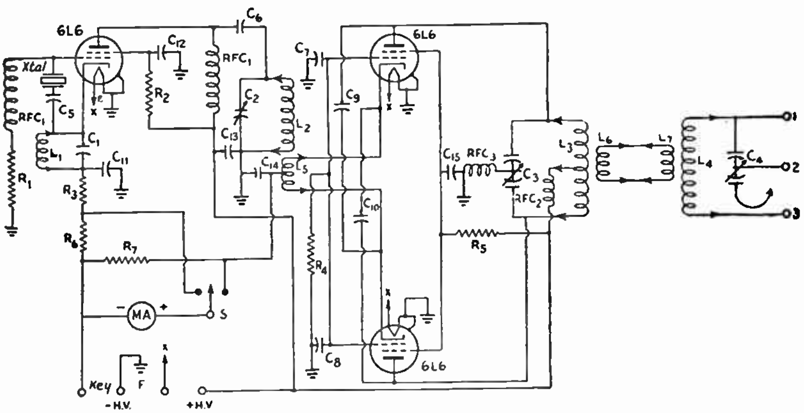


Fig. 1323 — Circuit diagram of the 90-watt push-pull 6L6 transmitter with built-in antenna coupler.

C_1 — 100-μμfd. mica.
C_2 — 250-μμfd. variable (National TMS-250).
C_3, C_4 — 250 μμfd. per section (Hammarlund MTCD-250-C).
C_5, C_6 — 0.001-μfd. mica.
C_7, C_8 — 50-μμfd. mica.
C_9, C_{10} — 10-μμfd. mica.
C_{11}, C_{12}, C_{13}, C_{14}, C_{15} — 0.01 μfd. paper.
R_1 — 0.1 megohm, ½-watt.
R_2 — 50,000 ohms, 2-watt.
R_3 — 500 ohms, 1-watt.
R_4 — 25,000 ohms, 1-watt.
R_5 — 12,000 ohms, 10-watt.
R_6, R_7 — 25 ohms, 1-watt.
MA — 0–300 milliammeter.
S — S.p.d.t. switch.
RFC_1 — 2.5-mh. r.f. choke, 100-ma.
RFC_2 — 1-mh. r.f. choke, 300-ma. (National R300).
RFC_3 — V.h.f. parasitic choke (Ohmite Z-1).

L_1 * — For 1.75-Mc. crystals: 32 turns No. 24 d.s.c., close-wound.
For 3.5-Mc. crystals: 10 turns No. 22, 1 inch long; 100-μμfd. mica condenser mounted in form, connected across winding.
For 7-Mc. crystals: 6 turns No. 22, ⅝-inch long.

L_2 * — For 1.75- and 3.5-Mc. bands — 38 turns No. 18 d.c.c. close-wound.
For 3.5- and 7-Mc. bands — 20 turns No. 18, 1⅝ inches long.
For 7- and 14-Mc. bands — 9 turns No. 18, 1½ inches long.

L_3 ** — B & W JCL series coils, dimensions as follows:
1.75 Mc. — 60 turns No. 24, 2⅛ inches long.
3.5 Mc. — 44 turns No. 20, 2⅛ inches long.
7 Mc. — 26 turns No. 16, 2⅛ inches long.
14 Mc. — 16 turns No. 16, 1⅞ inches long.

L_4 *** — B & W JVL series coils, dimensions as follows:
1.75 Mc. — 56 turns No. 24.
3.5 Mc. — 40 turns No. 20.
7 Mc. — 24 turns No. 16.
14 Mc. — 14 turns No. 16.
14 Mc. (series) — 8 turns No. 16.

L_5 — 1.75- and 3.5-Mc. bands — 20 turns, centertapped, No. 24 e., close-wound, wound close to bottom of L_2 on same form.
3.5- and 7-Mc. bands — 14 turns, centertapped, No. 22 e., close-wound, wound ⅛-inch from bottom of L_2 on same form.
7- and 14-Mc. bands — 8 turns, centertapped, No. 20 e., close-wound, wound ⅛-inch from bottom of L_2 on same form.

L_6, L_7 — 3 turns at center of L_3 and L_4.

* All wound on Hammarlund 1½-inch diameter 4-prong forms.
** All 1½ inches in diameter.
*** All 1⅞-inch diameter, 2¼ inches long. Dimensions are approximate for parallel tuning for the band indicated. For series tuning, the coil for the next-higher frequency band is approximately correct.

A Three-Stage 100-Watt Transmitter for Five Bands

The three-stage transmitter shown in Figs. 1324, 1326 and 1327 is designed to use a single 1000-volt 100-ma. tube such as the 1623, 809, HY40, or higher-voltage tubes at reduced ratings, in the output stage.

Referring to the circuit diagram of Fig. 1325, a 6L6, operating at a plate voltage of 400 but at reduced input, is used in the Tri-tet oscillator circuit. A potentiometer in the screen circuit provides a means of varying the screen voltage and, ultimately, the excitation to the final amplifier. The HY65 buffer-doubler circuit is capacitively coupled to the oscillator. This second stage makes it possible to obtain excitation for the final amplifier in a third band from a single crystal, operation in the second band being available by doubling frequency in

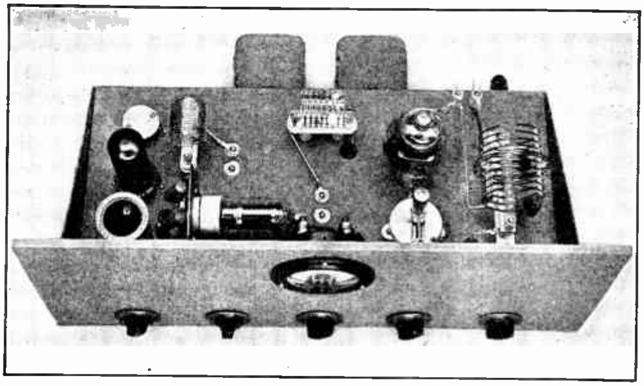


Fig. 1324 — All controls for the 100-watt five-band transmitter are below the chassis level. From left to right, they are the oscillator screen-voltage potentiometer, the oscillator plate-tank condenser, the buffer-doubler plate-tank condenser, the meter switch and the final-amplifier plate-tank condenser. The panel is of standard rack width and is 8¾ inches high.

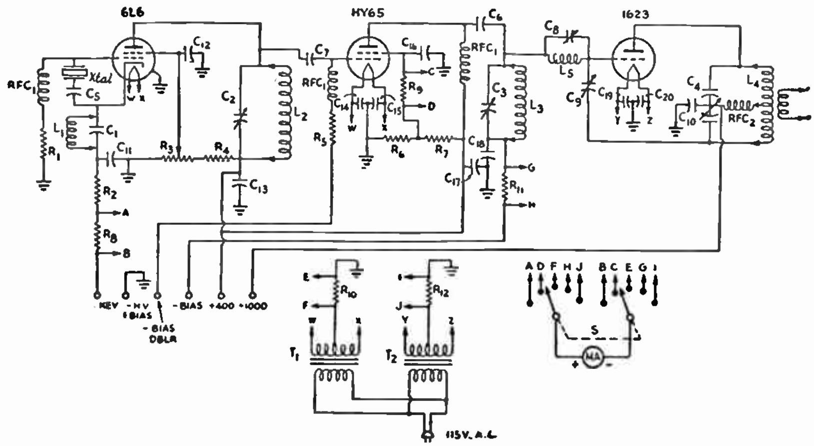


Fig. 1325 — Wiring diagram of the three-stage five-band 100-watt transmitter for 1000-volt operation.

C_1 — 100-μμfd. mica.
C_2, C_3 — 150-μμfd. variable (National ST-150).
C_4 — 100 μμfd. per section, 0.05-inch spacing (Hammarlund HFBD-100-C).
C_5, C_6 — 0.001-μfd. mica.
C_7 — 100-μμfd. mica.
C_8 — 6–60-μμfd. mica trimmer (two National M-30 in parallel).
C_9 — Neutralizing condenser (National NC-800).
C_{10} — 0.001 μfd., 5000 volts test.
C_{11}, C_{12}, C_{13}, C_{14}, C_{15}, C_{16}, C_{17}, C_{18}, C_{19}, C_{20} — 0.01-μfd. mica.
R_1 — 0.1 megohm, ½-watt.
R_2 — 300 ohms, 1-watt.
R_3 — 20,000-ohm 10-watt potentiometer (Mallory E2OMP).
R_4 — 25,000 ohms, 10-watt.
R_5 — 50,000 ohms, 1-watt.
R_6 — 20,000 ohms, 10-watt.
R_7 — 10,000 ohms, 10-watt.
R_8, R_9, R_{10}, R_{11}, R_{12} — 25 ohms, 1-watt.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 1-mh., 300-ma. r.f. choke (National R-300U).
S — Double-gang, 5-circuit switch (Mallory 3226J).
T_1, T_2 — Filament transformer, 6.3-volt, 3 amperes (UTC S-55).

L_1 — 1.75-Mc. crystals — 32 turns No. 24 d.s.c., close-wound.
3.5-Mc. crystals — 9 turns No. 22, 1 inch long; 100-μμfd. mica in form, connected across winding.
7-Mc. crystals — 6 turns No. 22, ⅝-inch long.
All on Hammarlund 1½-inch diameter forms.

L_2, L_3 — 1.75 Mc. — 56 turns, 1¼-inch diameter, 1¾ inches long, 54 μhy. (National AR80, no link).
3.5 Mc. — 28 turns, 1¼-inch diameter, 1½ inches long, 15 μhy. (National AR40, no link).
7 Mc. — 14 turns, 1¼-inch diameter, 1¼ inches long, 4.2 μhy. (National AR20, no link).
14 Mc. — 8 turns, 1¼-inch diameter, 1½ inches long, 1.25 μhy. (National AR10, no link).
28 Mc. — 4 turns, 1-inch diameter, ¾-inch long, 0.5 μhy. (National AR5, turns close, no link).

L_4 — 1.75 Mc. — 40 turns No. 18, 2½-inch diameter, 2½ inches long, 78 μhy. (B & W 160 BCL). An 80-μμfd. fixed air padder (Cardwell JD-80-OS) is placed in right-rear corner of chassis and attached to coil with flexible leads and clips.
3.5 Mc. — 32 turns No. 16, 2½-inch diameter, 2¾ inches long, 39 μhy. (B & W 80 BCL).
7 Mc. — 20 turns No. 14, 2-inch diameter, 2½ inches long, 12 μhy. (B & W 40 BCL).
14 Mc. — 8 turns No. 14, 2-inch diameter, 2 inches long, 2.5 μhy. (B & W 20 BCL). One removed turn from each end.
28 Mc. — 4 turns No. 12, 2-inch diameter, 1¾ inches long, 0.7 μhy. (B & W 10 BCL). One turn removed from each end.

L_5 — 5 turns No. 14, ½-inch diameter, ½-inch long.

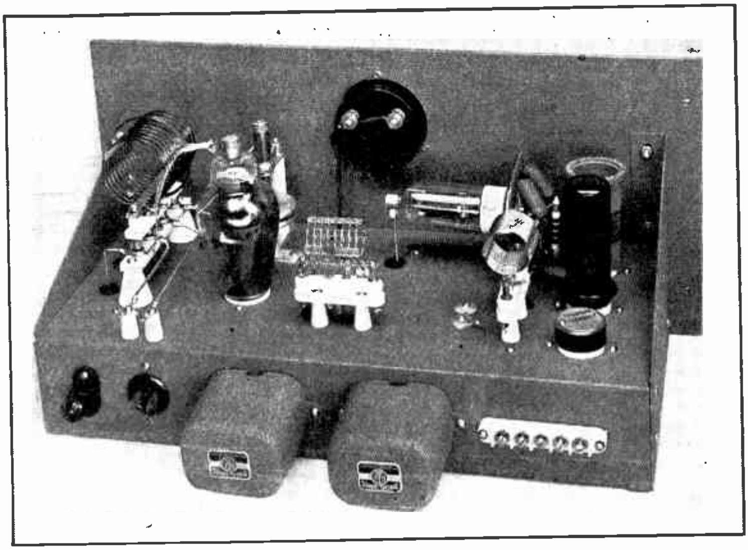


Fig. 1326 — On top of the chassis of the 100-watt transmitter, the cathode coil, L_1, the 6L6 and the crystal are in line at the right-hand end. The HY65 is mounted horizontally on a small panel which also provides mounting space for the filament and screen by-pass condensers, the coupling condenser, C_7, the grid leak, R_5, and the grid choke. L_2 is just to the left of the 6L6 and to the right of C_2 underneath. L_3 is in the center at right angles to L_2 and L_4 and just to the rear of C_3 underneath. The 1623 socket is submounted to lower the plate terminal. The neutralizing condenser, C_9, is directly in front of the tube. RFC_2 is just to the left of L_4. The two filament transformers are mounted on the rear edge.

the oscillator itself. Parallel plate feed is used in the second stage to permit series grid feed to the final amplifier, thereby avoiding the probability of low-frequency parasitic oscillations.

The neutralized final amplifier is directly coupled to the driver stage. C_8 and L_5 form a trap for v.h.f. parasitic oscillations.

The meter switch, *S*, shifts the milliammeter to read oscillator cathode current, driver screen current, driver cathode current, final-amplifier grid current and final-amplifier cathode current. The individual filament transformers permit independent metering of the cathode currents of the last two stages.

Power supply — This transmitter is designed to operate from the combination 1000-volt and 400-volt plate supply shown in Fig. 1329. Both fixed bias of 75 volts for the HY65 and cut-off bias for the final amplifier may be obtained from the unit shown in Fig. 1350. For the 1623 tube, resistors R_2 and R_3 should be 6000 ohms and 7000 ohms, respectively.

Tuning — Coils for the desired output frequency, consistent with the crystal frequency, should be plugged in the various stages, bearing in mind that frequency may be doubled in the plate circuit of the oscillator and again in the second stage, if desired. It should also be remembered that the selection of the cathode coil, L_1, depends upon the crystal frequency and not necessarily the output frequency of the oscillator, the same cathode coil being used for both fundamental and second-harmonic output from the crystal stage. Since much better efficiencies can be obtained with the HY65 operating as a straight amplifier, it is advisable to avoid doubling in this stage.

The first two stages should be tested first, with all voltages applied except the plate voltage for the final amplifier. Tuning the oscillator to resonance, with the key closed, should cause a slight dip in cathode current accompanied by an abrupt rise in the screen and cathode current of the second stage.

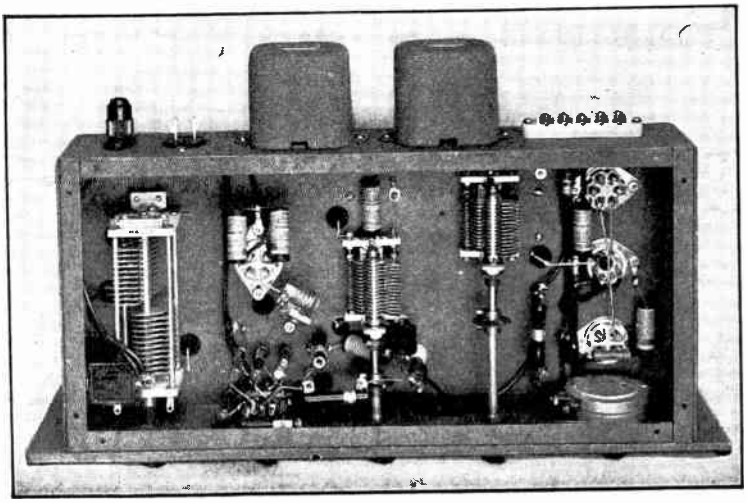


Fig. 1327 — Underneath the 8 × 17 × 3-inch chassis of the 100-watt transmitter. C_2 to the right and C_3 in the center are insulated from the chassis by polystyrene button insulators. C_4 to the left also is insulated and is spaced from the chassis to bring all shafts at the same level. Leads to the coils immediately above the tank condensers pass through large grommeted clearance holes. Meter-shunt resistances are soldered directly to the switch terminals. R_3 at the right is insulated from the chassis by extruded bakelite washers. The v.h.f. parasitic trap is suspended in the amplifier grid lead to the left of C_3. Insulating couplings are required for C_2 and C_3.

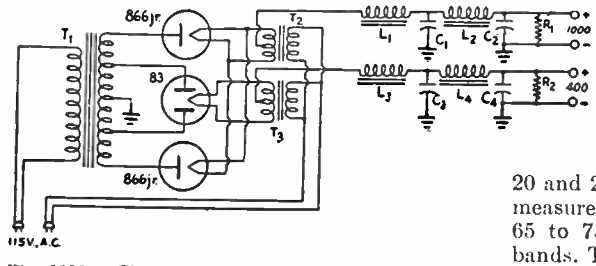


Fig. 1328 — Circuit diagram of the combination 1000- and 400-volt power supply for the 100-watt transmitter.

C_1, C_2 — 2-μfd. 1000-volt paper (Mallory TX805).
C_3 — 4-μfd. 600-volt electrolytic. (C-D) 604).
C_4 — 8-μfd. 600-volt electrolytic. (C-D 608).
L_1, L_3 — 5/20-henry swinging choke, 150-ma. (Thordarson T-19C39).
L_2, L_4 — 12-henry smoothing choke, 150-ma. (Thordarson T-19C46).
R_1 — 20,000 ohms, 75-watt.
R_2 — 20,000 ohms, 25-watt.
T_1 — High-voltage transformer, 1075 and 500 volts r.m.s. each side, 125- and 150-ma. simultaneous current rating (Thordarson T-19P57).
T_2 — 2.5 volts, 5 amperes (Thordarson T-19F88).
T_3 — 5 volts, 4 amperes (Thordarson T-63F99).

Tuning the HY65 plate circuit to resonance should produce a good dip in cathode current, with a simultaneous reading of maximum grid current to the final amplifier.

The amplifier should then be neutralized and tested for parasitic oscillation. The latter is done by shifting the final-amplifier plate-voltage lead to the 400-volt tap and turning off the bias supply. No plate voltage should be applied to the exciter stages. C_4 is then varied through its entire range for several settings of C_3. If at any point a change in the final-amplifier cathode current is observed, C_8 should be adjusted to eliminate it. During this process, plate voltage should not be applied long enough to cause appreciable heating of the tube.

Normal operating voltages may now be replaced and the final amplifier tuned up in the usual manner. A plate current of 100 ma. will indicate normal loading of the final amplifier. (Plate current will be the difference between grid and cathode currents under operating conditions.) With all stages tuned and the amplifier loaded normally, the oscillator cathode current should run between 16 and 30 ma., HY65 screen current between 6 and 11 ma., HY65 cathode current between 45 and 70 ma., HY65 grid voltage between 125 and 260 volts, oscillator screen voltage between 100 and 250 volts, and HY65 screen voltage between 210 and 250 volts, exact values depending upon whether the stage is operating at the fundamental or doubling frequency. Excitation should be adjusted to keep the amplifier grid current between 20 and 25 ma., when the grid voltage should measure 130 to 150 volts. Power output of 65 to 75 watts should be obtainable on all bands. The oscillator circuit may be arranged for optional v.f.o. input as shown in Fig. 1372, if desired.

If the output stage is to be plate-modulated, the plate voltage should be reduced to 750. Operating data for suitable tubes of other types will be found in the tables in the Appendix.

Complete 100-Watt 5-Band Transmitter

The transmitter of Fig. 1324 may be combined in a standard rack with other units to form a complete transmitter. Plate voltage for oscillator and driver as well as for the final-amplifier stage may be obtained from the duplex power supply shown in Fig. 1329. Bias voltage for both driver and final-amplifier stages may be obtained from the combination unit shown in Fig. 1350, with fixed bias for the HY65 being taken from the VR75-30 branch. A suitable antenna tuner is the one shown in Fig. 1319. The larger variable condensers should be used. The total height of the various units combined is 29¾ inches, allowing a 7-inch panel for the bias-supply unit.

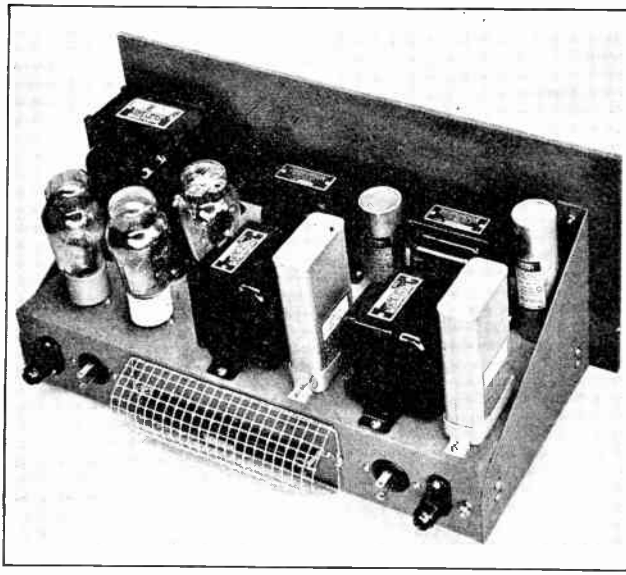


Fig. 1329 — This power supply makes use of a combination transformer and dual filter system delivering 1000 volts at 125 ma. and 400 volts at 150 ma., simultaneously. The circuit diagram is given in Fig. 1328. The 1000-volt bleeder resistor is mounted on the rear edge of the chassis, with a protective guard made of a piece of galvanized fencing material to provide ventilation. Millen safety terminals are used for the two high-voltage terminals. Ceramic sockets should be used for the 866 Jrs. The chassis measures 8 × 17 × 3 inches and the standard rack panel is 8¾ inches high.

A Two-Stage 200-Watt Beam-Tube Transmitter

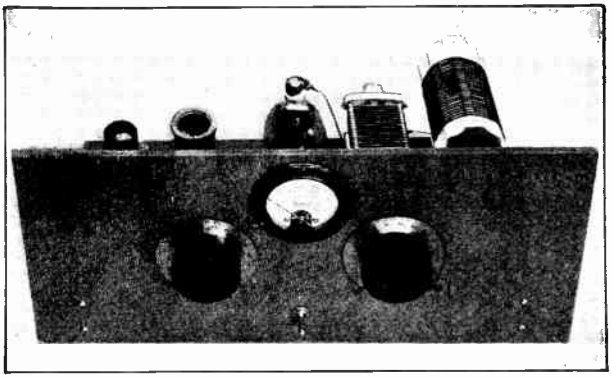


Fig. 1330 — **Front view of the 200-watt beam-tube transmitter.**

The simplicity of the 200-watt transmitter shown in Figs. 1330, 1332 and 1333 will appeal to many amateurs. As the circuit of Fig. 1331 shows, a 6L6 Tri-tet oscillator supplies excitation at either the crystal fundamental frequency or its second harmonic for the HY67 in the output stage. Since the latter is a screened tube, no neutralizing is required. Parallel feed in the oscillator circuit permits mounting C_2 on the chassis without insulation. The milliammeter may be switched to read either oscillator or amplifier cathode current. R_5 in series with the screen prevents parasitic oscillation.

Power supply — A 300-volt supply is required for the plate of the oscillator and the screen of the amplifier. This voltage, as well as fixed biasing voltage for the amplifier, may be obtained from the combination unit in Fig. 1315, using the components shown for 300-volt output. The supply shown in Fig. 1334 will furnish plate voltage for the amplifier.

Tuning — The simplicity of the circuit makes tuning easy. With a cathode coil, L_1, appropriate for the crystal in use, and an oscillator plate coil, L_2, which covers the crystal frequency with C_2 near maximum, the oscillator is tuned either to the fundamental frequency, near the maximum of C_2, or to the second harmonic, near the minimum capacity of C_2, by the customary plate-current dips. The key should not be kept closed for prolonged periods during this adjustment unless the 300-volt lead to the screen of the amplifier is disconnected. If plate-current dips for both fundamental and harmonic are not found, add or subtract turns from L_2 so that the range will be centered to cover both.

The amplifier stage should be tuned to resonance, with the proper coil for the desired

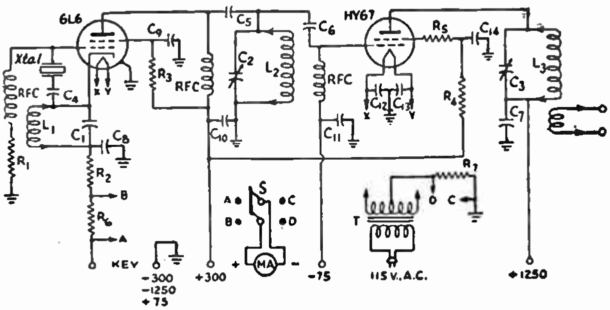


Fig. 1331 — **Circuit diagram of the two-stage 200-watt beam-tube transmitter.**

C_1 — 100-$\mu\mu$fd. mica.
C_2 — 300-$\mu\mu$fd. variable (National TMS-300).
C_3 — 250-$\mu\mu$fd. variable, 0.045-inch spacing (National TMK-250).
C_4, C_5 — 0.001-μfd. mica.
C_6 — 100-$\mu\mu$fd. mica.
C_7 — 0.001-μfd. mica, 5000-volt rating.
C_8, C_9, C_{10}, C_{11}, C_{12}, C_{13}, C_{14} — 0.01-μfd. paper.
MA — D.c. milliammeter, 0-300-ma. scale.
R_1 — 0.1 megohm, ½-watt.
R_2 — 500 ohms, 1-watt.
R_3 — 50,000 ohms, 10-watt.
R_4 — 2000 ohms, 10-watt.
R_5 — 50 ohms, 1-watt.
R_6 — 25 ohms, 1-watt.
R_7 — 25 ohms, 10-watt.
RFC — 2.5-mh. r.f. choke.
S — Double-pole double-throw toggle switch.
T — Filament transformer, 6.3 volts, 6 amperes (Thordarson T-19F98).
L_1 — For 1.75-Mc. crystals — 32 turns No. 24 d.s.c., close-wound.
For 3.5-Mc. crystals — 10 turns No. 22, 1-inch long; C_1 is mounted in form, connected across winding.
7-Mc. crystals — 6 turns No. 22, ⅝-inch long.
L_2 — 1.75- and 3.5-Mc. bands — 30 turns No. 20 e., 1½ inches long.
3.5- and 7-Mc. bands — 15 turns No. 18 e., 1½ inches long.
7- and 14-Mc. bands — 6 turns No. 18 e., ⅞-inch long.
All above coils wound on Hammarlund 1½-inch diameter coil forms.
L_3 — 1.75-Mc. band — 32 turns No. 18 d.c.c., 3½ inches long.
3.5-Mc. band — 20 turns No. 12, 3 inches long, turns wound in successive grooves.
7-Mc. band — 9 turns No. 12, 1⅜ inches long, turns wound in successive grooves.
14-Mc. band — 6 turns No. 12, 1¾ inches long, turns wound in alternate grooves.
All above wound on National XB-10A 2½-inch diameter coil forms. The form for the 1.75-Mc. coil is covered with a sheet of cardboard before winding. Number of link turns is adjusted for proper loading.

Fig. 1332 — The cathode coil, oscillator tube and crystal sockets of the 200-watt beam-tube transmitter are mounted in a row at the right. The socket for the HY67 is sunk an inch below the chassis level to shorten the plate lead, and the bottom portion of the tube is shielded with a section from a Hammarlund PTS tube shield. The amplifier plate tank condenser is insulated from the chassis by means of National polystyrene button-type insulators placed at the three mounting feet. The filament transformer and other small components are underneath the chassis, as may be seen in Fig. 1333. Power output of 130 to 150 watts on c.w. can be obtained on any of the bands covered. If the amplifier is to be plate-and-screen modulated, the input should be reduced to 1000 volts at 150 ma.

output frequency in place, after switching the meter to read the HY67 cathode current. The amplifier may then be link-coupled to an antenna tuner, such as the one shown in Fig. 1336, and loaded in the usual way. As a matter of fact, it is preferable to tune the amplifier with the load connected, after one has become accustomed to the tuning procedure, so as to limit screen heating.

Under normal conditions, the oscillator cathode current will run between 35 and 40 ma. when tuned to resonance in any band, while the cathode current of the amplifier should be about 225 ma. when fully loaded. This total cathode current will include screen current of about 30 ma. and grid current of about 20 ma. The oscillator screen voltage should be between 175 and 200 volts, while the amplifier screen voltage will run about 240 volts with the amplifier tuned and loaded. Power output of 130 to 150 watts should be obtainable on any of the four bands covered.

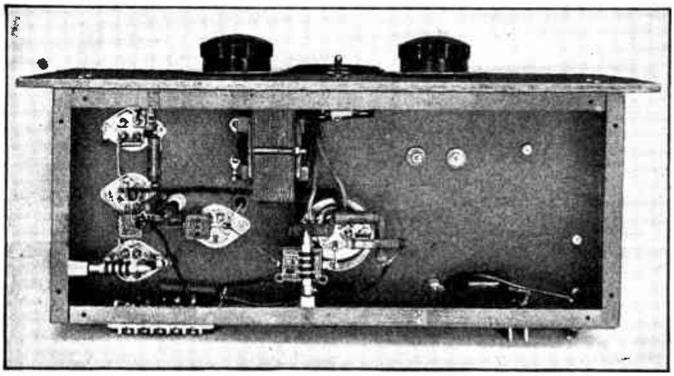


Fig. 1333 — Underneath the 3 × 7 × 17-inch chassis of the two-stage 200-watt beam-tube transmitter.

If desired, the oscillator circuit may be arranged for v.f.o. input as shown in the diagram of Fig. 1372. If the amplifier is to be plate-screen modulated, the input should be reduced to 1000 volts, 150 ma.

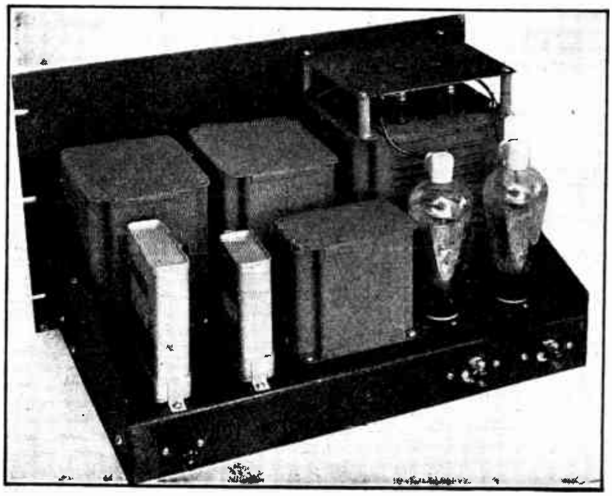


Fig. 1334 — This power supply unit delivers 830, 1060 or 1250 volts at 250 ma. The required voltage is selected by taps on the secondary. Ripple is only 0.25 per cent and the regulation is about 10 per cent. The transformer terminal board is covered with a panel mounted on pillars at the four corners. Insulating caps are provided for the tube plate terminals. A Millen safety terminal protects the high-voltage connection. The chassis measures 11 × 17 × 2 inches and the panel size is 10½ × 19 inches. The circuit is the same as that in Fig. 1349, the following components being used:

C_1 — 2-μfd. 1500-volt (Aerovox Hyvol).
C_2 — 4-μfd. 1500-volt (Aerovox Hyvol).
L_1 — Input choke, 5-25 henrys, 300 ma., 90 ohms (UTC S34).
L_2 — Smoothing choke, 15 henrys, 300 ma., 90 ohms (UTC S33).
R — 25,000 ohms, 100-watt.
T_{r1} — 1500-1250-1000 volts r.m.s. each side, 300-ma. d.c. (UTC S47).
T_{r2} — 2.5 volts, 10 amperes, 10,000-volt insulation (UTC S57).

ℭ Antenna Tuner for Medium Power

The antenna tuner shown in Fig. 1336 will usually be satisfactory for amplifiers operating at plate voltages not in excess of 1250 volts.

The two condensers are mounted from the panel by means of insulating pillars taken from National GS-1 insulators, which are fastened to the end plates with small sections of machine screws from which the heads have been cut. The variable link coil is mounted between the two rear end plates. The size of the coil is varied by short-circuiting turns, using clips which are attached to the condensers with flexible leads. As shown by the circuit diagram, Fig. 1335, the condensers are connected in parallel when the second pair of clips connects each rotor to the stator of the opposite condenser. The feeders are connected to the two large stand-off insulators mounted on the panel.

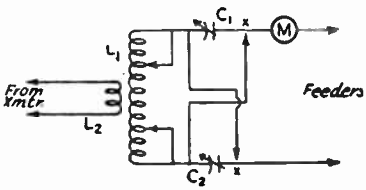


Fig. 1335 — Circuit diagram of the link-coupled antenna-tuning unit for use with medium-power transmitters.

C_1, C_2 — 100-μμfd. variable, 0.07-inch spacing (National TMC-100).

L_1 — 22 turns No. 14, diameter 2¾ inches, length 4 inches (Coto with variable link).

L_2 — 4 turns, rotating inside L_1.

M — R.f. ammeter, 0-2.5-ampere range for medium-power transmitters.

ℭ Complete 200-Watt Beam-Tube Transmitter

The units of Figs. 1330, 1334 and 1336 may be combined with that of Fig. 1315 to form a complete transmitter which will occupy a total height in a relay rack of 31½ inches. Plate voltage for the oscillator and screen and bias supply for the HY67 are obtained from the unit of Fig. 1315 (values for 300-volt output), which may be mounted on a 7-inch panel. Plate voltage of 1250 for the HY67 is obtained from the power-supply unit shown in Fig. 1334.

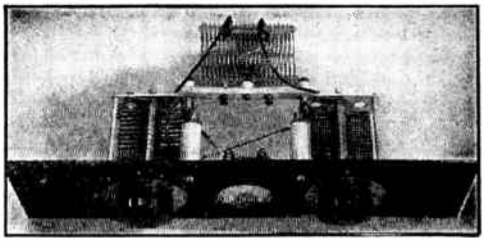


Fig. 1336 — A link-coupled antenna-tuning unit for use with resonant feed systems and medium-power amplifiers. The inductance, with variable link, is mounted on the condenser frames. Clips are provided for changing the number of turns and for switching the condensers from series to parallel. The panel is 5¼ × 19 inches.

ℭ A Push-Pull Amplifier for 200 to 500 Watts Input

Figs. 1337, 1339 and 1340 show various views of a compact push-pull amplifier using tubes of the 1500-volt 150-ma. class, although the design is also suitable for use with tubes of the 1000-volt 100-ma. class. With the lower plate voltages, a plate tank condenser with a spacing between plates of 0.05 inch and smaller tank coils may be used.

The circuit, shown in Fig. 1338, is quite conventional, with link coupling at both input and output. C_{11} and C_{12} are plug-in fixed air capacitors for the 1.75-Mc. band, to eliminate the necessity for an unduly large variable tank condenser to cover this one band. The tuned circuits, L_3C_6 and L_4C_5, are traps important for the prevention of v.h.f. parasitic oscillations. The 100-ma. meter may be shifted between the grid and cathode circuits for reading either grid current or cathode current. When shifted to read cathode current, the meter is shunted by a resistor, R_2, which multiplies the scale reading by five. This resistor is wound with No. 26 copper wire, the length being determined experimentally to give the desired scale multiplication.

Construction — The mechanical arrangement shown in the photographs results in a compact unit requiring a minimum of panel space. The tank condenser is mounted on the left-hand partition (Fig. 1339) at a height

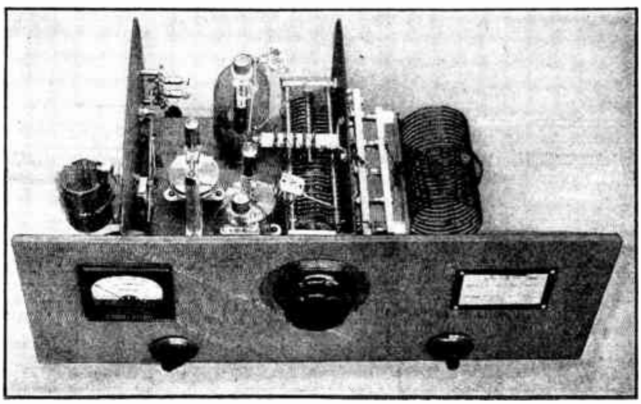


Fig. 1337 — A general view of the compact 450-watt push-pull amplifier, showing the front panel and top-of-chassis arrangement. Mounted on a standard relay rack, the height is only 7 inches and the depth 9 inches. Grid and plate tank circuits are isolated from each other by the double shielding partitions. On the panel are the 0–100 ma. milliammeter, which is switched to read current in all circuits, the plate-tank tuning dial, and a chart giving coil and tuning data. The small knob at the left below is the grid-circuit tuning control, while the one to the right is for the meter switch. The tube sockets are mounted adjacent to the stator terminals of the plate-tank condenser, C_2, in the center, with the neutralizing condensers between, providing short leads.

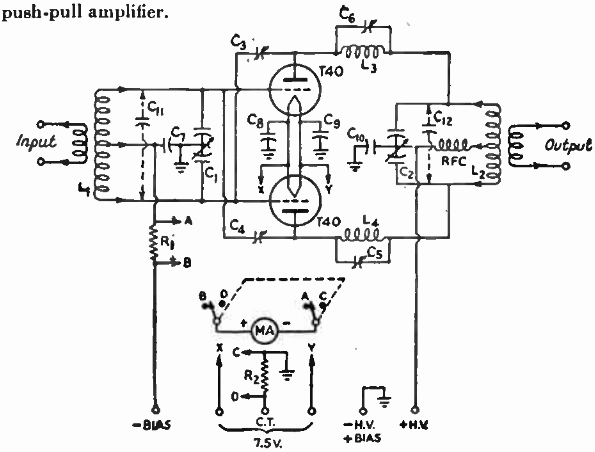


Fig. 1338 — Circuit diagram of the 450-watt push-pull amplifier.

C_1 — 100-μμfd. per section, 0.03-inch spacing (Hammarlund HFAD-100-B).
C_2 — 100-μμfd. per section, 0.07-inch spacing (Hammarlund HFBD-100-E).
C_3, C_4 — Neutralizing condensers (National NC-800).
C_5, C_6 — 3-30-μμfd. mica trimmers (National M-30).
C_7, C_8, C_9 — 0.01-μfd. mica.
C_{10} — 0.001-μfd. mica, 7500-volt rating (Aerovox 1653).
C_{11} — 50-μμfd. air padder (for 1.75 Mc.) 0.05-inch spacing (Cardwell EO-50-FS).
C_{12} — 50-μμfd. air padder (for 1.75 Mc.), 0.125-inch spacing (Cardwell JD-50-OS).
R_1 — 25 ohms, 1-watt.
R_2 — Meter-multiplier resistance for 5-times multiplication, wound with No. 26 wire.
RFC — 1-mh. r.f. choke (National R-154U).
MA — Milliammeter, 100-ma.
L_1 — B & W JCL series, dimensions as follows: *
1.75 Mc. — 60 turns No. 24, 2⅛ inches long.
3.5 Mc. — 44 turns No. 20, 2⅛ inches long.
7 Mc. — 26 turns No. 16, 2⅛ inches long.
14 Mc. — 14 turns No. 16, 1⅞ inches long (remove 2 turns from B & W coil).
28 Mc. — 6 turns No. 16, 1⅞ inches long (remove 2 turns from B & W coil).
L_2 — B & W TCL series, dimensions as follows: **
1.75 Mc. — 28 turns No. 12, 4¾-inch diameter, 4¼ inches long.
3.5 Mc. — 26 turns No. 12, 3½-inch diameter, 4½ inches long.
7 Mc. — 22 turns No. 12, 2½-inch diameter, 4½ inches long.
14 Mc. — 10 turns No. 12, 2½-inch diameter, 4¼ inches long, remove one turn from each end.
28 Mc. — 4 turns ⅛-inch copper tubing, 2½-inch diameter, 4½ inches long. Remove one turn from each end.
L_3, L_4 — 4 turns No. 14, ½-inch diameter, ¾-inch long.

* All 1½-inch diameter, 3-turn links.
** All coils fitted with 2-turn links.

which brings its shaft down 2⅝ inches from the top of the panel. The plate tank-coil jack bar is mounted centrally with the condenser on spacers which give a ½-inch clearance between the strip and the partition. The socket for the plate padder, C_{12}, is mounted in the lower rear corner of the left-hand partition. C_{10} is mounted with a small angle on the partition under the center of C_2. Leads from both ends of the rotor shaft are brought to one side of C_{10} for symmetry.

The two tube sockets are mounted in a line through the center of the chassis and at opposite ends of the plate tank condenser. They are spaced about one inch below the chassis on long machine screws. The neutralizing condensers are placed between the two tubes, so that the leads from the plate of one tube to the grid of the other are short. The r.f. choke is mounted just above the tank condenser.

The right-hand partition is cut out at the forward edge to clear the meter. This cut-out can be readily made with a socket punch and a hacksaw. The socket for the grid tank coil is mounted 4½ inches behind the panel, just above the chassis line. The grid-circuit padder, C_{11}, is fitted with banana plugs which mount in jacks set in the right-hand partition just behind the grid coil. The jacks are insulated from the metal by being mounted in National polystyrene button insulators which have been drilled out to fit.

The grid tank condenser, C_1, is mounted under the chassis without insulation. Large clearance holes, lined with rubber grommets, are drilled for connecting wires which must be

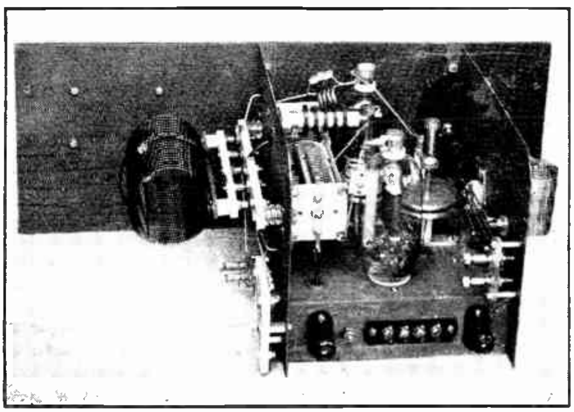


Fig 1339 — All components of the 450-watt push-pull amplifier are assembled around a small metal chassis 7 × 2 × 9 inches deep. The partitions are standard 6½ × 10-inch interstage shields. The plate tank condenser is mounted on the left-hand partition. The plate tank-coil jack-bar is mounted centrally, opposite the condenser on spacers which give ½-inch clearance between the strip and the partition. The socket for C_{12} is mounted in the lower rear corner of the left-hand partition. C_{10} is mounted with a small angle bracket on the partition under the center of C_2. The socket for the grid tank coil is mounted just above the chassis line. The grid-circuit padder, C_{11}, is fitted with banana plugs which mount in jacks set in the partition behind the grid coil. The jacks are insulated from the metal partition by polystyrene button-type insulators. Millen safety terminals are used for the external high-voltage plate and bias connections.

run through the chassis or partitions. The parasitic traps are made self-supporting in the plate leads from the tank condensers to the tube caps. The panel is placed so that the plate tank-condenser shaft comes at the center. The meter switch is mounted to balance the knob controlling C_1.

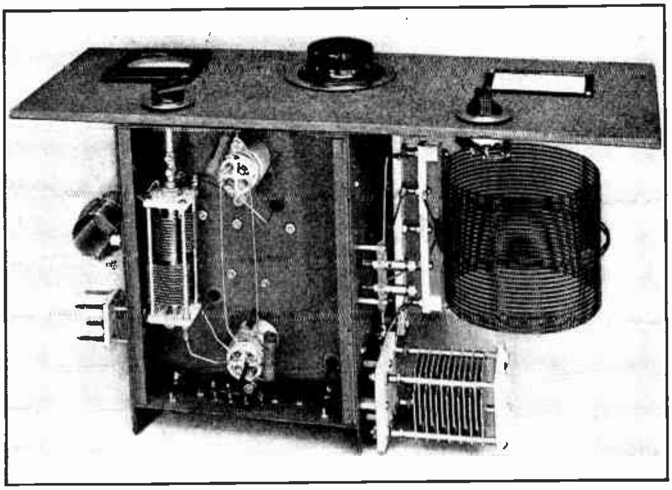


Fig. 1340 — Bottom view of the 450-watt push-pull amplifier, showing the position of the grid tank condenser between the two submounted tube sockets. The two air padding condensers, C_{11} and C_{12}, are in place for 1.75-Mc. operation.

Power supply and excitation — The T40 tubes shown in the photographs operate at a maximum plate voltage of 1500 for c.w. work. For this, the unit shown in Fig. 1348 is suitable. The supply shown in Fig. 1351, minus the VR-tube branch, will provide the biasing voltage required for plate-current cut-off. R_2 should have a resistance of 2500 ohms and R_3 of 1500 ohms. A filament transformer delivering 7.5 volts at 5 amperes also will be required; it may be mounted on the bias-supply chassis, if desired. The exciters of Figs. 1311 or 1307 will furnish adequate excitation.

Tuning — After the amplifier has been neutralized, a test should be made for parasitic oscillation. The bias should be reduced until the amplifier draws a plate current of about 100 ma. without excitation. With C_1 adjusted to various settings, C_2 should be varied through its range and the plate current watched closely for any abrupt change. Any change will indicate oscillation, in which case C_5 and C_6 should be adjusted simultaneously in slight steps until the oscillation disappears. Unless the wiring differs appreciably from the original, complete suppression will be obtained with the two condensers at full capacity. Changing bands should have no effect upon this adjustment.

With normal bias replaced, the amplifier should now be tuned up and the excitation adjusted so that a grid current of 60 ma. is obtained with the amplifier fully loaded. Full loading will be indicated when the cathode-current meter registers 360 ma., which includes the 60-ma. grid current. Under these conditions the biasing voltage should rise to 150 volts, dropping to about 70 volts without excitation when the plate current will fall to almost zero.

If the amplifier is to be plate-modulated, the plate voltage should be reduced to 1250 and the loading decreased to reduce the plate current to 250 ma. The same bias-supply adjustment will be satisfactory for this type of operation but excitation may be reduced to give a grid current of 40 ma., bringing the total cathode current to 290 ma. The antenna tuner shown in Fig. 1336 or the pi-section network of Fig. 1342 may be used.

Operating conditions for tubes of other characteristics will be found in the Appendix.

A Pi-Section Antenna Coupler

The photograph of Fig. 1342 shows the constructional details of a pi-section type antenna coupler. The wiring diagram appears in Fig. 1341. All parts are mounted directly on the panel using flathead machine screws. The condensers each are supported on three ceramic pillars from National type GS-1 stand-off insulators. A ¾-inch 6–32 machine screw is inserted in one end of each pillar and turned tight. The head of the screw is then cut off with a hacksaw and the protruding quarter-inch or so is threaded into the mounting holes in the end plate of the condenser. The shaft is cut off about ¼ inch from the frame and fitted with a Johnson rigid insulated shaft coupling (No. 252). Since the coupling will extend beyond the stand-off insulators, a ¾-inch clearance hole should be cut in the panel for each shaft. Alternatively, metal washers could be used between the panel and each pillar to extend the mounting.

Each coil form is supported on 1½-inch cone insulators. The two high-voltage blocking condensers, C_3, also are mounted on pillars from GS-1 stand-offs. A copper clip on a flexible lead, connected permanently to one end of each coil, permits adjustment of the coil inductance by short-circuiting turns.

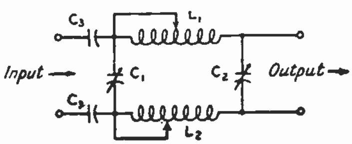


Fig. 1341 — Diagram of the pi-section antenna coupler.
C_1–C_2 — 300-μμfd. variable, 0.07-inch spacing (National TMC-300).
C_3 — 0.01-μfd. mica., 5000-volt rating.
L_1, L_2 — 26 turns No. 14, 2½-inch diameter, 3½ inches long (National XR10A form wound full).

Output connections are made to the two terminal insulators at the right, while input connections are made to the terminals of the two voltage-blocking condensers. When single-wire output is desired, the output terminal connected to the condenser rotors is grounded and the coil in that side short-circuited by the clip and lead.

Under most circumstances the components specified will work satisfactorily with transmitters of 400 or 500 watts input, operating at plate voltages up to 1500. For higher power, the condensers should have greater spacing and the coils should be wound with No. 12 or larger wire. Couplers for lower power may be made using smaller components of equal values.

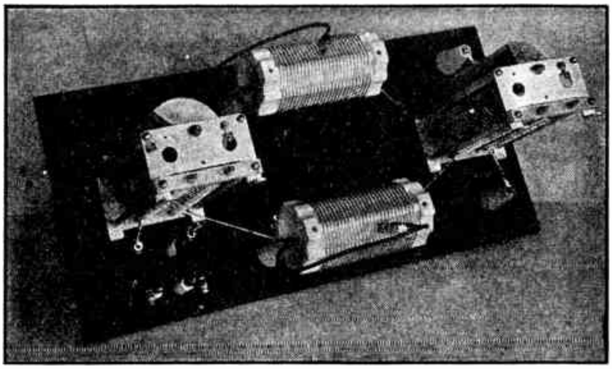


Fig. 1342 — Pi-section type antenna coupler. All parts are mounted on a Presdwood panel 8 × 19 inches. The circuit is given in Fig. 1341.

Complete 300- to 400-Watt Compact Plug-In Coil Transmitter

The compact exciter and amplifier units of Figs. 1307 and 1337 may be combined as a complete transmitter. Plate and filament supply for the exciter may be obtained from the unit of Fig. 1318. Plate voltage for the amplifier may be obtained either from the unit of Fig. 1334 or that of Fig. 1348. A 7.5-volt 5-ampere filament transformer may be combined on a 5¼-inch panel with the unit of Fig. 1350 (minus the VR75-30 branch), which will furnish bias for the amplifier. A 45-volt "B" battery will be required for biasing the 807.

Suitable antenna tuners are those of Figs. 1336, 1352 or 1342. The height of all units, including a 5¼-inch meter panel is 49 inches.

A 450-Watt Band-Switching Amplifier

The photographs of Figs. 1343, 1344, 1346 and 1347 illustrate a 450-watt push-pull band-switching amplifier capable of handling a power input of 450 watts at 1500 volts for c.w. operation or 375 watts with plate modulation. While the type T55 is shown, any of the comparable triodes in the 1000- or 1500-volt class, such as the 809, T40, HY40, RK35, UH50, 808, 812, RK51 or 35T, may be used in a similar arrangement.

The circuit is shown in Fig. 1345. Band-switching is accomplished by short-circuiting turns of both plate and grid coils by means of tap switches. Any three adjacent bands may be covered in this manner. By plugging in another pair of coils, a second set of three adjacent bands may be covered. Thus the 1.75-, 3.5- and 7-Mc. bands may be covered with one pair, 3.5, 7 and 14 Mc. with another pair, and 7, 14 and 28 Mc. bands with a third pair.

A plug-in fixed air condenser is required for the plate circuit for the 1.75-Mc. band. The plug-in jack base is mounted under the chassis and is wired to the lowest-frequency switch points, so that the condenser is automatically connected across the coil when the switch is turned for the 1.75-Mc. band. When the coil covering this band is not used, the fixed condenser should be removed; it may be omitted entirely if operation in this band is not desired. The grid circuit likewise requires padding at 1.75 Mc., but here a 15-$\mu\mu$fd. condenser may be connected permanently across the fourth set of switch contacts, which are

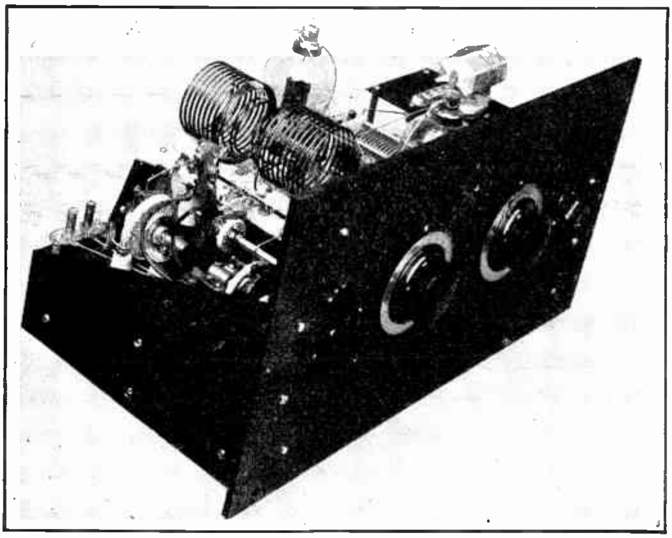


Fig. 1343 — A 450-watt band-switching amplifier. The panel size is 10½ × 19 inches. The large dials on the panel control the plate and grid tank condensers. The uppermost of the two small knobs to the left is for adjusting the variable-link output coupling, while the lower knob is for the plate band-switch. The grid band-switch knob is to the right. All controls should be well insulated.

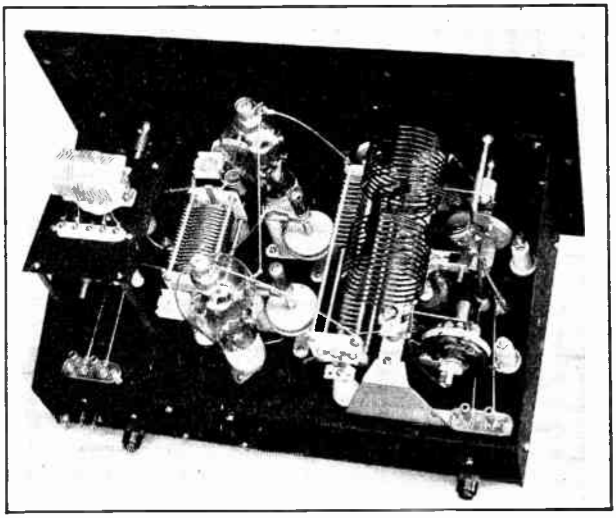


Fig. 1344 — Rear view of the 450-watt amplifier. The plate tank-coil jack bar at the right is mounted on brackets 2⅞ inches high so that the variable-link shaft will clear the switches. These are mounted on 1-inch cone insulators after their brackets have been revamped to bring the shafts 1½ inches above the chassis. The units are spaced so as to be central with the jack-bar terminals. The shafts are coupled with a section of ⅜-inch bakelite shaft fitted with brass reducing couplings at each end. The two feed-through insulators are for connections to the padder-condenser jack base underneath. The tank condenser is mounted on 1½-inch cone insulators. The plate r.f. choke and a feed-through insulator for high-voltage line are placed beneath the jack bar. The grid tank condenser is mounted to bring its shaft even with that of the plate condenser. The grid switch is mounted on insulators to balance the plate switch. The grid coil mounting is elevated over the switch. The tubes and the two neutralizing condensers are placed symmetrically between the two tank circuits.

not used for other bands. C_9L_3 and $C_{10}L_4$ are parasitic traps to eliminate v.h.f. parasitic oscillations. Fixed-link coupling is used at the input, with variable-link output coupling.

Coils — The plate-tank coils listed under the circuit diagram are of a special series designed primarily for use with a multi-section tank condenser. They are provided with four extra plugs which are used, in this case, for the short-circuiting taps. The coil covering 7, 14 and 28 Mc. requires slight alteration, however. Two turns on each side of center are cut free from the supporting strips and left self-supporting; otherwise, the coil heating which usually occurs at 28 Mc. may be sufficient to ruin the base strip. At the same time, these two turns on each side should be reduced in diameter to 1⅞ inches. This may be done quite readily by unsoldering the central ends, twisting the turns to the smaller diameter, and cutting off the excess wire. While the lower-frequency taps may be soldered, it is advisable to use clamps on the wire for the 28-Mc. taps. Johnson coil clips are suitable for this purpose.

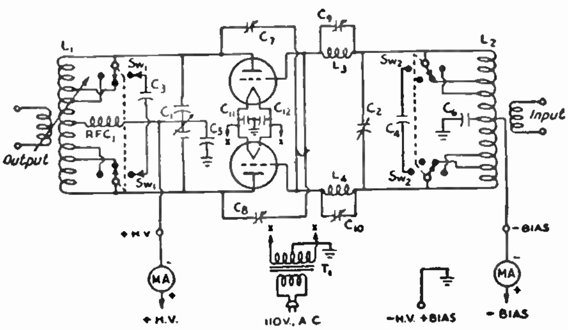


Fig. 1345 — Circuit diagram of the 450-watt amplifier.

C_1 — 100 μμfd. per section, 0.07-inch plate spacing (Hammarlund HFBD-100-E).
C_2 — 150 μμfd., 0.05-inch plate spacing (Hammarlund HFB-150-C).
C_3 — 50-μμfd. fixed air padder for 1.75 Mc., spacing 0.17 inch or greater (Cardwell JCO-50-OS).
C_4 — 15-μμfd. padder for 1.75 Mc., 0.05-inch spacing (see text) (Hammarlund HFA-15-E).
C_5 — 0.001-μfd. 7500-volt mica (Aerovox 1623).
C_6 — 0.01-μfd. paper.
C_7, C_8 — Neutralizing condenser (National NC800).
C_9, C_{10} — Isolantite mica trimmer, 20–100-μμfd. (Mallory CTX954).
C_{11}, C_{12} — 0.01-μfd. paper.
RFC_1 — 1-mh. r.f. choke, 600-ma. (National R154).
Sw_1 — Ganged sections of Ohmite BC-3 band-change switch.
Sw_2 — Ganged sections of Mallory 162C Hamband switch.
T_1 — 7.5-volt 6-ampere filament transformer (Thordarson T-19F94).

L_1 — For 1.75-, 3.5- and 7-Mc. bands — 60 turns No. 16, 5⅜ inches long, 2½-inch diameter, tapped at the 7th and 16th turn each side of center (B & W TVH-160) (90 μhy., tapped each side of center at 7/30 and 8/15 of the total turns in each half).
For 3.5-, 7- and 14-Mc. bands — 38 turns No. 14, 5¼ inches long, 2½-inch diameter, tapped at the 4th and 9th turn each side of center (B & W TVH-80 35 μhy., tapped each side of center at 2/19 and 9/38 of the total turns in each half).
For 7-, 14- and 28-Mc. bands — 24 turns No. 12, 5¼ inches long, 2½-inch diameter, tapped at 2nd and 5th turns each side of center (see text for alterations) (B & W TVH-40) 13 μhy., tapped each side of center at approximately ⅙ and 1/5½ of the total turns in each half.
L_2 — For 1.57-, 3.5- and 7-Mc. bands — 52 turns, 2 inches long, 1½-inch diameter, tapped at 9th and 17th turns each side of center. (Coto CS-160C) (56 μhy., tapped each side of center at 9/26 and 17/26 of the total turns in each half).
For 3.5-, 7- and 14-Mc. bands — 26 turns, 1½ inches long, 1½-inch diameter, tapped at 5th and 9th turns from each side of center. (Coto CS80C) (17 μhy., tapped each side of center at 5/13 and 9/13 of the total turns in each half).
For 7-, 14- and 28-Mc. bands — 16 turns 1⅞ inches long, 1½-inch diameter, tapped at 1st and 3rd turns each side of center. (Coto CS40C) (5 μhy., tapped each side of center at ⅛ and ⅜ of the total turns in each half).
L_3, L_4 — 8 turns No. 12, ½-inch inside diameter, 1⅛ inches long.

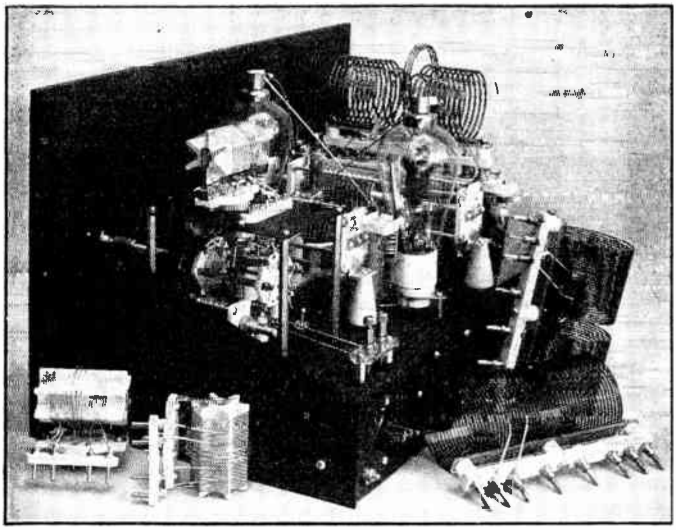


Fig. 1346 — A view of the grid-circuit end of the band-switching push-pull amplifier, showing the coil-switching arrangement and the 1.75-Mc. padder.

Grid coils with sufficient mounting pins being unobtainable the taps for the grid coils are brought out to a five-prong Millen coil-mounting bar (Type 40205). A plug-in socket for the bar is sub-mounted in back of the coil socket.

Wiring — All of the wiring, except for the power wiring underneath the chassis, is done with No. 14 tinned bus wire. Wherever possible, connections are made with short, straight pieces of wire running directly from point to point. Of most importance are the leads to the tube grids and plates. The leads to the tank condensers and those to the neutralizing condensers must be kept entirely separate; at no point should these leads be common. This practice helps in the prevention of parasitic oscillations. The grid by-pass condenser is mounted close to the grid-coil socket.

Fig. 1345 shows how d.c. milliammeters of suitable ranges may be connected for reading the grid and plate currents. These are not included in the unit, but may be mounted in a separate meter panel constructed as shown in Fig. 1381. The grid-current meter should have a 100 ma. scale, while the plate-current meter should have a range of 500 ma.

Tuning — Any one of the r. f. units shown in Figs. 1311, 1307 or 1324 will furnish sufficient excitation for this amplifier, the band-switching exciter of Fig. 1311 being recommended as an excellent companion unit.

Before excitation is applied, the two parasitic-trap condensers, C_9 and C_{10}, should be set at maximum capacity. With excitation applied and plate voltage off, grid current to the amplifier stage should run between 60 and 90 ma.

As the next step the amplifier should be neutralized, using the grid-current meter as a neutralization indicator. To test the amplifier for parasitic oscillation, the bias should be reduced to a point which will allow a plate current of 100 ma. or so to flow without excitation. This may be done by moving the biasing tap of the amplifier down toward the positive terminal of the bias supply. It is advisable to lower the plate voltage for this test, either by inserting a resistance of about 2500 ohms in series with the plate-voltage source or by inserting a 200-watt lamp in series with the primary winding of the plate transformer. The grid tank condensers should be set at various points while the plate tank condenser is swung through its range. The plate current should remain perfectly stationary while this is done. If a point is found where a sudden change in plate current takes place, C_9 and C_{10} should be adjusted, bit by bit, until the variation in plate current disappears. C_9 and C_{10} should be as close to

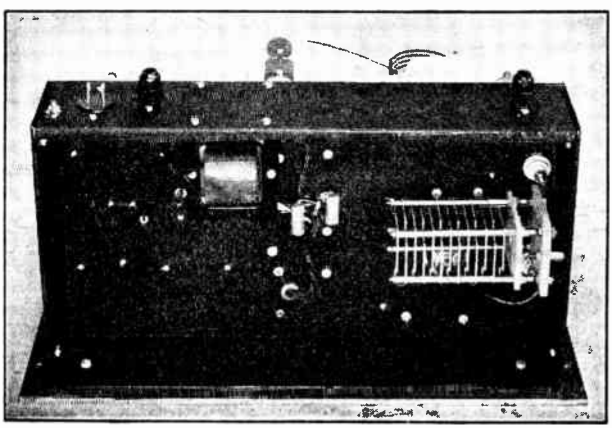


Fig. 1347 — Bottom view of the 450-watt band-switching amplifier. The chassis measures 10 × 17 × 3 inches. The plug-in air padding condenser for 1.75 Mc. is mounted so that it is an equal distance between top and bottom of the chassis. Filament by-pass condensers are soldered to the terminals of a fiber lug strip to which the filament transformer terminals are anchored. Millen safety terminals are used for bias and high-voltage output connections. A suitable 1500-volt plate-power unit for use with this amplifier is shown in Fig. 1348, the circuit diagram for which appears in Fig. 1349. The circuit diagram of a simple bias pack is shown in Fig. 1351. If this bias pack is used, the VR75-30 and the resistor, R_1, should be omitted and R_2 and R_3 made approximately 4000 ohms each for T55s. The two power-supply units may be combined on a single chassis.

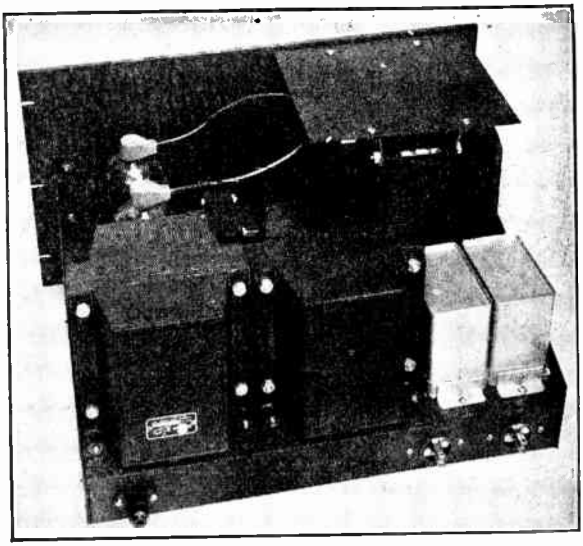


Fig. 1348 — This power supply delivers 1500 or 1250 volts at a full-load current of 425 ma., with 0.25 per cent ripple and regulation of 10 per cent. Voltages are selected by taps on the transformer secondary. The secondary terminal board is covered with a section of steel panel supported by brackets fastened underneath the core clamps and insulating caps are provided for the tube plate terminals. A special safety terminal (Millen) is used for the positive high-voltage connection. The panel is 10½ × 19 inches and the chassis size is 13 × 17 × 2 inches. The circuit for this supply is shown in Fig. 1349.

maximum capacity as it is possible to set them and yet eliminate the parasitic oscillation.

Normal biasing voltage may now be replaced and the amplifier tuned up and loaded. For c.w. operation, the output should exceed 300 watts when operated at the maximum rated input of 1500 volts, 300 ma. With plate modulation, the plate current should be reduced to 250 ma. and the output should exceed 250 watts. The amplifier will operate satisfactorily when the grid current is 40 to 70 ma. with the plate circuit loaded. The maximum rating of 80 ma. for the two tubes should not be exceeded.

Reference should be made to the vacuum-tube tables in the Appendix for data on the operation of other type of tubes.

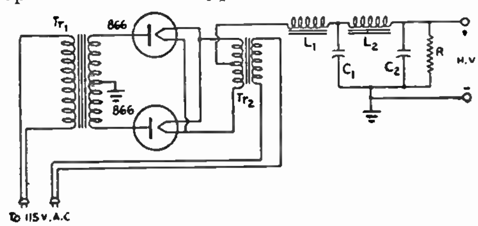


Fig. 1349 — Circuit diagram of the 1500-volt 425-ma. plate power supply for the band-switching amplifier.
C_1, C_2 — 4-μfd. 2000-volt paper (C-D TJU20040).
L_1 — 5–20 henrys, 500 ma., 75 ohms (Stancor C1405).
L_2 — 8 henrys, 500 ma., 75 ohms (Stancor C1415).
R — 20,000 ohms, 150-watt.
Tr_1 — 1820–1520-volts r.m.s. each side of center-tap, 500-ma. d.c. (Stancor type P6157).
Tr_2 — 2.5 volts, 10 amperes, 10,000-volt insulation (Stancor type P3025).
This circuit is also used for the 1250-volt supply shown in Fig. 1334 and the 2500-volt supply shown in Fig. 1362.

ℂ A Simple Combination Bias Supply

Fig. 1351 shows the circuit diagram of the simple transformerless bias unit, pictured in Fig. 1350, which may be used to supply cut-off bias voltages up to 100 volts or so. Through grid-leak action it will also provide the additional operating bias voltage required, if the resistor values are correctly proportioned. The circuit also includes a second branch, consisting of R_1 and a VR75-30 voltage-regulator tube, supplying regulated voltage. This branch may not be required in all cases, but will be found convenient in many applications for providing fixed cut-off or protective bias for a low-power stage independent of the main output voltage.

Adjustment — The voltage-divider resistances, R_2 and R_3, are combined in a single resistor with two sliding taps. One of these taps alters the total resistance by short-circuiting a portion of the resistance at the negative end, while the other adjusts the cut-off voltage. The method of determining the values of resistance in each section is as follows:

The bias section, R_3, is adjusted to equal the recommended grid-leak resistance for the tube or tubes in use. The value of resistance between the biasing tap and the short-circuiting tap is determined by the following formula:

$$R_3 = \frac{160 - E_{co}}{E_{co}} \times R_2,$$

where E_{co} is the voltage required for plate-

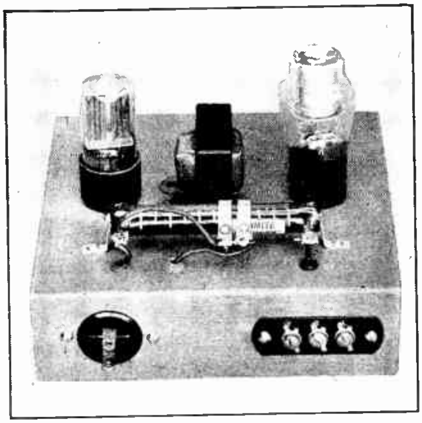


Fig. 1350 — A transformerless combination bias supply suitable for supplying bias for r.f. stages requiring 125 volts or less for cut off. A second branch, controlled by a VR75-30 regulator tube, provides 75 volts fixed bias for a second stage whose grid current does not exceed 20 ma. The unit above is constructed on a 7 × 7-inch chassis, although the components may easily be fitted into any spare space on another power-supply chassis. The regulated VR-tube branch may be omitted if not required. The circuit diagram is shown in Fig. 1351.

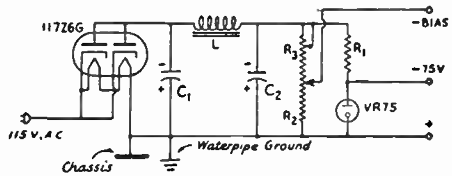


Fig. 1351 — Circuit diagram of the transformerless bias supply with voltage-regulated output shown in Fig. 1350.
C_1, C_2 — 16-μfd. 450-volt electrolytic.
L — 60-ma. replacement filter choke.
R_1 — 7500 ohms, 10-watt.
$R_2 + R_3$ — 15,000-ohm 50-watt wire-wound, with two sliders.
See text for details of adjustment and operation.

current cut-off. This value may be determined to a close approximation for triodes by dividing the plate voltage by the amplification factor of the tube. No supplementary grid-leak bias should be used in the stage being supplied by the pack.

The resistance in each section should be first set at the values determined by the formula. The biased amplifier should then be turned on, without excitation. If the plate current is not almost completely cut off, or at least reduced to a safe value, the biasing tap should be moved upward (in the negative direction). With the amplifier in operation and drawing rated grid current, the biasing voltage should be measured, using a high-resistance voltmeter. If the grid voltage is higher than that recommended in the tube operating tables, *both* the biasing tap and the short-circuiting tap on the upper section should be moved, bit by bit, toward the positive end until the correct operating bias is obtained. The bias voltage should then be measured again. A final adjustment may be necessary to again arrive at cut-off voltage without excitation.

Fig. 1350 shows the components assembled separately on a small chassis. They may, however, be combined with plate-supply components on a single chassis, since little additional space will be required.

It will be noticed in the circuit diagram that only one wire is shown connected to the power plug. The return connection for circuit is made through an actual ground connection to the chassis, to prevent possible short-circuit of the 115-volt line should the power plug happen to be incorrectly polarized when inserted.

ℂ A Wide-Range Antenna Coupler

The photograph of Fig. 1352 shows the constructional details of a wide-range antenna coupler suitable for use with high-power transmitters. Various combinations of parallel and series tuning, with high- and low-*C* tanks and high- and low-impedance outputs, are available. Diagrams of these various circuit combinations possible with this arrangement are given in Fig. 1353.

A separate coil is used for each band, and the desired connections for series or parallel tuning with high or low *C*, or for low-impedance output with high or low *C*, are automatically made when the coil is plugged in. Coil connections to the pins for various circuit arrangements are shown in Fig. 1353.

The tuning condenser specified, together with a set of standard plug-in transmitting coils, should cover practically all coupling conditions likely to be encountered.

Because the switching connections require the use of a central pin, a slight alteration in the B & W coil-mounting unit is required. The central link mounting unit should be removed from the jack bar and an extra jack placed in the central hole thus made available. The link assembly should then be mounted on a 2-inch cone insulator to one side of the jack bar.

Correspondingly, the central nut on each coil plug base must be removed and a Johnson tapped plug, similar to those furnished with the coils, substituted. An extension shaft may then be fitted on the link shaft and a control brought out to a knob on the panel.

The split-stator tank condenser is mounted by means of angle brackets on four 1-inch cone-type ceramic insulators, and an insulated flexible coupling is provided for the shaft.

If desired, the coils may be wound with fixed links on ceramic transmitting coil forms. The links should be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for seven connections.

The unit as described should be satisfactory for transmitters operating at a plate voltage of up to 1500 with modulation and somewhat more on c.w. For appreciably higher voltages, a tank condenser with larger plate spacing should be used.

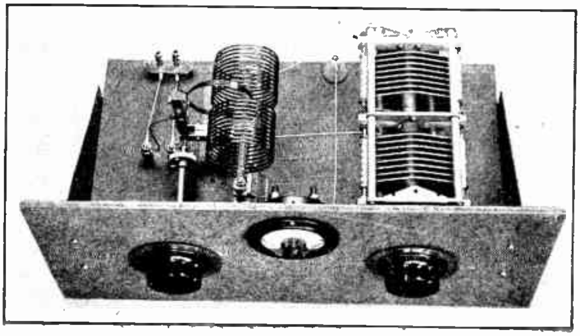


Fig. 1352 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel $8\frac{3}{4}$ × 19 inches in size. The variable condenser is a split-stator unit having a capacity of 200 μμfd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B & W TVL series. The r.f. ammeter has a 4-ampere scale. If desired, the coils may be wound with fixed links on standard transmitting ceramic forms. The links will have to be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for the seven plug-in connections required.

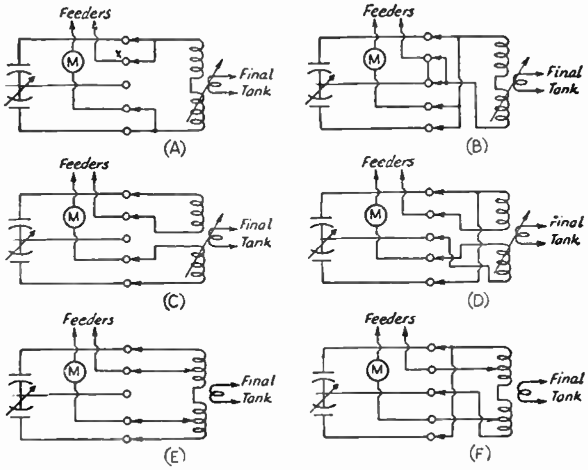


Fig. 1353 — Circuit diagram of the wide-range antenna coupler for use with the band-switching amplifier. A — Parallel tuning, low *C*. B — Parallel tuning, high *C*. C — Series tuning, low *C*. D — Series tuning, high *C*. E — Parallel tank, low-impedance output, low *C*. F — Parallel tank, low-impedance output, high *C*. For single-wire matched-impedance feeders, the arrangements of E or F would be used with a single tap instead of the double tap shown. For simple voltage-fed antennas, the arrangement of A would be used with the end of the antenna connected at "X." After the inductance required for each of the various bands has been determined experimentally, the connections to the coils can be made permanent. Then it will be necessary merely to plug in the right coil for each band, tune the condenser for resonance, and adjust the link for loading.

Complete 450-Watt Band-Switching Transmitter

The various units shown in Figs. 1311, 1316B, 1318, and 1343 through 1353, assembled together, form a complete high-power band-switching transmitter.

Heater, low-voltage plate and the 807 screen-voltage supply for the exciter may be obtained from the simplified 250-volt pack of Fig. 1316-B, while plate voltage for the 807 is furnished by the unit of Fig. 1318. Bias voltages for both amplifier and exciter are obtainable from the unit of Fig. 1350, while amplifier plate voltage is furnished by the unit of Fig. 1348. The units of Figs. 1316-B and 1350 may be combined in a single unit with a 7-inch panel. The addition of a 5¼-inch panel for the amplifier grid and plate meters and the antenna tuner of Fig. 1352 completes the transmitter.

The most logical arrangement for the units, from top to bottom, is as follows: (1) antenna tuner, (2) final amplifier, (3) meter panel, (4) exciter, (5) low-voltage and bias supplies, (6) 750-volt power supply, (7) high-voltage power supply. The combined height of all of these units will be 59½ inches.

Information on a suitable control circuit for such a transmitter will be found on pages 299–300.

Power supply and tuning — The plate power supply shown in Fig. 1362 may be used with this unit. Bias may be obtained from the unit shown in Fig. 1350. For this purpose, the VR75-30 branch may be omitted and a single resistor of 5000 ohms connected across the output of the pack, with the bias lead connected to the extreme negative end of the resistor.

The transmitter shown in Fig. 1324 should provide sufficient excitation. Fig. 1355 shows milliammeters connected in grid and plate leads. These meters are not included in the unit. They should be mounted on a separate well-insulated panel protected with a glass cover (see Fig. 1381).

An amplifier operating at high voltage should always, after neutralizing, be tuned up at reduced plate voltage. This may be obtained by connecting a lamp bulb in series with the primary of the plate transformer. Coupling between the exciter and the amplifier should be adjusted so that the grid current does not exceed 40 to 50 ma. with the amplifier tuned

A Single-Tube 500-Watt Amplifier

A single-tube amplifier which may be operated at inputs up to 500 watts at voltages as high as 3000 is shown in Figs. 1354, 1356 and 1357. The circuit, shown in Fig. 1355, is strictly conventional, with link coupling for both input and output circuits. While a Type 100TH tube is shown in the photographs, almost any other tube of similar physical size and shape which is designed to operate at plate voltages of 3000 or less may be used in a similar circuit arrangement.

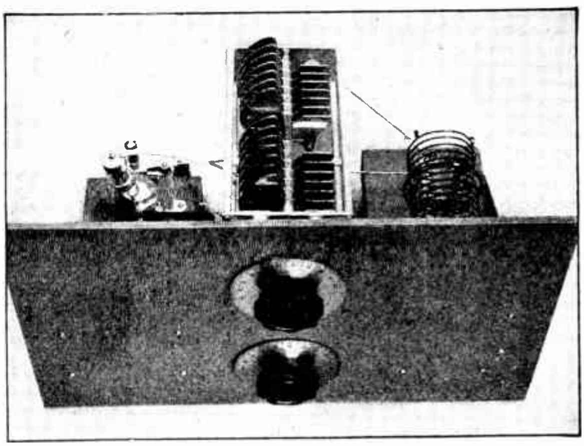


Fig. 1354 — A single-tube high-power amplifier for high-voltage inputs up to 500 watts. The standard rack panel is 12¼ inches high.

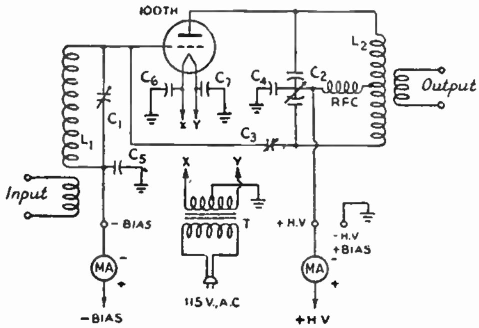


Fig. 1355 — Circuit of the 500-watt input amplifier.

C_1 — 250-$\mu\mu$fd. variable, 0.047-inch spacing (National TMK-250).
C_2 — 100 $\mu\mu$fd. per section, 0.171-inch spacing (National TMA-100-DA).
C_3 — Neutralizing condenser (National NC-800).
C_4 — High-voltage insulating condenser, 0.001-μfd. mica, 12,500-volt rating (Cornell-Dubilier 21A-86).
C_5, C_6, C_7 — 0.01-μfd. mica.
RFC — 1-mh. r.f. choke, 300-ma. (National R-300U mounted on GS-1 insulator).
MA_1 — Grid milliammeter, 100 ma.
MA_2 — Plate milliammeter, 300 ma.
T — Filament transformer — 5 volts, 8 amperes. (Thordarson T-19F84).
L_1 — 3.5 Mc. — 26 turns No. 16, 1½-inch diameter, 2⅛ inches long, 3-turn link (B & W JCL-40).
7 Mc. — 16 turns No. 16, 1½-inch diameter, 1⅞ inches long, 3-turn link (B & W JCL-20).
14 Mc. — 8 turns No. 16, 1½-inch diameter, 1⅞ inches long, 3-turn link (B & W JCL-10).
28 Mc. — 6 turns No. 16, 1½-inch diameter, 1½ inches long, 2-turn link (B & W JCL-10, 1 turn removed from each end).
L_2 — 3.5 Mc. — 26 turns No. 12, 3½-inch diameter, 4½-inches long, 2-turn link (B & W TCL-80).
7 Mc. — 22 turns No. 12, 2½-inch diameter, 4½-inches long, 2-turn link (B & W TCL-40).
14 Mc. — 12 turns No. 12, 2½-inch diameter, 4¼-inches long, 2-turn link (B & W TCL-20).
28 Mc. — 6 turns ⅛-inch copper tubing, 2½-inch diameter, 4½ inches long, 2-turn link (B & W TCL-10).

and loaded to the rated plate current of 167 ma. Power output of 225 to 300 watts should be obtainable on all bands at plate voltages from 2000 to 3000.

The tube tables in the Appendix should be consulted for data on the operation of other tubes suitable for use in this amplifier.

A Push-Pull 1-Kilowatt Amplifier

The push-pull amplifier shown in the photographs of Figs. 1359, 1360 and 1361 is capable of handling a power input of 1000 watts for c.w. operation or 900 watts with plate modulation.

The circuit is shown in Fig. 1358. Plug-in coils with fixed links are used in the grid circuit, while the output-coil mounting is provided with variable link coupling. L_3C_3 and L_4C_4 form traps against v.h.f. parasitic oscillation. Special multi-section plate tank condenser, C_2, provides a low minimum capacity for operation at the higher frequencies and the high maximum capacity needed for the lower frequencies.

Construction — The plate-tank tuning condenser is mounted on 1¼-inch ceramic cone insulators. The rotor is grounded through a high-voltage fixed condenser at the front end of the variable-condenser frame. The shaft is cut off and is fitted with a large Isolantite flexible shaft coupling. This is important, since the rotor is at high voltage. A panel-bearing assembly is fitted in the panel. The jack bar for the plate tank coil is mounted on a pair of angle brackets fastened to the condenser end plates. Two 300-ma. r.f. chokes in parallel are

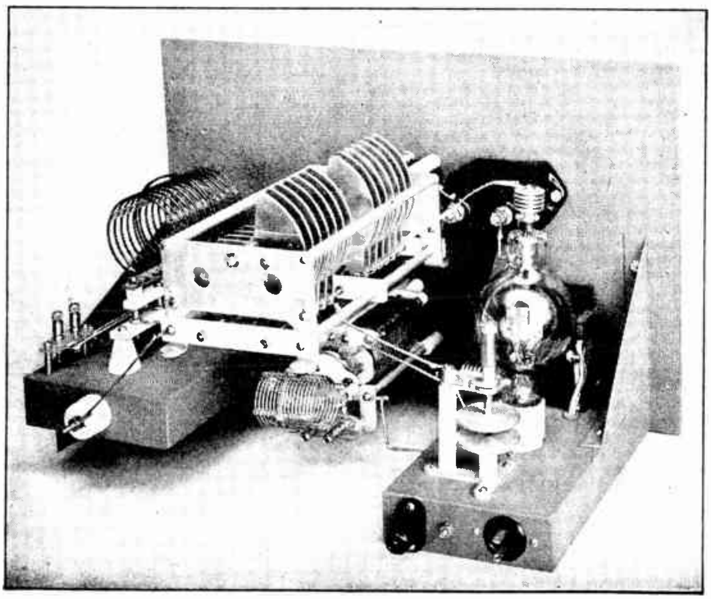


Fig. 1356 — Rear view of the high-power single-tube amplifier. The two tank condensers are mounted, one above the other, in the center of the panel by means of Isolantite pillars from stand-off insulators. Four National type GS-2 insulators are used to support the plate tuning condenser, while three type GS-1 insulators are used for the grid tuning condenser. Insulated flexible couplings and panel bearings are used on each shaft to insulate the controls. One of high break-down voltage rating should be used for the plate condenser, and *the panel bearings must be grounded!* The socket for the grid tank coil is mounted, using insulated spacers and a small metal plate as a base, on the rear end plate of C_1. Metal strips, also fastened to the end plate, support the input-link terminal strip. The insulating by-pass condenser, C_4, is mounted just to the right of C_2.

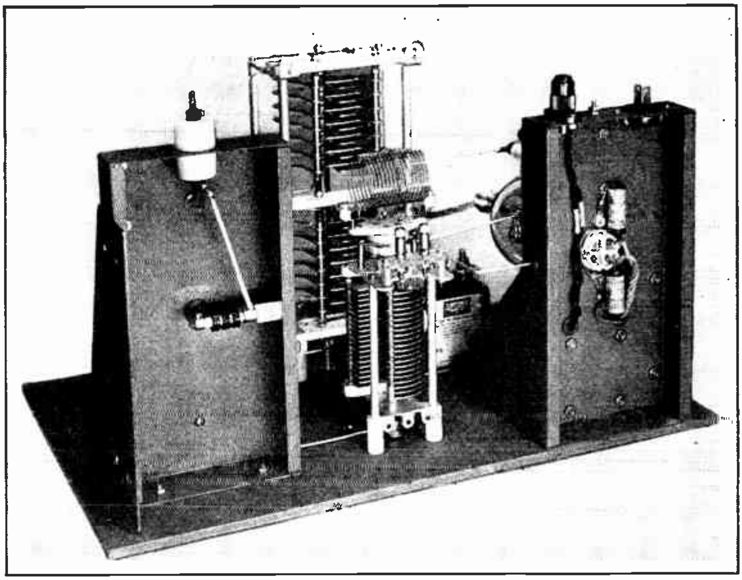


Fig. 1357 — Bottom view of the single-tube 500-watt amplifier. In the lower right-hand corner of the panel is fastened a chassis 9½ × 5 × 1½ inches, on which are mounted, in line, the filament transformer, the tube socket and the neutralizing condenser. A chassis of similar size to the left supports the plate tank coil and the output-link terminals. A large feed-through insulator in the rear edge of this chassis serves as the high-voltage terminal. In wiring the amplifier unit, the importance of well-spaced leads carrying high voltage cannot be stressed too greatly. It must be remembered that the arcing distances and break-down capabilities of voltages as high as 3000 are considerably greater than with the lower plate voltages more commonly used by amateurs.

used, one being connected between each condenser end plate and the center connections of the coil jack bar. The positive high voltage comes up through the chassis through a feed-through insulator at the rear of the condenser.

The grid tank condenser is mounted on ⅝-inch cone insulators topped with spacers which bring its shaft up level with that of the plate tank condenser. The two variable condensers are mounted with their shafts 3⅛ inches from the chassis edges. The jack bar for the grid tank coil is mounted on U-shaped brackets made from ½-inch brass strip, and these, in turn, are mounted on 2-inch cone insulators. The rotor of the grid tank condenser is grounded to the chassis at the center. The grid r.f. choke is mounted on a feed-through insulator carrying the biasing voltage up through the chassis. The grid by-pass condenser is soldered between the top of the r.f. choke and the rotor ground connection for the condenser.

The two tubes are mounted centrally with respect to the two tank condensers, the neutralizing condensers being placed between the tubes and the grid tank condenser. The sockets for the tubes are sub-mounted beneath the chassis on ⅝-inch spacers to lower the plate terminals. The parasitic-trap condensers and

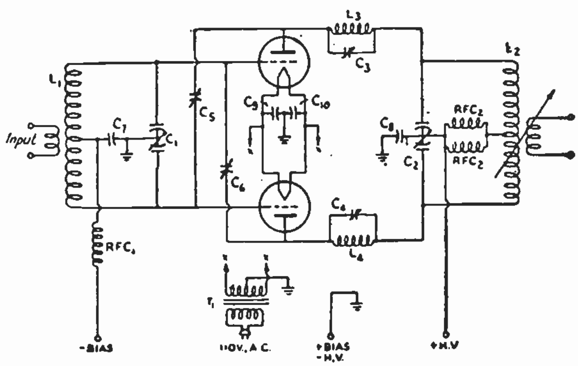


Fig. 1358 — Circuit diagram for the high-power 1-kilowatt input push-pull amplifier.

C_1 — 150 μμfd. per section, 0.05-inch spacing (Johnson 150FD20).
C_2 — Multi-section, maximum capacity 228 μμfd. per section, 0.84-inch spacing (Cardwell XE-160-70-XQ).
C_3, C_4 — 3–30-μμfd. mica trimmer condensers with Isolantite insulation (Millen 28030).
C_5, C_6 — Neutralizing condensers (Johnson N250).
C_7 — 0.01-μfd. 600-volt paper.
C_8 — 0.001-μfd. mica, 10,000-volt rating (Aerovox 1624).
C_9, C_{10} — 0.01-μfd. paper.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 1-mh. 300-ma. r.f. choke (National R-300).
T_1 — 10-volt 10-ampere filament transformer (Thordarson T-19F87).
L_1 — 1.75 Mc. — 42 turns No. 14, 3 inches long, 3½-inch diameter (110 μhy.)
3.5 Mc. — 32 turns No. 16, 2¾ inches long, 2½-inch diameter (40 μhy.) (B & W 80BL).
7 Mc. — 20 turns No. 14, 2½ inches long, 2-inch diameter (12 μhy.) (B & W 40BL).
14 Mc. — 10 turns No. 14, 2½ inches long, 2-inch diameter (3 μhy.) B & W 20BL).
28 Mc. — 6 turns No. 12, 2½ inches long, 2-inch diameter (1 μhy.) (B & W 10BL).
L_2 — 1.75 Mc. — 48 turns No. 14, 6¾ inches long, 3½-inch diameter (90 μhy.) (B & W 160HDVL).
3.5 Mc. — 32 turns No. 10, 6¾ inches long, 3½-inch diameter (40 μhy.) (B & W 80HDVL).
7 Mc. — 20 turns No. 8, 6¾ inches long, 3½-inch diameter (15 μhy.) (B & W 40HDVL).
14 Mc. — 8 turns No. 8, 4¾ inches long, 3½ inch diameter (3 μhy.) (B & W 20HDVL with one turn removed from each end).
28 Mc. — 4 turns 3/16-inch copper tubing or No. 4 wire, 5¼ inches long, 2⅜-inch inside diameter (0.8 μhy.) (B & W 10HDVL with one turn removed from each end).
L_3, L_4 — 6 turns No. 12, ½-inch inside diameter, ¾-inch long.

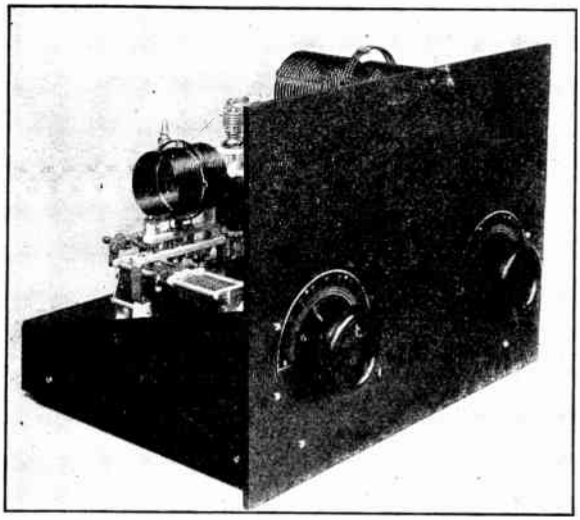


Fig. 1359 — The panel for the 1-kilowatt push-pull amplifier is 14 inches high and 19 inches wide. The chassis size is 13 × 17 inches.

Tuning — The only departure from ordinary procedure in tuning is that of adjusting the parasitic traps. The trap condensers, C_3 and C_4, should be set near maximum capacity, but not screwed up tight. After the amplifier has been neutralized, a bias voltage of about 22½ volts should be applied to the grid and the plate voltage applied through a 2500-ohm series resistance. With a pair of coils for any band plugged in, the plate current should not vary with any setting of the grid or plate condensers. If the plate current changes suddenly at any point, the trap condensers should be adjusted equally until the change disappears. The trap condensers should be set as near to maximum capacity as is possible consistent with parasitic suppression. If the r.f. wiring has been carefully duplicated, the initial adjustment of the parasitic traps as described above should be sufficient.

After the above adjustment is complete, excitation may be applied and the amplifier loaded. The high-capacity sections of the plate tank condensers are required only for the 3.5-Mc. band.

Grid current should run about 100 ma. on all bands, and the amplifier may be loaded until the plate current increases to 500 ma. The power output at 2000 volts on the plates should be approximately 750 watts.

coils are self-supporting and are fastened to the heat-radiating plate connectors.

The filament transformer is mounted underneath the chassis, and the filament by-pass condensers are wired in directly at the socket terminals. Millen safety terminals are provided for the positive high voltage and negative bias terminals. A male plug is set in the rear edge of the chassis for the 115-volt line connection to the filament transformer.

Power supply — A plate-supply unit suitable for this amplifier is shown in Fig. 1362. For bias, the unit shown in Fig. 1350 is suggested. The branch including the VR75-30 may be omitted and resistance values for R_2 and R_3 should be approximately 2000 and 2500 ohms, respectively. The transmitter shown in Fig. 1324 will furnish more than adequate excitation.

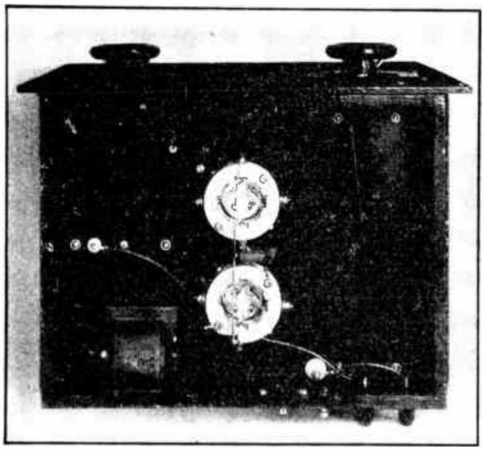


Fig. 1360 — The tube sockets in the 1-kilowatt amplifier are sub-mounted. The filament transformer is mounted close to the sockets.

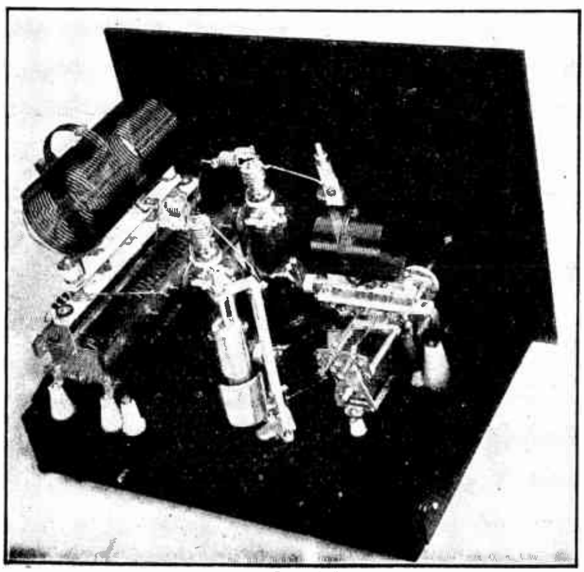


Fig. 1361 — Rear view of the 1-kw. amplifier, showing wiring and the placement of parts.

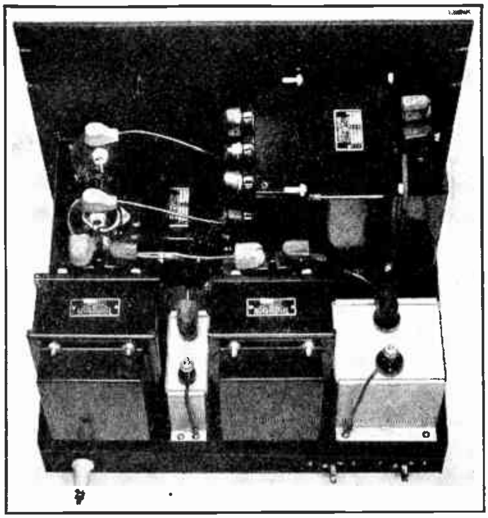


Fig. 1362 — This power supply unit delivers 2025 and 2480 volts at full-load current of 450 ma. with ripple of 0.5 per cent and regulation of 19 per cent. Voltages are selected by taps on the secondary. All exposed high-voltage terminals are covered with Sprague rubber safety caps and the tube plate terminals with moulded caps. The rectifier tubes are placed away from the plate transformer to avoid induction troubles. The panel is 14 × 19 inches and the chassis 13 × 17 × 2 inches. The exposed high-voltage terminal should be covered with a rubber-tubing sleeve. The circuit is the same as that shown in Fig. 1349, the components being as follows:

C_1 — 1-μfd. 2500-volt oil-filled (G.E. Pyranol).
C_2 — 4-μfd. 2500-volt oil-filled (G.E. Pyranol).
L_1 — Input choke, 5–20 henrys, 500 ma., 75 ohms (Thordarson T-19C38).
L_2 — Smoothing choke, 12 henrys, 500 ma., 75 ohms (Thordarson T-19C45).
R — 50,000 ohms, 200-watt.
Tr_1 — 3000–2450 volts r.m.s. each side of center, 500 ma. d.c. (Thordarson T-19P68).
Tr_2 — 2.5 volts, 10 amperes, 10,000-volt insulation (Thordarson T-64F33).

Note: The voltage regulation may be improved by the use of a lower valve of bleeder resistance, *R*, although at some sacrifice in maximum permissible load current.

Complete High-Power Transmitters

The 100-watt transmitter of Fig. 1324 may be used as a driver for either of the high-power amplifiers in Figs. 1354 and 1359. In addition to the power-supply units of Figs. 1350 and 1329 required for the exciter, a separate bias supply for the high-power amplifier will be necessary. A second bias-supply unit similar to that of Fig. 1350, minus the VR-tube branch, will be satisfactory. Plate voltage for either amplifier may be obtained from the large power-supply unit shown in Fig. 1362. The antenna tuner may be the one shown in Fig. 1352 with the substitution of a condenser of 0.1-inch plate spacing and coils of higher power rating.

For a combination using the amplifier of Fig. 1354, the combined heights of all units will be 66½ inches. If the push-pull amplifier of Fig. 1359 is used in the complete transmitter, the total height will be 68¼ inches.

A Variable-Frequency Exciter

The photographs of Figs. 1363, 1366 and 1367 illustrate the construction of a variable-frequency unit which is designed to take the place of the crystal as a frequency control in most of the common forms of crystal-oscillator circuits. The power output of the unit is approximately one and one-half watts, which is sufficient for this purpose, or for driving an 807. By means of plug-in coils, output at any frequency in the 1.75-, 3.5-, or 7-Mc. bands may be obtained.

Referring to the circuit diagram of Fig. 1364, a 6F6 is used in the e.c.o. circuit. Since the buffer stage provides adequate isolation, the use of a well-screened tube in the oscillator circuit is not a requirement. The cathode is connected to a feed-back winding, L_2, rather than to a direct tap on L_1, to make adjustment of feed-back less difficult. A high-*C* tank circuit is obtained by the fixed padders, C_1 and C_2, which are of the zero-drift type. Bandspread tuning is obtained by the split-stator condenser, C_3.

When coils 1 and 1A (see coil charts) are plugged in, the two sections of C_3 are connected in parallel and the output-frequency spread is 1750 to 2050 kc. to cover the 1.75-Mc. band or, through a doubler, the 3.5-Mc. band. Similarly, with coils 2 and 2A, the two sections of C_3 are in parallel and the output-frequency spread is 3500 to 4000 kc. to cover the 3.5-Mc. band.

When coils 1B and 1AB are plugged in, the sections of C_3 are in series and the output-frequency range is 1750 to 1825 kc. for obtaining, through doublers, the frequency ranges of 7000 to 7300 and 14,000 to 14,400 kc. Similarly, when coils 2B and 2AB are plugged in, the output-frequency range is 3500 to 3650 kc. for obtaining, through doublers, the same frequency ranges of 7000 to 7300 and 14,000

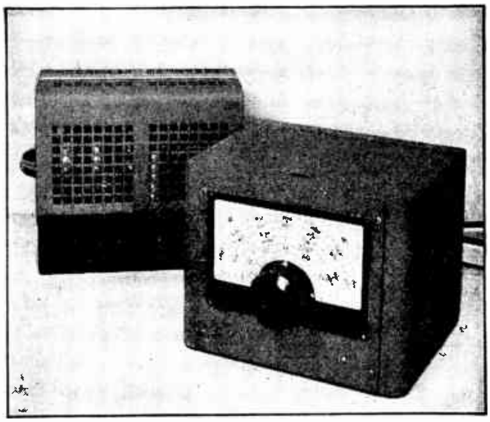


Fig. 1363 — The variable-frequency exciter is enclosed in an 8 × 8 × 10-inch Parmetal cabinet. The dial is the National type ACN, suitable for calibrating. The voltage-regulated power supply is mounted in an amplifier-foundation case with a 5 × 3 × 10-inch chassis.

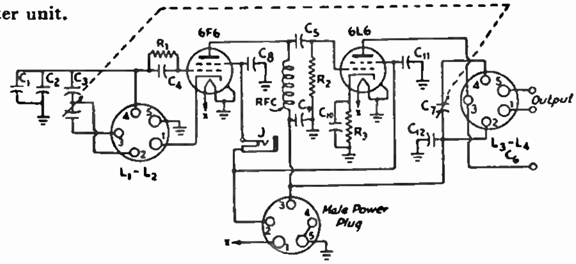


Fig. 1364 — Circuit diagram of the v.f.o. exciter unit.
C_1, C_2 — 300 μμfd. each, zero-drift type (Centralab 816Z).
C_3 — 140 μμfd. per section (Hammarlund MCD-140-S).
C_4 — 100-μμfd. mica.
C_5 — 250-μμfd. mica.
C_6 — 45–260-μμfd. mica trimmer (Hammarlund CTS-160).
C_7 — Approximately 65 μμfd. (Hammarlund MC-100-S with two stator and two rotor plates removed).
C_8, C_9, C_{10}, C_{11}, C_{12} — 0.01-μfd. paper.
J — Closed-circuit jack.
R_1 — 0.1 megohm, ½-watt.
R_2 — 0.1 megohm, ½-watt.
R_3 — 500 ohms, 1-watt.
L_1, L_2, L_3, L_4 — See Fig. 1365.

to 14,400 kc. The two sections of C_3 are also connected in series when coils 3 and 3A are plugged in, and the output-frequency range then becomes 7000 to 7300 kc. This is suitable for covering the 7-Mc. band and, through a doubler, the 14-Mc. band.

When coils 3B and 3AB are plugged in, only one section of C_3 is in use and the output-frequency range of 7000 to 7500 kc. is useful in obtaining, through doublers, the range of 28,000 to 30,000 kc.

Proper connections to C_3 are made automatically when each oscillator coil is plugged in, as shown in Fig. 1365.

Choke coupling is used between the oscillator and the 6L6 isolating stage. This stage is operated very close to Class-A conditions and is tuned to the second harmonic of the oscillator frequency. Thus, the oscillator operates at half the desired output frequency. The type 6L6 tube is used to take care of the unusually high dissipation resulting from this type of operation. The tuning of the output tank circuit is ganged with that of the oscillator. Tracking taps on the output coil, L_3, are required only for spreading the higher-frequency bands. Adjustable mica trimmers, C_6, are mounted in each coil form.

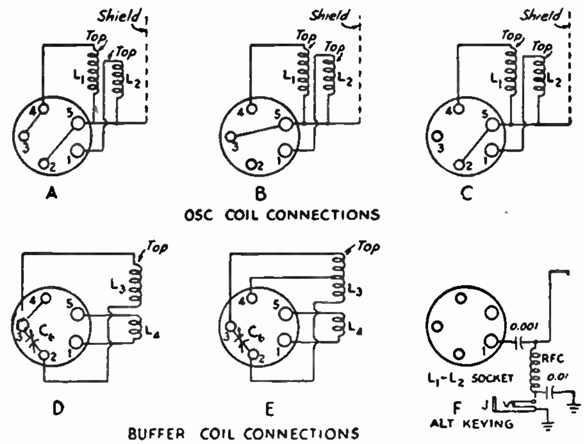


Fig. 1365 — Coil-form connections for the v.f.o. circuit of Fig. 1364. Connections shown at A are for coils 1 and 2. Those shown at B are for coils 3, 1B and 2B. Connections shown at C are for coil No. 3B. Buffer coils 1A and 2A should be connected as shown at D, while coils 3A, 1AB, 2AB and 3AB should be connected as shown at E. F shows the circuit for optional cathode keying instead of screen keying, as mentioned in text. *RFC* is an ordinary 2.5-mh. r.f. choke. Coil dimensions are as follows:

*Oscillator (L_1 and L_2)**

Coil No. 1 — (875 to 1025 kc.) — 47 turns No. 26 d.s.c., ⅞-inch long; 6 turns for L_2.
Coil No. 2 — (1750 to 2000 kc.) — 23 turns No. 20 d.s.c., 1¼ inches long; 2 turns for L_2.
Coil No. 3 — (3500 to 3650 kc.) — 14 turns No. 20 d.s.c., 1¼ inches long; 2 turns for L_2.
Coil No. 1B — (875 to 912.5 kc.) — 57 turns No. 26 d.s.c., 1⅛ inches long; 5 turns for L_2.
Coil No. 2B — (1750 to 1825 kc.) — 28 turns No. 20 d.s.c., 1 inch long; 2 turns for L_2.
Coil No. 3B — (3500 to 3750 kc.) — 13½ turns No. 20 d.s.c., 1-inch long; 2 turns for L_2.

* Wound on Millen 1-inch diameter forms, L_2 wound turn-for-turn over bottom end of L_1 in same direction.
** Wound on Hammarlund 1½-inch diameter forms, L_4 close-wound below L_3.

*Buffer Coils (L_3 and L_4)***

Coil No. 1A — (1750 to 2050 kc.) — 41 turns No. 24, 1¾ inches long; approximately 12 turns for L_4.
Coil No. 2A — (3500 to 4000 kc.) — 21 turns No. 18, 1½ inches long; approximately 6 turns for L_4.
Coil Co. 3A — (7000 to 7300 kc.) — 14 turns No. 18, 1½ inches long, tapped at 3 turns from bottom; approximately 4 turns for L_4.
Coil No. 1AB — (1750 to 1825 kc.) — 46 turns No. 24, 1¾ inches long, tapped at 19 turns from bottom; approximately 12 turns for L_4.
Coil No. 2AB — (3500 to 3650 kc.) — 24 turns No. 18, 1½ inches long, tapped at 9½ turns from bottom; approximately 6 turns for L_4.
Coil No. 3AB — (7000 to 7500 kc.) — 14 turns No. 18, 1½ inches long, tapped at 5 turns from bottom; approximately 4 turns for L_4.

To solve some of the difficulties often encountered in key-filtering an oscillator of this type, the oscillator stage is keyed in the screen circuit. This means that both sides of the key are at a potential of 150 volts above ground potential. It is, therefore, preferable to use a relay to isolate the key contacts from this voltage. Otherwise, due caution should be exercised. If preferred, cathode keying may be used as shown in Fig. 1365-F, but it is more difficult to obtain soft keying without introducing chirp with this system. With cathode keying, the screen connection will go directly to pin No. 2 on the power plug, eliminating the jack in the screen circuit.

A link winding, L_4, is provided for coupling the output of the exciter unit to the input of the amplifier stage which it is to drive.

Coils — Coil dimensions for several oscillator ranges are given in the coil table under Fig. 1365. Only those which suit the conditions under which the unit is to be operated need be constructed. This will depend upon the type of transmitter with which the unit is to be used. To begin with, only coils need be provided giving output in bands for which crystals, formerly used, are ground. For instance, if the oscillator stage to be driven is designed for 1.75-Mc. crystals only, coils need be wound for this band only. If the *transmitter* operates only in the 1.75-Mc. band, or, by doubling, in the 1.75- and 3.5-Mc. bands exclusively, only the 1.75-Mc. coils for the first bandspread range will be required. If, however, the transmitter is designed to cover the 7-Mc. band, as well as the lower-frequency bands, from a 1.75-Mc. crystal, coils for the second bandspread range also will be necessary to get full bandspread at 7 Mc. An examination of the coil-selection table will show what coils are required, depending upon the crystal frequency normally used to secure output in the desired band. If full bandspread at 7-Mc. and higher frequencies is not deemed necessary, the wide-bandspread coils for these frequencies need not be constructed.

The oscillator coils are wound on Millen one-inch diameter coil forms which are mounted in National PB-10 five-prong shielded plug-in bases. The feed-back coils, L_2, are wound over the bottom turns of L_1, and in the same direction. Connections to the base pins are given in Fig. 1365-A, B and C.

Fig. 1366 — Components for the v.f.o. exciter are assembled on a 7 × 7 × 2-inch chassis. The dual-section condenser is mounted by removing the shield between sections and fastening to the chassis with a single machine screw. The smaller condenser, C_7, is mounted on National polystyrene button insulators and metal spacers to insulate it from the chassis and bring its shaft in line with that of the dual condenser. It is reverse-mounted, with its tail shaft extension coupled to the tail shaft extension of the dual condenser to reduce the overall mounting space. The stop pin on the shaft must be removed. Leads from the tuning condensers to the sub-mounted coil sockets pass through the chassis via ½-inch holes lined with rubber grommets. The jack for the key, which must be insulated, and the male power connector mount in the side of the cabinet. The chassis is fastened firmly in place with long machine screws running through the chassis and the bottom of the cabinet. The terminals at the rear are for link-output connections, the binding post for capacity coupling.

The buffer coils are wound on Hammarlund 1½-inch diameter five-prong forms. The padding condensers, C_6, are mounted inside the coil forms, fastened in place with a 4-36 machine screw. Buffer coils for the higher-frequency ranges must be tapped as directed. One satisfactory way of making this tap is to drill a hole near the bottom of the form for a wire which may be brought outside from the pin to which the tap must be connected. The turn which is being tapped, as indicated in the table of coil dimensions, may be scraped and the tap wire soldered to this turn. Pin connections are shown in Fig. 1365-D and E.

COIL-SELECTION TABLE FOR VARIABLE-FREQUENCY UNIT *

Transmitter Output Freq.	*1.75 Mc.*	*3.5 Mc.*	*7 Mc.*	*14 Mc.*	*28 Mc.*
Crystal Freq. 1.75 Mc.	1 & 1A	1 & 1A	1B & 1AB	1B & 1AB	—
3.5 Mc.	—	2 & 2A	2B & 2AB	2B & 2AB	—
7 Mc.	—	—	3 & 3A	3 & 3A	3B & 3AB

* Numbers refer to coils in coil-dimension table.

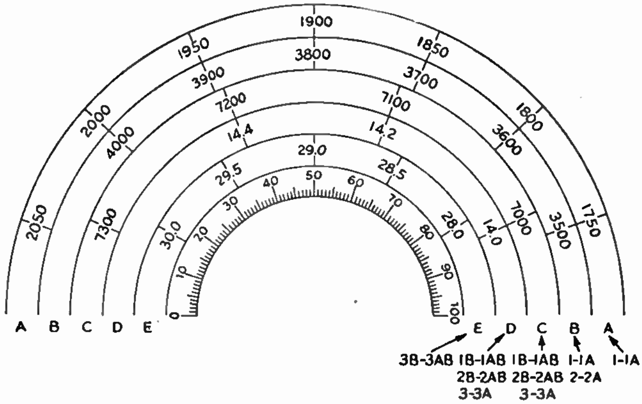


Fig. 1368 — Typical dial calibration for the v.f.o. unit. Notations at lower right indicate the calibrated ranges of the coil sets listed under Fig. 1365 and in the coil-selection table. Details of calibration are given in the text.

Tuning — Before an attempt is made to tune the circuits, the dropping resistor, R_2, in the power supply should be adjusted. This is done with any pair of coils plugged in and the key closed. Starting with maximum resistance, the slider should be adjusted, bit by bit, until the VR tubes ignite. As much resistance as possible should be left in the circuit consistent with the maintenance of reliable operation of the VR tubes. If the tubes ignite with maximum resistance in the circuit further adjustment will not be required, unless the output voltage of the pack used happens to be unusually high. If this is the case, the value of dropping resistance should first be increased until the VR tubes no longer ignite, and then brought back to the point where they just ignite.

The first step in adjusting the unit is to check the frequency range of the oscillator. It is probable that differences in wiring inductances and capacities will make it necessary to make slight alterations in the oscillator coil dimensions given in the table. Unless the construction differs widely from the original, however, no more than adjustment of the spacing of a few turns at the top of L_1 should be required.

Fig. 1367 — High-frequency connections underneath the chassis of the v.f. exciter unit are made with short, straight sections of heavy wire. The two zero-temperature padding condensers are soldered directly to the oscillator-coil socket. All components are mounted firmly with no opportunity to support mechanical vibration. Washers 1/16-inch thick are placed between the panel and the chassis to provide space for the lower lip of the cabinet opening.

If close calibration is desired, a 100-kc. frequency standard checked against WWV (see Chapter Twenty) or equivalent frequency-checking means should be provided. The approximate range of the oscillator coil under adjustment may be determined by listening to the exciter on a calibrated receiver. The 1.75-Mc. range of the receiver should be used for checking coil No. 1. The ranges of other coils may be checked with the receiver tuned to the 3.5-Mc. band, since the harmonics of 2000 to 2050 kc. are the only usable frequencies which fall outside this band.

If no signal is heard at any point in the band with any setting of the v.f.o. dial, run a wire from the receiver antenna post to a point near the oscillator coil. If it is still impossible to pick up the signal, it is possible that the oscillator may not be functioning. This can be verified by absence of rectified d.c. grid voltage between the 6F6 grid and ground. *One* turn should then be added to the feed-back winding. More than the single additional turn should not be required. If the winding is larger than is needed for reliable operation with the key closed, the 6F6 may continue to oscillate weakly even with the key open. This condition is to be avoided, of course, if break-in operation is contemplated.

When the oscillator is functioning satisfactorily, the spacing of the top turn or two of L_1 should be adjusted until the desired band is centered on the dial of the unit. This can be done by spreading a turn or two, as mentioned previously. The shield can should be replaced each time a check is made. When the adjustment is final, the turns should be cemented permanently in place. The v.f.o. unit should be warmed up thoroughly before making a permanent calibration.

The National ACN dial has imprinted scales for calibrating five ranges. Since the bandspread *ratio* is the same for the two lowest-frequency sets of coils, the oscillator coils for each of these ranges may be adjusted so that the 3.5-Mc. harmonics of the 1.75-Mc. range (1 and 1A) will coincide with the fundamental frequencies of the 3.5-Mc. range (2 and 2A) and one scale on the dial will serve for both calibrations. It is only necessary to adjust the oscillator coil of the 3.5-Mc. range so that the low-frequency end of the band falls at the same point as the second harmonic of 1750 of the 1.75-Mc. range falls when the 1.75-Mc. coils are plugged in. With similar adjustments, the 7-Mc. and 14-Mc. ranges of the coils 1B and 1AB, 2B and 2AB and 3 and 3A may be made to coincide. In the end there will be a single calibration on the dial for each band, and only five calibrations will be required for the complete set of coils listed in the coil table. A typical dial calibration is shown in Fig. 1368. Intermediate points may be marked in as desired. While the 14-Mc. band does not cover as much of the dial as do the other bands, nevertheless the bandspread is entirely adequate to enable accurate setting to zero-beat in this band.

With the oscillator ranges adjusted, the next step is to adjust the tracking of the buffer stage. A 6.3-volt (150-ma.) dial lamp with one or two turns of wire should be coupled to the output tank coil to act as an indicator. With the condenser gang set at minimum capacity, the padder, C_6, in the coil form should be adjusted for maximum brilliance of the lamp. The gang should now be turned to maximum capacity. If the lamp decreases in brilliance, readjust C_6, noting carefully whether an increase or decrease in capacity of C_6 is required to bring the lamp up to its original brilliance. (If the padders suggested in the parts table are used, and if they are mounted in the coil forms with their terminals downward, clockwise rotation of the adjusting screw will decrease capacity, while counter-clockwise rotation will increase capacity. If mounted with the terminals upward, the action will be reversed.) If an increase in the capacity of C_6 is required with coils having no bandspread tap, C_7 is not tuning fast enough and a turn should be added to L_3. If a decrease in the capacity of C_6 is required, a turn should be removed from L_3. On the tapped coils the tap should be moved upward a turn toward the top of L_3, if an increase in C_6 is required, or a turn downward toward the bottom of the coil, if C_6 is decreased.

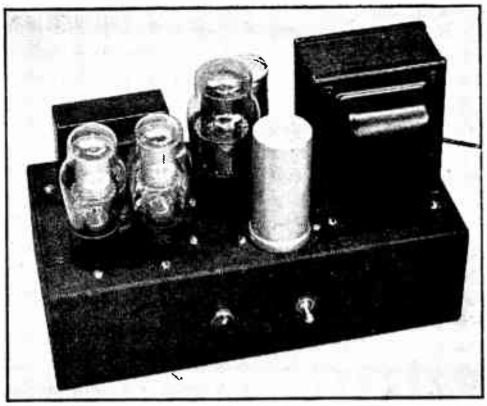


Fig. 1370 — Voltage-regulated power supply for the v.f. exciter unit. L_2 is mounted underneath the chassis.

After each adjustment of the coil, tracking should again be checked by adjusting C_6 for maximum brilliance with the condenser gang at minimum capacity and then checking at maximum capacity. These adjustments are simple and no trouble should be experienced in speedily arriving at the correct adjustments. When proper adjustments have been made, there should be no appreciable change in the brilliance of the lamp at any setting of the gang condenser.

If a check on plate currents is desired, meters may be inserted temporarily by opening up the wiring underneath the chassis. With correct adjustments of the tickler windings, L_2, the oscillator plate current should run between 12 and 15 ma. The buffer plate current should run at about 19 ma. with the key open and increase one milliampere or less with the key closed. Large changes in this plate current indicate that there are too many turns on L_2.

Power supply — The v.f.o. unit operates from the power supply shown in Fig. 1370 and whose circuit is shown in Fig. 1369. The two are connected with a length of five-conductor shielded battery cable fitted with a five-prong female connector at the unit and a similar male plug at the power-supply end. The shield is connected to pin No. 5 at each end. Almost any of the usual type of well-filtered receiver power supplies delivering 325 to 350 volts with a 50-ma. or better rating may be made to serve the purpose equally well, merely by the addition of the VR150-30 regulator tubes and the dropping resistor, R_2.

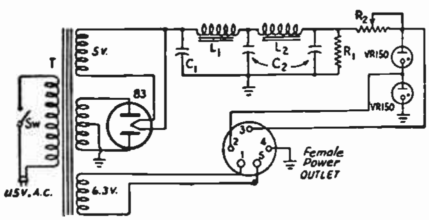


Fig. 1369 — Circuit diagram of the voltage-regulated power supply for the variable-frequency exciter unit.
C_1 — 8 μfd. 500-volt electrolytic (Mallory HD683).
C_2 — Dual-section 450-volt electrolytic, 40 μfd. per section, one section on each side of L_2 (Mallory FPD238).
L_1, L_2 — 15 henrys, 100 ma. (UTC R19).
R_1 — 25,000 ohms, 10-watt.
R_2 — 2500 ohms, 25-watt with slider.
T — Combination power transformer: 375 volts r.m.s. each side of center-tap, 100 ma.; 5 volts, 3 amperes; 6.3 volts, 6 amperes (UTC R12).
Sw — S.p.s.t. toggle switch.

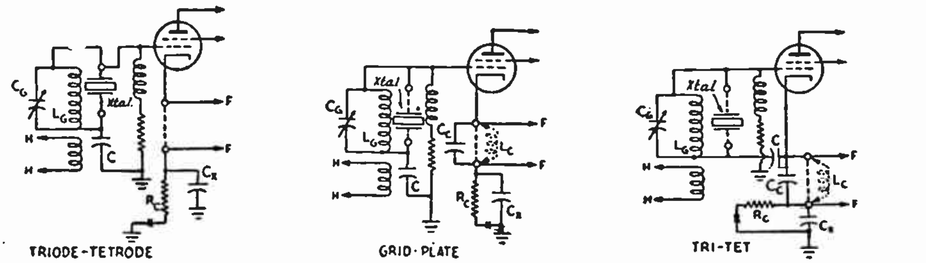


Fig. 1371 — Methods of coupling the output of the v.f.o. to crystal-oscillator stages of various types. See text for details. C is a mica condenser of 0.001 μfd. to prevent short-circuit of the grid leak. R_c and C_x are the usual oscillator cathode resistor and by-pass. C_c and L_c are the usual cathode-circuit tanks in the grid-plate and Tri-tet circuits. The v.f.o. link output is connected at H-H for harmonic operation and to F-F for fundamental operation. C_g is 100 $\mu\mu$fd. for the 1.75-Mc. band and 50 $\mu\mu$fd. for the 3.5- and 7-Mc. bands. Dimensions for L_g are as follows:

1.75-Mc. input — 64 turns No. 24 d.s.c., close-wound, 1½-inch diameter.

3.5-Mc. input — 40 turns No. 24, 1½-inch diameter, 1½-inches long.

7-Mc. input — 20 turns No. 18, 1½-inch diameter, 1½ inches long.

Link windings consist of 8, 6 and 5 turns respectively for the 1.75-, 3.5- and 7-Mc. bands, close-wound below L_g.

ℂ Feeding Crystal-Oscillator Stages

The output of the v.f.o. unit is sufficient to drive an 807 or similar type of tube. Such a stage may be link coupled to the exciter unit by means of L_4 or capacity coupled by connecting a small coupling condenser to the plate terminal of the 6L6. In the latter case, some readjustment of C_6 will be required to restore resonance, but retracking of the stage should not be necessary.

However, it is expected that the unit will be used more frequently to drive the crystal-oscillator stage of a crystal-controlled transmitter already in operation. While other methods of coupling between the crystal-oscillator stage and the v.f.o. unit may be devised, one satisfactory system which reduces the possibility of instability of the crystal-oscillator tube when coupled to the v.f.o. unit will be described in detail. Most crystal-oscillator stages are not sufficiently well-screened to permit operating the stage as a conventional straight amplifier with input and output circuits tuned to the same frequency. While the substitution for the crystal of a tuned circuit link-coupled to the output of the v.f.o. unit is the recommended method of coupling when the crystal stage is to be used as a frequency doubler, the stage will invariably break into oscillation if the same system is used for fundamental operation. One satisfactory method of preventing this is to switch the link line to the cathode circuit for fundamental operation. The practical application of this system is shown applied to several typical varieties of crystal-oscillator circuits in Fig. 1371.

In each case, a tank circuit, C_gL_g, tuned to the frequency of the crystal which it supplants, replaces the crystal when the stage is to be operated as a frequency doubler. The insertion of the condenser C is required to prevent short-circuit of the grid leak. The tank circuit is coupled to the output of the v.f.o. through a link line connecting at the points marked H-H. The openings indicated in the cathode circuits may be closed by a shorting bar. It is important to keep the shorting-bar leads as short as possible, otherwise there is danger of self oscillation even though the tuning of the grid and plate tanks may differ widely. In Tri-tet and grid-plate circuits, the cathode tanks must be shorted as indicated.

When the crystal stage is to be operated as a straight amplifier, the grid tank is removed, leaving the crystal position open. The link line from the v.f.o. is shifted to the points marked F-F and the cathode shorts indicated by the dotted lines removed. In Tri-tet or grid-plate

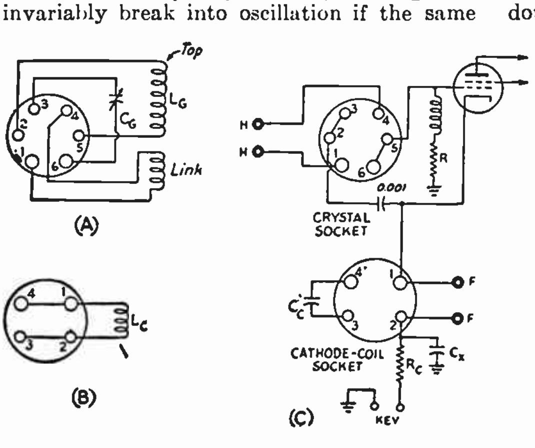


Fig. 1372 — Circuit arrangements for a plug-in coil system planned for most conveniently making connections in a Tri-tet oscillator circuit for optional crystal or v.f.o. operation. The grid tank for doubler operation is plugged into the same six-prong tube socket used by the crystal. The circuit at A shows the connections of the plug-in grid tank for frequency-doubler operation of the crystal stage with v.f.o. input. Values for L_g, C_g, and the associated link coils are given under Fig. 1371. B shows connections for the plug-in cathode coil, L_c, which is the usual Tri-tet cathode winding. C shows the adapter circuit complete with all socket connections. C_c is the usual Tri-tet cathode-tank condenser and R_c and C_x are the usual cathode resistor and by-pass condenser.

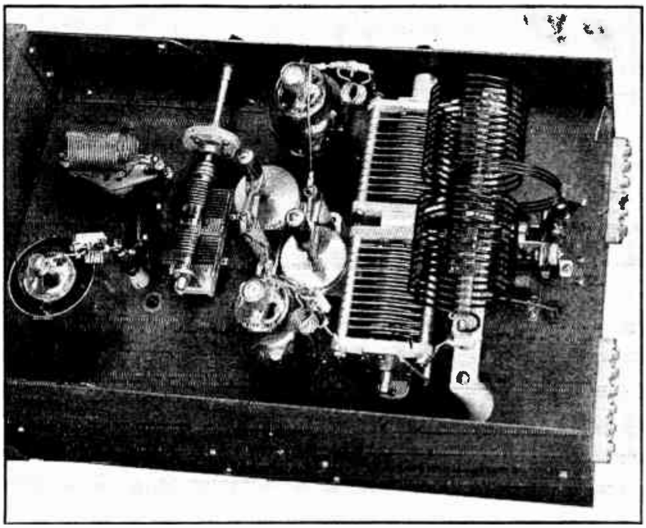


Fig. 1373 — Top view of the gang-tuned driver and push-pull amplifier designed to work with the v.f.o. unit of Fig. 1363. The chassis is elevated by 17 × 8-inch panels on each side. The 807 socket, which is mounted an inch below the chassis top on spacers, and the socket for the coupling transformer, L_1L_2, at the left-hand end of the chassis, are on either side of the bandspread condenser, C_2, underneath. The 807 padding condenser, C_1, is next to the right with an insulating coupling on its shaft which is 5½ inches from the left-hand end of the chassis. The shaft of the final-amplifier padding condenser, 5⅛ inches from the right-hand end of the chassis, is also fitted with an insulating coupling. The condenser is mounted on National polystyrene button insulators to bring its shaft level with that of C_1. The sockets for the 812s are at either end of C_5, with the neutralizing condensers between to make neutralizing leads short. The jack bar for the tank coil, L_3, is mounted on 2-inch cone insulators.

oscillators, the cathode inductances and preferably the cathode tuning condensers also must be removed. If a cathode resistor is used, the excitation should be introduced between the cathode and the junction of the cathode resistor and its by-pass condenser.

If the v.f.o. is to be keyed, the key terminals of the crystal stage must be shorted. A small amount of fixed bias may have to be connected between grid leak and ground to prevent excessive plate current when the key in the v.f.o. circuit is open. If break-in keying is not desired, the v.f.o. may be operated continuously and the crystal stage keyed in the usual manner.

Values for the substitute grid tank coil are given in Fig. 1371. A fairly-high L/C ratio has been chosen and, in most cases, any one band may be covered without retuning of the grid tank, if it is set to resonance in the middle of the band. The remainder of the transmitter will be tuned as usual.

The details of a convenient plug-in system which takes care of all connections in shifting from Tri-tet crystal operation (used in most of the transmitters described in this chapter) to either fundamental or doubler operation when using the v.f.o. unit are shown in Fig. 1372. The grid tank for doubler operation is plugged into the same six-prong tube socket used for the crystal. Link connections to the v.f.o. are made through pin jacks H-H. A short-circuiting wire connects pin jacks F-F into the cathode circuit. The leads from the cathode-coil socket to these jacks and the shorting wire should be kept as short as possible. The cathode coil should be removed from its socket.

For fundamental operation with the v.f.o. unit, the tank is removed from the grid-circuit socket and the shorting wire removed from F-F, to which the link line from the v.f.o. is then shifted.

For crystal operation, the crystal is plugged into the grid circuit between prongs 6 and 3, or between 5 and 2, and the cathode coil is plugged into its socket, automatically connecting in the cathode condenser, C_c. The v.f.o. link line must be disconnected. Similar combinations may be worked out for other oscillator circuits.

A Gang-Tuned 450-Watt Push-Pull Amplifier and Driver

Figs. 1373, 1374 and 1376 show a gang-tuned unit which may be added to the v.f.o. unit of Fig. 1363. As shown in Fig. 1375, it consists of a push-pull amplifier and a driver stage, the tuning controls of which are coupled to the tuning shaft of the v.f.o. unit. Once adjusted for any given band, the entire transmitter can be tuned with the single dial of the v.f.o. unit.

The two stages are coupled inductively with the tuning condensers connected across the grid winding. The use of inductive coupling solves the problem of balanced excitation to the amplifier without the dual tuning controls required with link coupling. C_1 and C_3 are the tank condensers, used for setting the circuits to the desired band. C_2 and C_4 are the band-tuning condensers. The two stages are adjusted for tracking by varying the portion of the coils across which C_2 and C_4 are connected.

The trap circuits, L_4C_5, L_5C_6 and L_6C_7, are for the suppression of v.h.f. parasitics.

The milliammeter may be switched to read 807 cathode or screen current, amplifier grid current, or amplifier cathode current.

Coils — While homemade coils of equivalent dimensions may be substituted, it may be found more convenient to alter manufactured coils. The National coils suggested for L_1 should be obtained minus the links and mountings. Stripped, it will be found that these coils fit snugly inside the B & W coils used for L_2, and that the plastic strips on each coil hold

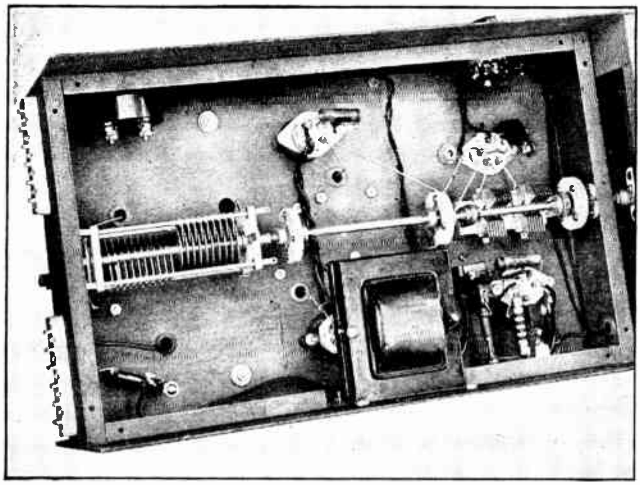


Fig. 1374 — Bottom view of the gang-tuned unit. The final amplifier bandspread condenser, C_4, is mounted as far to the left as possible, on National polystyrene button insulators stacked to bring the shaft level with that of the driver handspread condenser, C_2, to the right. The shafts of the two condensers are connected with flexible ceramic insulating couplings and also to the tail shaft of C_7 in the v.f.o. unit through a hole cut in the rear of the v.f.o. cabinet. C_2 is turned around so that its tail shaft couples to the shaft of the v.f.o. unit. The mounting hole of the condenser should come 2½ inches from the left-hand edge of the chassis. The shaft stop pin should be removed. The remaining below-chassis wiring is simple and direct. Heavy tinned wire is used for all r.f. leads. The filament transformer is mounted below the chassis at the center rear. Insulated or protected terminals are used for all external power supply connections.

them central to prevent short circuits between L_1 and L_2. The link winding should be removed from L_2. The free base-pins thus provided will serve for the connections to C_2. The tubular rivets at each end of the bottom spacing strip of the coil should be drilled or filed out, and ¾-inch 6–32 machine screws substituted. A Johnson banana plug is fastened at each end and the ends of L_1 are connected to these plugs.

In the chassis, on either side of the coil socket and directly below the banana plugs, a hole should be drilled. The one on the right-hand side should be ¼-inch in diameter, while the one on the left-hand side should be ½-inch in diameter. A jack to fit the banana plug should be placed in a National polystyrene button-type insulator with the shoulder filed off and the hole drilled out to fit the jack. This jack, mounted in the ¼-inch hole with the insulator as a spacer, then serves to make the ground connection for L_1. The ½-inch hole is for a second jack insulated from the chassis by a pair of button insulators which serves as the connection for the other end of L_1.

The B & W type TVH coils are selected not only because they are of the proper size for the power involved, but also because they are supplied with extra plugs which may be used for the ganging taps for C_4.

Both L_2 and L_3 require no bandspread taps for the 1.75-Mc. band; the plugs for the taps and those for the ends of the coils are simply tied together, connecting the bandspread and padding condensers in parallel for this band.

COIL-SELECTION TABLE FOR GANGED UNIT				
Band	*Osc.*	*Buffer*	*Driver*	*Final*
1.75 Mc.	No. 1	No. 1A	1.75 Mc.	1.75 Mc.
3.5 Mc.	No. 2	No. 2A	3.5 Mc.	3.5 Mc.
7 Mc.	No. 3	No. 3A	7 Mc.	7 Mc.
14 Mc.	No. 3	No. 3A	14 Mc.	14 Mc.

Combining Units

Fig. 1376 shows how the two units are joined together. The output of the v.f.o. and the input of the 807 driver stage are coupled capacitively, a short wire connecting the binding post in the v.f.o. unit with the coupling condenser, C_{10}, in the ganged unit. Large holes are made in the rear of the v.f.o. cabinet and the end of the chassis to clear a small National rigid shaft coupling. The height of the chassis should be adjusted so that the shafts of the two units line up perfectly. If the condenser gangs in each unit have been mounted as described, the shafts will be lined up when the bottom edge of the chassis is 2¼ inches above the bottom edges of the supporting panels.

The two units are fastened together with 7-inch triangular brackets, the tops of which have been cut off to fit, on each side of the chassis. The excitation lead to the grid of the 807 passes through a grommet-lined hole in the back of the v.f.o. cabinet.

Power-supply requirements are covered in the section of the complete gang-tuned transmitter which follows.

Tuning — If coil dimensions have been followed carefully, there should be little difficulty in lining up the various stages. The shaft couplings must be adjusted so that all condensers of the gang arrive at maximum or minimum capacity simultaneously. Coils should be plugged in the various stages for the desired band, using the coil-selection table as a guide.

With the tuning control set for the high-frequency edge of the band, the voltage-regulated supply and the bias supply should be turned on simultaneously. This will apply plate voltage to the v.f.o. unit and screen voltage to the 807. Using the 807 screen current as an indicator, the trimmer of the buffer stage in the v.f.o. unit should be lined up. Maximum screen current indicates resonance. The key should not be held closed for excessively long periods, to limit screen heating. Tuning to the low-frequency end of the band should show negligible change in screen current. Should there be evidence of poor tracking, the buffer stage can be brought into line again as discussed in the section describing the tuning of the v.f.o. unit.

Plate voltage may now be applied to the 807 and the stage tuned to resonance with C_1. A check should be made for parasitic oscillation, with a lamp of sufficient size to reduce the plate voltage to about half in series with the primary of the 750-volt transformer. At several settings of the v.f.o. unit C_1 should be varied throughout its range, carefully noting any change in cathode current which would indicate oscillation. An additional check may be made by touching a neon bulb to the plate of the 807. Should oscillation occur, C_5 should be adjusted until the oscillation is suppressed.

Turning now to the tracking of the driver stage, tuning C_1 to resonance should result in a showing of amplifier grid current. Again starting at the high-frequency end of the band, C_1 should be adjusted for maximum grid current. If there is a serious falling off of grid current as the unit is tuned to the low-frequency end of the band, a check should be made to determine if readjusting C_1 will bring the grid current back up. If it does not, the size of L_1 must be increased by one or two turns. If, however, retuning of C_1 shows the tuning to be off resonance at the low-frequency end of the band, it should be carefully noted whether an increase or a decrease in the capacity of C_1 is necessary to restore resonance. If an increase in C_1 is required, the taps of C_2 should be spread slightly farther apart; if a decrease is required, they should be brought closer together. After each check the tuning of the unit should be returned to the high-frequency end and realigned, before again checking the low-frequency end.

However, should the first check at the low-frequency end of the band show an increase in grid current over that obtained at the high-frequency end a turn or two should be removed from L_1, after which the tracking should again be checked as previous described.

With substantially constant grid current over the band, the amplifier may be neutralized in the usual manner. With the amplifier oper-

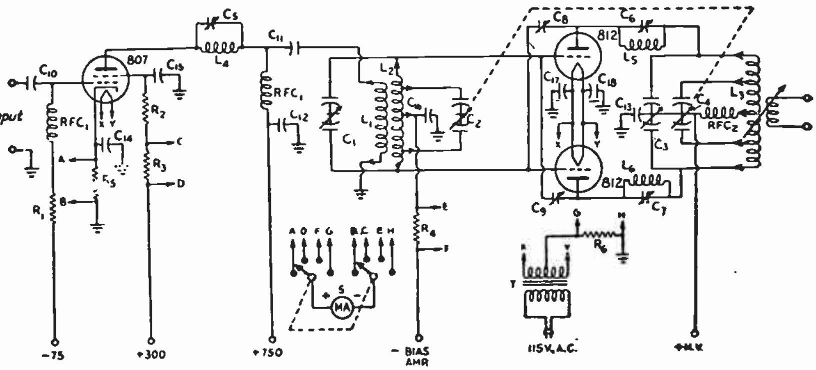


Fig. 1375 — Circuit diagram of the 450-watt gang-tuned driver and push-pull amplifier unit.

C_1 — 140 $\mu\mu$fd. per section (Hammarlund MCD-140-S).
C_2 — 100 $\mu\mu$fd. per section (Hammarlund MCD-100-S).
C_3 — 150 $\mu\mu$fd. per section, 0.07-inch spacing (Johnson 150ED30).
C_4 — 65 $\mu\mu$fd. per section, 0.07-inch spacing (Hammarlund HFBD-65-E).
C_5, C_6, C_7 — 3–30-$\mu\mu$fd. mica trimmer (National M-30).
C_8, C_9 — Neutralizing condensers (National NC-800).
C_{10} — 100-$\mu\mu$fd. mica.
C_{11}, C_{12} — 0.001-μfd. mica, 1000-volt rating.
C_{13} — 0.001-μfd. mica, 7500-volt (Aerovox 1623).
C_{14}, C_{15}, C_{16}, C_{17}, C_{18} — 0.01-μfd. mica.
MA — Milliammeter, 100-ma. scale.
R_1 — 25,000 ohms, 1-watt.
R_2 — 20,000 ohms, 10-watt variable.
R_3, R_4 — 25 ohms, 1-watt.
R_5 — Meter multiplier resistance, 2 times, wound with No. 26 wire.
R_6 — Meter multiplier resistance, 5 times, wound with No. 24 wire.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 500-ma. r.f. choke (Hammarlund CH500).
S — Two circuit, 4-contact switch (Mallory 3234J).
T — 6.3 volts, 10 amperes (Thordarson T-19F99).
L_1 — Mounted inside L_2:
1.75 Mc. — 45 turns No. 24, 1¼-inches diameter, 1⅜ inches long (National AR80, unmounted, 11 turns removed).
3.5 Mc. — 22 turns No. 22, 1¼-inch diameter, 1¼ inches long (National AR40, unmounted, 6 turns removed).
7 Mc. — 14 turns No. 20, 1¼-inch diameter, 1¼ inches long (National AR20, unmounted).
L_2 — 1.75 Mc. — 58 turns No. 24, 1⅝-inch diameter, 2¼ inches long, taps at ends of coil (B & W JCL-160, no link).
3.5 Mc. — 28 turns No. 22, 1⅝-inch diameter, 1½ inches long, taps at 3 turns from each end (B & W JCL-80, no link, 8 turns removed from each end).
7 Mc. — 18 turns No. 16, 1⅝-inch diameter, 1½ inches long, taps at 6 turns from each end (B & W JCL-40, no link, 5 turns removed from each end).
L_3 — 1.75 Mc. — 60 turns No. 16, 5½ inches long, 2½-inch diameter, ⅝-inch space at center for link, taps at ends of coil (B & W TVH-160).
3.5 Mc. — 38 turns No. 14, 5¼ inches long, 2½-inch diameter, ¾-inch space at center for link, taps at 3¾ turns from each end (B & W TVH-80).
7 Mc. — 24 turns No. 12, 5¼ inches long, 2½-inch diameter, ¾-inch space at center for link, taps at 7¾ turns from each end (B & W TVH-40).
L_4 — 5 turns No. 14, ⅜-inch diameter, 1 inch long.
L_5, L_6 — 4 turns No. 14, ⅝-inch diameter.

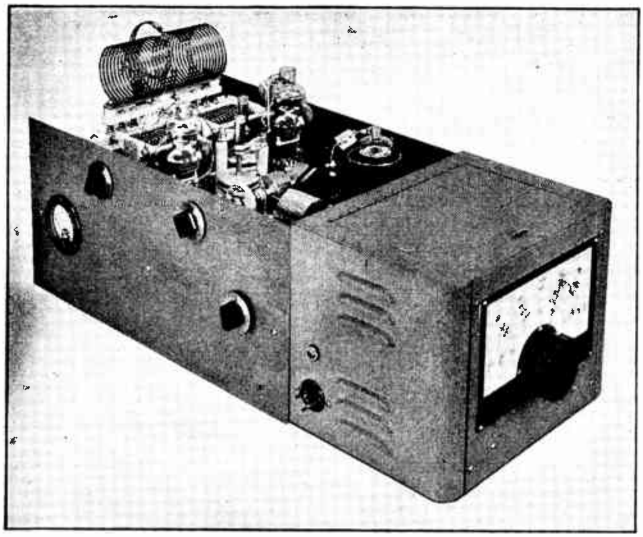


Fig. 1376—The v.f.o. unit of Fig. 1363 combined with the gang-tuned driver and push-pull final amplifier. The two units are fastened together with 7-inch triangular brackets, the tops of which have been cut off to fit, on each side of the chassis. The excitation lead to the grid of the 807 passes through a grommet-lined hole in the back of the v.f.o. cabinet. The milliammeter and the meter switch are placed on the panel to balance each other at opposite ends of the chassis. Holes for these components must be cut in the chassis edge. The control at the left is for setting the final-amplifier padder or band-setting condenser, C_3, while the control to the right is for the driver padder, C_1. The 10 × 17 × 3-inch chassis is elevated approximately 2¼ inches by supporting it on panels 8 inches high running the length of the chassis. The clearance and assembly holes through the panel and chassis should be made slightly oversize to permit accurate adjustment of the chassis height for lining up the tuning-condenser shafts.

ating at reduced plate voltage, a check similar to that described for the 807 stage should be made to eliminate any tendency toward parasitic oscillation. For several settings of the ganged control, C_3 should be varied throughout its range. If oscillation occurs, C_6 and C_7 should be adjusted in equal steps until it ceases.

Still operating at reduced plate voltage, the amplifier should be loaded with a lamp bulb of 150 to 200 watts connected to the output link. C_3 should be adjusted for resonance at the high-frequency end of the band. Tuning across the band should now show no appreciable change in power input or output. If a check, by retuning C_3 at the low-frequency end of the band, shows the stage to be off resonance, a note should be made as to whether an increase in the capacity of C_3 or a decrease is necessary to restore resonance. If an increase is required, the taps of C_4 should be spread slightly, while a decrease in C_3 indicates that the taps of C_4 should be brought slightly closer together. Again, each adjustment of tracking should be followed by realigning at the high-frequency end of the band before making a check on the new adjustment at the low-frequency end.

If coil dimensions have been followed carefully these tracking adjustments should not be required. They are described to take care of cases in which the constructor may have gone astray at some point, or in which the design has been changed to suit other requirements. Naturally, the adjustments for the higher-frequency bands must be made in smaller steps than those required for the lower-frequency bands.

At the plate voltages recommended, the screen current, when lining up the v.f.o. output stage, should run between 5 and 10 ma. Cathode current to the driver stage when tuned and loaded should be between 70 and 100 ma., while grid current to the final amplifier should exceed 50 ma. with the amplifier loaded to the rated plate current of 300 ma. at 1500 volts. Under operating conditions the driver screen voltage should run close to 250 volts. When correctly adjusted, the power output across any of the three bands should remain constant at 300 watts.

For 'phone operation with plate modulation, the input to the final amplifier should be reduced to 250 ma. at 1250 volts.

The tube tables in the Appendix should be consulted for the operating conditions of other tubes should they be used in the final amplifier.

Complete Variable-Frequency Gang-Tuned Transmitter

Fig. 1376 shows the two units of Figs. 1363 and 1373 combined for gang tuning. The voltage-regulated supply of Fig. 1370 may be used to furnish screen voltage for the 807 by bringing out a tap from the junction of resistors R_1 and R_2. The unit of Fig. 1350 will furnish biasing voltages for both 807 and final amplifier. The voltage-divider resistance of the bias unit should be adjusted with 4000 ohms in the R_2 portion and 4000 ohms in the R_3 portion. Plate voltage for the 807 may be obtained from the unit of Fig. 1318, while the unit of Fig. 1348 will furnish plate voltage for the amplifier. A suitable antenna tuner is shown in Fig. 1352.

To facilitate rapid setting of the band-set condensers, their dials may be provided with scales upon which the correct setting for each band is marked.

Similarly, to simplify antenna tuning and make it possible to adjust the antenna without putting a signal on the air, the antenna-tuner dial may be provided with a scale which may be calibrated in terms of receiver- or v.f.o.-dial settings. Since antenna tuning should not be critical, the dial need be calibrated for only several scattered points throughout each band.

A Practical Vacuum-Tube Keyer

Fig. 1378 shows a practical vacuum-tube keyer unit. The circuit diagram is shown in Fig. 1377. T_1, the rectifier, with C_1 and R_1 form the power-supply section for producing the blocking voltage necessary for cutting off the keyer tubes. With only R_2 in the circuit and Sw_2 in the open position, there will be no lag. As Sw_2 is turned to introduce more capacity in the circuit, the keying characteristic is "softened" at both make and break. Adding resistance by turning Sw_1 to the right affects the "break" only. The use of high resistances and small capacities results in small demand on the power supply and makes the key safe to handle.

As many 45s may be added in parallel as desired. The voltage drop through a single tube varies from 90 volts at 50 ma. to 52 volts at 20 ma. Tubes in parallel will reduce the drop in proportion to the number of tubes. If rated voltage is important in the operation of the keyed circuit, the drop through the keyer tubes must be taken into account and the transmitter voltage boosted to compensate for the drop.

If desired, a greater angle of lag can be obtained by using a rotary switch with more points and additional resistors and condensers. Suggested values of capacity, in addition to C_2 and C_3, are 0.001 and 0.002 μfd. From R_2, resistors of 2, 3 and 5 megohms may be added.

When connecting the output terminals of the keyer to the circuit to be keyed, care must be used to connect the grounded output terminal to the negative side of the keyed circuit.

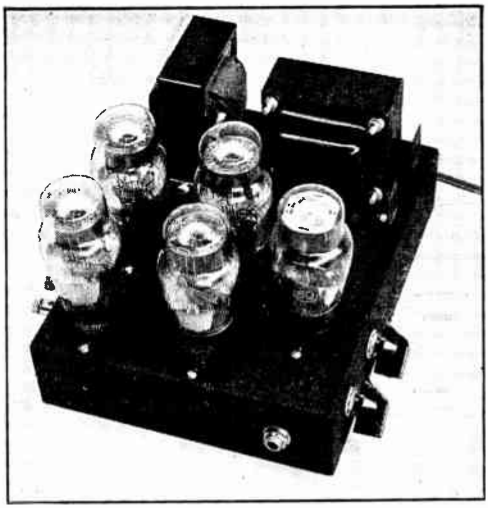


Fig. 1378 — A vacuum-tube keyer, built up on a 7 × 9 × 2-inch chassis with space for four or more keyer tubes and the power-supply rectifier. The resistors and condensers which produce the lag are mounted underneath, controlled by the knobs at the right. The jack is for the key, while terminals at left are for the keyed circuit.

Rack Construction

Most of the units described in the constructional chapters of this *Handbook* are designed for standard rack mounting. The assembly of a selected group of units to form a complete transmitter is, therefore, a relatively simple matter. While standard metal racks are available on the market, many amateurs prefer to build their own less expensively from wood. With care, an excellent substitute can be made.

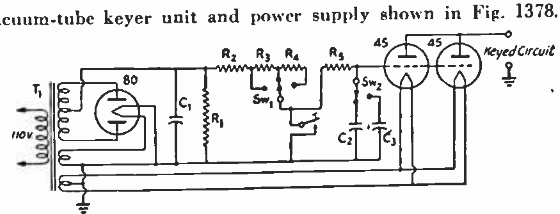
The plan of a rack of standard dimensions is shown in Fig. 1379. The rack is constructed entirely of 1 × 2-inch stock of smooth pine, spruce or redwood, with the exception of the trimming strips, *M*, *N*, *O* and *P*. Since the actual size of standard 1 × 2-inch stock runs appreciably below these dimensions, a much sturdier job will result if pieces are obtained cut to the full dimensions.

The main vertical supporting members of the wooden rack each is comprised of two pieces (*A* and *B*, and *I* and *J*) fastened together at right angles. Each pair of these members is fastened together by No. 8 flathead screws, with heads countersunk.

Before fastening these pairs together, pieces *A* and *J* should be made exactly the same length and drilled in the proper places for the mounting screws, using a No. 30 drill. The length of pieces *A*, *J*, *B* and *I* should equal the total height of all panels required for the transmitter plus *twice* the sum of the thickness and width of the material used. If the dimensions of the stock are exactly 1 × 2 inches, then 6 inches must be added to the sum of the panel heights. An inspection of the top and bottom of the rack in the drawing will reveal the reason for this. The first mounting hole should come at a distance of ¼ inch plus the sum of the thickness and width of the material from either end of pieces *A* and *J*. This distance will be 3¼ inches for stock exactly 1 × 2 inches. The second hole will come 1¼ inches from the first, the third ½ inch from the second, the fourth 1¼ inches from the third and so on, alternating 1¼ inches from the third and so on, alternating spacings between ½ inch and 1¼ inch (see detail drawing D, Fig. 1379). All holes should

Fig. 1377 — Wiring diagram of the practical vacuum-tube keyer unit and power supply shown in Fig. 1378.

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.003-μfd. mica.
C_3 — 0.005-μfd. mica.
R_1 — 0.25 megohm, 1-watt.
R_2 — 50,000 ohms, 10-watt.
R_3, R_4 — 5 megohms, 1-watt.
R_5 — 0.5 megohm, 1-watt.
Sw_1, Sw_2 — 3-position 1-circuit rotary switch.
T_1 — 325-0-325 volts, 5 volts and 2.5-volts (Thordarson T-13R01).



be placed ⅜ inch from the inside edges of the vertical members.

The two vertical members are fastened together by cross-member *K* at the top and *L* at the bottom. These should be of such a length that the inside edges of *A* and *J* are exactly 17½ inches apart at all points. This will bring the lines of mounting holes 18¼ inches center to center. Extending back from the bottoms of the vertical members are pieces *G* and *D* connected together by cross-members *L*, *Q* and *E*, forming the base. The length of the pieces *D* and *G* will depend upon space requirements of the largest power supply unit which will rest upon it. The vertical members are braced against the base by diagonal members *C* and *H*. Rear support for heavy units placed above the base may be provided by mounting angles on *C* and *H* or by connecting these members with cross-braces as shown at *F*.

To finish off the front of the rack pieces of ¼-inch oak strip (*M*, *N*, *O*, *P*) are fastened around the edges with small-head finishing nails. The heads are set below the surface and the holes plugged with putty or plastic wood.

The top and bottom edges of *M* and *O* should be ¼ inch from the first mounting holes, and the distance between the inside edges of the vertical strips, *N* and *P*, 19⅟₁₆ inches.

To prevent the screw holes from wearing out when panels are changed frequently, ½ × ⅟₁₆, or ⅟₃₂ inch iron or brass strip may be used to back up the vertical members of the frame.

The outside surfaces should be sandpapered thoroughly and given one or two coats of flat black, sandpapering between coats. A finishing surface of two coats of glossy black "Duco" is then applied, again sandpapering between coats. It is very important to allow each coat to dry thoroughly before applying the next, or sandpapering.

Since the combined weights of power supplies, modulator equipment, etc., may total to a surprising figure, the rack should be provided with rollers or wheels so that it may be moved about when necessary after the transmitter has been assembled. Ball bearing roller-skate wheels are suitable for the purpose.

Standard metal chassis are 17 inches wide. Standard panels are 19 inches wide and multiples of 1¾ inches high. Panel mounting holes start with the first one ¼ inch from the edge of the panel, the second 1¼ inches from the first, the third ½ inch from the second, the fourth 1¼ inches from the third, and the distances between holes from there on alternated between ½ inch and 1¼ inches. (See detail D, Fig. 1379.) In a panel higher than two or three rack units (1¾ inch per unit), it is common practice to drill only sufficient holes to provide a secure mounting. All panel holes should be drilled ⅜ inch in from the edge.

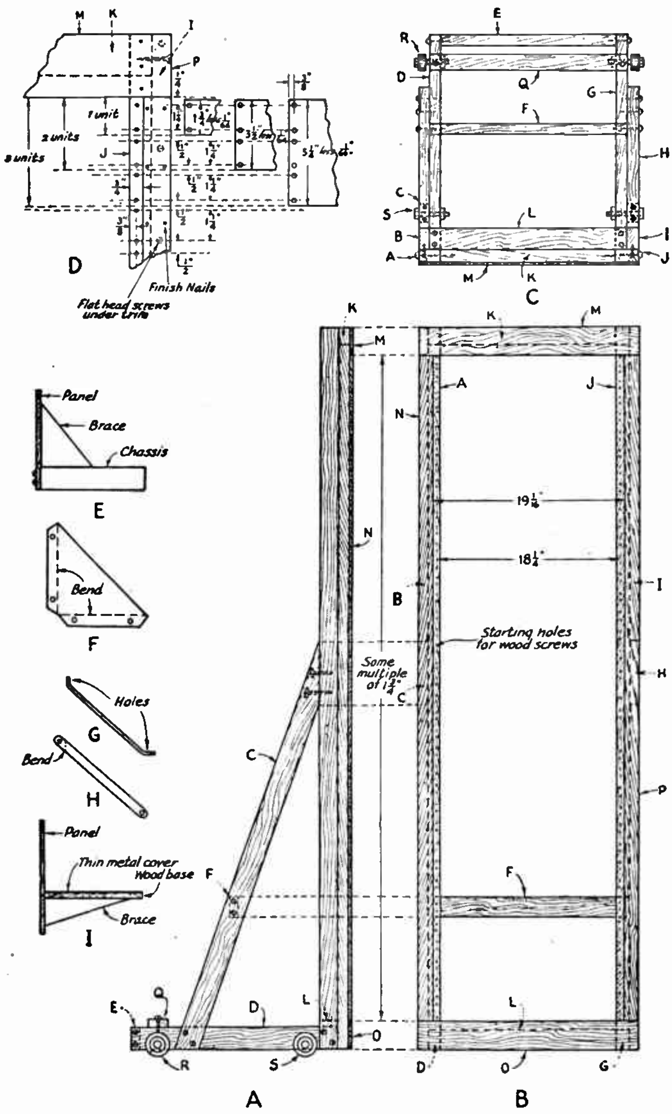


Fig. 1379 — The standard rack. A — Side view. B — Front view. C — Top view. D — Upper right hand corner detail. E — Panel and chassis assembly. F, G, H — Various types of panel brackets. I — Substitute for metal chassis.

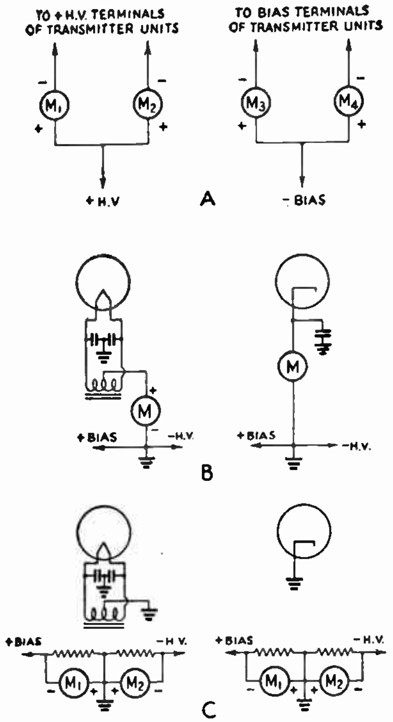


Fig. 1380 — Various methods of connecting milliammeters in grid and plate currents. A — High-voltage metering. B — Cathode metering. C — Shunt metering.

Metering

Various methods of metering are shown in Fig. 1380. A shows the meters placed in the high-voltage plate and bias circuits. M_1 and M_2 are for plate current and M_3 and M_4 for grid current. When more than one stage operates from the same plate-voltage or bias-voltage supply, each stage may be metered as shown. If this system of metering is used, the meters should be mounted so that the meter dials are not accessible to accidental contact with the adjusting screw. One method of mounting is shown in Fig. 1381, where the meters are mounted behind a glass panel.

When plate milliammeters are to be mounted on metal panels, care must be taken to see that the insulation is sufficient to withstand the plate voltage. Metal-case instruments should not be mounted on a grounded metal panel if the difference in potential between the meter and the panel is to be more than 300 volts; bakelite-case instruments can be used under similar circumstances at voltages up to 1000. At higher voltages than these an insulating panel should be used.

The placing of meters at high-voltage points in the circuit may be overcome by the use of the connections shown in Fig. 1380-B and -C. The disadvantage of the arrangements at B is that the meter reads total cathode current and the grid and plate currents cannot be metered individually. This disadvantage is overcome in C, where the meters are connected across low resistances in the grid and plate return circuits. M_1 reads grid current and M_2 plate current. The parallel resistors should have a value of not less than 10 to 20 times the resistance of the meter, and should be of sufficient power rating so that there will be no possibility of resistor burn-out. If desired, the resistance values may be adjusted to form a multiplier scale for the meter (see Chapter Twenty). The same principle is used in the meter-switching system shown in Fig. 1382.

Meters may also be shifted from one stage to another by a plug-and-jack system, but this system should not be used unless it is possible to ground the frame of the jack or unless a suitable guard is provided around the meter jacks to make personal contact with high voltages impossible in normal use of the plug.

Control Circuits

Proper arrangement of controls is important if maximum convenience in operation is to be attained. If the transmitter is to be of fairly high power, it is desirable to provide a special service line leading directly from the public utility meter board to the operating room. This line should be run in conduit or BX cable, and the conductors should be of ample size to carry the maximum load without undue voltage drop. The line should be terminated with an enclosed entrance switch, properly fused.

Fig. 1383 shows the wiring diagram of a simple control system. It will be noticed that,

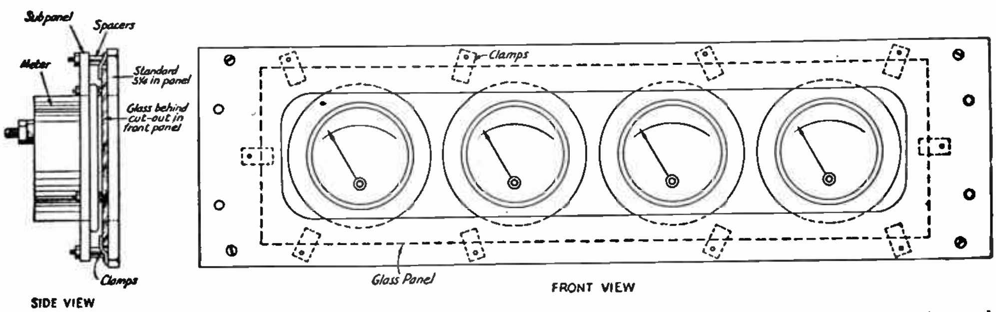


Fig. 1381 — Safety panel for meters. The meters are mounted in the usual manner on an insulating sub-panel spaced back of a glass-covered opening in the front panel. The glass is fastened in place with metal clamps or tabs, fastened to the front panel with small screws or pins. The front panel is of standard rack size, 19 × 5¼ inches.

because the control switches are connected in series, none of the high-voltage supplies can be turned on until the filament switch has been closed, and that the high-power plate supply cannot be turned on until the low-power plate supply switch has been closed. Furthermore, the modulator power cannot be applied until the final-amplifier plate voltage has been applied. SW_5 places a 100- to 300-watt lamp, L_p, in series with the primary winding of the high-voltage plate transformer for use during the process of preliminary tuning and for local c.w. work. The final amplifier should first be tuned to resonance at low voltage and SW_5 then closed, short-circuiting the lamp. Experience will determine what the low-voltage plate-current reading should be to have it increase to the full-power value when SW_5 is closed, so that the proper antenna-coupling and tuning adjustments may be made.

Preferably, SW_3 should be of the non-locking push-button type which remains closed only so long as pressure is applied. A switch of this type provides one of the simplest and most effective means of protection against accidents from high voltage. In the form which is usually considered most convenient, it consists of a switch, located underneath the operating table, which may be operated by pressure of the foot. When used in this manner the operator must be in the operating position, well removed from danger, before high voltage can be applied. If desired, SW_{3a} may be wired in parallel on the *front* of the transmitter panel, so that it can be used while tuning the transmitter. SW_3 also should be of the push-button type.

In more elaborate installations, and in remote control systems where the transmitter is located some distance from the operating position, similarly arranged switches may be used to control relays whose contacts serve to perform the actual switching at the transmitter.

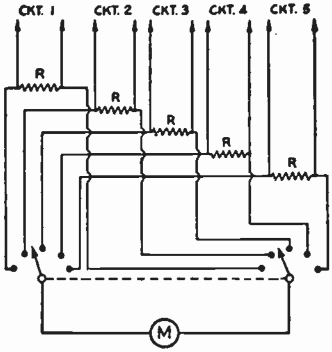


Fig. 1382 — Method of switching a single milliammeter to various circuits with a two-gang switch. The control shaft should be well insulated from the switch contacts, and should be grounded. The resistors, *R*, should have values of resistance ten to twenty times the internal resistance of the meter; 20 ohms will usually be satisfactory.

Two strings of utility outlets, one on each side of the entrance switch, are provided for operation of the receiver and such accessories as the monitor, lights, electric clock, soldering iron, etc. Closing the entrance switch should close those circuits which place the station in readiness for operation. SW_2 and SW_4 are normally closed and SW_3 is normally open. When SW_1 is closed upon entering the operating room, the transmitter filaments are turned on as also is the receiver, which should be plugged into line No. 2. With SW_4 closed (as well as SW_5 and SW_6), SW_3 performs the job of turning all plate supplies on and off during successive periods of transmission and reception.

All continuously operating accessories, such as the station clock, should be plugged into line No. 1. This is so that they will not be turned off when SW_1 is opened. Line No. 1 is of use also for supplying the soldering iron, lights, etc., when it is desired to remove all voltage from the transmitter by opening SW_1.

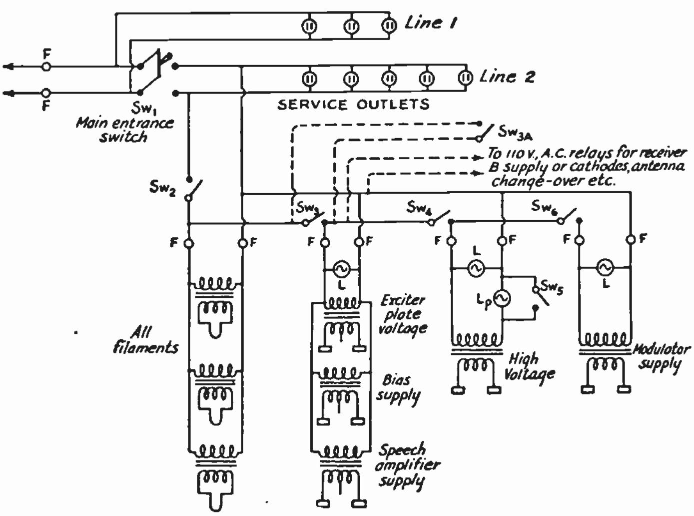


Fig. 1383 — A station control system. No high-voltage supply can be turned on until the filament switch has been closed; the high-power plate supply cannot be turned on until the low-power plate supply switch has been closed; and modulator power cannot be applied until the final-amplifier plate voltage has been applied. With all switches except SW_3 closed, SW_3 serves as the main control switch. SW_1 — Enclosed entrance switch. SW_2 — Filament switch. SW_3 — Low plate voltage and main control switch, preferably of the push-button type which remains closed only so long as pressure is applied. SW_4 — High plate-voltage switch. SW_5 — Low-power and tune-up switch short-circuiting L_p. SW_6 — Modulator plate-voltage switch. *F* — Fuse. *L* — Warning light. L_p — 100- to 300-watt voltage-reducing lamp.

Chapter Fourteen

Modulation Equipment

To PROVIDE the modulating power necessary in radiotelephone communication, audio power amplifiers or modulators are required. The units described in this chapter have been designed to give the required power output as simply and economically as possible, while still observing good design principles.

In many respects the arrangement of components is less critical in audio than in r.f. equipment; nevertheless, certain principles must be observed if difficulties are to be avoided. The selection of suitable modulation equipment for any of the transmitters in the preceding chapter is not difficult, if the fundamental principles of modulation as described in Chapter Five are understood. If the transmitter is to be plate-modulated and the power input to the modulated stage is to be of the order of 100 watts or higher, a Class-B modulator invariably will be selected. A pair of modulator tubes of any type capable of the required power output may be used. The tables in this chapter give the necessary information on the most popular tube types. The grid driving-power requirements for the modulator stage also are given, so that from this point on the speech amplifier tube line-up can be selected according to the principles outlined in Chapter Five.

The apparatus to be described is representative of current design practice for speech amplification, with units to provide the various output levels required to drive high- and low-power Class-B modulators. In some cases the power output of these amplifier units will be sufficient to modulate low-power transmitters directly, without additional power amplification. Also, practically any of the speech amplifiers shown can be used to grid-modulate transmitters up to the highest power input permitted in amateur transmitters.

Speech-amplifier equipment, especially voltage amplifiers, should be constructed on metal chassis, with all wiring kept below the chassis to take advantage of the shielding afforded. Exposed leads, particularly to the grids of low-level high-gain tubes, are likely to pick up hum from the electrostatic field which usually exists in the vicinity of house wiring. Even with the chassis, additional shielding of the input circuit of the first tube in a high-gain amplifier usually is necessary. In addition, such circuits should be separated as much as possible from power-supply transformers and chokes and also from audio transformers operating at fairly high power levels, to prevent magnetic coupling to the grid circuit which might cause hum or audio-frequency feed-back.

If a low-level microphone such as the crystal type is used, the microphone, its connecting cable, and the plug or connector by which it is attached to the speech amplifier, all should be shielded. The microphone and cable usually are constructed with suitable shielding. The cable shield should be connected to the speech amplifier chassis, and it is advisable — as well as frequently necessary — to connect the chassis to a ground such as a water pipe. Heater wiring should be kept as far as possible from grid leads, and either the center-tap or one side of the heater transformer secondary winding should be connected to the chassis. In a high-gain amplifier the first tube preferably should be of the type having the grid connection brought out to a top cap rather than to a base pin, since in the latter type the grid lead is exposed to the heater leads inside the tube and hence will pick up more hum. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

A 10-Watt Class-B Modulator for Low-Power Transmitters

A receiving-tube modulator, with a speech amplifier for either crystal or carbon microphones, is shown in Figs. 1401–1403, inclusive. It is suitable for modulating transmitters of 20 watts input or less, such as the low-power equipment frequently used on the very-high frequencies. Type 6A6 tubes are used throughout in the audio circuits, although any equivalent twin triode such as the 6N7 could be substituted. An inexpensive power supply is included, so that the unit is complete and ready for connection to the transmitter.

Fig. 1403 shows the circuit diagram of the

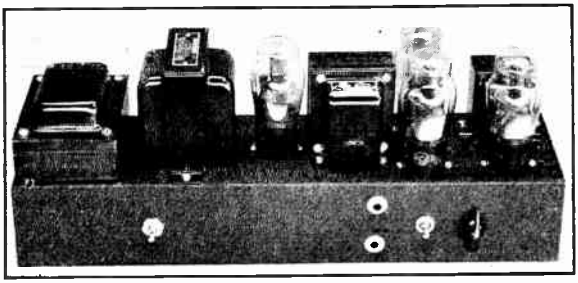


Fig. 1401 — A 10-watt audio unit complete with power supply. Three dual-triode 6A6 tubes provide a four-stage amplifier with Class-B output. Any of the popular types of microphones may be used.

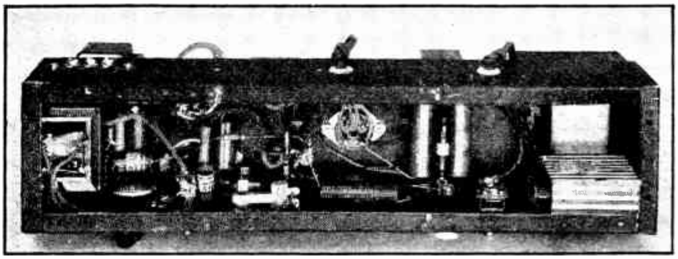


Fig. 1402 — The below-chassis wiring is visible in this view of the 10-watt modulator. The microphone input leads are kept short to reduce hum pick-up.

speech amplifier-modulator. One section of the first 6A6 is used as the input amplifier for a crystal microphone, the other half being a second speech-amplifier stage. Carbon microphones, which need less gain, are transformer-coupled to the second section of the first 6A6. The type of jack shown at J_2 in the circuit diagram must be installed if a double-button carbon microphone is to be used. J_2 may be the same as J_1 if a single-button microphone is to be used exclusively.

The gain control is connected in the grid circuit of the second section of the first 6A6 tube, which is resistance-coupled to the driver. The driver tube, also a 6A6, has its two sections connected in parallel.

The modulation transformer specified is designed to work between 6A6 plates and a 6500-ohm load; the impedance ratio actually used will, of course, depend on the load into which the modulator will work. A milliammeter can be connected across the shunt resistor, R_1, provided to measure the Class-B plate current.

The power supply is of the condenser-input type. Using the components specified, it will deliver 350 volts at 90 ma. A switch in the transformer center-tap lead is used for turning the plate voltage on and off without affecting the filament supply.

The power transformer is submounted at the left end of the chassis. Next to it is the filter choke, L_1, followed by the rectifier tube and T_3, the modulation transformer. The driver tube is at the extreme right, with T_2, the driver transformer, behind it. The Class-B tube is to the rear and in line with the speech-amplifier tube. For convenience in wiring, the audio tube sockets should be mounted with the filament prongs facing the right-hand end of the chassis.

The plate-voltage switch is on the front of the chassis toward the left in Fig. 1401. The microphone switch, gain control and microphone jacks are grouped at the right.

The bottom-view photograph, Fig. 1402, shows the layout for the components mounted below the chassis. T_1 is mounted at the left end. Wiring to the driver tube socket and the transformer secondary winding should be completed before the transformer is bolted in place, since it is difficult to reach the connecting points with a soldering iron afterwards. Short leads between the gain control, the microphone switch and the tube socket can be obtained by making the gain-control contacts face toward the switch, as shown in the photograph.

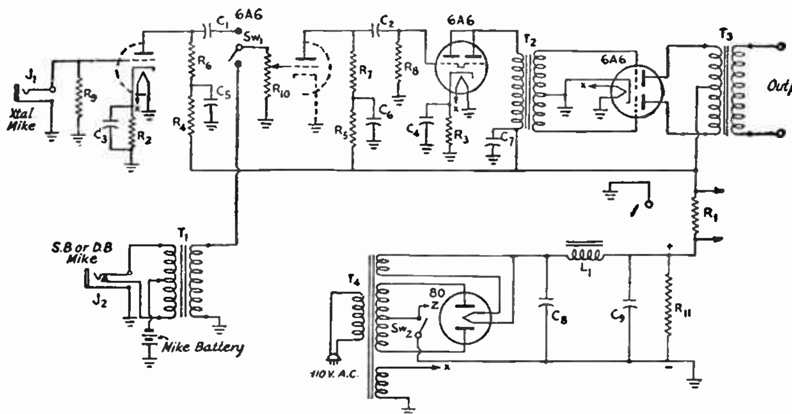


Fig. 1403 — Circuit diagram of the complete 10-watt Class-B audio modulator system for low-power transmitters.

C_1, C_2 — 0.1-μfd. 600-volt paper.
C_3, C_4 — 10-μfd. 50-volt electrolytic.
C_5, C_6, C_7, C_8, C_9 — 8-μfd. 450-volt electrolytic.
R_1 — 25 ohms, ½-watt.
R_2, R_3 — 900 ohms, 1-watt.
R_4, R_5 — 50,000 ohms, ½-watt.
R_6, R_7 — 0.25 megohm, ½-watt.
R_8 — 1 megohm, ½-watt.
R_9 — 5 megohms, ½-watt.
R_{10} — 500,000-ohm volume control.
R_{11} — 25,000 ohms, 10-watt.
Sw_1 — S.p.d.t. toggle switch.
Sw_2 — S.p.s.t. toggle switch (see text).
J_1 — Closed-circuit jack for crystal microphone.
J_2 — 2- or 3-circuit jack for single-button or double-button carbon microphone.
T_1 — S.b. or d.b. microphone transformer (Stancor A-4351).
T_2 — Driver transformer, parallel 6A6 plates to 6A6 Class-B (Stancor A-4216).
T_3 — Output transformer, 6A6 Class-B to 6500-ohm load (Stancor A-3845).
T_4 — Power transformer, 700-0-700 volts, 90 ma.; 5 volts at 3 amperes; 6.3 volts at 3.5 amperes.
L_1 — Filter choke, 5 henrys, 200 ma., 80 ohms (Thordarson T-67C49).

The compact microphone battery (Burgess type 3A2) will be held securely in place without brackets or clips if it is wedged in between the bottom of the power transformer and the lips on the bottom of the chassis. A 3-volt battery is sufficient for most carbon microphones, and low current frequently will give better speech quality. The 115-volt a.c. and the meter leads (rubber-covered lamp cord) enter the chassis through rubber grommets. A three-contact terminal strip is located at the right end of the base (left end in the bottom view). One of the contacts on this terminal strip is for an external ground connection and the other two are connected to the modulation-transformer output winding.

The actual measured power output of the unit shown in the photographs is 11 watts, as recorded at the point where distortion just begins to be noticeable. This order of audio power output is ample for modulating a low-power transmitter operating with 20 watts or so input to the final stage.

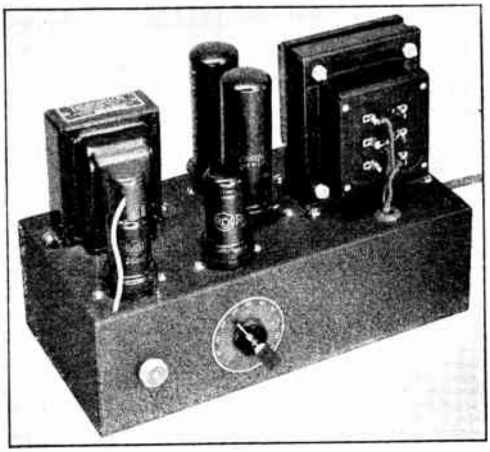


Fig. 1404 — A low-cost speech-amplifier or low-power modulator unit with a maximum audio output of 20 watts. The 6J7 is at the left near corner of the chassis, with the 6J5 to its right, just above the volume control.

ℂ A 20-Watt Speech Amplifier or Modulator

The amplifier shown in Figs. 1404–1406 will deliver audio power outputs up to 20 watts (from the output transformer secondary) with ample gain for ordinary communications-type crystal microphones. Class-AB 6L6s are used in the output stage, preceded by a 6J5 and a 6J7 preamplifier.

The unit is built up on a 5 × 10 × 3-inch chassis, with the parts arranged as shown in the photographs. About the only constructional precaution necessary is to use a short lead from the microphone socket (a jack may be used instead of the screw-on type, if desired), and to shield thoroughly the input circuit to the grid of the 6J7. This shielding is necessary to reduce hum. In this amplifier, the 6J7 grid resistor, R_1, is enclosed along with the input jack in a National type J-1 jack shield, and a shielded lead is run from the jack shield to the grid of the 6J7. A metal slip-on shield covers the grid cap of the tube.

To realize maximum power output, the "B" supply should be capable of delivering about 145 ma. at 360 volts. A condenser-input supply of ordinary design (Chapter Eight) may be used, since the variation in plate current is relatively small. The current is approximately 120 ma. with no input signal and 145 ma. at full output. If an output of 12 or 13 watts will be sufficient, R_9 and R_{10} may be omitted and all tubes fed directly from a "B" supply giving 270 volts at approximately 175 ma.

The output transformer shown is a universal modulation type suitable for coupling into the plate circuit of a low-power r.f. amplifier (input 40 watts maximum for 100 per cent modulation) for plate modulation. For cathode modulation, the r.f. input power that can be modulated can be determined from the data in

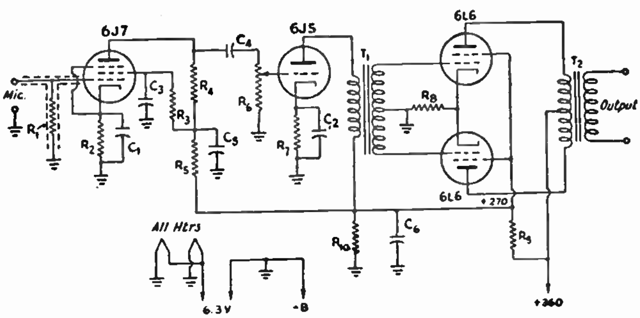


Fig. 1405 — Circuit diagram of the low-cost speech amplifier or modulator capable of power outputs up to 20 watts.

C_1, C_2 — 20-μfd. 50-volt electrolytic.
C_3 — 0.1-μfd. 200-volt paper.
C_4 — 0.01-μfd. 400-volt paper.
C_5, C_6 — 8-μfd. 450-volt electrolytic.
R_1 — 5 megohms, ½-watt.
R_2 — 1300 ohms, ½-watt.
R_3 — 1.5 megohms, ½-watt.
R_4 — 0.25 megohm, ½-watt.
R_5 — 50,000 ohms, ½-watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1-watt.
R_8 — 250 ohms, 10-watt.
R_9 — 2000 ohms, 10-watt.
R_{10} — 20,000 ohms, 25-watt.
T_1 — Interstage audio transformer, single plate to p.p. grids, ratio 3:1 (Thordarson T-57A41).
T_2 — Output transformer, type depending on requirements. A multi-tap modulation transformer (Thordarson T-19M14) is shown.

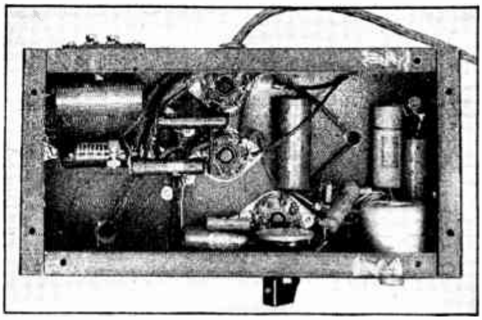


Fig. 1406 — Bottom view of the 20-watt speech amplifier or modulator chassis. The most important constructional point is complete shielding of the microphone input circuit up to the grid of the 6J7 first amplifier.

Chapter Five. The amplifier may also be used for grid-bias modulation with the transformer specified. If the unit is to be used to drive a Class-B modulator, it is recommended that the Class-B tubes be of the zero-bias type rather than a type requiring fixed bias. A suitable output transformer must be substituted for this purpose; data may be found in transformer manufacturers' catalogs.

The frequency response of the amplifier is ample for the range of frequencies encountered in voice communication. It may be extended for high-quality reproduction of music by using higher-priced audio transformers.

A 40-Watt Output Speech Amplifier or Modulator

The 40-watt amplifier shown in Figs 1407–1409 resembles in many respects the 20-watt amplifier just described. The first two stages are, in fact, identical in circuit and construction. To obtain the higher output, however, it is necessary to drive the 6L6s into the grid current region (Class AB_2 operation), so that a driver stage capable of furnishing sufficient power is required. A pair of transformer-coupled 6J5s in push-pull is used for this purpose, inserted between the single 6J5 stage and the push-pull 6L6s. Decoupling is provided (R_9 and C_5) to prevent motorboating because of the higher over-all gain of the amplifier.

A 6 × 14 × 3-inch chassis is used for the 40-watt amplifier. The photographs show the arrangement of parts. As in the case of the 20-watt unit, complete shielding of the microphone input circuit is essential. The amplifier has ample gain for crystal microphones.

This unit may be used to plate-modulate 80 watts input to an r.f. amplifier. For cathode modulation, the input that can be modulated will depend upon the type of operation chosen,

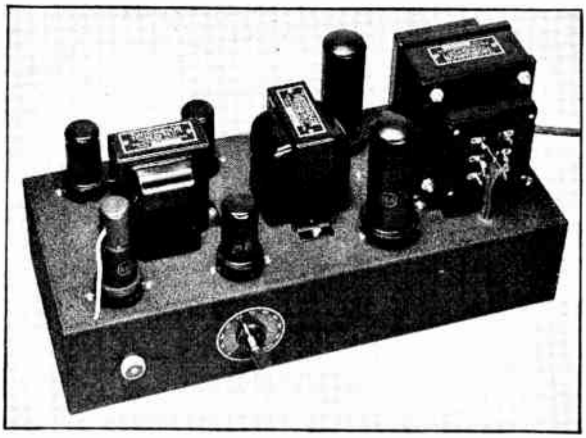


Fig. 1407 — A 40-watt speech amplifier or modulator of inexpensive construction. The 6J7 and first 6J5 are at the front, near the microphone socket and volume control, respectively. T_1 is behind them, and the push-pull 6J5s are at the rear of the chassis behind T_1. T_2, in the center, the push-pull 6L6s, and T_3 follow in order to the right.

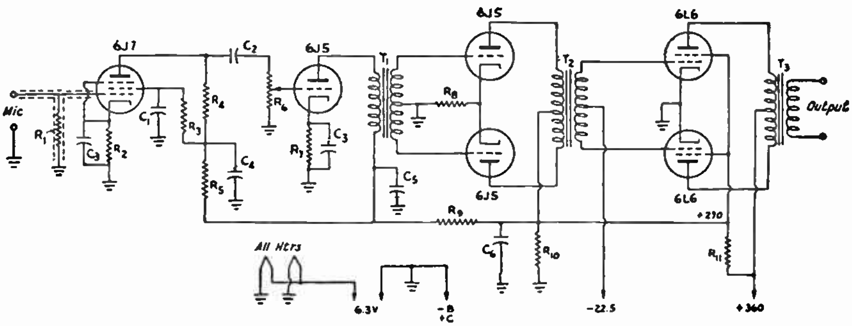


Fig. 1408 — Circuit diagram of the Class AB_2 push-pull 6L6 40-watt output speech amplifier or modulator.

C_1 — 0.1-μfd. 200-volt paper.
C_2 — 0.01-μfd. 400-volt paper.
C_3 — 20-μfd. 50-volt electrolytic.
C_4, C_5, C_6 — 8-μfd. 450-volt electrolytic.
R_1 — 5 megohms, ½-watt.
R_2 — 1300 ohms, ½-watt.
R_3 — 1.5 megohm, ½-watt.
R_4 — 0.25 megohm, ½-watt.
R_5 — 50,000 ohms, ½-watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1-watt.
R_8 — 750 ohms, 1-watt.
R_9 — 12,000 ohms, 1-watt.
R_{10} — 20,000 ohms, 25-watt.
R_{11} — 1500 ohms, 10-watt.
T_1 — Interstage audio, single plate to p.p. grids, 3:1 ratio (Thordarson T-57A41).
T_2 — Driver transformer, p.p. 6J5s to 6L6s Class AB2 (Thordarson T-84D59).
T_3 — Output transformer, type depending on requirements. A multi-tap modulation transformer (Thordarson T-19M15) is shown.

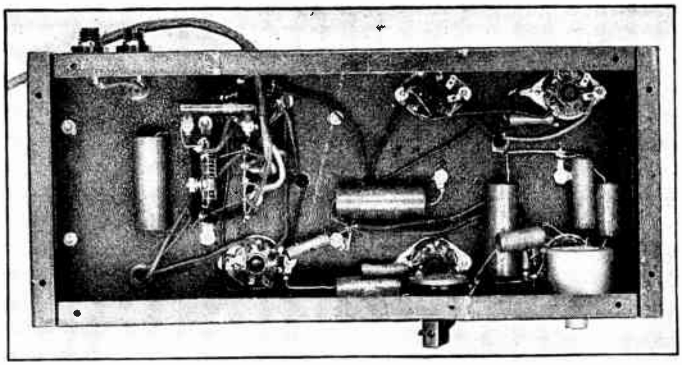


Fig. 1409 — Underneath the chassis of the 40-watt speech amplifier-modulator.

as described in Chapter Five; with 55 per cent plate efficiency in the r.f. stage, for instance, the input may be of the order of 200 watts, making an allowance for the small amount of audio power taken by the grid circuit.

A high-power Class-B modulator can be driven by the unit; data on suitable modulator tubes are given later in this chapter. Zero-bias tubes should be used, because they present a more constant load to the 6L6s than do relatively low amplification-factor tubes which require fixed bias for Class-B operation. A suitable Class-B driver transformer should be substituted for the universal modulation transformer shown.

The power supply should have good voltage regulation, since the total "B" current varies from approximately 140 ma. with no signal to 265 ma. at full output. A heavy-duty choke-input plate supply should be used; general design data will be found in Chapter Eight. The heater requirements are 6.3 volts at 3 amperes. Bias for the 6L6 stage is most conveniently supplied by a 22.5-volt "B" battery bloc; a small-sized unit will be satisfactory, since no current is drawn.

ℂ A Push-Pull 2A3 Amplifier with Volume Compression

Ideally, a Class-B modulator should be driven by an amplifier having exceptionally good voltage regulation, to minimize distortion (see Chapter Five). For average amateur work, the 6L6 amplifiers just described will give entirely satisfactory results as drivers for Class-B stages when operated well within their capabilities, especially with zero-bias Class-B tubes. However, somewhat better performance can be secured by using triode drivers, especially when the grid power requirements of the Class-B stage are modest enough to make the use of triodes such as the 2A3 practicable.

The amplifier shown in Figs. 1410–1412, inclusive, has an output (from the transformer secondary) of 6 watts with negligible distortion, and thus is suitable for driving Class-B stages of 100 to 250 watts output.

The amplifier also incorporates an automatic volume-compression circuit to maintain a high average percentage of modulation (Chapter Five). This feature is often of considerable value in practical communications work where interference conditions require maximum carrier power level to transmit intelligence successfully. Volume compression overcomes to some extent the general tendency of even the best operators to accentuate or otherwise vary the syllabic intensity. This is particularly true when talking close to the microphone, under which conditions slight movements of the head will cause a change in the modulation level.

A practical audio volume compression circuit functions much like the r.f. automatic gain control familiarly employed in superheterodyne receivers (§ 7-13).

In Fig. 1412, the side amplifier and rectifier, combined in the 6SQ7 tube, rectifies a portion of the voice current. The rectified output of this circuit is filtered and applied to the Nos. 1 and 3 grids of a pentagrid amplifier tube, thereby varying its gain in inverse proportion to the signal strength. With proper adjustment, an average increase in modulation level of about 7 db. can be secured without exceeding 100 per cent modulation on peaks.

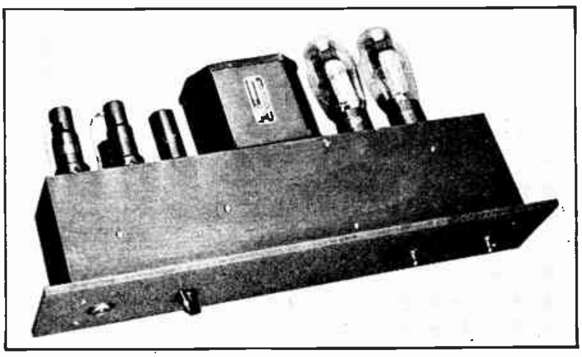


Fig. 1410 — A push-pull 2A3 speech amplifier having an output of approximately 6 watts. A volume-compression circuit is included.

The amplifier proper consists of a 6J7 first stage followed by a 6L7 amplifier-compressor. The 2A3 grids are driven by a 6N7 self-balancing phase inverter. The operation of the 2A3s is purely Class A, without grid current.

The amount of compression is controlled by means of the potentiometer, R_{20}, in the grid circuit of the 6SQ7. A switch, S_1, is provided to short-circuit the rectified output of the compressor when normal amplification is required.

The construction of the amplifier

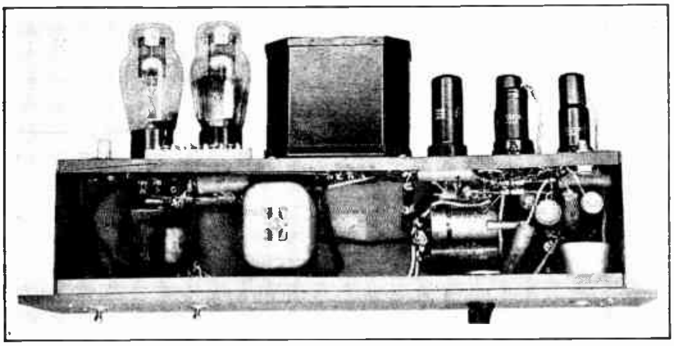


Fig. 1411 — Bottom view of the push-pull Class-A 2A3 speech amplifier with automatic volume compression. The circuit diagram is shown in Fig. 1412.

resembles that of the unit shown in Fig. 1401, the tubes and output transformer being mounted on the rear edge of a 17 × 4 × 3-inch chassis to save panel height in relay-rack mounting. Looking at the amplifier from the front, the 6J7 first amplifier is in the upper left corner, with the 6L7 to its right. The 6SQ7 is below the 6L7. The 6N7 is followed by the output transformer, the latter being placed in the middle of the chassis in order to assist in distributing the weight evenly. The 2A3 tubes and the power-supply and audio output terminals are at the right.

In the underneath view the input circuit is at the left, the grid risistor, R_1, and the microphone connector socket being shielded to minimize hum pick-up by the National JS-1 jack shield. The lead to the 6J7 grid is shielded, as are also the top caps of this tube and the 6L7. The volume compressor control, R_{20}, is mounted beside the 6J7, and is screwdriver adjusted; a midget control should be used, since the space is rather limited. The other parts are mounted as close as possible to the points in the circuit to which they connect. The filament transformers should be kept well separated from the wiring in the low-level stages, particularly that of the microphone input and the low-level grid circuits.

Adjustment of the compressor control is rather critical. First set R_{20} at zero and adjust the gain control, R_6, for full modulation with the particular microphone used. Then advance the compressor control until the amplifier just "cuts off" (output decreasing to a low value) on peaks; when this point is reached, back off the compressor control until the cut-off effect is gone but an obvious decrease in gain follows each peak.

Because of the necessity for filtering out the audio-frequency component in the rectifier output, there will be a slight delay (amounting to a fraction of a second) before the decrease in gain "catches up" with the peak. This is caused by the time constant of the circuit, and so is unavoidable.

When a satisfactory setting is secured, as indicated by good speech quality with a definite reduction in gain on peaks, the gain control, R_6, should be advanced to give full output with normal operation. Too much volume compression, indicated by the cut-off effect following each peak, is definitely undesirable, and the object of adjustment of the compressor control should be to use as much compression as possible without danger of over-compression.

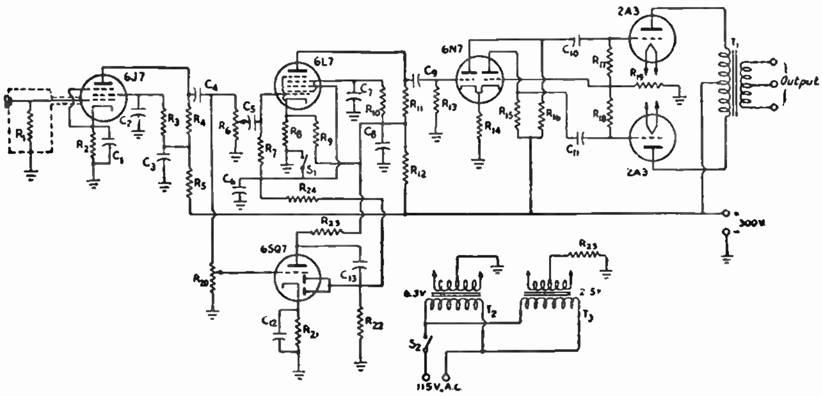


Fig. 1412 — Circuit diagram of the Class-A 2A3 volume-compression speech amplifier.

C_1, C_{12} — 10-μfd. 50-volt electrolytic.
C_2, C_4, C_5, C_6, C_9, C_{10}, C_{11}, C_{13} — 0.1-μfd. 400-volt paper.
C_3, C_8 — 8-μfd. 450-volt electrolytic.
C_7 — 0.5-μfd. 400-volt paper.
R_1 — 5 megohms, ½-watt.
R_2, R_8 — 1200 ohms, ½-watt.
R_3, R_7 — 2 megohms, ½-watt.
R_4, R_{13}, R_{22}, R_{24} — 0.5 megohm, ½-watt.
R_5 — 50,000 ohms, ½-watt.
R_6, R_{20} — 0.5-megohm variable.
R_9 — 0.25 megohm, 1-watt.
R_{10}, R_{11}, R_{23} — 0.1 megohm, ½-watt.
R_{12} — 10,000 ohms, ½-watt.
R_{14} — 1500 ohms, ½-watt.
R_{15}, R_{16} — 0.1 megohm, 1-watt.
R_{17}, R_{18}, R_{19} — 0.25 megohm, ½-watt.
R_{21} — 5000 ohms, ½-watt.
R_{25} — 750 ohms, 10-watt.
T_1 — Output transformer to match pp. 2A3s to Class-B grids. (UTC PA-53AX).
T_2 — Filament transformer, 6.3 volts, 2 amperes.
T_3 — Filament transformer, 2.5 volts, 5 amperes.

TABLE I — RESISTANCE-COUPLED VOLTAGE AMPLIFIER DATA

Data are given for a plate-supply of 300 volts; departures of as much as 50 per cent from this supply voltage will not materially change the operating conditions or the voltage gain, but the output voltage will be in proportion to the new voltage. Voltage gain is measured at 400 cycles; condenser values given are based on 100-cycle cut-off. For increased low-frequency response, all condensers may be made larger than specified (cut-off frequency in inverse proportion to condenser values provided all are changed in the same proportion). A variation of 10 per cent in the values given has negligible effect on the performance.

High-frequency cut-off with pentodes is approximately 20,000 cycles with a plate resistor of 0.1 megohm, 10,000 cycles with 0.25 megohm, and 5000 cycles with 0.5 megohm. With triode amplifiers, the high-frequency cut-off is well above the audio range.

	Plate Resistor Megohms	Next-Stage Grid Resistor Megohms	Screen Resistor Megohms	Cathode Resistor Ohms	Screen By-pass μfd.	Cathode By-pass μfd.	Blocking Condenser μfd.	Output Volts (Peak)[2]	Voltage Gain[3]
6A6, 6N7, 53	0.1	0.1	—	1150[1]	—	—	0.03	60	20
		0.25	—	1500[1]	—	—	0.015	83	22
		0.5	—	1750[1]	—	—	0.007	86	23
	0.25	0.25	—	2650[1]	—	—	0.015	75	23
		0.5	—	3400[1]	—	—	0.0055	87	24
		1.0	—	4000[1]	—	—	0.003	100	24
	0.5	0.5	—	4850[1]	—	—	0.0055	76	23
		1.0	—	6100[1]	—	—	0.003	94	24
		2.0	—	7150[1]	—	—	0.0015	104	24
6C5 (also 6J7, 6C6, 57, 6W7, 7C7 as triodes)[4]	0.05	0.05	—	2100	—	3.16	0.075	57	11
		0.1	—	2600	—	2.3	0.04	70	11
		0.25	—	3100	—	2.2	0.015	83	12
	0.1	0.1	—	3800	—	1.7	0.035	65	12
		0.25	—	5300	—	1.3	0.015	84	13
		0.5	—	6000	—	1.17	0.008	88	13
	0.25	0.25	—	9600	—	0.9	0.015	73	13
		0.5	—	12,300	—	0.59	0.008	85	14
		1.0	—	14,000	—	0.37	0.003	97	14
6C6, 6J7, 6W7, 7C7, 57 (pentode)	0.1	0.1	0.44	500	0.07	8.5	0.02	55	61
		0.25	0.5	450	0.07	8.3	0.01	81	82
		0.5	0.53	600	0.06	8.0	0.006	96	94
	0.25	0.25	1.18	1100	0.04	5.5	0.008	81	104
		0.5	1.18	1200	0.04	5.4	0.005	104	140
		1.0	1.45	1300	0.05	5.8	0.005	110	185
	0.5	0.5	2.45	1700	0.04	4.2	0.005	75	161
		1.0	2.9	2200	0.04	4.1	0.003	97	350
		2.0	2.95	2300	0.04	4.0	0.0025	100	240
6C8G (one triode unit)	0.1	0.1	—	2120	—	3.93	0.037	55	22
		0.25	—	2840	—	2.01	0.013	73	23
		0.5	—	3250	—	1.79	0.007	80	25
	0.25	0.25	—	4750	—	1.29	0.013	64	25
		0.5	—	6100	—	0.96	0.0065	80	26
		1.0	—	7100	—	0.77	0.004	90	27
	0.5	0.5	—	9000	—	0.67	0.007	67	27
		1.0	—	11,500	—	0.48	0.004	83	27
		2.0	—	14,500	—	0.37	0.002	96	28
6F5, 6SF5, 7B4	0.1	0.1	—	1300	—	5.0	0.025	33	42
		0.25	—	1600	—	3.7	0.01	43	49
		0.5	—	1700	—	3.2	0.006	48	52
	0.25	0.25	—	2600	—	2.5	0.01	41	56
		0.5	—	3200	—	2.1	0.007	54	63
		1.0	—	3500	—	2.0	0.004	63	67
	0.5	0.5	—	4500	—	1.5	0.006	50	65
		1.0	—	5400	—	1.2	0.004	62	70
		2.0	—	6100	—	0.93	0.002	70	70
6F8G (one triode unit), 6J5, 6J5G, 7A4, 7N7	0.05	0.05	—	1020	—	3.56	0.06	41	13
		0.1	—	1270	—	2.96	0.034	51	14
		0.25	—	1500	—	2.15	0.012	60	14
	0.1	0.1	—	1900	—	2.31	0.035	43	14
		0.25	—	2440	—	1.42	0.0125	56	14
		0.5	—	2700	—	1.2	0.0065	64	14
	0.25	0.25	—	4590	—	0.87	0.013	46	14
		0.5	—	5770	—	0.64	0.0075	57	14
		1.0	—	6950	—	0.54	0.004	64	14
6L5G	0.05	0.05	—	1740	—	2.91	0.06	56	11[5]
		0.1	—	2160	—	2.18	0.032	68	12[5]
		0.25	—	2600	—	1.82	0.015	79	12[5]
	0.1	0.1	—	3070	—	1.64	0.032	60	12[5]
		0.25	—	4140	—	1.1	0.014	79	13[5]
		0.5	—	4700	—	0.81	0.0075	89	13[5]
	0.25	0.25	—	6900	—	0.57	0.013	64	13[5]
		0.5	—	9100	—	0.46	0.0075	80	13[5]
		1.0	—	10,750	—	0.4	0.005	88	13[5]

[1] Value for both triode sections, assuming both are working under same conditions. In phase inverter service, the cathode resistor should not be by-passed.
[2] Voltage across next-stage grid resistor at grid-current point.
[3] At 5 volts r.m.s. output.
[4] Screen and suppressor tied to plate.
[5] At 4 volts r.m.s. output.

TABLE I—RESISTANCE-COUPLED VOLTAGE AMPLIFIER DATA—*Continued*

	Plate Resistor Megohms	Next-Stage Grid Resistor Megohms	Screen Resistor Megohms	Cathode Resistor Ohms	Screen By-pass μfd.	Cathode By-pass μfd.	Blocking Condenser μfd.	Output Volts (Peak)[2]	Voltage Gain[3]
6R7, 6R7G, 7E6	0.05	0.05	—	1600	—	2.6	0.055	50	9
		0.1	—	2000	—	2.0	0.03	62	9
		0.25	—	2400	—	1.6	0.015	71	10
	0.1	0.1	—	2900	—	1.4	0.03	52	10
		0.25	—	3800	—	1.1	0.015	68	10
		0.5	—	4400	—	1.0	0.007	71	10
	0.25	0.25	—	6300	—	0.7	0.015	54	10
		0.5	—	8400	—	0.5	0.007	62	11
		1.0	—	10,600	—	0.44	0.004	74	11
6S7	0.1	0.1	0.59	430	0.077	8.5	0.0167	57	57[5]
		0.25	0.67	440	0.071	8.0	0.01	73	78[5]
		0.5	0.71	440	0.071	8.0	0.0066	82	89[5]
	0.25	0.25	1.7	620	0.058	6.0	0.0071	54	98[5]
		0.5	1.95	650	0.057	5.8	0.005	66	122[5]
		1.0	2.1	700	0.055	5.2	0.0036	76	136[5]
	0.5	0.5	3.6	1000	0.04	4.1	0.0037	52	136[5]
		1.0	3.9	1080	0.041	3.9	0.0029	66	162[5]
		2.0	4.1	1120	0.043	3.8	0.0023	73	174[5]
6SC7	0.1	0.1	—	750[1]	—	—	0.033	35	29
		0.25	—	930[1]	—	—	0.014	50	34
		0.5	—	1040[1]	—	—	0.007	54	36
	0.25	0.25	—	1400[1]	—	—	0.012	45	39
		0.5	—	1680[1]	—	—	0.006	55	42
		1.0	—	1840[1]	—	—	0.003	64	45
	0.5	0.5	—	2330[1]	—	—	0.006	50	45
		1.0	—	2980[1]	—	—	0.003	62	48
		2.0	—	3280[1]	—	—	0.002	72	49
6SJ7	0.1	0.1	0.35	500	0.10	11.6	0.019	72	67
		0.25	0.37	530	0.09	10.9	0.016	96	98
		0.5	0.47	590	0.09	9.9	0.007	101	104
	0.25	0.25	0.89	850	0.07	8.5	0.011	79	139
		0.5	1.10	860	0.06	7.4	0.004	88	167
		1.0	1.18	910	0.06	6.9	0.003	98	185
	0.5	0.5	2.0	1300	0.06	6.0	0.004	64	200
		1.0	2.2	1410	0.05	5.8	0.002	79	238
		2.0	2.5	1530	0.04	5.2	0.0015	89	263
6SQ7, 6B6G, 7B6, 2A6, 75	0.1	0.1	—	1900	—	4.0	0.03	31	31
		0.25	—	2200	—	3.5	0.015	41	39
		0.5	—	2300	—	3.0	0.007	45	42
	0.25	0.25	—	3300	—	2.7	0.015	42	48
		0.5	—	3900	—	2.0	0.007	51	53
		1.0	—	4200	—	1.8	0.004	60	56
	0.5	0.5	—	5300	—	1.6	0.007	47	58
		1.0	—	6100	—	1.3	0.004	62	60
		2.0	—	7000	—	1.2	0.002	67	63
6T7G	0.1	0.1	—	1950	—	2.85	0.0245	44	27[5]
		0.25	—	2400	—	2.55	0.0135	58	32[5]
		0.5	—	2640	—	2.25	0.008	64	33[5]
	0.25	0.25	—	3760	—	1.57	0.012	57	37[5]
		0.5	—	4580	—	1.35	0.0075	69	40[5]
		1.0	—	5220	—	1.23	0.005	80	41[5]
	0.5	0.5	—	6570	—	1.02	0.008	62	42[5]
		1.0	—	8200	—	0.82	0.0055	77	43[5]
		2.0	—	9600	—	0.70	0.004	86	44[5]
56, 76	0.05	0.05	—	2400	—	2.8	0.08	65	8.3
		0.1	—	3100	—	2.2	0.045	80	8.9
		0.25	—	3800	—	1.8	0.02	95	9.4
	0.1	0.1	—	4500	—	1.6	0.04	74	9.5
		0.25	—	6400	—	1.2	0.02	95	10.0
		0.5	—	7500	—	0.98	0.009	104	10.0
	0.25	0.25	—	11,100	—	0.69	0.02	82	10.0
		0.5	—	15,200	—	0.5	0.009	96	10.0
		1.0	—	18,300	—	0.4	0.005	108	10.0

[1] Value for both triode sections, assuming both are working under same conditions. In phase inverter service, the cathode resistor should not be by-passed.
[2] Voltage across next-stage grid resistor at grid-current point.
[3] At 5 volts r.m.s. output.
[4] Screen and suppressor tied to plate.
[5] At 4 volts r.m.s. output.

TABLE II—CLASS-B MODULATOR DATA

Class-B Tubes (2)	Fil. Volts	Plate Volts	Grid Volts App.	Peak A.F. Grid-to-Grid Voltage	Zero-Sig.[1] Plate Current Ma.	Max.-Sig.[1] Plate Current Ma.[2]	Load Res. Plate-to-Plate Ohms	Max.-Sig. Driving Power Watts[3]	Max.-Sig.[1] Power Output Watts[3]
RK59[4]	6.3	500	−17	64	16	90	15,000	0.9	30
HY60[7]	6.3	300	−22.5	63	75	120	5,000	4.0	22
		400	−22.5	57	75	120	6,000	3.0	30
HY65[5]	6.3	450	—	—	—	125	—	0.4	34
801-A/801	7.5	600	−75	320	8	130	10,000	3.0	45
HY31Z[4,5]	6.3	300	0	104	20	100	5,000	1.4	18
HY1231Z[4,5]	12.6	400	0	140	26	150	5,000	2.0	40
		500	0	131	36	150	7,000	1.8	51
815[8]	6.3	400	−15	60	22	150	8,000	0.36	42
		500[6]	−15	60	20	150	6,200	0.36	54
1624[7]	2.5	400	−16.5	77	75	150	6,000	0.4	36
		600	−25	106	42	180	7,500	1.2	72
HY6L6GX[7]	6.3	400	−25	80	100	230	3,800	0.35	60
		500	−25	80	100	230	4,550	0.6	75
TZ20	7.5	800	0	160	40	136	12,000	1.8	70
HY61/807[7]	6.3	400	−25	80	100	230	3,800	0.35	60
RK807	6.3	500	−25	80	100	230	4,550	0.6	75
		600	−30	80	60	200	6,600	0.4	80
HY69[5,7]	6.3	300	−25	106	60	150	4,000	0.25	30
HY1269[5,7]	12.6	400	−25	145	60	170	4,000	0.4	40
		600	−35	183	65	120	4,500	0.3	65
		500	−25	120	65	200	5,000	0.7	97
RK12	6.3	750	0	129	50	200	9,600	3.4	100
800	7.5	750	−40	320	26	210	6,400	6.0	90
		1000	−55	300	28	160	12,500	4.4	100
		1250	−70	300	30	130	21,000	3.4	106
HY30Z	6.3	600	0	171	18	180	6,000	Note 9	75
		750	0	167	22	190	8,000	"	95
		850	0	171	28	180	10,000	"	110
807[10]	6.3	400	−25	78	100	240	3,200	0.2	55
1625[10]	12.6	500	−25	78	100	240	4,240	0.2	75
		600	−30	78	60	200	6,400	0.1	80
		750[6]	−32	92	60	240	6,950	0.2	120
HK24	6.3	1000	−29	248	30	150	15,000	4.5	105
		1250	−42	256	24	136	21,200	4.2	120
809	6.3	500	−10	170	40	200	5,200	3.5	60
		750	−25	200	35	200	8,400	4.0	100
		1000[6]	−40	230	30	200	12,000	4.2	145
830-B	10	800	−27	250	20	280	6,000	5.0	135
		1000	−35	270	20	280	7,600	6.0	175
HY40Z	7.5	750	0	171	32	225	6,000	Note 9	110
		850	0	185	40	250	7,000	"	155
		1000	0	185	45	250	9,000	"	185
RK31	7.5	1000	0	141	25	230	11,000	3.7	160
		1250	0	141	35	220	18,000	4.4	190
808	7.5	1250	−15	240	40	230	12,700	7.8	190
		1500	−25	220	30	190	18,300	4.8	185
RK37	7.5	1250	−35	282	25	235	18,000	7.2	200
811	6.3	1250	0	140	48	200	15,000	3.8	175
		1500[6]	− 9	160	20	200	18,000	4.2	225
35T	5.0 to 5.1	1000	−22	—	—	—	7,200	—	150
		1250	−30	—	—	—	9,600	—	200
		1500	−40	—	—	—	12,800	—	230
TZ40[8]	7.5	1000	0	220	—	280	7,350	5.5	175
		1250	− 4.5	269	—	280	10,000	6.0	225
		1500	− 9	265	—	250	12,000	6.0	250
RK52	7.5	1250	0	180	40	300	10,000	7.5	250
203-A	10	1000	−35	310	26	320	6,900	10	200
		1250	−45	330	26	320	9,000	11	260
211	10	1000	−77	380	20	320	6,900	7.5	200
		1250	−100	410	20	320	9,000	8.0	260
838	10	1000	0	200	106	320	6,900	7.0	200
		1250	0	200	148	320	9,000	7.5	260
HK158	12.6	750	−25	300	50	330	4,500	17	155
		1250	−50	280	35	225	12,500	10	200
		2000	−90	340	30	180	3,200	10	265
HK54	5.0	1500	−45	300	40	198	16,800	5.0	200
		2000	−70	360	24	180	36,000	6.0	260
		2500	−85	360	20	150	40,000	5.0	275
HY51Z	7.5	850	0	148	48	300	5,000	Note 9	160
		1000	0	170	60	350	6,000	"	260
		1250	0	155	90	300	10,000	"	285
203-Z	10	1000	0	206	50	350	6,200	6.5	230
		1250	− 4.5	215	60	350	8,000	6.75	300
ZB120	10	1000	0	190	70	310	6,900	5.0	200
		1250	0	180	95	300	9,000	4.0	245
		1500	− 9	196	60	296	11,200	5.0	300

TABLE II — CLASS-B MODULATOR DATA — *Continued*

Class-B Tubes (2)	Fil. Volts	Plate Volts	Grid Volts App.	Peak A.F. Grid-to-Grid Voltage	Zero-Sig.[1] Plate Current Ma.	Max.-Sig.[1] Plate Current Ma.[2]	Load Res. Plate-to-Plate Ohms	Max.-Sig. Driving Power Watts[3]	Max.-Sig.[1] Power Output Watts[3]
8005	10	1250	−55	290	40	320	8,000	4.0	250
		1500[6]	−80	310	40	310	2,500	4.0	300
HF100	10 to 11	1500	−52	264	50	270	12,000	2.0	260
		1750	−62	324	40	270	16,000	9.0	350
805 RK57	10	1250	0	235	148	400	6,700	6.0	300
		1500	−16	280	84	400	8,200	7.0	370
828[11]	10	1700	−120	240	50	248	16,200	0	300
		2000	−120	240	50	270	18,300	0	385
25T	5.0	2000	−80	270	16	80	55,500	0.7	110
		1500	−55	230	21	94	33,700	0.8	90
		1000	−30	210	32	120	15,800	1.2	70
		750	−20	205	43	133	9,200	1.4	50
3C24	6.3	Same as 25T							
75T	5.0	1000	—	—	—	—	6,800	—	200
		1500	—	—	—	—	10,000	—	300
		2000	—	—	—	—	12,500	—	400
8003	10	1350	−100	480	40	490	6,000	10.5	460
100TH	5.0 to 5.1	2000	Bias adjusted for maximum rated plate dissipation under no-signal conditions Zero bias up to 1250 v. plate				16,000	May be driven by push-pull 6L6s	380
		2500					22,000		460
		3000					30,000		500
HD203-A	10	1500	−40	—	36	425	8,000	Note 12	400
		1750	−67	—	36	425	9,000		500
HK254	5.0	2000	−65	400	50	260	16,000	7.0	328
		2500	−80	420	50	248	22,000	7.0	418
		3000	−100	456	40	240	30,000	7.0	520
810	10	1500	−30	345	80	500	6,600	12	510
1627	5.0	2000	−50	345	60	420	11,000	10	590

[1] Values are for both tubes.
[2] Sinusoidal signal values; speech values are approximately one-half for tubes biased to approximate cut-off and 80 per cent for zero-bias tubes.
[3] Values do not include transformer losses. Somewhat higher power is required of the driver to supply losses and provide good regulation. Input transformer ratios must be chosen to supply required power at specified grid-to-grid voltage with ample reserve for losses and low distortion levels. Driver stage should have good regulation.
[4] Dual tube. Values are for one tube, both sections.
[5] Instant-heating filament type.
[6] Intermittent amateur and commercial service rating.
[7] Beam tube. Class AB_2. Screen voltage: 300.
[8] Beam tube. Class AB_2. Screen voltage: 125 at 32 ma.
[9] Driver: one or two 45s at 275 volts, self-biased (−55 volts).
[10] Beam Tube. Class AB_2. Screen voltage: 300 at 10 ma. Effective grid circuit resistance should not exceed 500 ohms.
[11] Pentode. Class AB_2. Suppressor voltage: 60 at 9 ma. Screen voltage: 750, 4/43 ma. at 1700 plate volts, 2/60 ma. at 2000.
[12] Can be driven by a pair of 2A3s in push-pull Class AB at 300 volts with fixed bias.

Class-B Modulators

Class-B modulator circuits are practically identical no matter what the power output of the modulator. The diagrams of Fig. 1313 therefore will serve for any modulator of this type that the amateur may elect to build. The triode circuit is given at A and the circuit for tetrodes at B. When small tubes with indirectly heated cathodes are used, the cathode should be connected to ground.

Design considerations for Class-B stages are discussed in Chapter Five, and data on the performance of various tubes suitable for the purpose are given in the accompanying tables. Once the requisite audio power output has been determined and a pair of tubes capable of giving that output selected, an output transformer should be secured which will permit matching the rated modulator load impedance to the modulating impedance of the r.f. amplifier. Similarly, a driver transformer should be selected which will properly couple the driver stage to the Class-B grids.

The plate power supply for the modulator should have good voltage regulation and must be well filtered. It is particularly important, in the case of a tetrode Class-B stage, that the screen-voltage power-supply source have excellent regulation, to prevent distortion. The screen voltage should be set as exactly as possi-

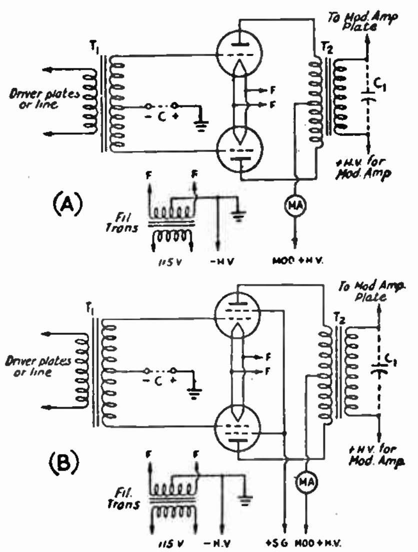


Fig. 1413 — Class-B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

ble to the recommended value for the tube.

In estimating the output of the modulator, it should be remembered that the figures given in the tables are for the tube output only, and do not include output-transformer losses. The efficiency of the output transformer will vary with its construction, and may be assumed to be in the vicinity of 80 per cent for the less expensive units and somewhat higher for higher-priced transformers. To be adequate for modulating the transmitter, therefore, the modulator should have a theoretical power capability about 25 per cent greater than the actual power needed for modulation.

The input transformer, T_1, may couple directly between the driver tube and the modulator grids or may be designed to work from a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the input to the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator, since the second transformer not only introduces additional losses but also further impairs the voltage regulation.

The bias source for the modulator must have very low resistance. Batteries are the most suitable source. In cases where the voltage values are correct, regulator tubes such as the VR75-30, VR105-30, etc., may be connected across a tap on an a.c. bias supply to hold the bias voltage steady under grid-current conditions. Generally, however, zero-bias modulator tubes are preferable, not only because no bias supply is required but also because the loading on the driver stage is less variable and consequently distortion in the driver is reduced.

Condenser C_1 in these diagrams will give a "tone-control" effect and filter out high-frequency side-bands (splatter) caused by distortion in the modulator or preceding speech-amplifier stages. Values in the neighborhood of 0.002 to 0.005 μfd. are suitable. Its voltage rating should be adequate for the peak voltage across the transformer secondary. The plate by-pass condenser in the modulated amplifier will serve the same purpose.

The photographs illustrate different types of construction which may be used for Class-B modulators. The actual placement of parts in filling the requirements of any given unit is not critical.

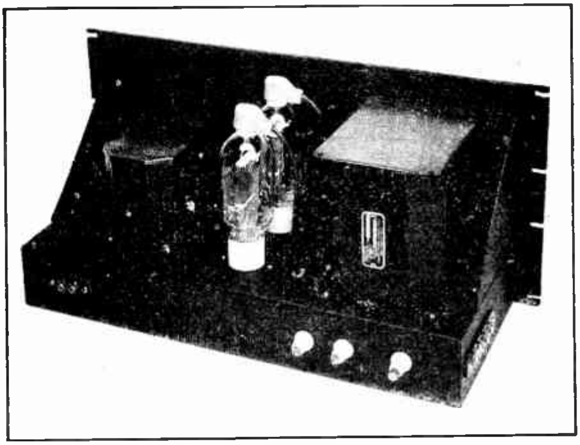


Fig. 1414 — A conventional chassis arrangement for low- and medium-power Class-B modulator stages. The mechanical layout in general follows the typical circuit diagrams given in Fig. 1313.

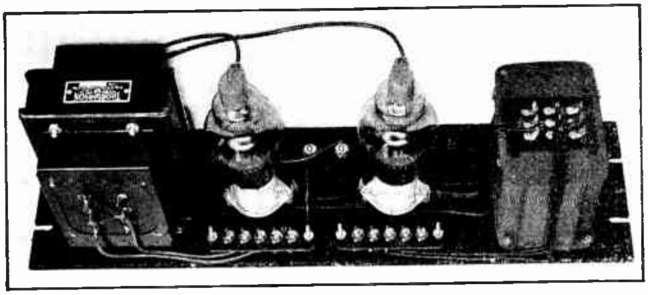


Fig. 1415 — Chassis-less construction for a low-power Class-B modulator. Small tubes and transformers capable of an audio output of the order of 100 watts can be mounted directly on the panel, eliminating the chassis.

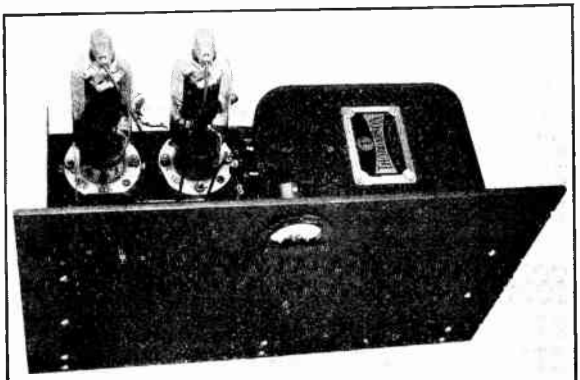


Fig. 1416 — A chassis arrangement for a higher-power Class-B modulator. This unit has the filament transformer for the tubes mounted on the chassis. Where the input transformer is included with the speech amplifier, less chassis space will be needed. The tubes are placed near the rear, where the ventilation is good. The plate milliammeter is provided with a small plate over the adjusting screw, to prevent touching the screw accidentally. A Presdwood panel was used for this modulator; with a metal panel, the meter should be mounted behind glass on a well-insulated mount (the meter insulation is not intended for voltages above a few hundred) or connected in the filament center-tap rather than in the high-voltage lead.

Chapter Fifteen

V.H.F. Receivers

In essential principles, modern receiving equipment for the 28- and 56-Mc. bands does not differ from that used on lower frequencies. In view of the higher frequency there are, of course, certain constructional precautions which must be taken to insure good performance. The 28-Mc. band serves as the meeting ground between those ordinarily termed "communications frequencies" and the very-highs, and it will be found that most of the receivers described in Chapter Twelve are capable of working on 28 Mc. In this chapter are described receivers and converters capable of good performance on 56 Mc. and higher.

Federal regulations require that transmitters working on all frequencies below 60 Mc. must meet similar requirements respecting stability of frequency and, when amplitude modulation is used, freedom from frequency modulation. It is thus possible to use receivers for 56-Mc. a.m. reception having the same selectivity as those designed for the lower frequencies. This order of selectivity is not only possible but desirable, since it makes possible a considerable increase in the number of transmitters which can work in the band without interference, as compared to broad-band receivers. Also, high selectivity greatly improves the signal-to-noise ratio, both in the receiver itself and in the response to external noise. This means that the effective sensitivity of the receiver can be considerably higher than is possible with non-selective receivers. Receivers for f.m. signals usually are designed with less selectivity so that they can accommodate the full swing of the transmitter, but, at least for 28- and 56-Mc. f.m. reception, the h.f. oscillator should be as stable as in a narrow-band a.m. receiver.

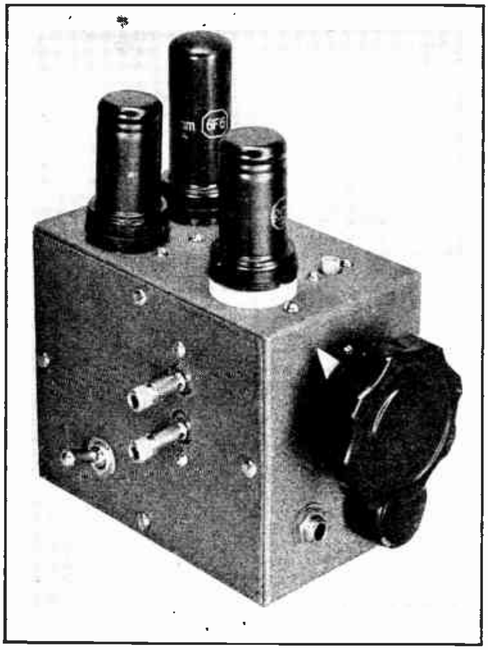


Fig. 1501 — The compact 112-Mc. receiver is built in a 3 × 4 × 5-inch metal box. Note the detector trimming condenser adjustment to the right of the 6J5 detector (front tube). The tuning control, headphone jack and regeneration control are on the front panel, the on-off switch and antenna binding posts on the side.

The superheterodyne type of receiver is used almost universally on frequencies below 60 Mc., because it is the only type of receiver that fulfills the above requirements for stability. A superheterodyne for a.m. reception and one for f.m. reception differ only in the i.f. amplifier and second detector, so the "converter" or high-frequency portion of the superheterodyne can be used for either a.m. or f.m. reception. Although superheterodynes can be built for 112-Mc. reception, the superregenerative type of receiver is much more widely used. The superregenerative receiver has the advantage of low cost and good sensitivity, although its selectivity does not compare with the superheterodyne type of receiver.

A superheterodyne receiver for 56-Mc. work should use a fairly high intermediate frequency so that image response and oscillator "pulling" will be reduced. At 56 Mc., for example, a difference between signal and image frequencies of 900 kc. (the difference when the i.f. is 450 kc.) is a very small percentage of the signal frequency, consequently the response of the r.f. circuits to the image frequency is very nearly as great as to the desired signal frequency. To get discrimination against the image equivalent to that obtained at 3.5 Mc. with a 450-kc. i.f. would require for 56 Mc. an i.f. 16 times as high, or about 7 Mc. if the circuit *Q*s were the same in both cases. However, the *Q* of a tuned circuit at 56 Mc. is not as high as at the lower frequencies, chiefly because the tube loading is considerably greater. As a result, still higher i.f.s are desirable, and a practical compromise is reached at about 10 Mc.

Since high selectivity cannot be obtained with a reasonable number of circuits at 10 Mc., the double superheterodyne principle is commonly employed. The 10-Mc. frequency is changed to an i.f. of the order of 450 kc. by a second oscillator-mixer combination. Thus the receiver has two intermediate frequencies, at both of which amplification takes place before the signal is finally rectified and changed to audio frequency.

Very few amateurs build complete 56-Mc. superhet receivers along these lines. General practice is to use a conventional superhet receiver to handle the 10-Mc. output of a simple frequency-converter. Thus a regular communications-type receiver — or even an all-wave broadcast receiver — can be used with excellent effect on 56 Mc. with the addition of a relatively simple and inexpensive converter unit. For those amateurs who have communications receivers, the construction of a good superheterodyne for 56 Mc. becomes a relatively simple matter.

From a practical aspect, superregenerative receivers may be divided into two general types. In the first the quenching voltage is developed by the detector tube itself — so-called "self-quenched" detectors. In the second, a separate oscillator tube is used to generate the quench voltage. The self-quenched receivers have found wide favor in amateur work. The simpler types are particularly suited for portable equipment where the apparatus must be kept as simple as possible. Many amateurs have "pet" circuits which are claimed to be superior to all others, but the probability is that the arrangement of their particular circuit has led to the use of correct operating conditions. Time spent in minor adjustment of values will result in a smooth-working receiver free from howling and irregular performance and is well worth the effort.

A Simple Superregenerative Receiver

One type of simple superregenerative receiver is pictured in Figs. 1501 to 1504. As shown in the wiring diagram, Fig. 1504, a 6J5 superregenerative detector is followed by resistance-coupled 6J5 and 6F6 audio stages. The circuit is quite conventional except for the use of inductive tuning in the detector and the use of resistance coupling throughout.

The receiver is built in a 3 × 4 × 5-inch metal box, with a 3 × 4-inch face serving as the panel. The panel controls are the tuning knob and the regeneration control, and the headphone jack is also mounted on the panel. The power cable plug is mounted at the rear of the box, as are the speaker terminals. The on-off switch and the antenna terminals are mounted on the left-hand side of the box.

The detector trimmer condenser, C_1, is fastened to the upper face of the box and can be adjusted from the top of the receiver. The quench-frequency r.f. choke, RFC_2, is supported off the under side of the upper face of the box by a long screw, with a brass sleeve over the screw furnishing sufficient spacing from the box. The r.f. choke is essential because the resistance-coupled amplifiers show but slight attenuation of the quench frequency, and the quench voltage will overload the output audio tube at rather low signal levels. When transformer coupling is used between the detector and first audio stage the transformer keeps most of the quench voltage out of the following stages and consequently the quench-frequency choke is not always necessary.

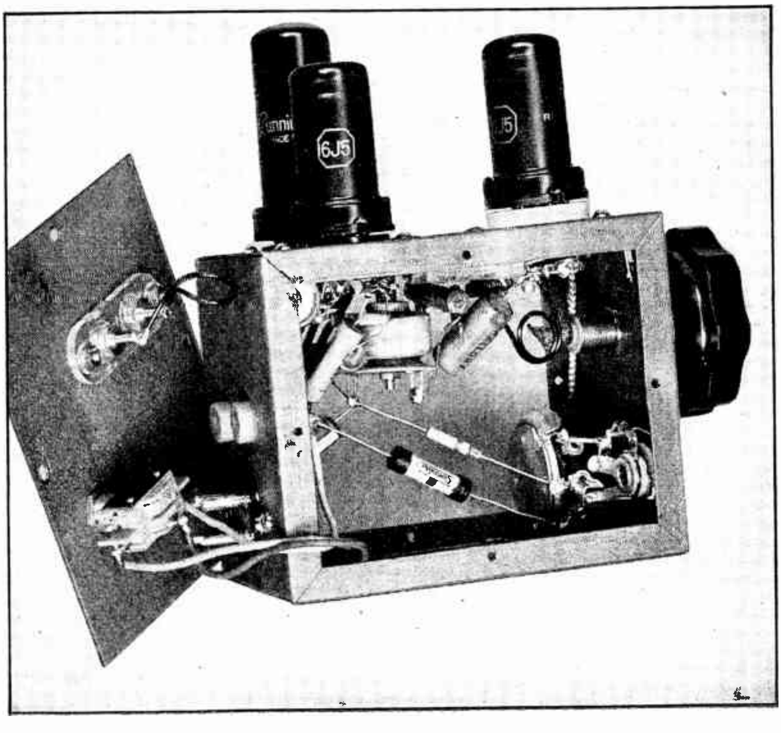


Fig. 1502 — Inside the small receiver, this left-hand view shows how the tuning-loop assembly and the send-receive switch are mounted on one of the side panels, together with the placement of some of the parts not visible in the other views. The power-supply plug and the loudspeaker binding posts may be seen at the rear of the chassis.

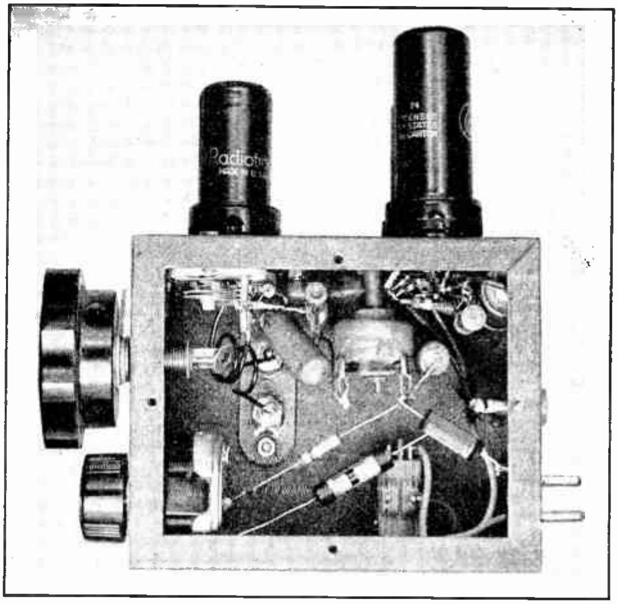


Fig. 1503 — A view from the right-hand side of the compact receiver.

The plate lead of the tube socket is connected to the rotor of the trimmer condenser by means of a short length of wire, and the coil L_1 is connected to the center of this wire and to the stator connection of the condenser. A length of ¼-inch shaft pushed through the shaft bearing will serve as a guide in soldering the coil in place, and the axis of the coil should make an angle of 45 degrees with the shaft.

The inductive tuning loop is a small copper washer cemented to the end of a ¼-inch shaft of insulating material (Lucite or bakelite). The end of the shaft is cut at an angle of 45 degrees to mount the washer at 45 degrees with respect to the axis of the shaft and, consequently, 180-degree rotation of the shaft turns the copper washer from a position coaxial with the coil to one at right angles to it. The copper washer, acting as a single shorted turn, decreases the effective inductance of the coil as it becomes more closely coupled and consequently tunes the system. The copper washer is made by drilling a ⅛-inch hole in a small piece of sheet copper and then cutting around the hole to form a washer of 7/16-inch outside diameter. The washer is fastened to the angled face of the shaft by Duco cement. Because the copper washer is larger than the shaft, the shaft must be pushed through the panel bearing from the inside of the box, but this can be done easily by loosening the panel bearing while sliding the shaft through. A fiber washer should be placed on the shaft before it is pushed through the panel bearing, and later cemented to the shaft to serve as a collar to prevent the shaft's pulling through the bearing.

The wiring of the unit requires only brief mention. A soldering lug at each socket furnishes a convenient ground for the components of that stage. All condensers and resistors are mounted by fastening directly to the sockets and other terminals, with the exception of the coupling condenser, C_6, one side of which must be run down to the headphone jack through an extra length of wire. The wires running to the toggle switch should be made of extra-length flexible wire so that the side of the box can be removed without unsoldering the wires to the switch. All wiring should be completed before L_1 and L_2 are put in place.

The detector coil is made by winding the wire around a ½-inch diameter drill or dowel as a former. The coil is then removed and the ends trimmed and bent until the coil can be soldered in place in proper alignment with the panel bushing used to support the tuning loop shaft.

It is easier to check the performance of the receiver before the tuning loop is added, and with the large trimmer condenser used there should be no difficulty in finding the 112-Mc. band.

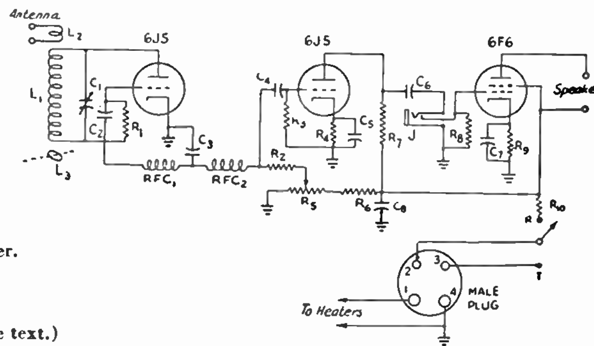


Fig. 1504 — Circuit diagram of the compact 112-Mc. superregenerative receiver.

C_1 — 25-μμfd. air trimmer (Hammarlund APC-25).
C_2 — 50-μμfd. midget mica.
C_3, C_4, C_6 — 0.01-μfd. 600-volt paper.
C_5, C_7 — 10-μfd. 25-volt electrolytic.
C_8 — 8-μfd. 450-volt electrolytic.
R_1 — 5 megohms, ½ watt.
R_2 — 25,000 ohms, ½ watt.
R_3 — 0.25 megohms, ½ watt.
R_4 — 1500 ohms, ½ watt.
R_5 — 50,000-ohms wire-wound potentiometer.
R_6, R_7 — 50,000 ohms, 1 watt.
R_8 — 0.1 megohms, ½ watt.
R_9 — 500 ohms, 1 watt.
R_{10} — 2000 ohms, 10-watt wire-wound. (See text.)
J — Closed-circuit jack.
S_1 — S.p.d.t. toggle.
L_1 — 1⅔ turns No. 14 enameled, ½-inch inside diameter, spaced ⅔ wire diameter.
L_2 — ¾ turn No. 14 e., ½-inch inside diameter.
L_3 — Tuning loop. See text.
RFC_1 — V.h.f.r.f. choke (Ohmite Z-1).
RFC_2 — 80-mh. r.f. choke (Meissner 19-2709).

The trimmer will be set at about two-thirds capacity if the coil is right. The detector should go into the hiss condition when the regeneration control is advanced not more than two-thirds of its travel. It is well to try different values of capacity at C_3, using the one which allows the detector to be worked at the minimum setting of the regeneration control without by-passing too much of the audio.

When the receiver is working and the tuning loop installed, the tuning range of the loop can be adjusted by moving the shaft in the panel bearing so that the loop is nearer to or farther from the coil. Moving the loop closer will increase the tuning range. It will be found that the tuning rate is slow when the loop is at right angles to the coil and becomes faster as the loop and coil become more nearly coaxial. It is therefore advisable to set the band and bandspread so that the receiver tunes from about 111.5 to 119 Mc., since this will spread the band over the main portion of the dial. When the shaft position which gives proper bandspread has been found, the fiber washer can be fastened to the shaft with Duco cement. When this is dry, the dial or knob can be attached to the outside end of the shaft. Play of the shaft in the bearing can be cured by slipping two metal washers and a half-slice of rubber grommet on the shaft before the dial is slipped on. The dial set screw should be tightened when the shaft is being pushed out from the inside, and the spring of the rubber grommet will then hold the collar (fiber washer) tightly against the inside of the panel bearing. A paper scale can be glued to the box and the megacycle and half-megacycle points marked on it, for ease in spotting stations and convenient resetting.

The antenna coupling should be adjusted with the antenna connected, and it should be made as tight as is consistent with some reserve in the regeneration control to take care of low voltages and other variables.

A Superregenerative Receiver for 112 and 224 Mc.

The receiver shown in Figs. 1505, 1506 and 1507 has very good sensitivity on both 112 and 224 Mc., although it is not free from radiation as is a receiver with an r.f. stage. However, for the amateur who wishes to experiment on these two v.h.f. bands this receiver will permit good reception at a minimum of expense. There is nothing unusual about the circuit; it is the familiar type of self-quenched superregenerative detector, followed by two stages of audio amplification.

The receiver is built on a 7 × 7 × 2-inch chassis. The dial is mounted in the center of the panel and is connected to the tuning condenser by a flexible bakelite coupling. The condenser is mounted on a metal bracket, cut out in the shape of a U to clear the stator connections of the condenser.

The socket for the plug-in coils is made using contacts taken from an Amphenol 78-7P miniature tube socket. They are obtained by squeezing the socket in a vise until the bakelite cracks, after which they can be easily removed. One contact is soldered to each of the tuning condenser connections and a third is soldered to a lug supported by one of the extra holes in the Isolantite base of the tuning condenser. In mounting the contacts they must all be at the same height, so that the plug-in coil will seat well on them. The band-set condenser, C_2, is mounted by soldering short strips of wire to the ends and then soldering these wires to the tuning condenser terminals.

The polystyrene tube socket for the 9002 is mounted on a metal bracket, which is placed close enough to the tuning condenser to allow a very short lead from the tuning condenser to the plate connection and just enough room between the rotor of the condenser and the grid connection of the tube for the grid condenser.

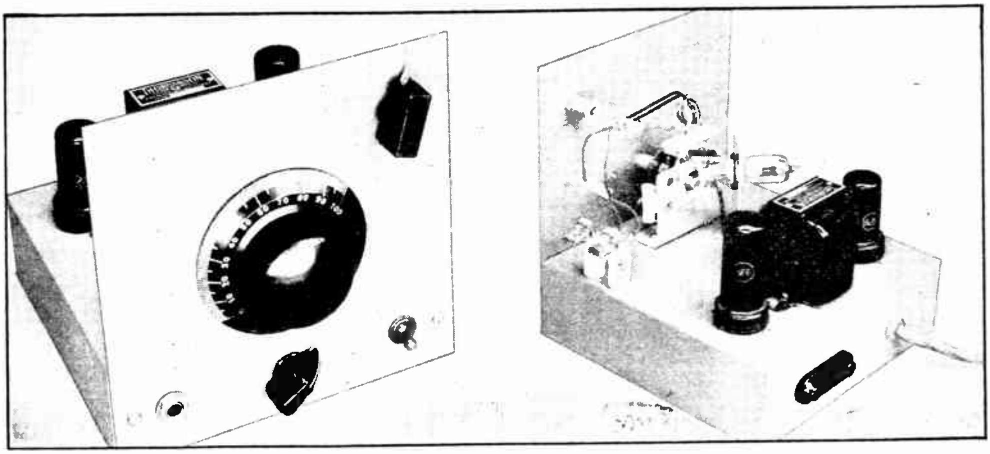


Fig. 1505 — Left — The panel of the two-band superregenerative receiver measures 7 inches square. The knob in the upper right-hand corner adjusts antenna coupling, while the knob below the tuning dial controls regeneration. *Right* — A rear view of the two-band superregenerative receiver, showing the variable antenna coupling and the placement of parts. Note the 224-Mc. coil in the foreground; the 112-Mc. coil is in the coil socket in this view.

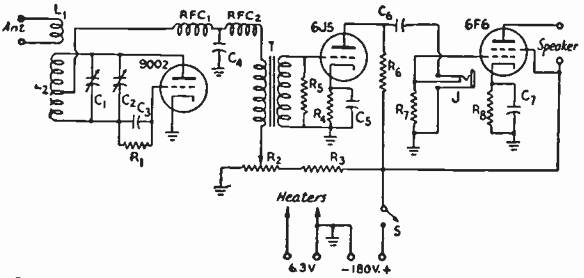


Fig. 1506 — Wiring diagram of the v.h.f. superregenerative receiver for 112 and 224 Mc.

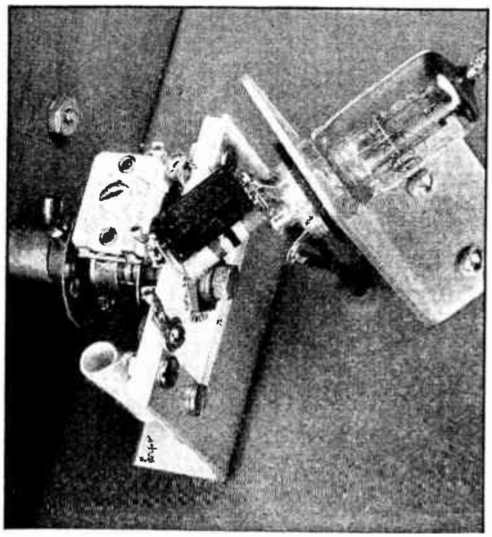
C_1 — Two-plate variable (National UM-15, 4 plates removed).
C_2 — 3-30-$\mu\mu$fd. mica trimmer.
C_3 — 50-$\mu\mu$fd. mica.
C_4 — 0.003-μfd. mica.
C_5, C_7 — 10-μfd. 25-volt electrolytic.
C_6 — 0.01-μfd. 400-volt paper.
R_1 — 10 megohms, ½ watt.
R_2 — 50,000-ohm wire-wound potentiometer.
R_3 — 0.1 megohm, 1 watt.
R_4 — 2500 ohms, ½ watt.
R_5, R_6, R_7 — 0.1 megohm, ½ watt.
R_8 — 500 ohms, 1 watt.
J — Closed-circuit jack.
S — S.p.s.t. toggle switch.
T_1 — Single plate to single grid audio transformer (Thordarson T-57A36).
L_1 — 1 turn No. 14 e., ⅜-inch inside diameter.
L_2 — 112 Mc.: 3 turns No. 18 e., ½-inch diameter, ¼-inch long. Tapped 1¼ turns from plate end.
224 Mc.: 2 turns No. 18 e., ¼-inch diameter, spaced over ½ inch. Tapped at center.
RFC_1 — 25 turns No. 24 d.c.c. close-wound, ¼-inch diameter.
RFC_2 — 8 mh. r.f. choke.

Heater and cathode leads are brought to the underside of the chassis through a rubber grommet.

The variable antenna coupling coil, L_1, is mounted on a polystyrene rod supported by a shaft bearing. The rod is prevented from moving axially in the bearing by cementing a fiber washer to the shaft and tightening the knob on the other side so that the shaft does not move too freely. The antenna coupling loop should be adjusted so that, as it is rotated, it will just clear the coils when they are plugged into the socket.

The coils are mounted on small strips of ⅛ inch polystyrene (Millen QuartzQ) as bases, which have three small holes drilled in them corresponding exactly to the tops of the coil sockets. Each coil is cemented to the strip with Duco cement at the points where the wire passes through the base. The No. 18 wire used for the coils will fit snugly in the sockets if the contacts are pinched slightly. A coil socket of this type allows very short leads to be used, and is about the only thing practical until some manufacturer brings out a commercial product along these lines. The coils are trimmed to the bands by spreading or squeezing the turns slightly by the procedure previously described. However, in this case the band-set condenser gives some further range of adjustment. In the receiver as described, it is screwed down fairly tightly for the 112-Mc. band and loosened about four revolutions for 224 Mc. If there are no good marker stations available by which frequency can be checked, an absorption frequency meter or the Lecher wire system described in Chapter Twenty may be used for spotting the bands.

Two factors which will be found to influence the sensitivity of the receiver are the value of C_4 and the degree of antenna coupling. It is rec-



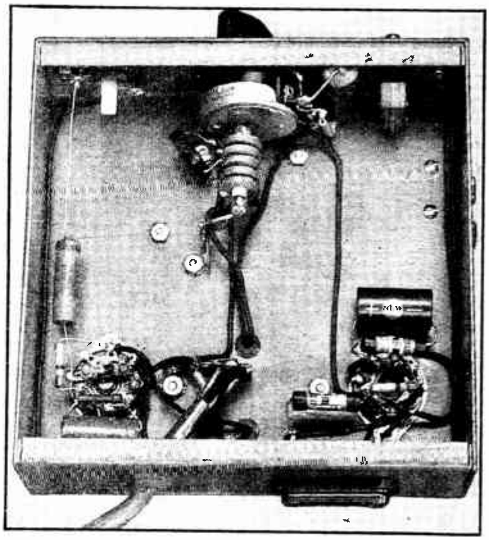


Fig. 1507 — *Left* — A close-up view of the tuning assembly, showing how the leads from the tuning condenser to the tube socket have been kept short and how the coil socket is mounted on the tuning condenser. Hidden by the grid condenser (the 50-$\mu\mu$fd. condenser so prominent in the picture), the plate terminal of the tube socket goes to a lug which has been added to the rotor of the tuning condenser. *Right* — The arrangement of parts under the chassis may be seen in this photograph. The 6J5 socket is at the left and the 6F6 socket is at the right, near the speaker terminals. The 8-mh. r.f. choke, seen just under the regeneration control at the top center, is supported by tie strips.

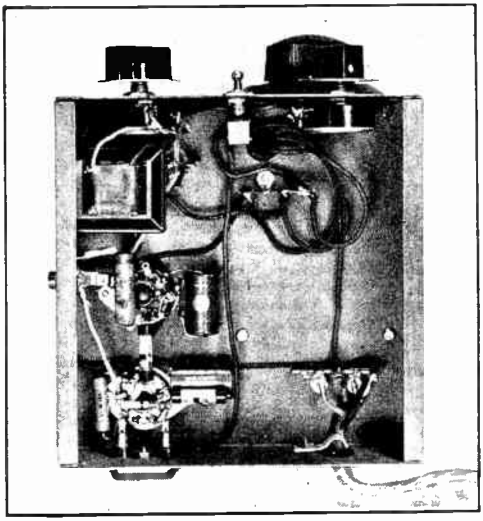


Fig. 1508 — *Left* — The 112-Mc. t.r.f. superregenerative receiver uses a 9001 in the r.f. stage, 9002 detector, 6J5 first audio, and 6F6 output stage. The knobs along the front are audio volume control (left) and the regeneration control. The rubber grommet on the side of the 3 × 4 × 5-inch box centers the screwdriver used for setting the detector band-set condenser; a similar one is provided on the other side for the r.f. band-set adjustment. Note the 'phone jack on the side; the speaker terminals are located at the rear. *Right* — A view under the chassis of the t.r.f. receiver shows the audio transformer and the arrangement of some of the other components. The three wires coming through the chassis to the right of the "B" + switch are the leads from the r.f. section of the receiver.

ommended that values of C_4 from 0.001 to 0.005 μfd. be tried. The antenna coupling will, of course, vary greatly with the setting of L_1 and with the type of antenna used, and it is well worth while to tune the antenna circuit and then vary the coupling with the panel control. Tight coupling usually will give better results than loose coupling and the coupling can be increased almost up to the point where it is no longer possible to make the detector oscillate, with no ill effects except increased radiation and possible QRM for other receivers in the vicinity.

No audio volume control was included in this receiver because the parts were held down to a minimum, but one could easily be added in place of the fixed grid resistor, R_7. In this receiver, the value of R_7 was adjusted until normal loudspeaker output was obtained; this value may be varied to meet any particular requirements.

◖ A T.R.F. Superregenerative Receiver

The receiver shown in Figs. 1508, 1509, 1510 and 1511 is practically identical to that described in the foregoing section, with the exception that a stage of tuned r.f. amplification and an audio gain control have been added. The 9001 tube used for the r.f. amplifier gives some slight gain, freedom from antenna effects, and — most important of all — prevents radiation from the receiver.

The arrangement of parts, as shown in the photographs, is convenient in that it results in a fully shielded receiver (except for the r.f. tubes) which is easy to work on. The r.f. unit can be demounted from the chassis and worked with separately; once adjusted, it can be replaced and left alone. The receiver is a one-band affair, but the only disadvantage in this is a lack of economy. The main chassis, which

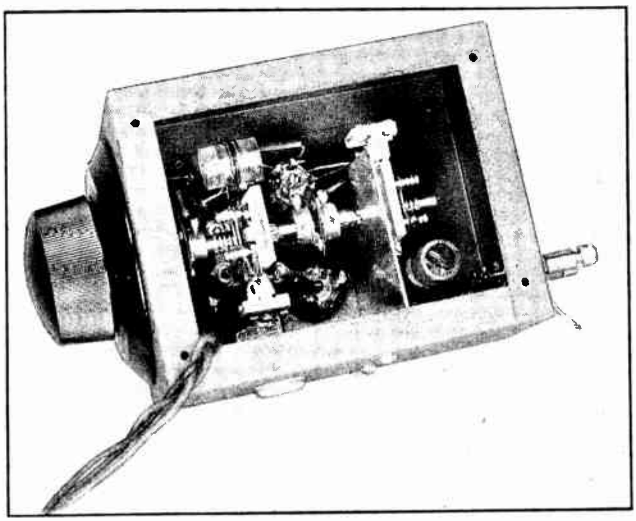


Fig. 1509 — The r.f. section of the 112-Mc. t.r.f. receiver removed from the chassis. The detector tuning condenser, C_3, is nearest the tuning dial and the detector socket is at the bottom of the picture. The interstage shield is fastened to the side of the box. The trimming loop for adjusting inductance can be seen on the r.f. coil, near the antenna posts at the right.

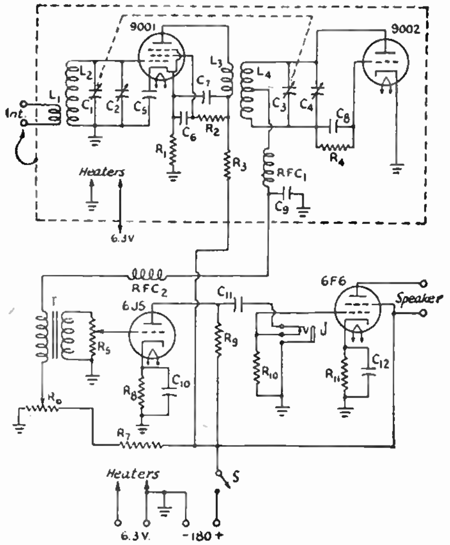


Fig. 1510 — Wiring diagram of the tuned r.f. superregenerative receiver for 112-Mc.

- C_1, C_3 — 2-plate midget variable (National UM-15 with 4 plates removed), ganged.
- C_2, C_4 — 3–30-μμfd. mica trimmer.
- C_5, C_6, C_7 — 500-μμfd. mica.
- C_8 — 50-μμfd. mica.
- C_9 — 0.003-μfd. mica.
- C_{10}, C_{12} — 10-μfd. 25-volt electrolytic.
- C_{11} — 0.01-μfd. 400-volt paper.
- R_1 — 200 ohms, ½ watt.
- R_2 — 0.25 megohm, ½ watt.
- R_3 — 10,000 ohms, ½ watt.
- R_4 — 10 megohms, ½ watt.
- R_5 — 0.5-megohm volume control.
- R_6 — 50,000-ohm wire-wound variable.
- R_7, R_9 — 0.1 megohm, 1 watt.
- R_8 — 2500 ohms, ½ watt.
- R_{10} — 0.5 megohm, ½ watt.
- R_{11} — 500 ohms, 1 watt.
- L_1 — 1⅔ turns No. 28 d.s.c. interwound between turns of L_2.
- L_2 — 2 turns No. 20 e., ¼-inch winding length. See text for trimming method.
- L_3 — 1½ turns No. 28 d.s.c. interwound between turns of L_4.
- L_4 — 2¼ turns No. 20 e., ¼-inch winding length. Tapped ½ turn from plate end. See text on how to trim.
- (L_1-L_2 and L_3-L_4 on National PRE-1 forms.)
- RFC_1 — V.h.f. r.f. choke (Ohmite Z-1).
- RFC_2 — Low-frequency choke (National OSR with windings in series, "B +" and "Gnd" connected together).
- J — Closed-circuit jack.
- S — S.p.s.t. toggle switch.
- T_1 — Single plate to single grid audio transformer (Thordarson T-13A34).

measures 7 × 7 × 2 inches, contains the audio components and the volume and regeneration controls. The r.f. portion is housed in a 3 × 4 × 5-inch box with everything but the dial and the antenna terminals mounted on the removable cover, enabling the builder to get at the parts easily. Only three leads are brought down from this box to the main chassis, and these are left long enough so that they do not need to be unsoldered when the box is removed from the chassis. A shield mounted on the side of the box helps to prevent coupling between the r.f. and detector coils. Holes on either side of the box allow the trimmer condensers to be adjusted when the receiver has been finally assembled.

As can be seen from the close-up view of the r.f. portion, the two tuning condensers and the two sockets are mounted on the removable top of the box and support all of the components. The trimmer condensers are soldered directly to the tuning condenser terminals and the coils are self-supported by their leads. A tie strip takes the leads that run out of the box and also serves as a convenient point to fasten RFC_1, C_9 and some of the other associated resistors and condensers. The leads are not quite as short in this arrangement as they are in the receiver of Fig. 1505, but that makes no practical difference because this receiver is built only for 112 Mc. and does not have to tune down to 230 Mc.

The coils are wound on small polystyrene forms. It is suggested that the No. 20 wire secondary coils be wound first. The plate tap for L_4 should be soldered, and the coils can then be doped. When the dope has hardened, the fine-wire coils can be more easily wound in between the turns and fastened with dope. The No. 20 wire leads run through holes in the forms, while dope only is used to keep the fine-wire coils secure. This has the advantage that the fine-wire coils can be trimmed by "peeling off" a small fraction of a turn at a time. The larger coils are trimmed by bringing the last half-turn back through the *inside* of the coil.

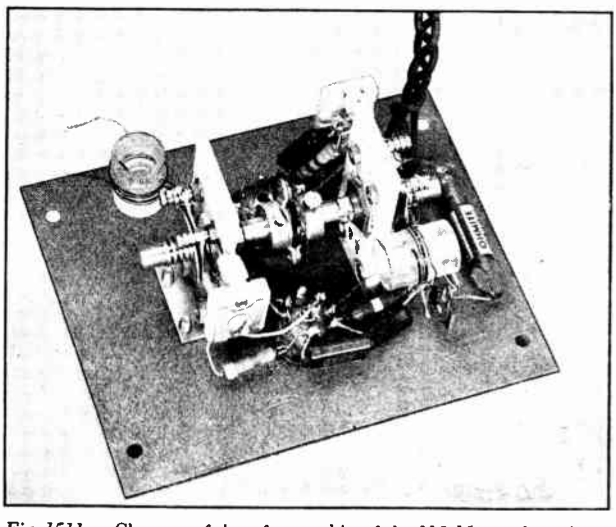


Fig. 1511 — Close-up of the r.f. assembly of the 112-Mc. t.r.f. receiver, showing arrangement of parts. The band-set condensers are mounted on the tuning condensers. The ends of the antenna coil (upper left) are soldered to the antenna posts after the assembly is mounted in the box.

By moving this half-turn around, the inductance of the coil can be adjusted over a range wide enough to allow the detector and r.f. circuits to track well over the whole band. This method of inductance trimming is described in detail in § 7-7.

The r.f. stage is trimmed by adjusting its trimmer condenser to the point where the regeneration control has to be set at a maximum. Either side of this point the control does not have to be advanced as far, indicating that the r.f. stage is not in resonance. When the r.f. and detector circuits are tracking properly, it will not be necessary to change the setting of the regeneration control more than 45° or so over the entire range. The bandspread can be increased by using less inductance and more trimmer capacity. With the coil specifications given, the band covers about 75 dial divisions.

A two-wire line from the antenna usually will prove best. It should be tried with one side grounded or not, to see which gives the better coupling. In one instance where a single-wire antenna was used some instability of the r.f. amplifier was traced to the antenna wire running too close to the detector tube, and it is recommended that the antenna wire or wires be run away in such a fashion that there is no chance for coupling of this type.

112-Mc. Superregenerative Superheterodyne Receiver

The ordinary 112-Mc. superregenerative receiver is not very selective, and in localities where several networks are operating within a relatively narrow frequency band it may be necessary to go to a more selective type of receiver to minimize interference. In addition, the radiation from a superregenerator can cause an annoying type of interference when station locations are fairly close together. Both these disadvantages can be overcome by using a superheterodyne. The well-known advantages of the superregenerator — simplicity, sensitivity and economy of tubes and components — can in large part be retained by using a superregenerative detector as the i.f. system of the superheterodyne. Since the intermediate frequency will be considerably lower than the signal frequency the selectivity will be increased in proportion, while the receiver as a whole is increased in complexity only by the addition of two tubes and relatively simple accompanying circuits.

A superheterodyne of this type is shown in Figs. 1512 to 1515. It uses commonly-available triodes in the r.f. and i.f. sections, and equivalent types such as the 6J5, 6C5, etc., can be substituted with little difference in performance. The i.f. in the receiver shown is approximately 26 Mc., a frequency low enough to give a worth-while increase in selectivity while still high enough to give good superregenerative operation.

A complete circuit diagram of the receiver is given in Fig. 1514. The audio section consists of a 6C5 and 6V6, and is conventional except for the fact that the headphone jack is shunted across the 6V6 grid circuit rather than being arranged to disconnect the audio signal from the speaker tube when the 'phones are plugged in. This allows others to hear what is going on while the operator excludes outside noise by wearing headphones.

The mixer input circuit consists of the tuned circuit, L_2C_1, coupled to the antenna or feeders by the one-turn coil, L_1. Since the grid-cathode portion of the tube acts as a diode rectifier, it takes energy from the tuned circuit; consequently a rough impedance match between the circuit and tube is brought about by tapping down on the coil. Voltage from the oscillator is capacity-coupled to the mixer through C_8.

The mixer plate circuit is tuned to the intermediate frequency, 26 Mc., by a 30-$\mu\mu$fd. mica trimmer, C_3, across the coil L_4. To prevent short-circuiting the d.c. plate voltage a blocking condenser, C_4, is incorporated in the tuned circuit. The plate coil, L_4, is inductively coupled to the superregenerative detector coil, L_5. For optimum sensitivity the mixer plate voltage should be in the vicinity of 20 to 25 volts. This voltage is obtained from a voltage divider, R_4R_5.

The local oscillator should have a fairly high-C tank circuit to insure stability, and the grid-leak resistance should be low enough to prevent squegging. A value of 10,000 ohms is satisfactory with the oscillator operating at approximately 100 volts on the plate.

The superregenerative detector circuit is similar to those previously shown. For optimum results different values for the grid leak, R_2, and the plate by-pass condenser, C_7, should be tried.

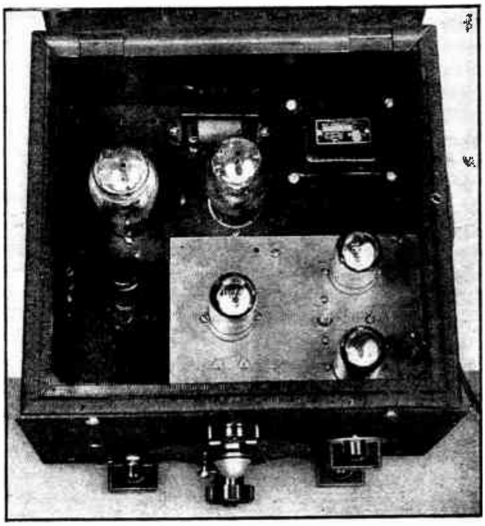


Fig. 1512 — A superregenerative superheterodyne receiver, designed for greater selectivity than can be obtained from the ordinary superregenerator. The audio section is along the left edge of the chassis and the self-contained power supply is at the upper right. The cabinet is approximately 9½ inches square and 7 inches deep.

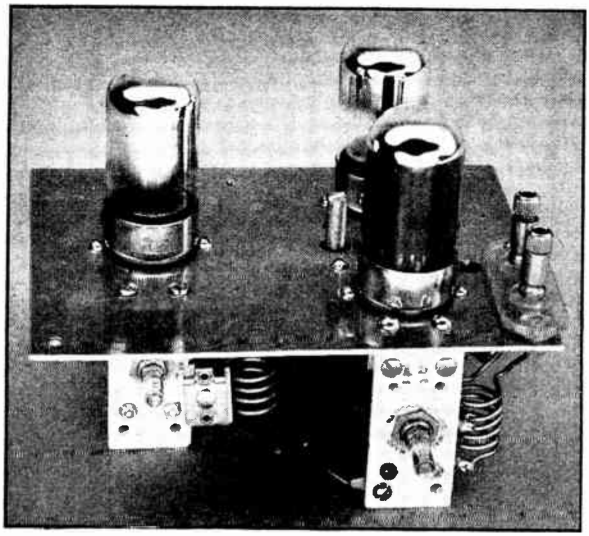


Fig. 1513 — The r.f. unit of the superregenerative superheterodyne, consisting of the mixer, high-frequency oscillator and superregenerative detector, is built on a small 4 × 6-inch metal subchassis. The hole between the mixer tube and antenna binding post assembly is for screwdriver adjustment of the i.f. transformer primary.

The r.f. section of the outfit is the only part which requires particular care in construction; as shown in the photographs, this is assembled on a small subchassis and can be removed as a unit from the receiver. The subchassis is a 4 × 6-inch piece of aluminum and mounts to the main chassis by means of three brass-rod pillars a little over 2¼ inches long. Viewed from the top, the oscillator is to the left, the mixer is in the right foreground, and the i.f tube is at the rear right. The oscillator is mounted somewhat back from the front to allow room for an insulated coupling and extension shaft, necessary because both ends of the tuning condenser are "hot." The mixer tuning condenser is mounted at the front of the subchassis so that its shaft can extend through the front panel. The tank coils in both the oscillator and mixer circuits are mounted directly on the condenser terminals.

The i.f. coils L_4 and L_5 are wound on a polystyrene form mounted between the mixer and detector as shown. The detector condenser, C_5, is air-tuned and is mounted on the subchassis. Both sides must be insulated from ground. The primary tuning condenser, C_3, is a mica trimmer mounted so that its adjusting screw is accessible through a hole in the subchassis. The movable plate should be connected to ground so that body capacity can be avoided in making adjustments. The plate choke, RFC_2, is supported at one end by the tap on L_5 and at the other by C_7, the grounded terminal of which is soldered to a lug fastened under one of the socket-mounting screws.

The sensitivity of the receiver depends a great deal on the amount of oscillator voltage injected into the mixer grid circuit. The sensitivity will be poor if the oscillator voltage is too small, but once enough is secured a further increase has relatively little effect. The "con-

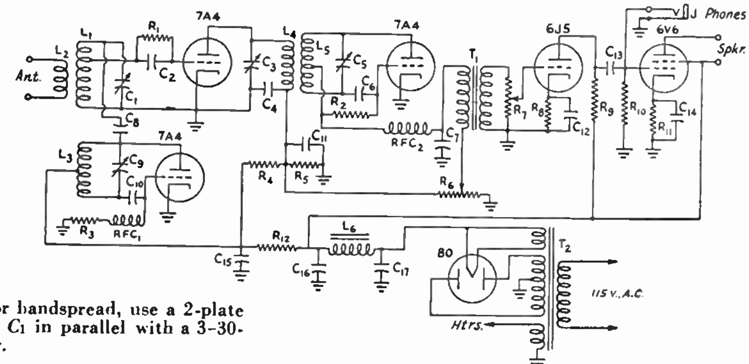


Fig. 1514 — Circuit diagram of the superregenerative superheterodyne.

C_1 — 3-plate midget, approximately 5 μμfd. (National UM-15 cut down.)
C_2 — 50-μμfd. mica.
C_3 — 3-30-μμfd. mica trimmer.
C_4 — 500-μμfd. mica.
C_5 — 35-μμfd. variable (Millen 20035).
C_6 — 100-μμfd. mica.
C_7 — 0.002-μfd. mica.
C_8 — See text.
C_9 — 15-μμfd. variable. For bandspread, use a 2-plate variable similar to C_1 in parallel with a 3-30-μμfd. mica trimmer.
C_{10} — 100-μμfd. mica.
C_{11} — 0.01-μfd. 400-volt paper.
C_{12}, C_{14} — 25-μfd. 25-volt electrolytic.
C_{13} — 0.1-μfd. 400-volt paper.
C_{15}, C_{16}, C_{17} — 8-μfd. 450-volt electrolytic.
R_1 — 2 megohms, ½ watt.
R_2 — 5 megohms, ½ watt.
R_3 — 10,000 ohms, ½ watt.
R_4, R_5 — 50,000 ohms, 1 watt.
R_6 — 50,000-ohm volume control.
R_7 — 0.5-megohm volume control.
R_8 — 2000 ohms, 1 watt.
R_9 — 50,000 ohms, 1 watt.
R_{10} — 0.5 megohm, ½ watt.
R_{11} — 750 ohms, 1 watt.
R_{12} — 15,000 ohms, 10 watt.
L_1 — 5 turns No. 14, ½-inch diameter, turns spaced slightly more than diameter of wire; grid tap 3½ turns from ground end.
L_2 — 1 turn, same diameter as L_1.
L_3 — 4 turns No. 12, ½-inch diameter, length ½ inch.
L_4 — 12 turns No. 24 d.s.c. on ½-inch form, close-wound.
L_5 — Same as L_4 but tapped at center, spaced ½ inch from L_4.
L_6 — 10-henry 50-ma. filter choke.
J — Open-circuit jack.
T_1 — Interstage audio transformer.
T_2 — Power transformer, 250 volts at 50-60 ma., with rectifier filament and 6.3-volt heater windings.
RFC_1 — 1½-inch winding of No. 30 s.c.c. on ¼-inch diameter form.
RFC_2 — 2.5-mh. r.f. choke.

denser" actually used is formed by fastening a machine screw in one of the mounting holes in the Isolantite frame of the tuning condenser, C_1, so that it is fairly close to one stator mounting post of the condenser. The small capacity thus provided (C_8) is ample for coupling when the oscillator is operated with about 100 volts on the plate.

In making preliminary adjustments, the first operation is to get the superregenerative detector on frequency, which is best done by listening for its "hash" output on a regular communications receiver tuned to the approximate intermediate frequency of 26 Mc. The mixer and detector should be in their sockets but the oscillator should either be out or its plate voltage disconnected. Set the detector in operation, adjusting R_6 until superregeneration occurs, and tune C_5 to put the detector on frequency. Then tune C_3 to resonance, which will be indicated by absorption of energy from the detector circuit so that R_6 has to be advanced to maintain superregeneration. Since tuning one circuit affects the tuning of the other it will be necessary to go through this process several times to maintain the desired frequency. If the primary circuit extracts too much energy, the superregenerative hiss will lose its smooth character and the sensitivity will be poor. This condition can be corrected either by loosening the coupling between the two coils or by detuning the primary sufficiently to obtain satisfactory operation.

To adjust the oscillator, set up a low-power 112-Mc. oscillator in the vicinity and then tune C_9 until a signal is heard. If the bandspread system with a mica padder is used, the movable plate on the padder should be fairly close to the fixed plate, but not squeezing the mica. The padding capacity required is of the order of 10 $\mu\mu$fd., depending upon the inductance of L_3 which will vary somewhat with the turn spacing. Some care must be used, because it is readily possible for harmonics of one or both of the oscillators to beat to produce a spurious i.f. signal. If several signals can be heard in adjusting the padder over its range, the strongest one should be chosen.

Once the signal is found, the mixer circuit should be adjusted to resonance. It is probable that the coil will have to be pruned or the turn spacing altered to bring the signal definitely on the tuning scale of C_1. The coil size also depends upon the position of the grid tap. By a process of cut-and-try the coil size and tap position which give maximum signal strength can eventually be determined.

In practice the tuning of the receiver is effectively single control, since C_1 may be set at about the center of the band and the circuit will be broad enough to give good response at both ends. The regeneration control is not affected by tuning of the r.f. circuits except when the primary condenser, C_3, is set very close to the critical point with L_4 and L_5 overcoupled. Under these conditions any small change either in loading or capacity can throw the detector into or out of oscillation, and since the mixer is a triode such changes can be reflected from the r.f. circuits through the grid-plate capacity of the tube. Under optimum i.f. adjustments this does not occur.

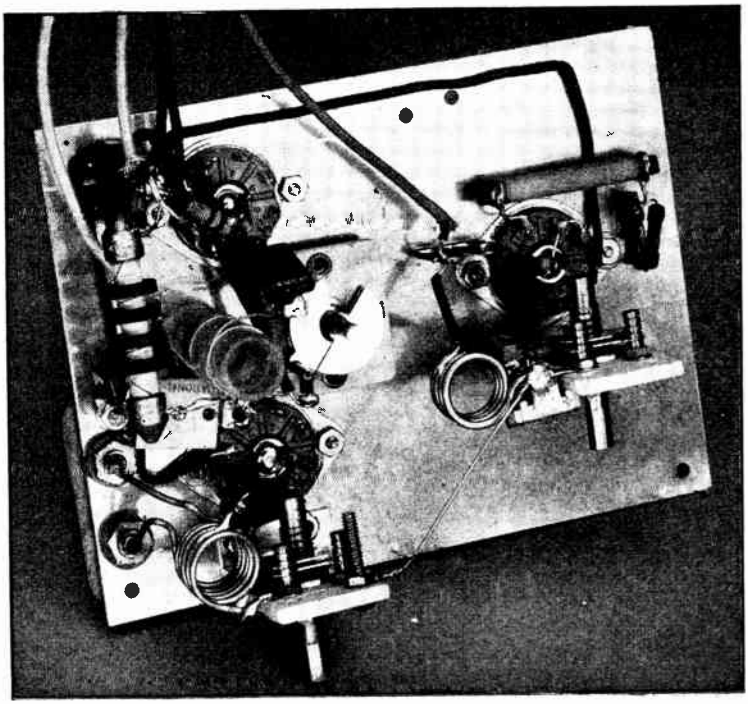


Fig. 1515 — A bottom view of the superheterodyne r.f. unit. The mixer circuit is at the lower left, with the detector above and oscillator to the right. Note the machine screw (on the mixer tuning condenser) which forms one "plate" of the oscillator-mixer coupling condenser. The mica padders used in the oscillator and i.f. circuits are convenient and inexpensive, but air padders would be preferable for long-time stability.

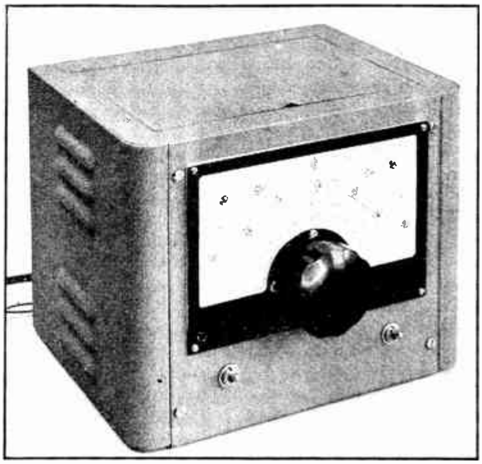


Fig. 1516 — This 2½- and 5-meter converter, complete with self-contained power supply, is mounted in an 8 × 8 × 10-inch cabinet. Plug-in coils give bandspread coverage of the 56- and 112-Mc. amateur bands.

V.H.F. Converters

For the amateur who already possesses a high-frequency communications receiver or even a reasonably good all-wave broadcast receiver of adequate sensitivity and stability which is capable of tuning to either 5 or 10 Mc. there is little or no necessity for building an elaborate separate v.f.h. receiver, particularly for operation on the 56-Mc. band. It is not only easier but often more satisfactory to build a v.h.f. converter which, in conjunction with the already existing receiver, will constitute a double superheterodyne (§7-8). This is particularly true if the receiver has controllable or broad-band selectivity to permit reception of the less stable signals on the higher frequency bands.

The output transformer for such a converter should be designed to tune to an i.f. of either 5 or 10 Mc. (the higher frequency being preferable for operation on bands above 56 Mc.), with a low-impedance secondary. The output signal from the converter is coupled through a low-impedance shielded line to the input circuit of the communications receiver, in much the same manner as link coupling is used between stages in a transmitter. The r.f. and mixer circuits of the receiver are tuned to the same frequency as the output transformer — 5 or 10 Mc. — which then becomes the first i.f. Thereafter the receiver dial remains untouched, all tuning being done with the converter. The output volume, however, is adjusted to a suitable level by means of the gain control on the receiver into which the converter is working.

A High-Performance Converter for 56 and 112 Mc.

The converter shown in Figs. 1516, 1517, 1518, 1519 and 1520 uses the new 9000-series "button" tubes, which are quite similar electrically to "acorn" tubes but are somewhat easier to handle. As may be seen from the diagram in Fig. 1518, a 9001 r.f. stage is transformer-coupled to a 9001 mixer. The h.f. oscillator is a 9002 and it is capacity-coupled to the mixer grid through C_{15}. The output circuit (C_{14}, C_{16} and L_7) is tuned to 10.2 Mc., approximately, although the converter could be made to work into some other i.f. with suitable changes in the output circuit and the oscillator coil, L_5L_6, constants.

As indicated in the diagram, the screen and plate by-pass condensers are returned to one cathode lead (the one to which the suppressor is connected) while the other lead is grounded through a condenser to serve as the grid return. In the mixer plate circuit a low-drift mica condenser, C_{14}, is connected directly from plate to cathode, to short-circuit the signal-frequency component in the plate circuit. This condenser is part of the i.f. tuned circuit, and its capacity must be taken into account in calculating the inductance required at L_7.

The mixer and r.f. tuned circuits are made as low-C as is possible under the circumstances; the use of plug-in coils unavoidably introduces some stray capacity that would not be present if the circuits were made to operate on one frequency only. The tuning condensers are cut

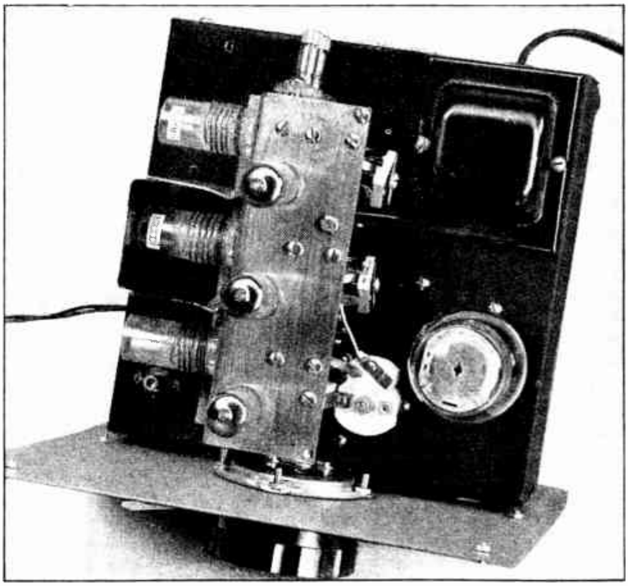


Fig. 1517— A top view of the converter, showing arrangement of tubes and coils. The shaft projecting through the main chassis at the lower left is the i.f. transformer tuning control. The power transformer is submounted so it does not interfere with adjustment of the r.f. trimmer.

down to two plates each, and have just about enough capacity range to cover the 56-Mc. band with a little to spare. The trimmers are mica units operated at nearly minimum capacity, so that the mica is a negligible factor in the operation of the condenser; for all practical purposes, the dielectric is purely air. The L/C ratio compares favorably with those commonly attained in acorn receivers.

The oscillator circuit is of the grid-tickler type, with the tuned tank in the plate circuit. The tuned circuit is made higher-C than the signal-frequency circuits to improve the stability, and as a consequence somewhat more tuning capacity is needed to cover the frequency range. The tuning condenser is a 15-μμfd. unit cut down to three plates and the trimmer is a 25-μμfd. air-dielectric unit. The oscillator and mixer circuits are coupled through a small homemade condenser, C_{15}, tailored to give suitable injection of oscillator voltage into the mixer grid circuit.

The oscillator is tuned to the low side of the signal frequency on both 56 and 112 Mc., to give slightly better oscillator stability. A VR105-30 voltage-regulator in the power supply adds further to the stability of the oscillator.

The power-supply part of the circuit needs no comment, except to explain that a separate filament transformer was used only because no suitable small plate transformer was available with a 6.3-volt heater winding.

The "chassis" on which the converter is assembled is a piece of sheet copper, somewhat less than 1/16 inch thick, 5½ inches long, and bent as shown in the photographs. The width on top is 1¾ inches, the height 2¼ inches, and the bottom lip, for fastening to the main chassis, is ¾-inch wide. The tubes are mounted on top near the bent edge, allowing just enough room to insert the socket mounting ring, and are 1¾ inches apart, center to center, with the r.f. tube 1⅜ inches in from the rear edge. The coil sockets are mounted on the side, ¾ of an inch down from the top, so that the connections between socket prongs and the tuning condenser terminals can be made without additional wires. The spacing is such that the lead from the stator connection on the condenser to the grid prong on the tube socket is only about ¼-inch long.

In building an assembly of this type, it is a practical necessity to do all the wiring before the tuning condensers are mounted. The inside view gives some idea of the arrangement of by-pass condensers; the chief consideration in placing them is to eliminate leads, insofar as possible. Each stage has its own ground point, which, in the case of the r.f. and mixer stages, is on the side of the chassis directly below the tube socket and the length of the cathode by-pass condenser away from it. The screws which hold the ground lugs in place are threaded into the copper, and on the outside also help support the vertical interstage shields. The oscillator ground is also on the side but close to the cathode pin, which is grounded directly; the plate by-pass condenser, C_{13}, is brought to the same point. In the other two stages the ground leads from the tuned circuits are ⅜-inch wide strips of thin copper, this being used in preference to wire to reduce the inductance.

For electrostatic shielding between the r.f. and mixer stages, two baffle plates are used. One small plate, not visible in the photograph, is fastened to the side of the chassis directly opposite the tube socket and is soldered to the shield cylinder in the center of the socket. It effectively shields the grid wiring from the plate circuit, and is about an inch square. Since

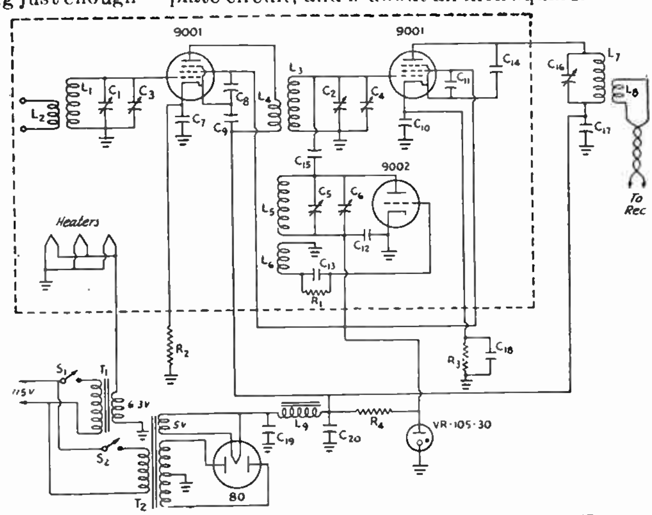


Fig. 1518 — Circuit diagram of the high-performance plug-in coil 56–112-Mc. converter using 9000 tubes.

C_1, C_2 — Approximately 5-μμfd. variable (National UM-15 cut down to 2 plates).
C_3, C_4 — 3–30-μμfd. mica trimmer.
C_5 — Approximately 8-μμfd. variable (National UM-15 cut down to 3 plates).
C_6 — 25-μμfd. air trimmer (Hammarlund APC-25).
C_7–C_{12} — 500-μμfd. midget mica.
C_{13} — 100-μμfd. mica.
C_{14} — 50-μμfd. silvered mica.
C_{15} — Oscillator-mixer coupling condenser (see text).
C_{16} — 25-μμfd. air trimmer (Hammarlund APC-25).
C_{17} — 0.002-μfd. mica.
C_{18} — 0.01-μfd. 400-volt paper.
C_{19}, C_{20} — 8-μfd. 450-volt electrolytic.
R_1 — 50,000 ohms, ½ watt.
R_2 — 1200 ohms, ½ watt.
R_3 — 10,000 ohms, ½ watt.
R_4 — 6000 ohms, 10 watt.
L_1–L_6 — See coil table, p. 291.
L_7 — 18 turns No. 22 e., close-wound on ⅝-inch diameter form.
L_8 — 8 turns similar to L_7, at ground end of L_7.
L_9 — Filter choke, 8 henrys, 55 ma. (Thordarson T-14C62).
T_1 — Filament transformer, 6.3 volts, 1.2 amperes (Stancor P-6134).
T_2 — Power transformer, 560-0-560 volts, 30 ma. (Thordarson T-60R49).
S_1, S_2 — S.p.s.t. toggle switch.

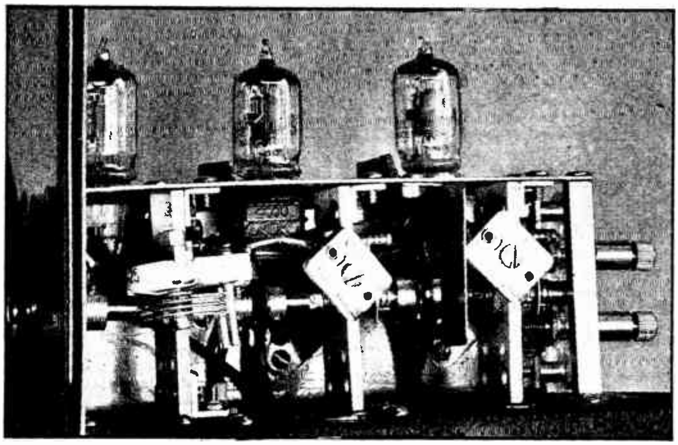


Fig. 1519 — **Inside the converter unit, showing arrangement of the tuning condensers. The layout is quite compact, with leads kept as short as possible.**

it crosses the tube socket and should be placed as close to it as possible, care must be taken to see that the socket prongs are bent away so they cannot touch it. The other shield is almost all on the outside and is used chiefly to prevent electrostatic coupling between the r.f. and mixer trimmer condensers, which are mounted on the sides of the tuning condensers. A transverse shield plate completely boxing off the two stages would be better, but it is an awkward job mechanically in view of the necessity for assembling the condenser gang.

No shielding is required between the mixer and oscillator; in fact, the stray coupling is too small to give good frequency conversion. The trimmer condenser is supported from the top of the chassis by a small bracket made from brass strip, bent to such a size that the rotor connection of the trimmer comes right at the rotor spring on the tuning condenser, where the two are soldered together. A small strip of copper is soldered between the two sets of stator plates, using the soldered mounting on top of the trimmer for its connection. The coupling condenser is a small piece of copper bolted to the trimmer end plate and bent to face the other soldered mounting. The separation is about a sixteenth of an inch.

The vertical shield plates between the coils are 2⅜ × 1⅜ inches, with bent-over edges to fasten to the side of the chassis. To complete the magnetic shielding the end of the mixer coil must be boxed in, which is done by a piece of copper in the shape of a shallow U, held in place simply by making it fit tightly between the vertical shields. This piece must be removable for changing the mixer coil.

Care must be used in making soldered connections on the polystyrene sockets and forms, since the material will soften with the application of heat. Have all surfaces well cleaned before attempting to solder the connections, and heat the lugs only enough to get a good joint.

The bottom view shows the arrangement of the power supply and the i.f. output circuit. The transformer for the latter is wound on a National PRE-3 polystyrene form. It is mounted on a bracket to keep it about equally spaced from the top of the chassis and the bottom of the cabinet in which the chassis fits. The various a.c. and d.c. supply connections from the converter are brought to lug strips, as shown; cathode resistors for the r.f. and mixer stages are mounted where they are readily accessible for trying different values. The power-supply parts are arranged to fit in the remaining space. The rubber feet at the rear of the chassis give a little space for circulation of air, since a fair amount of heat is developed by the transformers and regulator tube.

A few mechanical points should be given consideration in assembling the tuning condensers. The screw-on shafts are likely to come loose with use unless they can be anchored in some way, and soldering is about the simplest scheme. The heat tends to cause the lubricant to run out of the shaft bearing, however. Another important point is to get the shafts of the three condensers lined up accurately so that the rotors turn freely. Any twist, particularly at the oscillator condenser shaft, will tend to bend the rotor out of line slightly with respect to the stator, and, since the twisting depends upon the direction of rotation, this means that the assembly will have bad backlash. For the same reason, the dial must be lined up accurately with the condenser shafts. Line up the shafts to run as true as possible and fix the stators where they want to come on the chassis, using shims if necessary.

Alignment of the converter will involve some cut-and-try, using the coil specifications given in the table as a guide. It is best to line up the set with the 56-Mc. coils first before tackling the 112-Mc. band. The first step is to make the oscillator cover the proper range, the

COIL DATA

Band	Coil	No. of Turns	Wire Size	Length Inches	Remarks
112 Mc.	L_1	1 11/16	18	1 5/16	
	L_2	1⅛	24		⅛″ from L_1
	L_3	1 11/16	18	1 5/16	
	L_4	1⅞	24	⅛	⅛″ from L_3
	L_5	¾	18		
	L_6	1⅛	24		⅛″ from L_5
56 Mc.	L_1	4⅝	18	⅜	
	L_2	2⅞	24	⅛	⅛″ from L_1
	L_3	4½	18	7/16	
	L_4	2⅞	24	⅛	⅛″ from L_3
	L_5	3⅝	18	⅜	
	L_6	2⅞	24	5/32	⅛″ from L_5

All coils wound on ¾-inch diameter forms (Amphenol type 24-5H, 5-prong).

object being to spread the band over about 75 per cent of the dial scale. With the 10.2-Mc. i.f., the oscillator range, to cover 56 to 60 Mc., will be from 45.8 to 49.8 Mc.; this may be checked on another receiver, if available. If not, probably it will be necessary to use actual signals in the band for the purpose, which also will involve having at least the mixer hooked up. With the circuit specifications given, the oscillator padding condenser should be set at about half-scale. The inductance of L_5 may be adjusted by closing up or opening out the turn spacing, which can be done within limits without moving the ends of the coil. Once the right range is secured, the turns should be cemented in place. An alternative method of adjustment is to make the coil slightly large at first and then cut down the inductance with a shorted turn of wire, slid along the coil form.

The oscillator tickler, L_6, should be adjusted to give stable oscillation without squegging. Squegging is evidenced by a whole series of signals instead of one and can be cured by reducing the feed-back, either by using a smaller number of tickler turns or by moving the tickler further away from the plate coil. Incidentally, the oscillator should deliver a steady d.c. note when heard on another receiver. For this check to mean anything, the receiver used must introduce no modulation on incoming signals.

Once the oscillator range is set, the mixer should be lined up to match. To do this, place the r.f. tube in its socket but connect a resistor of a few hundred ohms from its grid to ground, instead of using L_1. The mixer primary, L_4, must be in place, since it will have some effect on the tuning range of L_3C_2. Connect the r.f. output leads to the doublet posts on the communications receiver, set the latter to 10.2 Mc. and adjust C_{16} for maximum hiss, with the oscillator tube out of its socket. Then replace the tube and, with the oscillator set for 56 Mc., adjust the trimmer, C_4, for maximum hiss; reset the oscillator to 60 Mc. and readjust C_4. If more capacity is needed at C_4 for maximum hiss, the inductance of L_3 is too large; if less, L_3 is too small. Make an appropriate small change in the inductance and try again, continuing the process until C_4 peaks at the same setting at both ends of the band. Adjustment of the inductance of L_3 may be accomplished by the means described above.

When this process is finished, C_4 should be well in the air-dielectric portion of its range. Should the movable plate be close to the mica, L_3 is considerably too small. However, this would be accompanied by reduced tuning range on C_2, and it is doubtful if high padding capacity would permit full band coverage.

The r.f. stage is aligned in just the same way as the mixer circuit. The initial alignment should be done with nothing connected to the antenna posts. Should oscillation occur, reduce the size of L_4 until the stage is stable. Some slight trace of regeneration may remain, as indicated by an exaggerated peaking in the r.f. stage, but this will disappear with any sort of antenna load on the r.f. tuned circuit.

The procedure for the 112-Mc. coils is similar to that for 56 Mc. It is desirable to adjust the oscillator coil so that the trimmer, C_6, does not need resetting when changing bands.

A Simple 56/112-Mc. Converter

The converter shown in Figs. 1521, 1522, 1523, 1524, 1525 and 1526 uses a 1232 loktal tube for the mixer and a 7A4 for the h.f. oscillator. Although its sensitivity is not quite as good as that of the converter just described, it affords a simple converter for the v.h.f. range that will prove perfectly satisfactory. By grouping the tuning condenser, coil and tube socket closely together, it is a relatively simple matter to achieve low-enough circuit capacities to work readily on 112 Mc. As can be seen from Fig. 1523, the grid of the 1231 mixer is tapped down on the coil to reduce the loading on the circuit and obtain a better gain in the stage. The plate-tickler circuit in the oscillator permits the cathode to be grounded directly, causing a minimum of hum on the signal.

The oscillator tuning condenser is a 15-$\mu\mu$fd. condenser from which several plates have been removed, and this is paralleled by a 35-$\mu\mu$fd. band-set condenser. With this type of bandspread system the converter can be set to the desired frequency band and the mixer condenser turned to the point where the noise is greatest; thereafter the tuning is all done with the small oscillator condenser. When a signal has been tuned in the mixer can be peaked again, but this is not usually necessary over the range of the bandspread condenser. Pulling of the oscillator circuit by the mixer tuning is slight, because of the loose coupling.

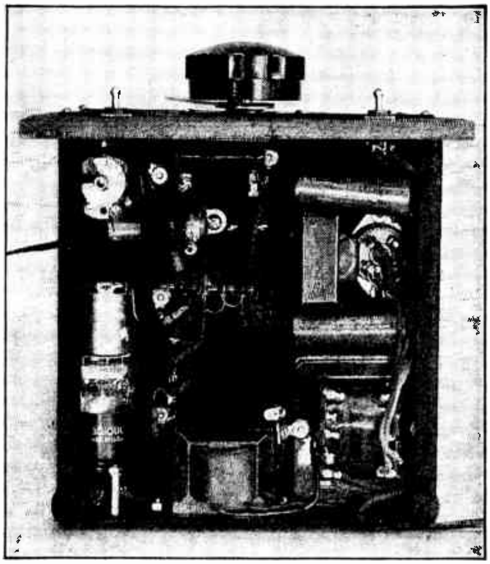


Fig. 1520 — The converter power supply occupies the right-hand section of the chassis in this bottom view. The i.f. output section is in the upper left-hand corner.

The chassis is made of 1/16-inch thick aluminium; other metals could be substituted, if necessary. The panel is $5\frac{1}{2} \times 8$ inches, but could be trimmed to 6 inches long. The extra length was included to put the dial in the center of the panel and also to provide room for possible future switches for shifting to various i.f. amplifiers. The chassis itself is built from a piece of 5¼-inch wide metal, bent to form a top 3⅞ inches wide and a back 4 inches deep. A ½-inch lip is bent down from the top to fasten the chassis to the panel. The two sides are made by forming shallow Us (with ½-inch sides) to fit between the panel and the back of the chassis. A shield is fitted under the chassis, making the oscillator compartment 2½ inches wide. This shield mounts the oscillator tuning condenser and the National TPB Victron through-bushing which serves as a coupling condenser between oscillator and mixer.

The coil forms are the small ¾-inch diameter Amphenol type made of polystyrene. The coil sockets, also of polystyrene, mount simply by drilling a hole and sliding the retainer rings over the sockets. The tube sockets are mounted in the same fashion. For short leads, the oscillator socket should be mounted with the slot towards the rear and the mixer socket with the slot towards the left-hand side.

As mentioned before, the oscillator tuning condenser, C_3, is mounted on the shield partition, and the band-set condenser, C_2, is mounted on the right-hand side of the chassis. The band-set condenser is insulated from the metal by fiber washers, so that there is only one ground point to the chassis for the oscillator circuit — that through the oscillator tuning condenser. The mixer tuning condenser, C_1, is mounted on the right-hand side of the chassis and grounds the mixer circuit at that point.

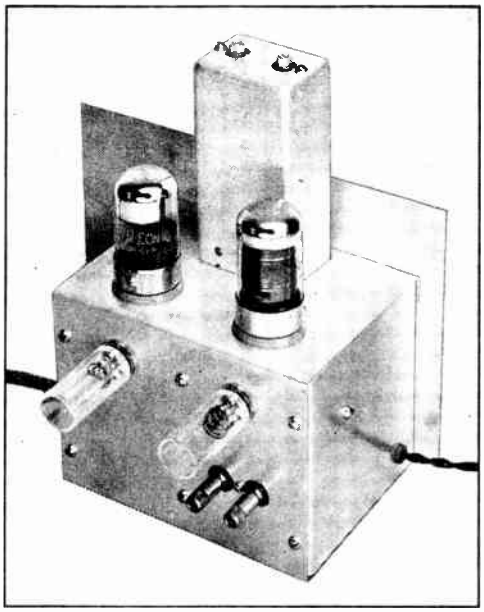


Fig. 1522 — A rear view of the converter, showing the plug-in coils and the antenna terminals. The cable leading off at the left goes to the power supply; the twisted pair on the right carries the output to the i.f. amplifier.

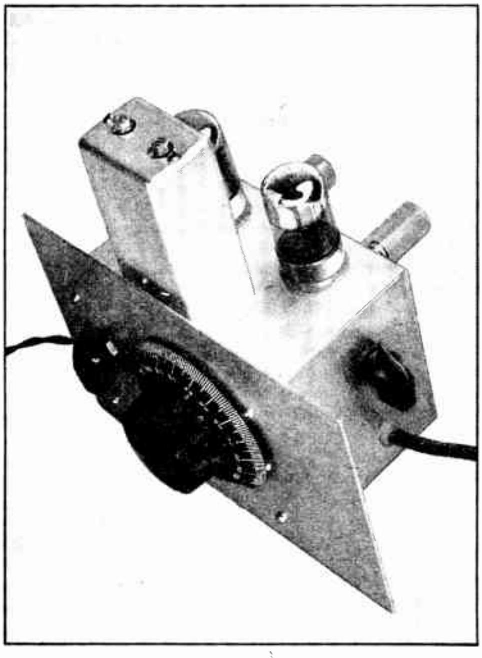


Fig. 1521 — The simple converter uses a 7A4 oscillator and a 7G7/1232 mixer. The panel dial is for oscillator tuning; the knob is the mixer tuning control. Knob on side adjusts the oscillator band-set condenser.

The oscillator tuning condenser and the mixer tuning condenser are fastened to their respective panel controls through insulated couplings, to avoid duplication of grounds.

The panel and sides should be left off until all of the wiring that can be done without them has been finished. Heater leads, ground connections, by-pass condensers, and resistors all should be put in before the sides and panel are attached. Do not hold the soldering iron on the polystyrene socket lugs for longer than is necessary to start the solder flowing or the socket contacts will loosen from softening of the polystyrene. A small, pointed soldering iron is best. A lead is run from the grid of the 7G7/1232 to the through-bushing on the partition but no connection is made on the oscillator side, since the capacity between the bushing and the oscillator leads is sufficient for coupling. All r.f. leads and the leads for the by-pass condensers must be kept short and direct.

The coil for the 56–60-Mc. range is wound in the usual manner on the outside of the coil forms. No trouble should be had in finding the 56-Mc. amateur band, since the tolerance on the range of this coil is fairly wide. The only care necessary is to prevent the pins from loosening up in the forms because of the heat when soldering. The wire should be well cleaned and a spot of flux used on the tip of the pin. No attempt should be made to flow solder on the pin and wire; a drop of solder picked up by the iron can be held against the pin for just an instant, long enough to solder wire and pin together. If the pin loosens up or moves out of place, it can be heated again slightly (by holding the soldering iron against it) and held in the proper position with long-nosed pliers. When the metal (and coil form) cools, it should be as solid as ever. If it isn't it doesn't matter too much, since the form still can be plugged in the socket without difficulty.

The coils for the 112–116-Mc. range are wound *inside* the coil forms. The forms can first be sawed through near the base and the coils adjusted by spreading the turns. When the necessary adjustments have been made the coil form can be refastened to the base by cement, thus avoiding danger of the coil being injured by handling.

The usual rule must be followed with the oscillator coil; i.e., if both grid and plate coils are wound in the same direction, the grid and plate connections should come off at opposite ends (in this case, the outside ends).

The connections for the oscillator coil, looking at the form from the bottom, are (starting with the widely spaced pin and going clockwise): plate, ground, "B"+, grid, blank. In the same manner, the mixer-coil connections are: grid, tuning condenser, antenna, antenna, ground. Both mixer and oscillator coil sockets are mounted with the odd pin at the top.

If connections have been kept short, no trouble should be experienced in making the oscillator oscillate on any of the ranges. For the 112-Mc. band, the oscillator band-set condenser setting will be at minimum capacity. For 56 Mc. it should be set at about mid-scale, varying slightly with the i.f. used.

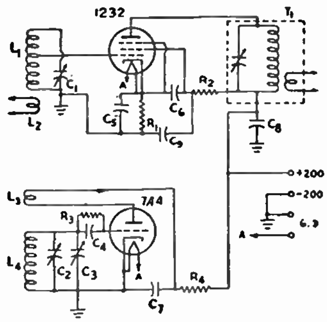
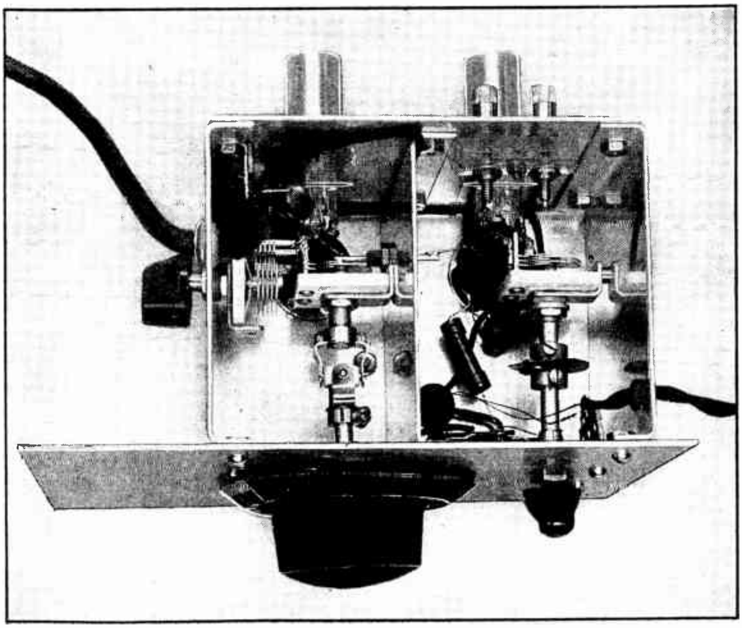


Fig. 1523 — Circuit diagram of the v.h.f. converter.

C_1 — 15-μμfd. midget variable (Hammarlund HF-15).
C_2 — 35-μμfd. midget variable (Hammarlund HF-35).
C_3 — 10-μμfd. midget variable (Hammarlund HF-15 with one stator and one rotor plate removed).
C_4 — 100-μμfd. midget mica.
C_5, C_6, C_7 — 500-μμfd. midget mica.
C_8, C_9 — 0.01-μfd. 600-volt paper.
R_1 — 500 ohms, ½-watt.
R_2 — 125,000 ohms, 1-watt.
R_3 — 20,000 ohms, ½-watt.
R_4 — 10,000 ohms, 1-watt.
T_1 — 3 Mc.: 75 turns No. 30 d.s.c., close-wound; coupling coil 20 turns No. 30 d.s.c., close-wound.
5 Mc.: 45 turns No. 30 d.s.c., close-wound; coupling coil 14 turns No. 30 d.s.c., close-wound.
L_1 — 112 Mc.: 2¼ turns No. 20 e., ⅜-inch diameter, spaced wire diameter. Grid tap is ¾-turn from top.
56 Mc.: 4½ turns No. 20 e., ¾-inch diameter, spaced over ½ inch. Grid tap is 1½ turns from top.
L_2 — 112 Mc.: 3 turns No. 20 e., ¼-inch diameter, close-wound one wire diameter below cold end of L_1.
56 Mc.: 3 turns No. 24 e., close-wound ⅛ inch below L_1.
L_3 — 112 Mc.: 1 turn No. 20 e. ¼-inch diameter, 3 wire-diameters below L_4.
56 Mc.: 1½ turns No. 24 e. close-wound ⅛ inch below L_4.
L_4 — 112 Mc.: ⅞ turn No. 20 e., ⅜-inch diameter.
56 Mc.: 1¾ turns No. 20 e., spaced over ¼ inch winding length.

Fig. 1524 — A view underneath the v.h.f. converter. Note that in the oscillator section (on the left) the band-set and tuning condensers butt into each other for short leads. The band-set condenser, C_2, is insulated from the side panel by washers, and the oscillator circuit is grounded to the chassis at only one point — through the tuning condenser. The Victron through-bushing which serves as the coupling condenser between mixer and oscillator can be seen on the partition just above the oscillator tuning condenser. The bushing connects to the 7G7/1232 grid on one side and is blank on the other. The i.f. transformer, T_1 (not visible in this view), is built in a Hammarlund ETU shielded unit. Both sections of the trimmer assembly condenser are used, in parallel.



The converter is coupled into the i.f. amplifier through a low-impedance link. This requires that the input transformer in the i.f. amplifier be modified by winding a number of turns about the grid coil as a link coil. Alternatively, a duplicate of the output transformer, T_1, can be substituted for the first transformer in the i.f. amplifier. If a receiver is used for the i.f. amplifier, the output leads go to ground and to the grid cap of the mixer tube in the receiver, replacing the regular grid lead.

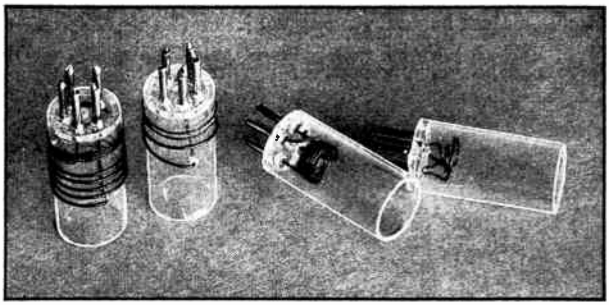


Fig. 1525—The 112-Mc. coils (right) are wound self-supporting inside the coil forms, while the 56-Mc. coils are wound in the usual manner.

Antennas for use with the converter present the same problem that they do with any v.h.f. receiver; the best one for the service required is the one to use. Information concerning various types of v.h.f. antennas will be found in Chapter Ten. A little experimenting with the antenna coil, L_2, may help in giving a better match to the antenna system; the dimensions given are average values which should work out about right for low-impedance line input.

If signals are weak, the trouble probably can be accounted for by too much or too little oscillator voltage reaching the mixer. This can be adjusted over a considerable range by moving the tickler coil, L_3, closer to or farther away from L_4. However, the adjustment will not be found to be too critical.

For maximum stability, a voltage-stabilized power supply should be used (§8-9).

◖ F.M. I.F. Amplifiers

As was pointed out earlier in this chapter, an f.m. superheterodyne receiver differs from an a.m. receiver in that the intermediate-frequency amplifier pass-band is wider and a limiter and discriminator are used instead of a second detector. The front end of an f.m. receiver is conventional, and any of the converters described can be used. The f.m. i.f. amplifier may be either the i.f. amplifier of an f.m. broadcast receiver (of which there are several on the market) or one can readily be built for the purpose by the amateur. If the i.f. system of the f.m. broadcast receiver is used, the intermediate frequency should be learned so that the output of the converter can be tuned to this frequency and coupled to the grid of the mixer tube of the receiver.

The i.f. amplifiers of most f.m. broadcast receivers are designed to operate on a frequency in the vicinity of 5 Mc. although earlier models may be found with i.f.s as low as 3 Mc. In a few instances higher i.f.s of the order of 8 to 10 Mc. have been used. If the existing output transformer in the converter does not tune to the required frequency, add or remove enough turns from the coil to enable it to be tuned to the receiver i.f.

For operation on the 112-Mc. band or higher, a double superheterodyne set-up may may be desirable if the f.m. receiver i.f. is lower than 5 Mc. A simple 6K8 oscillator-mixer, may be used as an intermediate converter on 10 or 20 Mc.

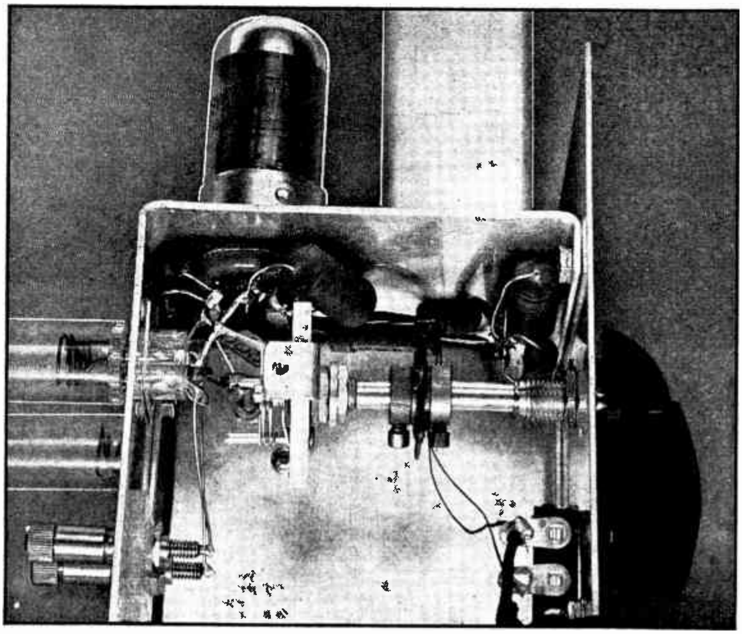


Fig. 1526 — The mixer circuit can be seen in this view of the 56/112-Mc. converter with the side panel removed. The tie strip at the lower right takes the output leads from the i.f. transformer. The interstage through-bushing can be seen just to the left of and under the tuning condenser, with a wire from it running to the 1232 grid terminal.

A 5-Mc. F.M. I.F. System

The i.f. amplifier shown in Figs. 1527, 1528 and 1529 is a broad-band combination affair working on 5 Mc. which can be used for either f.m. or a.m. reception merely by switching the connection to the grid lead of the first audio tube from across the discriminator load (for f.m.) to the limiter grid resistor (for a.m.).

With the converters described (or any combination capable of working into a 5-Mc. amplifier), this system can be used for the reception of a.m. and f.m. signals in the 43-Mc. band, a.m. and f.m. amateur signals in the 56-Mc. band, or f.m. and a.m. signals in the 112-Mc. band. When operators of 112-Mc. stations using modulated oscillators reduce the modulation percentage and thus bring the frequency deviation down to a reasonable range, the system constitutes an excellent receiver for the reception of modulated-oscillator transmissions. When operated with reduced modulation even the smallest transceiver will sound many times better; moreover, audio power will be saved, as well.

As may be seen from Fig. 1528, the two stages of high-gain amplification using 6AC7/1852 tubes are unconventional only in that resistors are used across the transformer windings to widen the pass band, and no gain control is included. No means of controlling gain is required, because it is always desirable to work the stages preceding the limiter at their highest level.

The limiter stage uses a 6SJ7, with provision through a variable resistor, R_{18} to control the plate and screen voltage to set the limiting action to meet operating conditions. The use of a grid leak and condenser, R_{16} and C_7, together with the low screen and plate voltages allows the tube to saturate quickly, even at low signal levels, and the tube wipes off any amplitude modulation (including noise) and passes only frequency modulation. For a.m. reception, the audio system is switched by Sw_1 to the grid leak, R_{16}, and the grid and cathode of the tube are used as a diode rectifier to feed the audio system. The jack, J, in series with the grid leak, is used for plugging in a low-range milliammeter so that the limiter current can be read. The limiter-current indication is invaluable in aligning the amplifier, and the meter can be used as a tuning meter during operation.

The discriminator circuit uses a 6H6 double diode in the conventional circuit. Audio from the discriminator (or from the limiter stage, in a.m. reception) is fed through the volume control, R_{25}, into a two-stage audio amplifier using a 6SF5 and 6F6 output pentode. The resistor, R_{11}, and condenser, C_{12}, in the audio input circuit, serve as a combined r.f. filter and compensating network to attenuate the higher audio frequencies. This is necessary when listening to 43-Mc. broadcast stations, since nearly all use "pre-distortion" (accented higher frequencies). A 0.01-μfd. condenser across the output terminals will give further high frequency compensation, if necessary.

The power supply uses a two-section filter, and an outlet socket is provided so that the converter power cable can be plugged in. A VR150-30 regulator tube is used for additional stability with changes in line voltage. The regulator tube is not absolutely necessary, and, if desired, it may be omitted.

The amplifier is built on a 7 × 9 × 2-inch chassis. Reference to Figs. 1527 and 1529 will show the location of the parts on the chassis. After all holes have been drilled the sockets and the transformer should be fastened in place on the chassis, leaving off the variable resistors, switches, binding posts, jack and chokes until after most of the wiring has been done.

If low-impedance input coupling is to be used, as with a converter removed some distance from the amplifier, the first i.f. transformer must be modified. A link winding is made by first winding a short half-inch wide strip of paper over the cardboard tubing used as a former in the i.f. transformer. Eleven turns of No. 30 d.s.c. wire are then close-wound flat over the center of the paper ring. Holding the wire in place with a finger, paint the coil with Duco cement to secure the turns in place. When the cement has dried, slip the coil off the form. The plate and "B"+ wires may be removed from the trimmer condenser in the

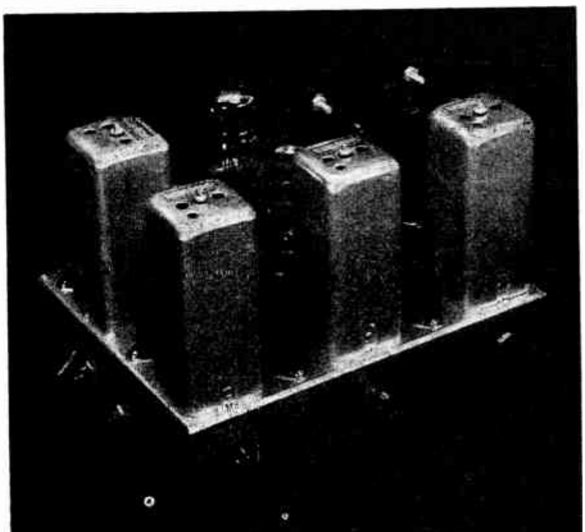


Fig. 1527 — A 5-Mc. f.m./a.m. amplifier, complete with power supply. Controls on the front, from left to right, are the audio volume control, "B"+ switch, and the limiter control. The f.m./a.m. switch is on the end. The jack beside it is for the limiter-current meter.

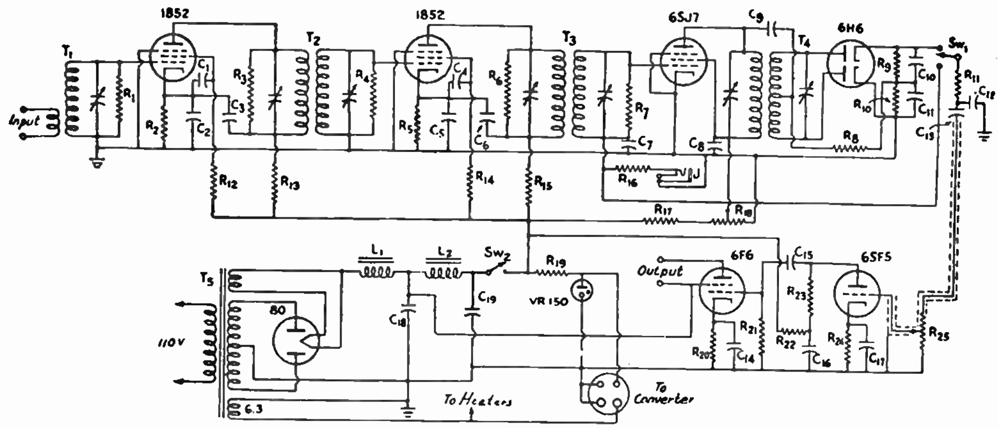


Fig. 1528 — Wiring diagram of the broad-band 5-Mc. frequency-modulated/amplitude-modulated i.f. amplifier.

C_1, C_2, C_3, C_4, C_5, C_6, C_8, C_{13}, C_{15} — 0.01-μfd. 600-volt paper.
C_7, C_{10}, C_{11} — 100-μμfd. midget mica.
C_9 — 50-μμfd. midget mica.
C_{12} — 0.001-μfd. midget mica.
C_{14}, C_{17} — 10-μfd. 25-volt electrolytic.
C_{16}, C_{18}, C_{19} — 16-μfd. 450-volt electrolytic.
R_1, R_4 — 55,000 ohms, ½-watt.
R_2 — 200 ohms, ½-watt.
R_3, R_6 — 50,000 ohms, ½-watt.
R_5 — 300 ohms, ½-watt.
R_7 — 40,000 ohms, ½-watt.
R_8, R_{11}, R_{22} — 75,000 ohms, ½-watt.
R_9, R_{10}, R_{16} — 150,000 ohms, ½-watt.
R_{12}, R_{14} — 60,000 ohms, ½-watt.
R_{13}, R_{15} — 100 ohms, ½-watt.
R_{17} — 25,000 ohms, 10-watt wire-wound.
R_{18} — 3000-ohm wire-wound potentiometer.
R_{19} — 5000 ohms, 10-watt wire-wound.
R_{20} — 500 ohms, 1-watt.
R_{21}, R_{23} — 250,000 ohms, ½-watt.
R_{24} — 5000 ohms, ½-watt.
R_{25} — 500,000-ohm volume control.
T_1 — 5-Mc. i.f. input transformer (see text) (Millen 67503).
T_2, T_3 — 5-Mc. f.m. interstage i.f. transformer (Millen 67503).
T_4 — 5-Mc. f.m. discriminator transformer (Millen 67504).
T_5 — 350-0-350-volt 90-ma. power transformer with 6.3- and 5-volt filament windings.
L_1 — 9-henry 85-ma. filter choke (Thordarson T-13C29).
L_2 — 10-henry 65-ma. filter choke (Thordarson T-13C28).
Sw_1 — Selector switch (Yaxley 32112-J).
Sw_2 — S.p.s.t. toggle switch.
J — Closed-circuit jack.

transformer, and the wires from the plate coil to the trimmer condenser disconnected. By unwinding and cutting off a turn or two of paper from the inside of the paper ring, the 11-turn coil can be slipped easily over the grid coil and fastened in position so that it covers the ground end of the grid coil. A piece of paper between the grid coil and the ground lead will avoid any possibility of this lead shorting against the turns of the coil when the paper ring is slipped in place. The two ends of the link coil are brought out the bottom of the shield can, later to be wired to the input terminals of the amplifier unit.

It is possible to use the transformer by merely running the plate lead to the mixer tube in the converter, but this makes it less convenient to use the converter with other i.f. amplifiers since it would require soldering and unsoldering wires each time the change was made. Furthermore, the long lead to the mixer tube would increase the chances for stray pick-up of signals in the vicinity of 5 Mc.

The screen by-pass condensers, C_1, C_4 and C_8, are mounted across the sockets so that they act as partial shields between the plates and grids of the single-ended tubes. Tie-points are used wherever needed for mounting the resistors and condensers. It is recommended that the 6AC7/1852, 6SJ7 and 6H6 stages be wired first, so that the leads carrying r.f. can be made as short and direct as possible. The rest of the leads may be filled in wherever convenient. The wires from the audio volume control, R_{25}, are shielded by a length of flexible copper braid. Whenever convenient, spare pins on sockets are used to support resistors, condensers, etc.

The two variable resistors mounted on the front of the chassis will not clear the spade bolts projecting down from the i.f. transformers unless about ⅛ inch is cut off the bolts before mounting the transformers in place. The input terminal strip (Millen 33002) is mounted on the outside of the chassis so that the contacts will clear the limiter control. A handy connector for plugging into this input terminal can be made from an old five-prong tube base or coil form, sawing across the base and removing the two correctly spaced pins and their supporting strip of bakelite.

With a 5-Mc. signal source, preferably a signal generator, alignment of the amplifier is an easy matter. If no such source is available a simple e.c.o. can be built using an ordinary receiving pentode such as a 6K7, with the grid circuit on 2.5 Mc. and the plate on 5 Mc. Or, if a converter is available, tune the regular receiver to 5 Mc., couple in the converter and tune in a strong, steady signal. The converter output can then be transferred to the f.m./a.m. i.f. and the transformers aligned. This is done by plugging a 0–1 ma. meter into the jack, *J*, and tuning the trimmers of the transformers for maximum current. It may be necessary to hunt around a bit before the meter shows any indication, but once it starts to read the rest is easy. With a variable-frequency signal source the signal is swung back and forth until some indication is obtained, and then the amplifier alignment is completed. The exact frequency of alignment is unimportant provided every stage can be tuned *through* resonance, which means that each trimmer can be adjusted through a maximum reading of the tuning meter. With the resistors across the circuits, it will be found that the transformers tune somewhat broader than normal; the correct setting is in the mid-point of the broad region. Once the i.f. transformers, T_1, T_2 and T_3, are aligned, it should be possible to switch Sw_1 to a.m. reception and hear signals, or at least noise, provided the converter is on 56 or 43 Mc. There isn't much noise to be heard on 112 Mc.

The alignment procedure can be carried out with a loudspeaker connected to the 6F6 through an output transformer. If no speaker is used at this point, however, the output terminals should be shorted; otherwise, the 6F6 may be injured. The use of a meter for alignment is a practical necessity, and no attempt should be made to line up the amplifier by ear except possibly for only a very rough initial alignment.

If there is an f.m. broadcast station within range, adjustment of the discriminator transformer, T_4, is a simple matter. Switch the amplifier to a.m., plug in the proper coils in the converter, and tune in the f.m. station. Then switch the amplifier to f.m. and tune the trimmers on T_4 until the signal is heard again. This is best done with the audio gain almost open and the limiter control at about half-scale. The trimmers are best adjusted with an insulated tool, to reduce body capacity effects, and they should be adjusted until the b.c. signal is clearest and loudest. It will be found that the trimmers in the plate circuit will affect the volume mostly, while the trimmer in the grid circuit will have the greatest effect on the quality. During this period of adjustment the receiver should be kept tuned to the signal, as indicated by maximum limiter current. An audio output meter, if available, may be used to determine maximum audio output, but this is not an essential.

In the event there is no local f.m. broadcast station the only alternative is to line up the discriminator on an f.m. signal from an amateur station, or, as a last resort, from a 112-Mc. modulated oscillator. The disadvantage with the self-excited v.h.f. oscillator is that usually it is modulated too heavily and doesn't stay on one frequency long enough to allow the amplifier to be aligned accurately.

The final adjustment of the discriminator tuning can be checked by tuning in an a.m. signal. If the discriminator is properly tuned, the audio output (signal and noise) should practically disappear at the point where the signal, as indicated by limiter current, is a maximum. This is an indication that the discriminator characteristic crosses the axis at the mid-resonance point of the amplifier. Tuning the signal (by tuning the converter), it should be possible to understand the audio output at points either side of this minimum-volume setting. These points should appear symmetrically on either side of the minimum-volume point and should have about the same volume. Slight readjustment of the discriminator-transformer setting will accomplish this result.

When using the amplifier it could be noted that a.m. signals appeared to be louder than those from f.m. stations, comparing audio volume-control settings on stations showing equal limiter current. This did not indicate that the amplifier was not working properly or that more audio is obtained from an a.m. signal than from an f.m. signal of similar strength. It was, however, an indication that the discriminator characteristic could have more slope to it and not have the peaks so far apart. With improved discriminator-transformer construction now to be expected, this apparent shortcoming will disappear.

The performance of the amplifier on a.m. reception could be improved somewhat by the inclusion of a.v.c. on the two 1852 tubes, taking the a.v.c. voltage from the limiter grid leak through the usual filter circuit. However, this was considered an unnecessary refinement because the amplifier was intended to be used primarily on f.m. reception. The amplifier should be run "wide open" on f.m. reception.

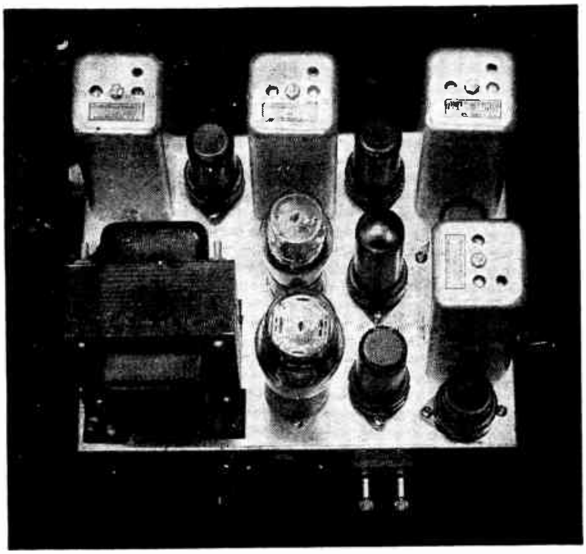


Fig. 1529 — **A top view of the f.m./a.m. amplifier. Along the rear, from left to right, are the input transformer, first 1852 tube, interstage transformer, second 1852 tube, and second interstage transformer. In the second row of tubes, from right to left, are the 6SJ7 limiter, 6F6 audio output and VR150-30 voltage regulator. At the right front is the discriminator transformer, with the 6H6 detector below it. To the left of the 6H6 is the 6SF5 first audio. Output terminals, power socket, and 115-volt line cord are on the lower edge.**

Chapter Sixteen

V.H.F. Transmitters

THE very-high frequency region is generally considered to have its lower frequency limit in the vicinity of the 28-Mc. band, and it is also in about this region that it becomes desirable to adopt more compact methods of construction and to select tubes with particular care. As the frequency becomes higher the length of connecting leads becomes more important, because a length of a few inches may represent a considerable fraction of the operating wavelength. Tube interelectrode capacities, as well as the usual stray capacities, must be given particular attention. Unduly high shunt capacity in the circuit not only may reduce the efficiency but also will ultimately set the upper limit of frequency at which the transmitter can be made to work. For best results at very-high frequencies, tubes designed to operate well in that region must be used. All of these considerations indicate the advisability of building separate r.f. equipment for transmission at very-high frequencies, rather than attempting to adapt for v.h.f. use a transmitter primarily designed for operation at ordinary frequencies.

Transmitter stability requirements for operation in the 56-Mc. band are the same as for the lower-frequency bands. Above 112 Mc. there are no restrictions as to frequency stability except that the whole of the emission must be confined within the band limits. Modulated-oscillator type transmitters therefore can be used above 112 Mc. and are, in fact, used practically exclusively above 224 Mc., since few available tubes will operate satisfactorily as amplifiers at this high frequency. However, up to 60 Mc. methods similar to those employed in the transmitters described in Chapter Thirteen can be used.

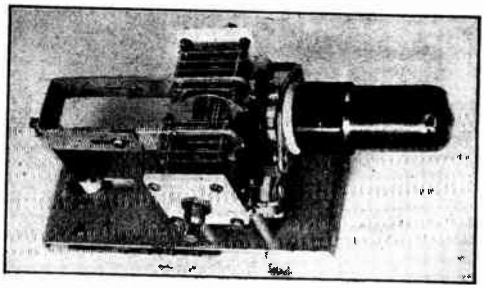


Fig. 1601 — A stable 112-Mc. oscillator using a flat copper strip parallel line capacity-loaded resonant circuit in a conventional ultraudion circuit. The line is an extension of the stator assembly of the variable condenser.

Most of the 56-Mc. transmitters shown in this chapter are crystal controlled, for use with amplitude modulation. However, they can be adapted for f.m. by replacing the crystal with excitation from an f.m. oscillator similar to that described in Figs. 1634, 1635 and 1636. Higher-powered transmitters can be built by adding amplifiers to the units shown, basing the design on the medium-power amplifier-driver unit shown in Figs. 1631, 1632 and 1633.

A "Square-U" Tank 112-Mc. Oscillator

The simple self-excited oscillator in Figs. 1601–1603 consists of a 6V6GT in a conventional ultraudion circuit. Because of the unique tank-circuit construction employed, however,

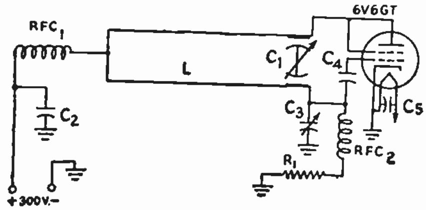


Fig. 1602 — Circuit diagram of the 112-Mc. oscillator.
C_1 — 40-μμfd. per section variable (Cardwell).
C_2 — 3–30-μμfd. mica trimmer.
C_3 — 25-μμfd. silvered mica.
C_4 — 500-μμfd. midget mica.
R_1 — 15,000 ohms, ½ watt.
RFC_1 — 45 turns No. 22 e. closewound on ⅜-inch diameter form.
RFC_2 — 40 turns No. 24 d.s.c. closewound on ¼-inch diameter form.
L — See text and photographs.

it delivers an order of performance superior to that of ordinary oscillators of this general type, combining good frequency stability with comparatively high efficiency.

The tuned circuit is made up of a short length of flat parallel-conductor transmission line, *L*, together with a balanced variable condenser, C_1. The line is made of annealed pure copper strip which has been silver-plated to reduce contact resistance and to prevent oxidation. The ends of this strip are incorporated in the condenser assembly, each constituting one plate in one of the stator assemblies of C_1.

This type of line, together with the method of connection to the tuning condenser described, possesses several advantages. At the very-high frequencies, because of skin effect, the r.f. currents tend to follow the shortest path, — that along the inner edge of the loop perimeter. The use of flat strip "on edge" provides considerably greater area on the inner side than with any other shape of conductor. The greater the useful area, the lower the resistance — and, therefore the higher the Q, resulting in improved stability and efficiency.

Because a relatively low L/C ratio is required for stability, the effective length of the line is only a small fraction of a wavelength. Thus the current at the junction between the line and condenser is high and the connection must have low r.f. resistance. The area of contact with the condenser plates must be as large as possible. Ordinary terminals, connecting at only one point are not effective in this respect because the current instead of flowing through the condenser plates, goes around and over them — thus greatly increasing the path and consequently the resistance. One method of achieving this, as illustrated on page 367, is to make contact simultaneously with all plates in the stator assembly. In this oscillator, the problem is solved by making the terminal ends of the inductance portion of the tank an integral part of the condenser. This is done by removing a stator plate in each assembly and proportioning and spacing the conductors so that each end of the copper strip replaces one of these plates.

To preserve the effectiveness of the tank circuit, the ceramic octal socket for the 6V6 is so mounted as to provide the shortest possible leads between the tank and the tube prongs. The plate terminal on the socket is soldered directly to one end of the copper strip, while the grid condenser connects to the other end with leads less than ¼-inch long over-all. The cathode terminal, one side of the heater, and the No. 1 pin are grounded by a quarter-moon-shaped bit of sheet copper.

The r.f. chokes, the dimensions of which appear under Fig. 1602, are wound on ⅜-inch ceramic forms, the size of the windings being determined experimentally for optimum performance over the range from 100 to 150 Mc.

Under load the frequency stability remains excellent, even at 140 Mc. Total drift during the warm-up period does not exceed 0.06 per cent, and the frequency shift, with a variation in plate voltage corresponding to 75 per cent amplitude modulation, is less than 2 kc.

Output coupling may be by means of an adjustable tap along the plate side of the line or by a hair-pin loop proportioned to fit inside the square-U tank line.

In operation, the only adjustments required are to tune the oscillator to the desired frequency by C_1 and to set the output coupling and the excitation control, C_2 for maximum stability and output. The output may be determined by connecting a 5- or 10-watt light bulb in series with a small variable condenser and a loop of wire which will tune to the operating frequency. Watching its brilliance while adjusting C_2 will readily show the optimum setting. Maximum stability is obtained when C_2 is set at the largest capacity which will give

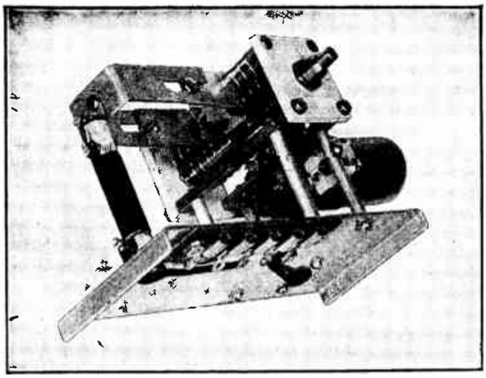


Fig. 1603 — Another view of the parallel-line 6V6GT oscillator, showing the plate and grid r.f. chokes, the feed-back condenser, C_2 (at right, grounded to one of the pillars supporting C_1), and the pin-jack terminal strip with extra terminals for making meter connections.

good output. First increase its capacity to the point where the output drops off and then decrease it just to the point where the output comes back to normal. As the capacity is decreased still more the output should be reduced somewhat. If the plate current tends to be unduly high it can be regulated by changing the value of grid leak, larger values giving lower plate current and vice versa.

The frequency may be checked by means of Lecher wires or a frequency meter. Such equipment is described in Chapter Twenty.

When the frequency of an oscillator-type transmitter is given a final check, the antenna should be connected and the antenna coupling adjusted for normal operation. This is necessary because the frequency will be affected by the antenna coupling, so that a measurement made without an antenna (or with a dummy antenna) will not necessarily hold when the actual antenna and transmission line are coupled to the transmitter. The resonance indicator should be connected in series with the transmission line. If a flashlight bulb will not light under these conditions, a more sensitive resonance indicator consisting of a 112-Mc. tuned circuit, a crystal detector and a low-range milliammeter should give a satisfactory indication. Such a device also is described in Chapter Twenty.

The measurement procedure involves very few additional operations. Tune the meter to resonance as indicated by maximum milliammeter reading, then move it as far as possible from the transmitter while still getting a reading of the order of 25 per cent of maximum. Couple the loop at the end of the Lecher wires to the coil and take a trial setting of the shorting bar. The resonance point will be evidenced by a sharp dip in the meter reading. Slow variations as the bar is slid along mean simply that some detuning of the circuit is occurring. The actual dip will be quite pronounced; the bar should not have to be moved more than a fraction of an inch to go completely through resonance.

Once the resonance point is identified, loosen the coupling until the dip is just a slight downward kick in the reading. From this point on, the measurement procedure is the same as before. By this method it is possible to avoid detuning of the oscillator by the Lecher wires, some amount of which usually takes place even with loose coupling when the line is coupled to the oscillator itself. This occurs because of the necessity for abstracting an appreciable amount of energy from the circuit to get a good resonance indication from a flashlight lamp or similar device. With the crystal detector, it is possible to work at least a foot or two from even a low-power oscillator.

A Grid-Stabilized 815-Tube 112-Mc. Transmitter

The transmitter shown in Figs. 1604, 1605 and 1606 uses an 815 dual beam-power tube in a grid-stabilized oscillator circuit which can be used at an input of 60 watts with good efficiency. The circuit (Fig. 1606) is the basic tuned-grid tuned-plate circuit except that a linear tank instead of a coil and condenser is used in the grid circuit. By tapping the grids down on the line the loading is light, and consequently the line retains its high Q. The 815 does not have high enough grid-plate capacities to give all of the necessary feed-back, and some additional capacity must be added between plate and grid in both sections of the tube. This is easily done by running two short lengths of wire from the plate terminals to near the grid lines (C_F in Fig. 1605).

The transmitter is mounted on a 3 × 4 × 5-inch metal box which houses the filament transformer and the various fixed condensers, resistors, and the r.f. choke. The grid line is made of ½-inch copper tubing and is supported a half inch from the box by three feed-through insulators, which also serve as convenient connectors to the grids of the tube and to the grid leak. The open ends of the parallel tubes are fitted with 3-inch lengths of ⅜-inch diameter tubing which can be moved in and out to adjust the frequency of the oscillator. The extensions are held securely in place by set screws in holes tapped through the wall of the ½-inch outer tubing.

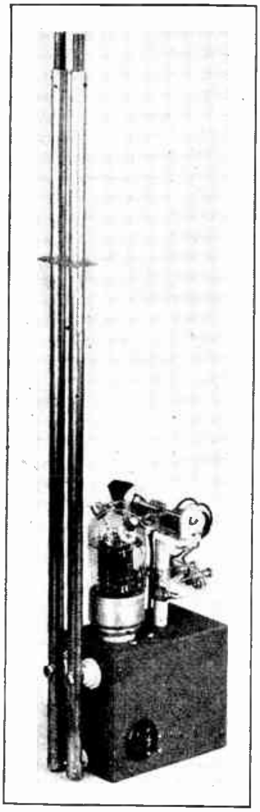


Fig. 1604 — The grid-stabilized 815 112-Mc. transmitter is mounted on a 3 × 4 × 5-inch box. Frequency is changed by adjusting the length of the grid lines by sliding the inner tubes in and out. The power-supply cable plug is mounted on the side of the box.

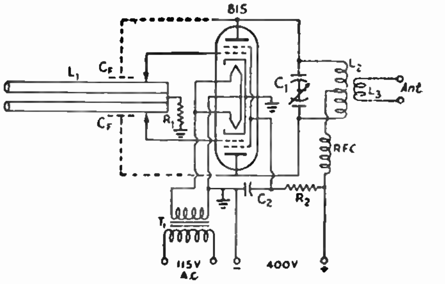


Fig. 1605 — Circuit of the grid-stabilized transmitter.

C_1 — 15-μμfd. dual variable condenser (Hammarlund HF-15X).
C_2 — 0.002-μfd. mica.
C_F — Feed-back condenser (see text).
R_1 — 15,000 ohms, 1-watt.
R_2 — 25,000 ohms, 10-watt.
L_1 — ½-inch diameter copper tubing, 23 inches long, spaced 1 inch on centers; grids tapped 2¾ inches from shorted end.
L_2 — 2 turns No. 12 e., 1-inch diameter, turns spaced to occupy ¾ inch.
L_3 — 2 turns No. 12 e., ¾-inch diameter, turns spaced to occupy ¼ inch.
RFC — V.h.f. r.f. choke (Ohmite Z-1).
T_1 — 6.3-volt 2-ampere filament transformer (Thordarson T-19F81).

The plate tuning condenser is supported by a 3-inch steatite pillar, which also serves as a guide for the sliding variable antenna coupling arrangement. Two large 866-type plate clips are slid over the pillar, and the antenna binding-post assembly (National FWB) is fastened to these clips with short lengths of No. 12 wire. By sliding this entire assembly up and down on the pillar, the antenna coupling can be set to any value desired.

There is nothing unusual about the tuning of the transmitter, aside from the adjustment of the feed-back condensers. This can best be done with a dummy load, such as a 25-watt lamp bulb, connected to the antenna terminals. The lead from the grid leak, R_1, to ground should be opened and a 0-10 milliammeter connected in the circuit. With plate voltage applied, the plate tuning condenser should be rotated for maximum output as indicated by the brilliancy of the lamp.

The grid current should be between 3.5 and 5 ma. at this point. A higher value than this indicates that there is too much feed-back, and the feed-back capacity should be reduced by trimming off a short length of the wire or by moving it away from the grid lines. This adjustment is not critical, but it should be made before the transmitter is put on the air. After the proper feed-back adjustment is found, the antenna can be coupled to the transmitter and modulation

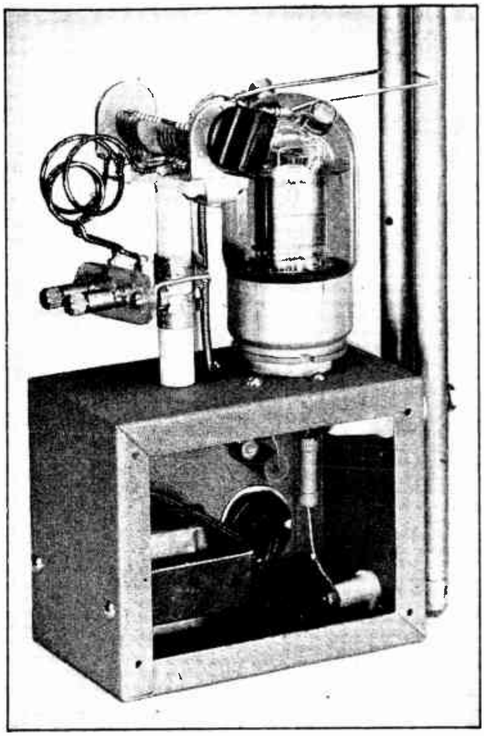


Fig. 1606 — A close-up view of the 815 transmitter, showing how the antenna coupling is changed by sliding the antenna-coil support on the insulating pillar. Note the wires for feed-back control running from the plates of the 815 close to the grid lines. The filament transformer and the various resistors and by-pass condensers are mounted in the box, shown here with one side removed.

applied. The antenna coupling may be increased until the plate current is 150 ma.; the grid current then should be between 3.5 and 5 ma.

The power supply is required to deliver slightly over 165 ma. at 400 volts, and the modulator must supply at least 30 watts to modulate the oscillator fully. The unit shown in Fig. 1407, using a pair of 6L6s in Class AB_1, will be satisfactory for the modulator.

ℂ Parallel-Line Push-Pull 112- and 224-Mc. Oscillators

Figs. 1607–1609 show a low-powered push-pull oscillator using a linear tank circuit of the "tuned-plate tuned-cathode" variety, which gives reasonably good stability and efficiency on 112 and 224 Mc. Using HY615 tubes, the unit is capable of about five watts output at 112 Mc. and somewhat less at 224 Mc.

The transmitter is built on a 2½ × 4½ × 15 inch chassis. The sockets for the tubes are oriented so that the plate caps face the left-hand end of the chassis. The plate lines are held together rigidly by a sturdy copper strap at the shorted end and by a polystyrene bar at a high-potential point near the tubes.

The cathode lines are mounted underneath the chassis. At one end they are connected directly to the cathode terminals on the tube sockets, the other ends being strapped together and supported by a small ceramic pillar. The heater leads, of rubber-covered hook-up wire, go through these lines as shown in Fig. 1608.

The cathode line is equipped with a sliding clamp-type shorting bar, while two similar bars are installed on the plate line. These are made of brass strip, bent around the parallel rods. Machine screws and nuts hold the clamp firmly in place.

In preliminary adjustment for 112-Mc. operation, the cathode bar is set near the end of the line most distant from the tube sockets. Both plate bars are first set at the shorted end of the line. To provide protection for the tubes, a 1000- or 2000-ohm protective resistor should be connected in series with the plate supply lead. A 0–100 or equivalent milliammeter also is placed in series with the high voltage lead. With heater power applied, check for oscillation as indicated by a dip in plate current when one of the conductors is touched with an insulated screwdriver or a neon bulb. If no oscillation is apparent, the effective length of the cathode line should be extended by moving the sliding bar toward the shorted end.

The frequency of the oscillator may be adjusted to resonance, using Lecher wires or equivalent measuring means, by moving the primary sliding bar toward the tubes until the desired operating frequency is reached.

For 224 Mc. operation the primary bar is advanced about six inches toward the tubes. The secondary bar is also moved, until oscillations occur. Some shortening of the effective length of the cathode line may be necessary. If the resulting frequency is not the desired operating frequency, adjust the two plate shorting bars simultaneously until the correct points for maximum output at the desired frequency have been determined.

The function of the primary shorting bar is to establish a length of line which, with the self-capacity of the tubes, will be correct for the desired frequency. The secondary shorting bar establishes an isolating quarter-wave section

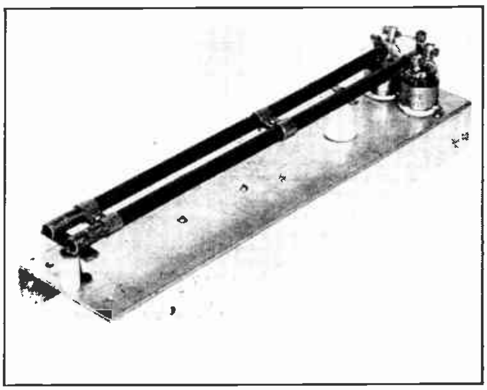


Fig. 1607 — A parallel-line push-pull HY615 oscillator which operates on either 112 or 224 Mc. by changing the positions of the primary and secondary shorting bars.

equivalent to an r.f. choke. The position of the secondary bar has some effect upon the frequency of oscillation, although the spacing between the two bars need not bear an exact quarter-wave relationship.

Fig. 1610 shows the construction of similar tuned-plate tuned-filament 112-Mc. transmitter for medium-power tubes. Fundamentally, the circuit is the same as that of Fig. 1609 except for the changes made necessary by the use of directly-heated tubes. With conventional tubes the efficiency is about 50 per cent.

Fig. 1610 shows the arrangement of the plate circuit on top of the chassis, which is 4½ inches wide, 15 inches long, and 2½ inches deep. A small 15-μμfd. variable condenser is connected across the filament line. There is no tuning condenser for the plate line; while one may be used if desired, for best efficiency the line should be kept as long as possible.

The high-voltage connection, brought through an insulator in the chassis is shown in Fig. 1610 just to the left of the supporting insulator. The antenna coupling link, L_3, is made from small-diameter copper tubing; its length should be adjusted to give the desired loading.

The tuned filament line is located underneath the chassis. Each pipe is soldered to, and partly supported by, a filament prong on one of the tube sockets. The shorted end of the line is held in place by a metal pillar, which also makes the ground connection to the chassis. An insulated wire is fed through each pipe and connected to the other filament prong on the appropriate socket. These wires are connected at the shorted end and the filament voltage applied between this common connection and ground.

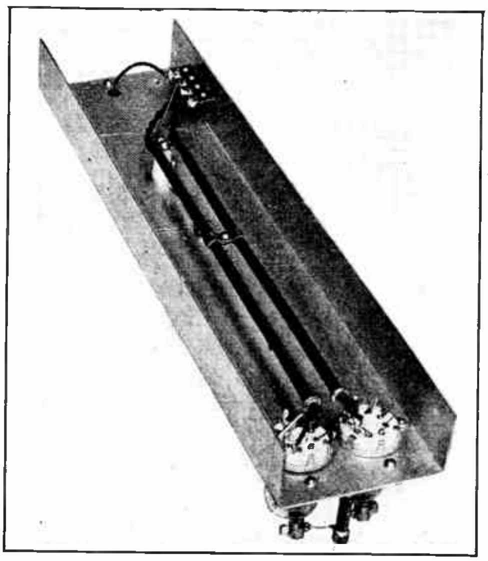


Fig. 1608 — Underneath view of the 112/224-Mc. linear oscillator showing placement of the cathode lines.

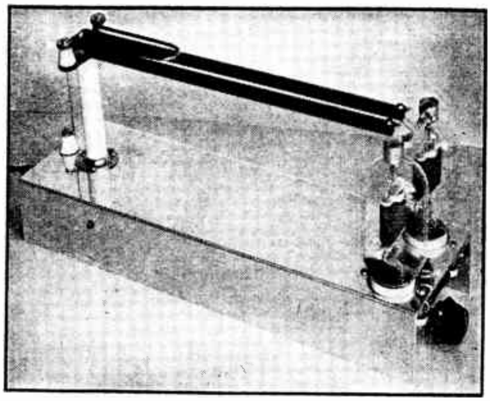


Fig. 1610 — This transmitter operates efficiently with conventional tubes at 112 Mc. A slider is used for frequency adjustment. Hairpin coupling link is at the left.

C_1, the filament-line tuning condenser, rests on the insulated portions of the sockets and is securely mounted by two small aluminum brackets which fit under the socket-mounting screws. Care must be taken to prevent ground-

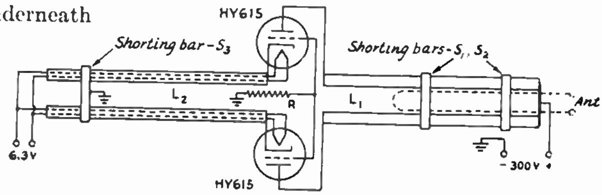


Fig. 1609 — Circuit diagram of the push-pull 112-Mc. oscillator.
R_1 — 5000 ohms, 10-watt.
L_1 — Plate line: ½-inch o.d. copper tubing, length 12 inches, spaced diameter of tubing.
L_2 — Cathode line: ¼-inch diameter copper tubing, length 10 inches, spaced ½ inch.

ing of the condenser plates. A short connection is made between the two grid prongs on the sockets; the grid resistor, R_1, runs from the center of this connection to ground.

Tuning is similar to that already described for the low-power transmitter. The setting of C_1 which gives minimum plate current is not, however, the adjustment at which the oscillator delivers maximum output. A lamp-bulb dummy antenna coupled to the pipes will show that, as the condenser setting is slightly altered, the plate current will rise and the output will increase. The current should not be allowed to exceed 200 ma. at full load.

Other tubes than the T40s shown have been used quite successfully in this circuit, including types 809, T20, RK32, RK11, RK12, 811, and TZ40. Still others of similar construction and ratings probably also would function satisfactorily. Tubes like the HK24 and 35T will work well at 224 Mc.

A modulator capable of delivering 100 watts of audio power is required when the transmitter is operated at 200 watts input. A pair of HY30Zs in Class B is recommended. The power supply must furnish 200 ma. at 1000 volts.

A Simple Transmitter for 224 Mc.

At frequencies higher than about 150 Mc., considerable difficulty is found in getting good performance with tubes other than those designed expressly for v.h.f. operation. However, there are several inexpensive v.h.f. tubes available to the amateur that will perform well on 224 Mc. The transmitter in Figs. 1611-1613 shows how one of these types, the HY75, may be put to work in a basically very simple oscillator circuit.

The transmitter is built on a 3½ × 6½-inch piece of ¼-inch Presdwood supported by two strips of 1 × 2-inch wood. A rectangular hole is cut in the center of the Presdwood to accommodate the tuning condenser, which is supported by two metal pillars at one end. The tuned circuit consists of two lengths of ¼-inch copper tubing, 3½ inches long, which are supported at one end by two feed-through insulators. The ends of the screws in the feed-through insulators are sweated into the ends of the tubing, and the tuning condenser is connected to two lugs right at this point. Connections from the tubing to the grid and plate terminals on the tube are made by ½-inch lengths of flexible shield braid. The filament chokes, the plate r.f. choke, and the grid leak are mounted under the chassis.

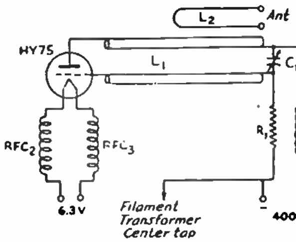


Fig. 1611 — Wiring diagram of the 224-Mc. oscillator.
C_1 — 100-μμfd. midget variable (National UM-100).
R_1 — 5000 ohms, 10-watt wire-wound.
L_1 — ¼-inch copper tubing, 3½ inches long, spaced ½-inch on centers.
L_2 — 2-inch loop of No. 16 bare wire.
RFC_1 — V.h.f. r.f. choke (Ohmite Z-1 or Z-0).
RFC_2, RFC_3 — 10 turns No. 18 e., close-wound on ½-inch diameter, self-supporting.

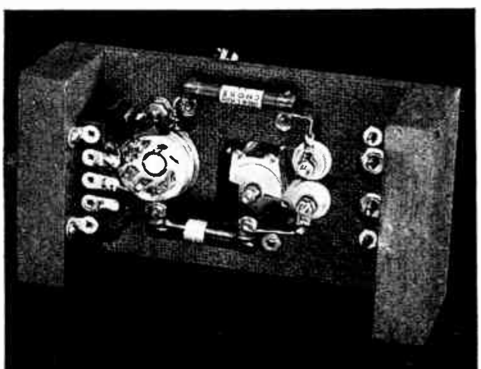


Fig. 1612 — The r.f. chokes and the grid leak are mounted under the chassis of the 224-Mc. transmitter. The power-supply cable is brought through a hole in the side piece to a tie strip mounted on the left-hand side.

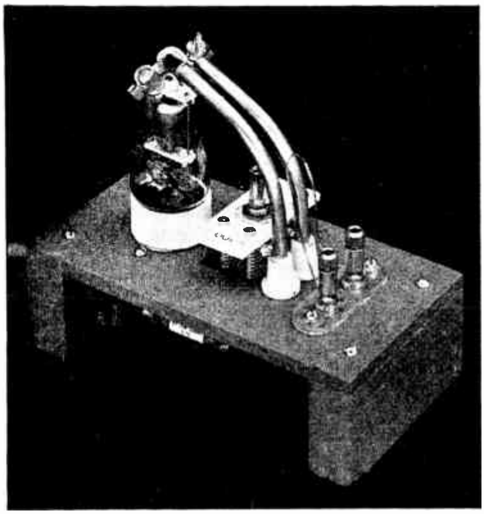


Fig. 1613 — A 224-Mc. transmitter using an HY75 tube. A rectangular clearance hole in the chassis allows the tuning condenser to be mounted for shortest leads. The condenser is adjusted by an insulated screw driver.

The antenna-coupling circuit consists of a loop of wire parallel to the copper tubing and terminating in the antenna binding posts. Coupling is varied by simply moving the hairpin loop nearer to or farther away from the copper-tubing line.

The transmitter should first be tested with a dummy load. A 10-watt lamp bulb is excellent for the purpose. The load is connected to the antenna posts and the power supply turned on. If everything is connected properly the lamp will light, its brilliancy depending upon the tightness of coupling and the setting of C_1. It will be found that the output is greater towards the maximum-capacity end of the range of C_1. The frequency coverage of the transmitter should be checked, using Lecher wires or a frequency meter, to make sure that it will cover the desired range. The coverage can be adjusted slightly by changing the separation of the copper-tube conductors; if this does not effect enough change, the lines will have to be made either shorter or longer, as required. The tuning condenser is adjusted by means of an insulated screw driver.

The transmitter requires a plate power supply delivering 60 ma. at 400 volts, and the modulator unit should be capable of furnishing 12 watts of audio power. The 6A6 modulator described in Chapter Fourteen will be quite adequate for the purpose.

Because of its small size, a transmitter of this type can be built as a unit into a rotatable antenna for the 224-Mc. band, if desired. It is desirable not to run a feed line for any great distance at this frequency, because of the possibility of radiation from the line.

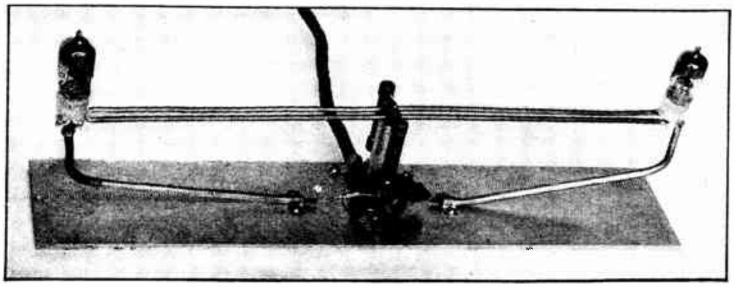


Fig. 1614 — The "teeter-totter" oscillator is useful for experimental purposes. It will function at frequencies up to 600 Mc. The typical model shown at the left operates at 224 Mc. The 9002 tube sockets are connected to symmetrical grid, plate and cathode lines.

A Symmetrical V.H.F. Oscillator

The "teeter-totter" oscillator, so termed because of its appearance, is a logical constructional development of the push-pull oscillator circuit of Fig. 431-E, using cathode lines instead of r.f. chokes.

The oscillator pictured in Fig. 1614 operates on 224 Mc., employing 9002 tubes. Using these tubes and the same general circuit and structural arrangement, by proportionately shortening the length of the lines it is capable of functioning up to 600 Mc. or higher.

Inasmuch as there is no convenient means for tuning or adjusting the length of the lines in this circuit, the lengths must be made to fit with all dimensions adjusted to the desired operating frequency. An important precaution is to make each pair of lines of identical dimensions and to make the entire layout fully symmetrical, so as to avoid unbalance to ground at any point. This procedure eliminates the not-infrequent difficulty arising from excessive unbalanced currents to ground, which occur even in apparently balanced symmetrical push-pull circuits and lower the upper frequency limit of oscillation.

One consideration in preserving circuit symmetry is to use tubes of similar characteristics. Unbalance in this connection, assuming all other elements of the circuit are symmetrical, may be determined by comparing r.f. voltage, as shown by a neon bulb or other indicator, at points equidistant from the feed point on both grid and plate lines.

To minimize the effects of any slight remaining dissymmetry, which does not reduce efficiency but may encourage spurious oscillation, voltage is fed through an r.f. choke wound with No. 16 wire or larger on a 1000-ohm resistor form.

The terminals of the polystyrene tube sockets are soldered directly to the ends of the lines. This operation is one requiring some care in construction. The best method is to split the end of the ⅛-inch copper tubing in such a manner that the extended socket terminal can be bent over and inserted and crimped in place. Before attaching the terminal, the end of the line should be well tinned. The lines and the two associated sockets may then be assembled and supported in a vise or similar holding fixture, whereupon the application of heat for brief duration from a moderately hot iron will flow solder between the terminals and the tubing without misalignment of the socket terminals. Even with this practice, it is wise to insert a dummy tube in the socket before attempting to solder the terminals.

The tubes and sockets are supported only by the parallel-wire plate and grid lines and the coaxial cathode lines with which they are associated. This type of mounting does not have good inherent mechanical stability, and vibration effects must be minimized. On 224 Mc. auxiliary support can be provided in the form of polystyrene rods or blocks arranged to reinforce the socket mountings at the ends of the line. If this arrangement is employed near the maximum upper frequency limit of the tubes, however, such insulation will impair the efficiency of the oscillator. In this case the lines should be anchored tightly, with solid, unvarying supports and brackets at the feed points, and the entire unit mounted on a cushioning base.

With 300 volts on the plates of the 9002s, the combined plate current should run about 40 ma. The actual output will vary with loading, but a watt or two should be available over a moderate range of load impedance.

The cathode lines have the cathode on one side of the heater circuit connected to the tubing and the heater return circuit made through a rubber-insulated wire running inside the copper conductor. Both cathode leads of the 9002s are connected in parallel. Similarly, the plate line is connected to the dual plate leads in each tube, in parallel.

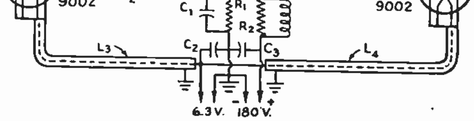


Fig. 1615 — "Teeter-totter" oscillator circuit diagram.
C_1 — 0.001-μfd. mica.
C_2, C_3 — 500-μμfd. midget mica.
R_1 — 25,000 ohms, 1-watt (see text).
R_2 — 1000 ohms, 1-watt (wound with 10 turns No. 16 e.).
L_1, L_2 — Grid and plate lines, ⅛-inch copper tubing, 12½ inches long, spaced approximately ¼ inch.
L_3, L_4 — Cathode lines, 3/16-inch copper tubing, 8½ inches long; heater return conductor is No. 16 rubber-covered hook-up wire.

A Coaxial-Line Oscillator

The coaxial oscillator of Figs. 1616–1617 illustrates one form of construction which may be used for a wide-range tunable oscillator with good frequency stability in the 200–300-Mc. region. These characteristics may be retained as low as 60 Mc., the added range being obtained by adding tuning capacity.

Factors contributing to the frequency stability of this oscillator include the high initial Q (approximately 5000) of the frequency-controlling coaxial grid tank and the fact that this Q is maintained at a high value by tapping the grid down on the line. Further improvement results from temperature vs. frequency compensation afforded by the C_1C_2 combination.

C_1 is a disc-type condenser mounted inside the top end of the outer conductor, which is made of 17 S-T aluminum tubing, 3 inches in diameter and 10½ inches long. The inner conductor is ¾-inch copper tubing, 8½ inches long. The "stator" of C_1, a 1½-inch circle cut from 3/32-inch aluminum sheet, is attached to the end of the inner conductor, while the "rotor" is a similar disc connected electrically to the outer conductor. The spacing between the two is varied by a threaded shaft on the rotor disc, supported by a bushing with a matching thread in the top plate of the outer conductor. With 1½-inch discs, adjustable over a maximum distance of 1 inch, the effective capacity range is somewhat greater than 1 to 10 μμfd.

The temperature-compensating operation of the C_1C_2 combination is based on the fact that C_2, like any variable condenser of conventional construction, has a slight positive temperature/frequency coefficient, as have the sum total of the other circuit components, including the tube. C_1, on the other hand, is so constructed as to have a negative temperature characteristic. The aluminum-pipe outer conductor has a temperature coefficient of approximately 25, as against 18 for the copper-tubing inner conductor. As temperature increases the outer conductor therefore expands more rapidly than does the inner, lifting the "rotor" plate away from the stator and thereby decreasing the capacity of C_1.

By properly apportioning the net capacity between C_1 and C_2, at any given frequency a substantial degree of compensation can be achieved. Assuming a total tube and circuit capacity of approximately 3 μμfd., resonance at 224 Mc. occurs with both C_1 and C_2 near their minimum capacity settings.

The reactive but non-resonant output circuit, comprised by the flat strip plate loop and the associated by-pass condenser, C_5, avoids the interaction commonly experienced with the normal tuned-grid tuned-plate circuit. It is adjusted by splitting the strip and slotting one portion, as shown in Fig. 1614. Toward the high-frequency end of the range the inductance of this loop together with the tube and circuit capacity approach resonance. At lower frequencies it serves merely as a reactive load, comparable to an r.f. choke.

For operation much below 200 Mc., the inductance of the plate loop must be increased. Alternatively a conventional coil and condenser or a resonant-line output circuit could be provided.

The plate loop is ½-inch wide, cut from 0.04-inch sheet copper. At the lower end a 1-inch square tab forms part of the plate-circuit bypass condenser. This tab is insulated from the chassis, which constitutes the central grounded

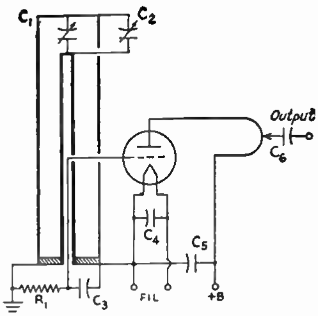


Fig. 1616 — Circuit of the high-Q coaxial-line oscillator.

C_1 — See text.
C_2 — 15-μμfd. variable for 224 Mc. (Cardwell ZT-15-AS); 35 μμfd. for 112 Mc. (Cardwell ZR-35-AS); 100 μμfd. for 60 Mc. (ZU-100-AS).
C_3, C_4 — 500-μμfd. midget mica.
C_5 — See text.
C_6 — 50-μμfd. 2500-volt mica.
R_1 — 50,000-ohm 10-watt wire-wound semi-variable.

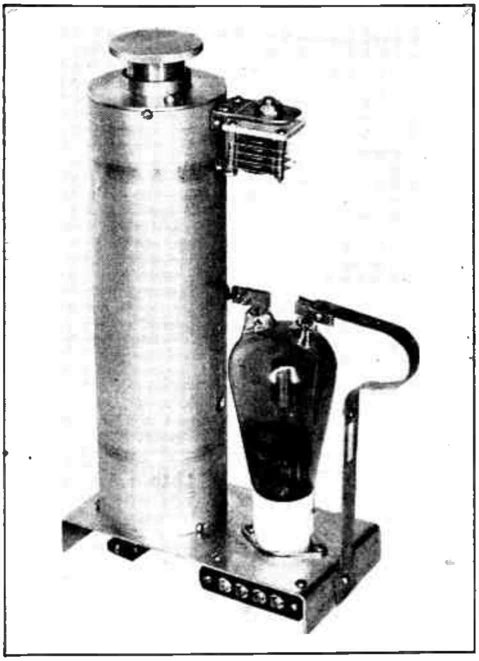


Fig. 1617 — A coaxial-line oscillator with a high-Q capacity-loaded frequency-controlling grid circuit. The plate output circuit is the flat-strip loop at the right.

plate, by two sheets of 2 or 3 mil sheet mica. A machine screw passing through the plate, the mica and an enlarged hole in the chassis connects to a second 1-inch square copper plate inside the chassis, also mica insulated, the combination forming a 3-plate mica by-pass condenser of negligible inductance.

At a plate voltage of 750, the 304-B tube shown is capable of delivering 40 watts output at 112 Mc., representing about 40 per cent efficiency, and 25 watts at 224 Mc. with 30 per cent efficiency. The grid leak, R_1, requires careful adjustment; between 15,000 and 20,000 ohms should be satisfactory. Too high a value of resistance will result in squegging.

Additional feed-back may be required to sustain oscillation at the lower frequencies. This can be provided by adding small metal tabs to the grid and plate connectors, to increase the grid-plate capacity. Only as much feed-back capacity as is necessary should be used, to avoid disturbing the frequency stability.

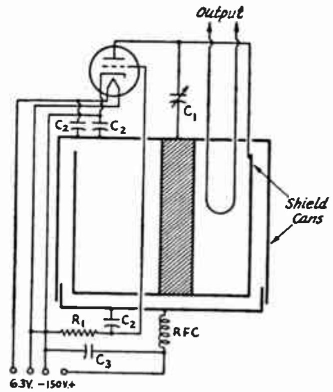


Fig. 1619 — Circuit diagram of the "pot" oscillators.
C_1 — 20-μμfd. midget variable (Hammarlund MC-20-S).
C_2 — 100-μμfd. midget mica (see text).
C_3 — 500-μμfd. midget mica (see text).
R_1 — 10,000 ohms, 2-watt carbon.
RFC — 20 turns No. 22, self-supporting, ¼-inch diameter; turns spaced one wire diameter.
Tank Circuit Dimensions:

		Diameter	*Length*
112 Mc.	— Outer cylinder:	3 inches	3¾ inches
	Inner "	2½ "	2¾ "
	Central rod:	¾ "	3½ "
224 Mc.	— Outer cylinder:	3 "	3 "
	Inner "	2½ "	2½ "
	Central rod:	¾ "	2¾ "
400 Mc.	— Outer cylinder:	2 "	3 "
	Inner "	1½ "	1½ "
	Central rod:	½ "	1¾ "

ℂ Pot Oscillators

The diagram shown in Fig. 1619 is that of a tuned-plate oscillator, using a "pot" tank (§ 2-12) and inductive coupling to the grid, used in the oscillators pictured in Fig. 1608.

The plate of the tube is connected to the open end of the inner cylinder, which corresponds to one end of a conventional coil-condenser parallel circuit, and the grid coupling is obtained by running a "tickler" wire up through the inner cylinder. Changing the position of this wire in relation to the rod changes the coupling, and hence the excitation.

Power is coupled out of the circuit by a hairpin loop of wire running parallel to the rod and in the plane of a radius. Pushing the loop further into the pot increases the coupling.

In the 112-Mc. oscillator the tuning condenser is connected between the plate of the tube and the outer shield can, or, in effect, across the tuned circuit. Since the pot is connected directly to the plate, the whole pot is at the d.c. plate voltage above ground. As a safety measure, the whole outer surface of the pot may be given several coats of clear lacquer.

The only other components in the oscillator circuit are the heater and cathode by-pass condensers, C_2, the grid leak, R_1, an r.f. choke, and the plate by-pass condenser, C_3.

With the exception of those in the 112-Mc. oscillator, all by-pass condensers are constructed by clamping thin mica sheet between copper plates or tabs and the grounded metal chassis, as previously described on page 339. For additional capacity, the strip may be held in place by placing another piece of sheet mica on the "hot" piece and holding the entire assembly in place by the topmost grounded strap, which can be attached to the chassis by machine

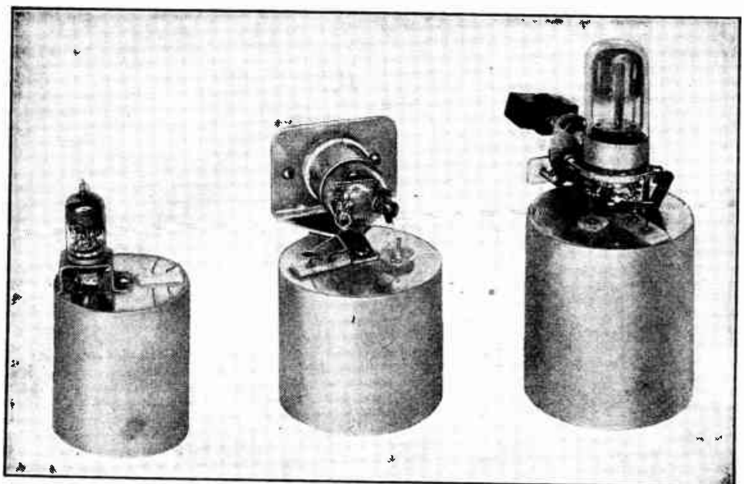


Fig. 1618 — Oscillators for 400, 224, and 112 Mc. employing "pot"-type high-Q lumped-constant tank circuits. The 112-Mc. oscillator uses a 7A4; a HY615 is used on 224 Mc., and a 9002 on 400 Mc. The latter oscillator unit is useful not only for low-power transmission but as the local oscillator in u.h.f. and s.h.f. superheterodyne receivers.

screws at either end. If a lead is required, it should be a continuation of the by-pass strip.

The grid lead and the coupling loop are insulated from the cans by cementing small strips of ⅛-inch polystyrene sheet to the can after drilling clearance holes to pass the grid lead and the coupling loop. It is advisable to fasten the by-pass condensers and the heater and cathode leads at the socket before fastening the socket to the bracket.

The bottom of the outer shield can is enclosed with a cover, with the exception of the 400-Mc. oscillator. This bottom cover supports the assembly.

The 112-Mc. oscillator is resonated by the midget condenser, C_1. The higher frequency oscillators are used at their natural resonant frequency to avoid added loss. If means of adjusting frequency over a limited range is desired, it may be achieved in the form of a rotatable vane or paddle, arranged to increase or decrease the coupling between the central pillar and the wall of the inner shield can. This vane, which may be made of copper or brass and mounted on the top of the outer can by a dial shaft-and-bushing assembly, acts as a short-circuited turn coupled to the inductance on the same principle as described on page 314.

The signal from these oscillators is remarkably stable. Drift is negligible after the first five or ten seconds. Load changes and voltage changes alter the frequency only a matter of a kilocycle or so. Heavy plate modulation of the oscillator will result in some frequency modulation, but even in this case the stability surpasses that of other self-excited oscillators. Body-capacity effects occur when the operator's hand approaches "hot" points of the circuit, but the outer can itself can be touched without throwing the beat note off more than a few kilocycles, and the same is true of the power-supply cable running from the terminal strip to the power supply. V.h.f. r.f. chokes inserted in each power lead at the terminal strip will remove any residual r.f. in the cable.

A flash-light bulb may be used as a dummy load to measure the output, a midget variable condenser being placed in series with the loop and bulb and tuned for maximum brilliance. It should be possible to overcouple the oscillator (as evidenced by double-peak tuning) without throwing it out of oscillation. The output of the 112-Mc. oscillator should be about one-quarter to one-half watt for an input of 2 watts (150 volts, 13 ma.). This represents 15 per cent efficiency, which can be considered good for an ordinary receiving tube at 112 Mc. The 2-watt input is based on a plate voltage of 150, as established by a VR-150-30 regulator tube. Plate-voltage regulation removes the "burbles" that appear in a self-excited oscillator operating on an unregulated supply.

The input can be run somewhat higher, although inputs above 4 watts (200 volts, 20 ma.) are not recommended for continuous operation.

U.H.F. Oscillators

Amateur radio after the war will have new frontiers of exploration and development opened to it in the region above 300 Mc. — the ultrahigh-frequency region. While enormous successes already have been achieved there, much development remains to be performed, particularly in the direction of adapting conventional techniques employing ordinary negative-grid tubes to function with reasonable efficiency and adequate frequency stability.

Figs. 1620 through 1623 show two acorn-tube oscillators which illustrate the type of construction commonly employed in this frequency region. The oscillator of Fig. 1621 is designed to work at approximately 400 Mc., while that of Figs. 1622–1623 — also shown, complete with antenna and reflector system, in Fig. 1814 — operates at 700–750 Mc. Both oscillators use the half-wave parallel-line circuit shown in Fig. 1620, and both employ coaxial-type antennas which are constructed as a permanent part of the oscillator.

The half-wave parallel-rod type of oscillator is used because of its unfailingly consistent operation and because of its ability to transfer a relatively large percentage of its power output into the antenna, as compared to other oscillators designed for higher stability. Although the stability of the parallel-rod oscillator is relatively poor, in many cases the lack of stability in a microwave oscillator lies not so much in the electrical design as in the mechanical construction. With this in mind, the oscillator should be required to tune smoothly, be shielded from external objects to minimize stray radiation, and have a mechanically rigid antenna system.

A length of 2-inch copper pipe is the basic structure for each oscillator. This pipe serves the dual purpose of effectively shielding the parallel rods and of forming a solid mechanical support for everything connected with the oscillator. It also provides an excellent ground for radio-frequency by-pass condensers at any point on its surface. Shielding of the parallel rods aids stability by eliminating all hand-capacity effects, and also allows perfect by-passing for the power leads.

The "socketless" mounting method used for the acorn tubes, as shown in the photographs, indicates the precautions that must be taken to prevent losses. All connections should be direct, and no insulation should be used to support the leads if they are at voltage loops. The tuning mechanism further illustrates the necessary low-loss construction.

The tuning system employed may be compared to the usual variable condenser; the surfaces of the rods serve as the stator plates of the condenser, while the movable grounded copper strip, which is varied in its spacing from the parallel rods by a threaded screw through a bushing in the shield, is analogous to the rotor.

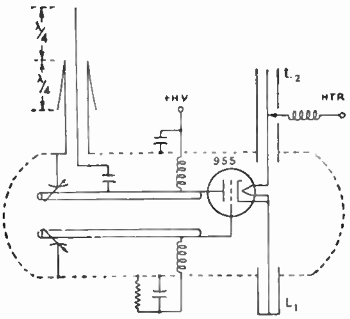


Fig. 1620 — Schematic of the shielded parallel-line microwave oscillator. The parallel resonant lines are made of 5/16-inch copper tubing, spaced 1/2-inch between centers. The lines are 4 inches long for 750–800 Mc., 10 inches long for 400 Mc. The closed filament line (L_1) for 750 Mc. is slightly longer than one quarter wave. L_2 is the length of L_1 plus a quarter wave. The 400-Mc. filament circuit is made of two trough lines, each 3/8-wave long; one line is insulated from the shell by a thin mica sheet.

Filament lines are not actually necessary for operation of the 955 tube on 400 Mc., since filament "chokes" would serve practically the same purpose. The lines are used in this case to simplify the design and to stabilize the mechanical construction, however. The use of such lines leaves no doubt as to the efficiency of the filament circuit, and, as the frequency is raised, their superiority over r.f. chokes becomes all the more pronounced.

Figs. 1621 and 1622 show two types of construction for the filament lines. The trough line of Fig. 1621 facilitates adjustment, since it is an easy matter to insert sliders between the trough and the inner conductor to adjust the electrical length for optimum performance. One trough line is fastened solidly to the shielding pipe, while the other is insulated from the pipe with thin mica sheet for the filament return connection.

Use of the coaxial line instead of the trough complicates the manner of adjustment. The line with the closed end, projecting to the right in Fig. 1622, is cut to approximate length and soldered in place with no means of adjustment, while the second line is made a quarter wave longer. The end of the longer line is at a voltage node and is left open, allowing the length of the inner conductor to be varied for filament-circuit adjustment. At the same time, it leaves one side of the filament insulated from the shield. The filament connection is made through an r.f. choke tapped at the current loop on the inner conductor.

The coaxial antenna consists of a quarter-wave radiator with a quarter-wave skirt attached to the outer conductor. The skirt in this case is made of sheet copper, bent so that the upper end fits tightly over the outside conductor, while the bottom flares out so that it will have a clearance of one-half to one inch from the coaxial line. Four or more quarter-wave wires or copper-tubing elements may be used in place of the skirt with practically the same results. In coupling the coaxial line to the oscillator, either inductive or capacitive cou-

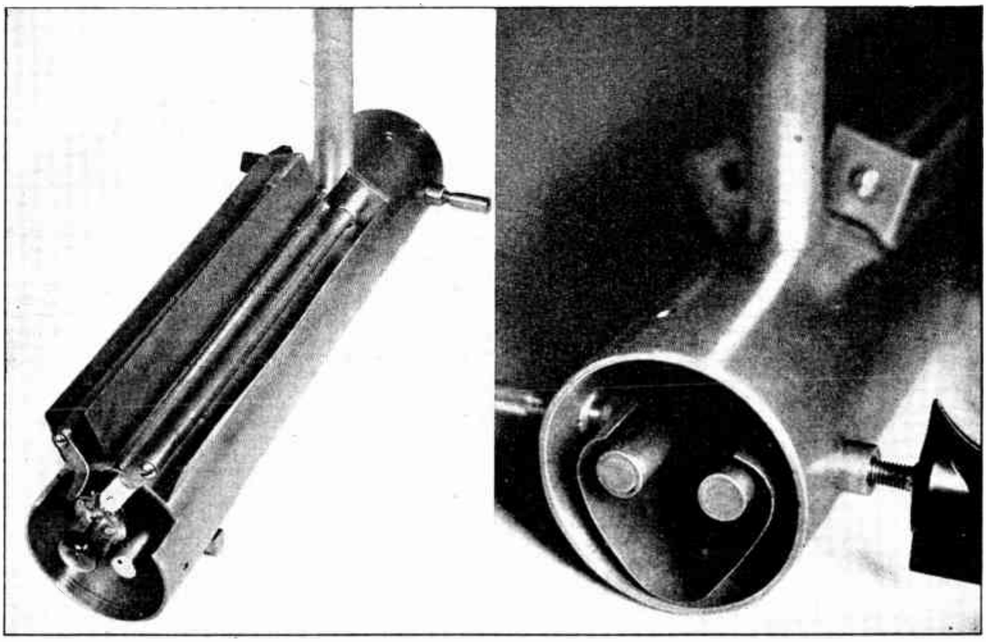


Fig. 1621 — *Left* — A general view of the 400-Mc. oscillator. The parallel-line tank circuit is mounted inside a 2-inch copper pipe which serves both as shield and as mounting base. The 955 acorn tube is mounted at the ends of the resonant line. Trough lines 3/8-wavelength long are used for tuning the filament circuit in this oscillator. The vertical pipe at the rear is the concentric line to the antenna. *Right* — The tuning mechanism of the 400-Mc. oscillator. Only one adjustment actually is used in tuning; the other balances the line for maximum output.

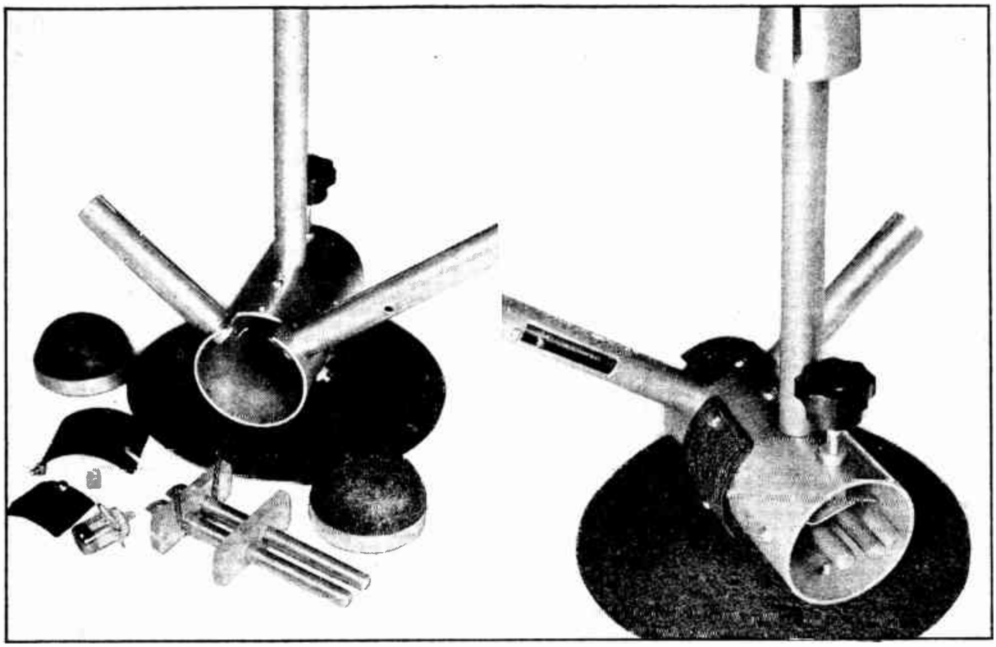


Fig. 1622 — *Left* — The 750-Mc. oscillator dis-assembled, showing the 2-inch outer shell with the filament lines and the radiator in place. The tuned-circuit parallel lines with their polystyrene insulators are in the foreground, with the feed chokes projecting upward at the terminal ends. The hemispherical end shields are of hammered copper, made to fit tightly inside the ends of the shell. *Right* — The tuning mechanism of the 750-Mc. oscillator. Turning the knob in its threaded bushing varies the spacing between the curved copper strip and the parallel-rod assembly. At the top may be seen the lower part of the "skirt" or lower quarter-wave section of the coaxial-type antenna.

pling may be used. Although ordinarily the inductive coupling is the most convenient, as the frequency is raised into the ultrahigh region the capacitive type of coupling appears to be the more satisfactory.

In the case of the 775-Mc. oscillator, there is sufficient coupling through the capacity between the rod and a copper strip lying parallel to the plate rod. This strip, a quarter of an inch wide and three-quarters of an inch long, is connected directly to the coaxial line.

The best method of adjusting the antenna coupling is through the use of a field-strength meter. A crystal detector in the center of a half-wave pick-up antenna, coupled through r.f. chokes to a 0-1 ma. meter as described in Chapter Twenty, will serve as a field-strength meter of ample sensitivity. With this type of indicator, good readings have been obtained at distances of four or five wavelengths from the transmitter with only a few watts input.

The 400-Mc. oscillator performs like a normal v.h.f. oscillator in every way. The radiated output is as good or better than that of the average 112-Mc. transmitter using the same power input, and the mechanical stability is excellent.

Since the 750–775-Mc. oscillator operates near the critical frequency of the 955 acorn tube it is somewhat more difficult to operate as a regenerative oscillator. The highest frequency at which oscillation is obtainable is approximately 800 Mc., but usable output is available at frequencies between 750–775 Mc.

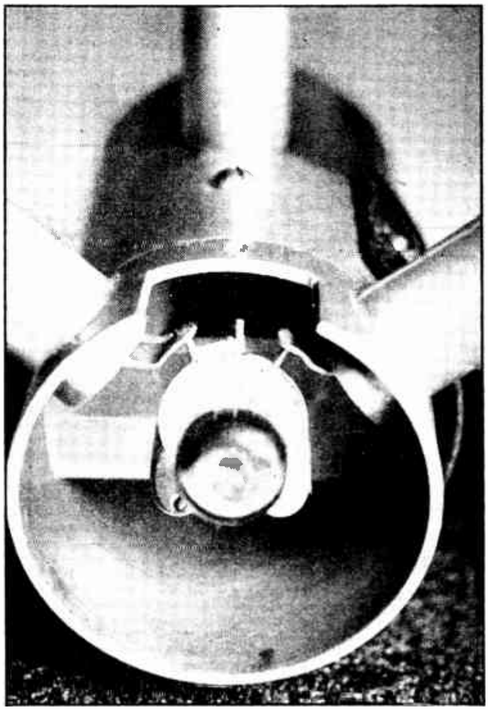


Fig. 1623 — End view of the 750-Mc. oscillator. The grid and plate pins of the 955 acorn tube are inserted in small holes in the ends of the parallel rods, with the filament connections being made through small spring connectors which also hold the tube in place. The cathode pin is strapped to one of the filament pins, as shown.

A 10-Watt 56-Mc. Transmitter

The inexpensive transmitter shown in Figs. 1624, 1625 and 1626 uses dual-triode 6A6-type tubes throughout. One section of the first tube is used as a crystal oscillator on 7 Mc. while the second half doubles to 14 Mc. The two sections of the second tube are used as 28- and 56-Mc. doublers, and the third tube is a push-pull final amplifier. Capacitive interstage coupling is employed throughout except between the 56-Mc. doubler and the final.

In the oscillator, parallel plate feed permits grounding the rotor plates of the tuning condenser. Cathode bias allows the tube to operate at low plate current; it is not necessary to obtain much power from the oscillator, since the excitation requirements of the first doubler are low.

The 14- and 28-Mc. doubler circuits are identical except for the cathode resistor, R_2, in the first doubler stage. The second doubler has no cathode bias, because as much output as possible is desirable to drive the 56-Mc. doubler. Parallel plate feed is used in both stages. The 56-Mc. doubler is series fed through an untuned plate coil. The coil is made nearly self-resonant to transfer maximum energy.

Meter switching with shunt resistors (R_7 through R_{12}) provides for measuring the plate current in each stage, although the meter is not incorporated in the transmitter itself.

The transmitter is built on a chassis measuring 3 × 4 × 17 inches. The oscillator and doubler tube sockets are mounted with the filament prongs toward the front of the chassis and the amplifier socket with its filament prongs facing the right end. The crystal socket and output terminals each are centered 1¾ inches in from the ends of the chassis. The second-doubler tuning condenser, C_3, is mounted in the center of the front wall of the chassis. The other variable condensers are located to the left and right, with ± 2-inch spacing between shaft centers. C_1, C_2 and C_3 are supported by the chassis wall, but C_4 and C_5 are mounted on small metal pillars from the upper side of the chassis. This mounting arrangement brings the shafts of C_4 and C_5 in line with the other three.

Wiring to the meter switch is simplified if the switch is located 6½ inches in from the output end. This point is also convenient to the supply ends of the plate chokes for the first three stages, so that these chokes can be mounted directly on the switch points. The shunt resistors should be soldered to the switch contacts before the switch is installed.

The filament transformer and crystal lamp are at the left end of the chassis, in the bottom view. The transformer should be kept as far as possible to the left so that it will not be near the r.f. circuits. The lamp is held firmly in the grommet by stiff leads soldered to its base. The plate-supply terminals are out of the way at the extreme left end of the base. Two positive terminals are provided, so that a modulator transformer secondary may be connected in the plate lead of the final amplifier.

The rest of the parts are mounted so r.f. leads will be short and direct, particularly in the last two stages. The grid connections in the amplifier should be made directly between the grid prongs of the socket and the stator plate terminals of the grid tank condenser. The plate prongs and the stator sections of C_5 should be cross-connected, so that the neutralizing condensers, C_6 and C_7, may be supported by the condenser lugs, as shown in Fig. 1625. This gives leads of negligible length and perfect symmetry, both of which contribute to good neutralizing. Trimmer-type condensers can be used for neutralizing since the neutralizing capacity required is small and the effective dielectric is mostly air. The output coupling coil has its ends soldered to lugs which are held in place by the feed-through terminals. The lugs will bend as the position of the coil is varied to change the coupling.

Each tank circuit will be in resonance when adjusted for minimum plate current to the tube with which it is associated. The current values should be 10, 18, 18 and 40 ma., in the order listed, for the first four stages. It is quite possible that the values will vary slightly in different layouts, but they should be approximately as given. Tuning of the various tanks should be adjusted to obtain maximum output from the 56-Mc. doubler, as indicated by maximum grid current in the final-amplifier grid leak, R_6. If no grid current is obtained, it is probably an indication that the coupling between L_4 and L_5 is either too tight or too loose; this coupling is quite critical,

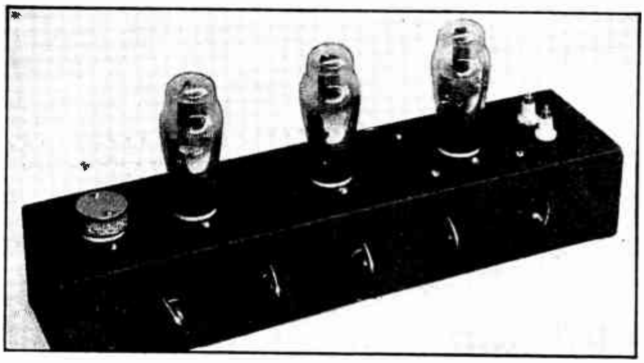


Fig. 1624 — In this front view of the 10-watt 56-Mc. transmitter the oscillator, doubler and amplifier tubes are from left to right. The crystal socket is at the left end of the chassis and the output terminals are at the right. The tuning controls are arranged in line along the front wall of the chassis. Plate-voltage terminals, meter switch, meter cord and 115-volt line cord, and the crystal-current bulb, mounted in a rubber grommet are at the rear.

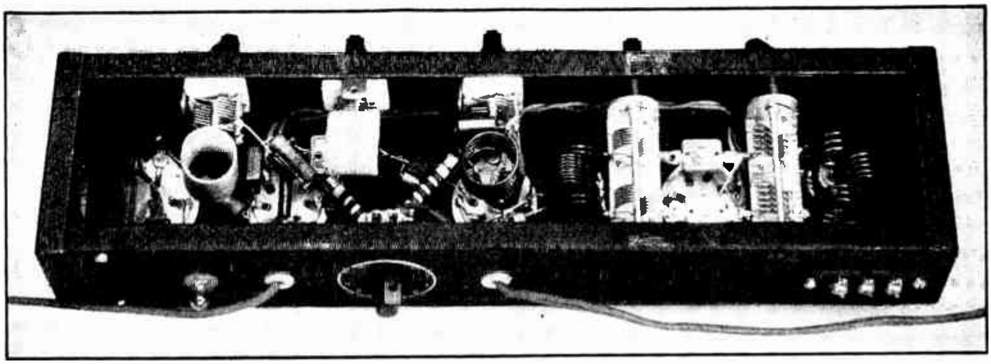


Fig. 1625 — This bottom view shows how the tuning condensers are mounted with respect to the tube sockets. The self-supporting coils mount directly on the tuning condensers. Filament transformer is in lower left-hand corner.

and therefore deserves careful adjustment. The amplifier grid current should be 25 ma. or more when the coupling is optimum. Each time the coupling is changed, the grid condenser, C_4, as well as the preceding tuning condensers should be readjusted.

After a grid-current indication is obtained, the amplifier should be neutralized. Plate voltage must be removed from the final amplifier but the rest of the circuits should be in normal operating condition. Start with the plates of the neutralizing condensers screwed up tight and then back off a full three turns on each condenser. This places the neutralizing capacities at approximately the correct values. Condenser C_5 is then rotated through resonance, which will be indicated by a kick in the grid current. Adjust the neutralizing condensers in small steps, turning both screws in the same direction and the same amount each time, until the grid current remains stationary when C_5 is rotated. This indicates complete neutralization. Retune the grid circuit after neutralization, so that maximum excitation will be secured; also recheck the coupling between L_4 and L_5, since neutralization will change the load on the driver somewhat.

Plate voltage may now be applied to the amplifier. When the plate tank is tuned to resonance, the plate current should fall to 20 or 25 ma. A load, such as an antenna or feeder system or a 10-watt lamp used as a dummy antenna, should be connected and the coupling adjusted until the plate current reaches the full-load value of 60 ma. The grid current will fall off to 10 ma. or so when the amplifier is loaded.

At the recommended input of 21 watts (60 ma. at 350 volts), the output as measured in a dummy antenna is something over 10 watts.

To modulate the transmitter 100 per cent, about 11 watts of audio power is required. The modulator output transformer must match an impedance of 5833 ohms (modulated-amplifier

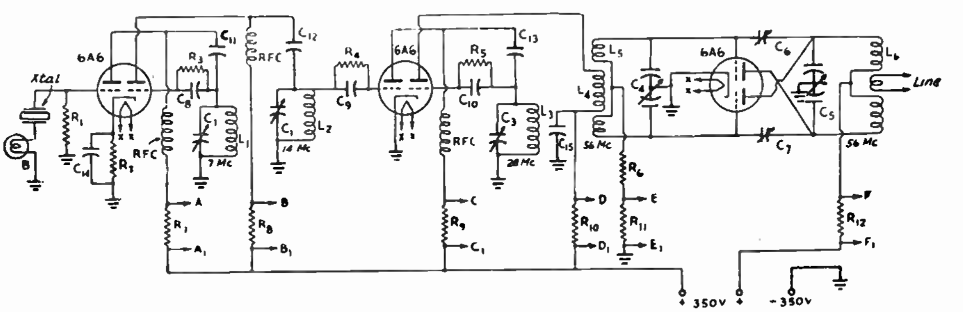


Fig. 1626 — Wiring diagram of the 10-watt 6A6 dual triode crystal-controlled 56-Mc. transmitter-exciter unit.

C_1 — 50-μμfd. variable (Hammarlund HF-50).
C_2 — 35-μμfd. variable (Hammarlund HF-35).
C_3 — 15-μμfd. variable (Hammarlund HF-15).
C_4 — 50-μμfd. per section dual variable (Hammarlund HFD-50).
C_5 — 15-μμfd. per section dual variable (Hammarlund HFD-15-X).
C_6, C_7 — 3-30-μμfd. mica trimmer (National M-30).
C_8, C_9, C_{10} — 100-μμfd. midget mica.
C_{11}, C_{12}, C_{13}, C_{14}, C_{15} — 500-μμfd. midget mica.
R_1 — 15,000 ohms, ½-watt.
R_2 — 500 ohms, 1-watt.
R_3, R_4, R_5 — 30,000 ohms, ½-watt.
R_6 — 1000 ohms, 1-watt.
R_7, R_8, R_9, R_{10}, R_{11}, R_{12} — 25 ohms, ½-watt.
RFC — 2.5-mh. r.f. choke (National R-100).
B — 60-ma. pilot bulb.
L_1 — 21 turns No. 22 d.s.c., close wound, 1-inch diameter.
L_2 — 11 turns No. 22 d.s.c., 1 inch long, 1-inch diameter.
L_3 — 6 turns No. 14, ¾ inch long, 1-inch diameter.
L_4 — 9 turns No. 14, ⅝ inch long, ¾-inch diameter.
L_5 — 2 turns No. 12 each side of L_4, 1-inch diameter, center opening ¾ inch. Turns spaced diameter of wire.
L_6 — 3 turns No. 12 each side of coupling link, ⅞-inch diameter, center opening ¾ inch. Turns spaced diameter of wire.
Link — 5 turns No. 12, ⅞-inch diameter, ½ inch long.

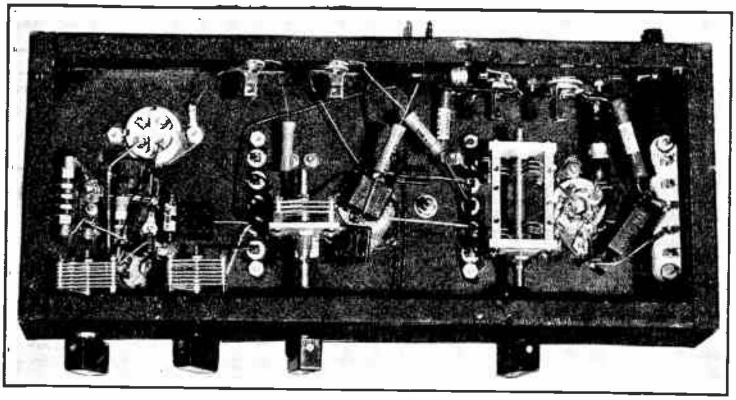


Fig. 1630 — A view underneath the chassis of the 815-tube transmitter. By-pass condenser leads are kept as short as possible, and r.f. leads are made with heavy wire.

parable transmitter, with a few minor modifications. If the oscillator plate coil, L_1, is wound as specified, the crystals should oscillate with condenser C_1 about half meshed. Coils L_2 and L_3 are next wound and plugged in, and a 0–10 milliammeter is connected in the grid circuit of the 7C5. With the crystal oscillating (as indicated by a neon bulb touched to the "hot" end of L_1, or by monitoring the signal in a receiver), C_2 and C_3 should be tuned for maximum grid current to the 7C5. There will be some interlocking of the tuning of these two circuits if the coupling between L_2 and L_3 is too tight, in which case the two coils should be moved in relation to each other until practically "one-spot" tuning is obtained. The grid current to the 7C5 should be 1.5 to 2 ma. with 275 volts on the plate of the 7N7.

Next, the 7C5 plate coil, L_4, should be wound, leaving off L_5 for the time being since L_4 is to be made self-resonant. If it is in resonance, as indicated by a neon bulb touched to the plate end, L_5 may be wound on; but if not, the turns should be pushed together or pulled apart until signs of r.f. can be seen. When L_5 is added the coupling should be made rather loose at first, since it will be found that there is more than enough drive available as indicated by the grid current to the 815. The coupling can be increased until the grid current is 6 or 7 ma. It will not be possible normally to obtain more than 4 to 5 ma. grid current to the 815 on 112 Mc., even with the two coils tightly coupled, but this value is sufficient to drive the tube. The plate current of the 7C5 will be between 35 and 40 ma.

After grid current has been obtained in the 815 stage (with no plate or screen voltage applied), the tank circuit should be tuned to resonance as indicated by a sharp flicker of the grid current. The plates of the neutralizing condensers should then be moved with respect to the tube until no flicker remains, indicating that the tube is neutralized. Neutralization should always be checked when shifting from one band to another, since an accidental jarring of the tube or some unbalance in the stage may affect the adjustment. If the stage is correctly neutralized, it will be possible to apply several hundred volts (more might injure the tube) to the screen and plate and, with no bias, the excitation off and the plate tank unloaded, be able to detect no signs of r.f. anywhere in the circuit with any setting of the tank condenser. Unless the stage is neutralized, it will be impossible to modulate the amplifier fully without distortion.

When the amplifier has been neutralized, plate voltage may be applied with the excitation on. The amplifier may then be loaded up to its rating of 150 ma. at 400 volts. About 30 watts of audio power will be required to modulate the 815 stage fully.

The linear tank is tuned by sliding the shorting bar (two National metal-tube grid clips soldered together) up and down the bar until resonance is indicated. The bar has plate voltage on it, and all tuning should be carefully done with an insulated screwdriver. The antenna coupling is by means of two similiar grid caps which can be moved up and down the lines until proper loading is obtained.

If f.m. is to be used on the 56- or 112-Mc. bands the frequency-modulated oscillator can be coupled in through several turns around L_1, with crystal removed from its socket. An f.m. oscillator is shown in Figs. 1634, 1635 and 1636.

More than enough excitation is available on the 28- and 56-Mc. bands, but there is no advantage in running the grid current above 4 or 5 ma. since the output will not increase and the linearity of the amplifier is good with only 3 to 4 ma. grid current. The r.f. output of the transmitter could not be measured accurately above 28 Mc. (where it was close to 40 watts) but, with the lamp loads used, it appeared to be about 35 watts on 56 Mc. and 30 watts on 112 Mc., at the rated input of 60 watts.

A 275-volt and a 400-volt power supply are required for this transmitter. Figs. 1315 and 1304 show suitable power supplies. A pair of 6L6s in Class AB will furnish enough audio power to modulate fully the 815; such a modulator unit is shown in Fig. 1404.

A 300-Watt 112- and 56-Mc. Driver-Amplifier

The driver-amplifier shown in Figs. 1631–1633 uses a pair of 35Ts or 35TGs in the output stage running at 200 ma. with 1500 volts on the plates. With another 35T or 35TG to provide driving power, the efficiency and performance is excellent. The exciter diagrammed in Fig. 1630 is a suitable driver, although any transmitter of comparable output will serve.

The push-pull output amplifier is built on a 5½ × 9½ × 1½-inch metal chassis. It and the driver-amplifier are mounted on a panel of Presdwood. Two panel brackets are used as supports for the amplifier chassis, one at each end of the chassis. The final tank plate tuning condenser, C_{12}, is mounted on one of these brackets and insulated from it by small steatite bushings. The grid tuning condenser, C_7, is mounted on the chassis and the rotor is left unconnected. Both grid and plate tuning condensers are adjusted by an insulated screwdriver through holes in the panel. In the case of the driver, dials on insulated extension shafts are used for tuning adjustments.

The final plate tank coil is mounted on a Millen 40205 plug base, and the corresponding socket is supported on the tank condenser by two small brass angles. The grid coil is mounted on a National PB-16 plug, and its XB-16 socket is raised above the chassis by small steatite stand-off insulators.

The neutralizing condensers, C_{10} and C_{11}, are mounted next to the tubes, immediately under C_{12}. They are ganged by an insulated flexible coupling linking the rotor shafts. If 35Ts are used, as shown in the photographs, the lower neutralizing condenser plates are connected through bushings running through the chassis to the grid terminals on the sockets. If 35TGs are used, it is not necessary to carry the grid leads through the chassis. In either event, the grid leads from the neutralizing

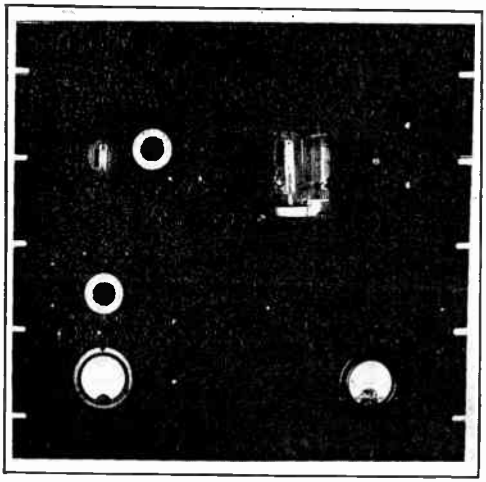


Fig. 1632 — The 300-watt 112- and 56-Mc. driver-amplifier is built on a Presdwood panel. Meters read driver cathode current and final amplifier plate current.

condensers must be crossed to provide the proper phasing for neutralization.

The final amplifier chassis is fastened to the panel by the brackets referred to above. A rectangular hole in the center of the panel allows the tubes to be observed during operation.

The driver stage is built up directly on the Presdwood panel. The tube socket is supported on an aluminum bracket approximately 3 inches square. The input and output tuned circuits are mounted at suitable locations be-

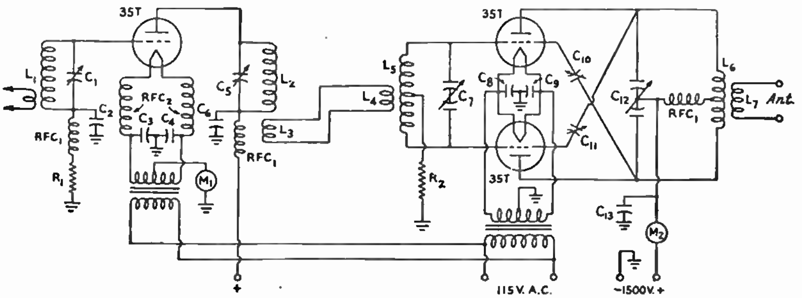


Fig. 1631 — Wiring diagram of the 300-watt output driver-amplifier for 112- and 56-Mc. operation.

C_1, C_5 — 15-μμfd. variable (Hammarlund HFA-15-E).
C_2, C_8, C_9 — 0.001-μfd. mica.
C_3, C_4 — 0.002-μfd. mica.
C_6, C_{13} — 0.001-μfd. mica, 5000-volt rating.
C_7 — 25-μμfd. per section (Cardwell ER-25-AD).
C_{10}, C_{11} — 1.5–4-μμfd. neutralizing condensers (Cardwell ES-4-SDI).
C_{12} — 35-μμfd. per section (Millen 13035).
R_1 — 50,000 ohms, 10-watt.
R_2 — 2500 ohms, 10-watt.
RFC_1 — V.h.f. r.f. choke (Ohmite Z-0).
RFC_2 — 9 turns No. 14 e., self-supporting, closewound on ⅜-inch diameter.
M_1 — 0-150 ma.
M_2 — 0-300 ma.
T_1 — 5-volt 4-ampere filament transformer (Thordarson T-63F99).
T_2 — 5-volts 8-ampere filament transformer (Thordarson T-19F84).

L_1 — 56 Mc. (28-Mc. input): 6 turns No. 12, 1½-inch diameter, spaced to occupy 2 inches. Link, 3 turns No. 14 outside ground end of coil.
112 Mc. (56-Mc. input): 4 turns No. 12, 1-inch diameter, spaced to occupy 2 inches. Link, 2 turns No. 14 outside ground end of coil.
L_2 — 56 Mc.: 4 turns No. 12, 1-inch diameter, spaced to occupy 2 inches. Link, 2 turns No. 14 inside coil.
112 Mc.: 2 turns No. 12, 1-inch diameter, ½-inch spacing. Link, 1 turn.
L_5 — 56 Mc.: 6 turns No. 12, ¾-inch diameter, spaced to occupy ⅞ inch. Link is 2 turns No. 14, 1½-inch diameter.
112 Mc.: 2 turns No. 12, ¾-inch diameter, spaced ¼-inch. Link is 1 turn, 1-inch diameter.
L_6 — 56 Mc.: 4 turns No. 12, 1½-inch inside diameter, spaced diameter of wire with ¾-inch gap in center to accommodate 3-turn swinging link of No. 14 wire.
112 Mc.: 2 turns No. 12, 1-inch inside diameter, spaced ½-inch. Link is 2 turns, 1-inch diameter.

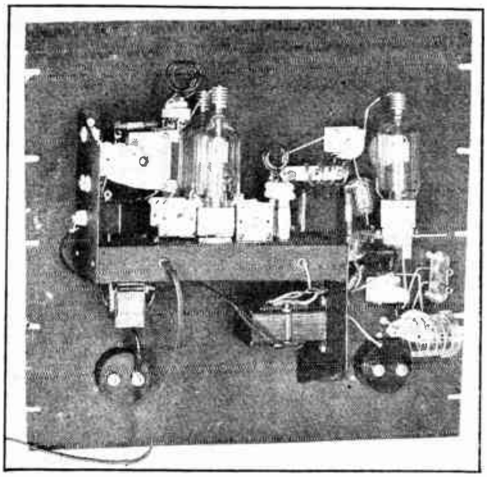


Fig. 1633 — Behind the panel of the 300-watt amplifier.

low and above this sub-base. Filament connections, via r.f. chokes which stabilize operation at 112 Mc., are made direct to a separate filament transformer mounted on the near panel bracket supporting the amplifier chassis.

The plug-in driver plate-coil socket base, which is also a National XB-16, together with its tuning condenser, C_5, are mounted on the panel in a position suited for convenient link coupling to the final amplifier grid circuit. C_5 must be of the type which has the rotor contact at the center of the shaft, to preserve circuit symmetry.

The driver grid-coil socket base, which in this case is a standard 5-prong ceramic coil socket, and C_1 are similarly mounted. Leads for the excitation input to the driver from the link on L_1 are brought to binding posts mounted on a polystyrene bracket at the grid end of the chassis.

The 35T driver is operated as a frequency doubler on both 56- and 112-Mc., requiring 28-Mc. input in the first instance and 56-Mc. excitation in the second.

Cathode current to the driver stage, as read on M_1, will be approximately 50 ma. for 56-Mc. output and 75 ma. for 112-Mc. operation, under typical conditions.

Sufficient excitation is available from the driver on 112 Mc. for a grid current of 55 ma. or more when the final plate circuit is loaded to 200 ma. at 1500 volts. The amplifier can be loaded to 225 ma. or higher if the grid current does not drop below this value with the increased plate loading. The plates of the tubes will run a dull red under normal operation; if the two tubes do not show the same color, it indicates unbalance in the circuit. However, with the amplifier laid out as shown, no trouble of this sort should occur.

At 300 watts plate input, approximately 150 watts of audio power will be required to modulate fully the output of the amplifier. A pair of HY40Zs is recommended for the modulator unit. Other appropriate combinations will be found in Table II of Chapter Fourteen.

A suitable power supply, delivering the necessary 200 ma. at 1500 volts for the output stage and 50–75 ma. for the driver, can be built along the lines of the supply described in Fig. 1348, substituting lower-current filter chokes and power transformer.

An F.M. Modulator-Oscillator Unit

Apart from the requirement for a means of varying the frequency of the oscillator output in accordance with the applied modulating voltage, the r.f. circuits of a frequency-modulation transmitter for amateur use differ little from standard v.h.f. practice, with the exception of the oscillator circuit. Any of the multistage exciter or amplifier units described in this chapter, therefore, may be adapted for use on f.m. as well as a.m.

Where a crystal- or e.c.o.-controlled transmitter for the 28-, 56- or 112-Mc. band is available, it is a relatively easy matter to disconnect the regular plate modulator and substitute for the crystal or e.c.o. the f.m. oscillator-modulator shown in Figs. 1634, 1635 and 1636. The r.f. output of the unit is intended to be fed through a link to a tuned circuit wound on a five-prong coil form which substitutes for the crystal-holder in the crystal oscillator. This tuned circuit is resonant at the same frequency as the output tank of the control unit, L_2C_3 in Fig. 1635, and is, in fact, identical in construction.

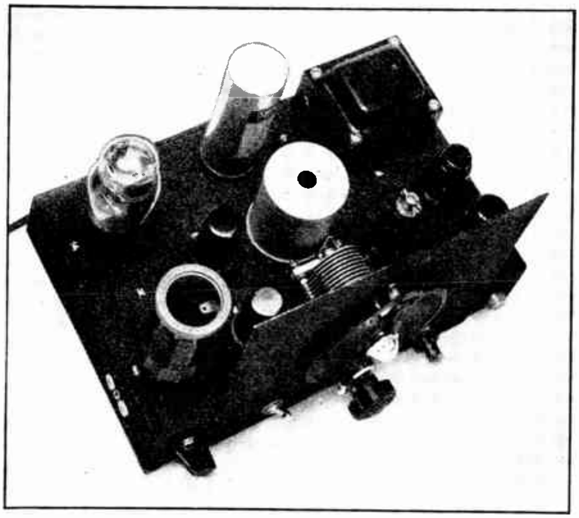


Fig. 1634 — This modulator-oscillator unit is used with normally crystal-controlled v.h.f. transmitters for frequency-modulated output. It contains a speech amplifier and power supply, so that no additional equipment is needed. The oscillator coil is in the round shield can in the center. The coil in the left foreground is the buffer output circuit. The speech amplifier and modulator are at the right, with the power supply along the rear. A 7 × 11-inch chassis is used.

In transmitters using triode oscillators, or pentode crystal oscillators in which the tubes are not well screened, it is advisable to use the crystal oscillator tube as a doubler rather than as a straight amplifier. If the transmitter uses a 7-Mc. crystal oscillator, for example, the output of the unit may be on 3.5 Mc. and the grid circuit of the ex-crystal tube also tuned to 3.5 Mc. This will avoid difficulty with self-oscillation in the ex-crystal stage. With a pentode oscillator it is possible to work straight through, provided the grid tank substituted for the crystal is tuned well on the high-frequency side of resonance, but this procedure is not advisable since it may make the modulation nonlinear. It is rather important that all circuits in the transmitter be tuned "on the nose" for best performance. Of course, if the crystal tube is a well-screened transmitting type it can be used as a straight amplifier.

With harmonic-type crystal oscillators the input frequency can be the same as that of the crystal, since the output frequency of the crystal tube is already a harmonic. In the Tri-tet oscillator, the cathode tank should be short-circuited; in the types using a cathode impedance to provide feed-back, this impedance also should be shorted. Care should be taken to avoid short-circuiting the grid bias, whether from a cathode resistor or grid leak. In the latter case this usually will mean that a blocking condenser (500 μμfd. or larger) should be connected between the "hot" end of the grid tank and the grid of the ex-crystal tube, with the grid lead (and choke) connected on the grid side of the condenser. Such a blocking condenser may be incorporated in the plug-in tank. The grid-tank tuning condenser may be a small air padder mounted in the coil form.

Where a suitable power supply and speech amplifier are already available, the lower part of Fig. 1635 can be omitted and only the oscillator, buffer and modulator units need be built. Transformer input to the modulator may be used where the available speech amplifier happens to have a low-impedance output circuit. The transformer and gain control connect

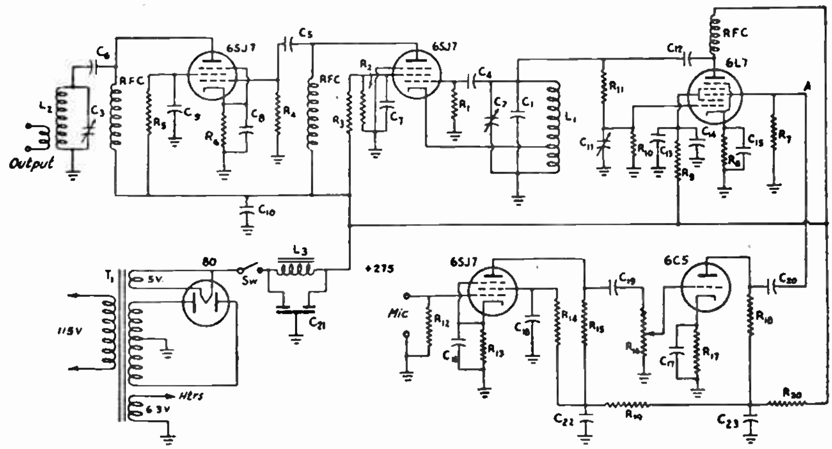


Fig. 1635 — **Circuit diagram of the f.m. control unit for use with normally crystal-controlled v.h.f. transmitters.**

C_1 — 150-μμfd. silvered mica for 7 Mc.; 650 μμfd. for 3.5 Mc.; 1150 μμfd. for 1.74 Mc.
C_2 — 100-μμfd. variable (National SE-100).
C_3 — 50-μμfd. variable (Hammarlund HF-50).
C_4 — 100-μμfd. mica.
C_5, C_{12} — 250-μμfd. mica.
C_6 — 0.001-μfd. mica.
C_7, C_8, C_9, C_{10}, C_{13}, C_{15}, C_{19}, C_{20} — 0.01-μfd. paper.
C_{11} — 3–30-μμfd. mica trimmer.
C_{14}, C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
C_{16}, C_{17} — 10-μfd. 25-volt electrolytic.
C_{18} — 0.1-μfd. 200-volt paper.
C_{21} — Dual 450-volt 8-μfd. electrolytic.
R_1 — 0.1 megohm, 1-watt.
R_2 — 25,000 ohms, 1-watt.
R_3, R_4, R_5, R_{11} — 50,000 ohms, 1-watt.
R_6, R_8 — 300 ohms, ½-watt.
R_7, R_{10} — 0.5 megohm, ½-watt.
R_9 — 30,000 ohms, 1-watt.
R_{12} — 5 megohms, ½-watt.
R_{13} — 900 ohms, ½-watt.
R_{14} — 1 megohm, ½-watt.
R_{15}, R_{19} — 0.25 megohm, ½-watt.
R_{16} — 0.5-megohm volume control.
R_{17} — 2000 ohms, ½-watt.
R_{18} — 50,000 ohms, ½-watt.
R_{20} — 0.15 megohm, 1-watt.
RFC — 2.5-mh. r.f. choke.
L_1 — 7 Mc.: 10 turns No. 18 e., length ¾ inch, 1-inch diameter, tapped 3rd turn from ground.
3.5 Mc.: 11 turns No. 24 e., length ¾ inch, 1-inch diameter, tapped 4th turn from ground.
1.75 Mc.: 21 turns No. 24 e., length 1 inch, 1-inch diameter, tapped 6th turn from ground.
L_2 — 14 Mc.: 10 turns No. 18.
7 Mc.: 20 turns No. 18.
3.5 Mc.: 40 turns No. 24.
1.75 Mc.: 75 turns No. 26.
All coils wound with enameled wire on 1½-inch diameter forms (Hammarlund SWF-4). 1.75-Mc. coil close-wound; others spaced to a length of 1½ inches.
Link 3 to 5 turns (not critical).
L_3 — Filter choke, 10 henrys, 40 ma.
T_1 — 250-0-250 volts, 40 ma.; 6.3 volts at 2 amperes; 5 volts at 2 amperes. (Thordarson T-13R11).
Sw — S.p.s.t. toggle switch.

Note: **Data for L_1 is subject to individual trimming for proper frequency coverage. Adjust inductance by changing turn spacing to bring low frequency end of band near maximum capacity on C_2. Coil specifications given apply to coil centered in round shield 2 inches in diameter and 2½ inches high. The 3.5- and 7-Mc. coils give full coverage of the 56–60-Mc. band with C_2 100 μμfd.; the 1.75-Mc. coil will cover approximately 57–60 Mc. with the same condenser.**

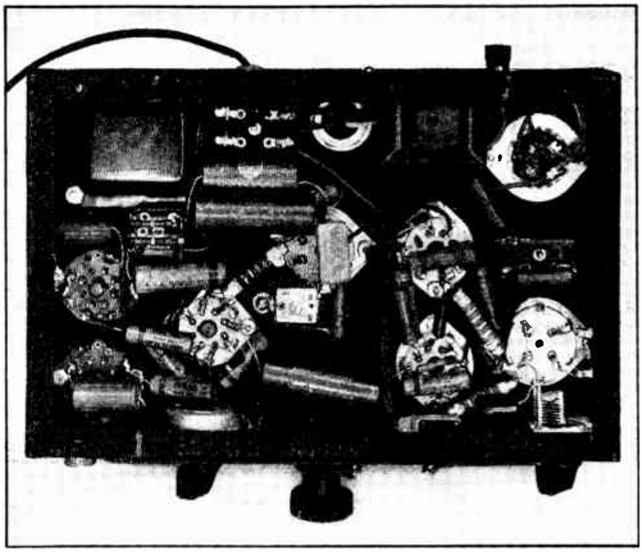


Fig. 1636 — In this bottom view, the r.f. section is at right and the audio at left. The oscillator socket is to the right of the coil socket in the center.

between ground and point "A" of Fig. 1635. R_7 being omitted. Any of the conventional methods may be used to couple the modulator to an available speech amplifier, with one precaution — if a high-impedance connection is used, the "hot" lead should be shielded to prevent hum pick-up.

If the transmitter to be used has a self-excited oscillator, electron-coupled or otherwise, a separate oscillator need not be built. The reactance modulator can be connected directly across the tank circuit of the oscillator. If the oscillator has too high a C/L ratio, not enough deviation may be obtained without distortion. It is advisable to use an L/C ratio in the oscillator comparable to those given in Fig. 1635.

A Complete 56-Mc. F.M. Transmitter

The transmitter shown in Figs. 1637, 1638 and 1639 will yield a frequency-modulated carrier output of approximately 7 watts on 56 Mc., using a plate power supply delivering 300 volts.

A reactance modulator stage, utilizing a 6SA7 reactance tube to modulate a 6F6 oscillator, is incorporated in the unit, along with a microphone input transformer. A single-button microphone is sufficient to drive the 6SA7, no additional speech amplification being required.

For complete flexibility and wider utility provision for alternative amplitude modulation may be made, as well. If it is desired to use amplitude modulation, the gain control on the reactance modulator should be set at zero and the necessary 6 watts of audio connected in series to the plate and screen lead of the 7C5 output amplifier. Used as an f.m. transmitter, the entire unit requires 300 volts at about 90 ma., making it ideal to run from a vibrator pack for portable/mobile work.

A single-button carbon microphone is transformer-coupled to the 6SA7 reactance modulator, which is connected across the tank circuit of the 6F6 e.c.o. A VR150-30 stabilizes the voltage across the oscillator and modulator and aids materially in keeping the mean frequency constant. The grid circuit of the e.c.o. tunes from 14 to 15 Mc. with a slight margin at either end of the tuning range, and the plate circuit of the e.c.o. is tuned to 28 Mc. by means of a self-resonant coil which is adjusted for maximum output by squeezing the turns together or spreading them apart. Once adjusted, it need not be touched for any change in tuning conditions. The 28-Mc. output of the e.c.o. drives a 7G7/1232 doubler to 56 Mc., which in turn drives the output amplifier.

With a 300-volt supply, the 7C5 final-amplifier grid current should be about 0.6 ma. under load for linear amplitude modulation. If f.m. is used exclusively, the grid current can be lower with no harmful effect other than a slight de-

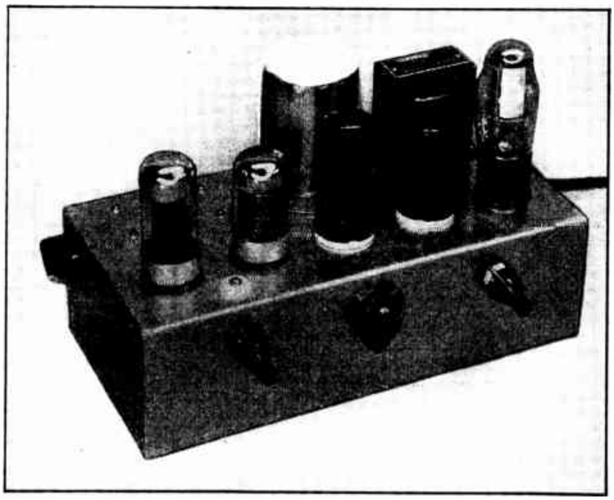


Fig. 1637 — The complete 56-Mc. f.m. transmitter has all r.f. components mounted under the chassis with the exception of the oscillator grid coil, which is housed in the shield can in the rear center of the chassis. The tubes, from left to right, are 7C5 output amplifier, 7G7 doubler, 6F6 e.c.o., 6SA7 reactance modulator and VR150-30 voltage regulator.

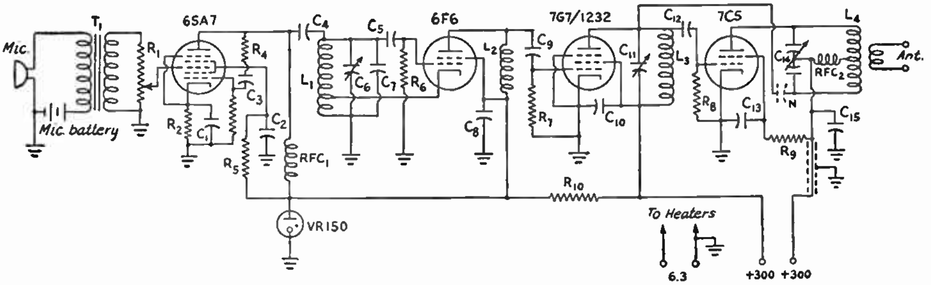


Fig. 1638 — Wiring diagram of the complete 56-Mc. 7-watt frequency-modulated transmitter.

C_1 — 0.01-μfd. 400-volt paper.
C_2 — 8-μfd. 450-volt electrolytic and 0.005-μfd. mica in parallel.
C_3 — 0.001-μfd. mica.
C_4 — 500-μμfd. mica.
C_5, C_9, C_{12} — 100-μμfd. mica.
C_6 — 15-μμfd. midget variable (Hammarlund HF-15).
C_7 — 25-μμfd. silvered mica.
C_8, C_{10}, C_{13} — 0.005-μfd. mica.
C_{11} — 35-μμfd. midget variable (Hammarlund HF-35).
C_{14} — 35-μμfd. per section dual variable (Cardwell ER-35-AD).
C_{15} — Two 500-μμfd. mica, one at each end of rotor of C_{14}.
N — Neutralizing condenser (see text).
R_1 — 100,000-ohm volume control.
R_2 — 750 ohms, ½-watt.
R_3 — 0.25 megohm (not marked in diagram).
R_4 — 50,000 ohms, ½-watt.
R_5 — 5000 ohms, ½-watt.
R_6 — 25,000 ohms, ½-watt.
R_7 — 0.1 megohm, ½-watt.
R_8 — 75,000 ohms, ½-watt.
R_9 — 5000 ohms, 1-watt.
R_{10} — 3000 ohms, 10-watt.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — V.h.f. choke (Ohmite Z-1).
T_1 — Microphone transformer (Thordarson T-58A37).
L_1 — 10½ turns No. 20 e., spaced to occupy 1 inch on a 1-inch diameter form; cathode tap 2½ turns up. Plugged into socket on chassis.
L_2 — 14 turns No. 20 e., spaced to occupy 1⅛ inches, 9/16-inch diameter; self-supporting (see text).
L_3 — 4 turns No. 20 e., ½-inch diameter, ¼-inch long.
L_4 — 6 turns No. 14 e., ¾-inch inside diameter, wound to occupy 1-inch length, with ⅜-inch gap in center for swinging link of 2 turns No. 14 e., same diameter.

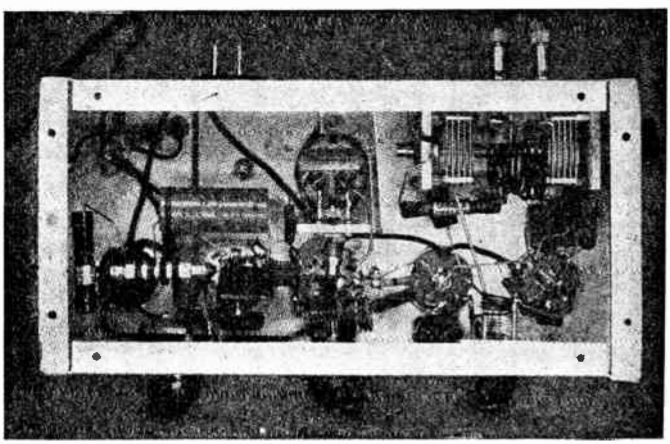
crease in output of the amplifier. The 7C5 final amplifier is plate neutralized by running a length of stiff wire from the plate side of the doubler tuning condenser over to a point near the open side of the final-amplifier split-stator tuning condenser. The capacity from this wire to the stator of the condenser may be adjusted to neutralize the final amplifier by cutting the end of the wire, a bit at a time, until the plate-tank tuning shows no reaction on the grid current (with both plate and screen voltage off).

No difficulty should be encountered in adjusting the transmitter other than setting the e.c.o. coils to the proper frequencies. The grid coil should be adjusted to cover the proper range with the reactance modulator tube in the circuit. The range can be varied by pushing the turns together or spreading them apart, while checking the resulting frequency on a calibrated receiver. The e.c.o. plate coil can best be adjusted by reading grid current to the final amplifier (by connecting a 0-1 ma. d.c. milliammeter between R_8 and ground) and adjusting L_2 until the 7C5 grid current is a maximum with the oscillator set at 14.5 Mc.

The plate current of the final amplifier will be about 45 ma. when the stage is properly loaded. The loading is varied by changing the position of the "swinging link" fastened to the antenna output binding posts.

When using f.m. the amount of deviation is controlled by the setting of the gain control, R_1. With the gain control wide open the deviation is over 30 kc. on 58.5 Mc., which is more than adequate for all purposes. When the receiving station does not have a regular f.m. receiver, the signal can be received on a conventional receiver by reducing the deviation at the transmitting end and tuning the signal off to one side of resonance at the receiving end.

Fig. 1639 — A view underneath the chassis of the 56-Mc. f.m. transmitter shows the volume control at the left, the oscillator control at the center, the doubler tuning control at the right, and the final amplifier tuning control at the side. The microphone connector is on the left side of the chassis, and the four-prong plug and flexible wire connect to the power supply and microphone battery, respectively. Note the shield between the final tuning condenser and the oscillator tuning condenser, to reduce reaction between the two circuits, and the wire running from the doubler tuning condenser to near the final tank condenser which is used as a neutralizing condenser (N in Fig. 1523). The output connects to two binding posts on a Victron strip.



Chapter Seventeen

The War Emergency Radio Service

Since the suspension of all amateur operating as such, civilian amateur radio licensees have found many practical operating and constructional opportunities in the War Emergency Radio Service. The WERS is a temporary wartime communication service established to aid in the protection of civilian life and property in the event of enemy attack or natural disasters. It shall be the purpose of this chapter to demonstrate the workings of the three categories of WERS (Civilian Defense, State Guard and Civil Air Patrol), and to show how the equipment and personnel of the stay-at-home radio amateur can best be utilized in this service.

In the original rules governing this service, issued in June, 1942, operation was authorized to two kinds of organizations, civilian defense and state guard. The civilian defense organization is under the supervision of the U. S. Office of Civilian Defense, Protection Branch, and is executed by the U. S. Citizens Defense Corps, an organization of enrolled civilian volunteers established within OCD. Civilian radio amateurs have played a greater part in CD-WERS (civilian defense WERS) than in either of the other two categories.

Practically every state now has its own state guard organization, a semi-military group established within the state to provide protection and a military reserve for duty within that state. SG-WERS (state guard WERS) is a service operated by the state for communications relating directly to the activities of the state guard or other equivalent officially recognized organizations.

In January, 1943, the FCC included a third category within this service, that of the Civil Air Patrol, a rapidly expanding organization of civilian volunteer pilots and air personnel, established originally within the Office of Civilian Defense but later absorbed into the U. S. Army Air Forces.

In each category an official is appointed by the licensee (a local municipality in the case of civilian defense stations, a state in the case of state guard, and a "wing" in the case of the Civil Air Patrol) to supervise the network, after thorough investigation and certification by the licensee. This official is called a "radio aide" in CD-WERS, and a "communications officer" in SG- and CAP-WERS. One license authorization may be issued to cover the operation of all fixed, portable, mobile and portable-mobile transmitters proposed to be used in a single coördinated system. Call letters are assigned with subnumbers, one number for each transmitter in the system; for example, WQRR-1 might be the control unit for a typical network, with WQRR-2 through WQRR-23 as subordinate units. Operation may be on any frequencies in the amateur 112-, 224- and 400-Mc. bands, using any normal type of emission, with a maximum input of 25 watts and certain stability specifications. WERS operation may take place only in those specific instances as defined hereafter, except for authorized practice blackouts and mobilizations and for weekly test periods. Station units may be operated only by the holders of WERS operator permits, available to any FCC operator licensee whose services are wanted by a WERS licensee, upon proper application with accompanying certification of the station licensee. The complete rules and regulations are contained in the publication, *A Manual for the War Emergency Radio Service*, obtainable from ARRL, West Hartford 7, Conn., without charge to ARRL members; 10¢ a copy to others.

The radio amateur has played an important part in all three categories of this service, although perhaps more so in CD-WERS than in the other two. There is no priority rating available for equipment to be used in this service, and much of the equipment was supplied by amateurs who had it on hand or who were able to build it from parts out of their junk boxes. Amateurs themselves serve as administrators, technicians and often as operators, as well as in training additional personnel to qualify for operating assignments. Since the service has expanded greatly in scope and magnitude since its origin amateurs now represent an increasingly smaller part of the participating personnel, but an increasingly more important part by reason of their technical ability and operating experience.

Civilian Defense Stations

WERS was organized primarily at the request of the Office of Civilian Defense, which wished to fit WERS units into its nationwide plans for protection against enemy attack. With the disruption or overloading of normal telephone and telegraph circuits, a CD-WERS unit provides emergency radio channels between the Citizens' Defense Corps control centers, air-raid warden posts, police, hospitals and other strategic points. They also are available for use during other emergencies endangering life, important property and public safety in connection with civilian defense or national security and, like all other WERS stations, at the request of any agency of the

United States Government during emergencies endangering the safety of life or property. Aside from those periods during which CD-WERS stations are activated for essential operation in connection with their duties, certain hours of the week have been set aside for testing and practice drilling purposes so that the personnel involved, consisting entirely of civilian volunteers who may or may not have had previous experience in radio communication, can be trained to execute their assigned duties promptly and efficiently.

Plan of organization — Under FCC regulations, a civilian defense station license is issued to any local government, such as a town, city, borough or county, but the radio aide has actual control of all units under the jurisdiction of the license. This radio aide is appointed by the licensee, who must certify, on a prescribed form, that he is of proved loyalty and known integrity. The licensee then leaves most of the responsibility for the local CD-WERS in his hands; thus it is essential that the position of radio aide be given to the best possible man available for the job.

For control, communication and air-raid warning purposes, the OCD plan of organization is based fundamentally on what is known as an *air-raid warning district.* Such a district usually contains several hundred square miles — say, for example, an area 20 or 25 miles square — and its boundaries were originally chosen in terms of telephone toll-line organization. Somewhere near the center is a *district warning center,* usually in the largest city of the area. At this center, air-raid warning and other signals are received from regional information centers, which derive their instructions from the Army defense commander. The d.w.c. has the duty of relaying certain of these signals and information to other communities in its warning area, known as *subcontrol centers.* Through its communication facilities, the warning district's defense-corps staff arranges for allocation of apparatus for fire-fighting, road clearance, etc., in the event of a heavy air raid concentrated only in parts of the area. Communication with the warning-district center is important, since CD-WERS then can take over district air-raid warning signals and other traffic if the need develops, and is assured of prompt notification should the WERS be ordered off the air by the defense commander.

The control and communications section of OCD desires the establishment of networks based on these warning districts where possible. It wishes the station license for the entire district to be held by the city in which the d.w.c. is located, with subnumbers assigned in blocks to other cities in the area. In the fictitious example shown in Fig. 1701, Centralia is the d.w.c. city of its area; it holds the license for the entire warning district, blocks of subnumbers being distributed to its communities according to their communications requirements. Some advantages of this arrangement are:

a) The important lines of civilian telephone communication for air-raid protection are paralleled by the radio network.

b) Available frequency channels can be apportioned to adjacent communities in such a manner as to keep interference at a minimum.

c) The station units have radio channels direct into the district warning center, and may thereby be notified immediately of any shut-down order.

d) Facilities, personnel and portable-licensed equipment can be exchanged among the various communities under the license at will without special FCC authorization.

This requires a communications set-up differing but slightly from individual community plans. Organization of a group of communities under the district plan requires the utmost coöperation among the communities involved, but, once such coöperation is pledged or found to be existent, plans can go forward smoothly. Here are the preliminary steps to be taken in such organization:

a) Each interested community in the district appoints its local radio aide.

b) Local radio aides and communications officials of all local defense councils get together and appoint a district radio aide. In most districts, it has been found that a district radio aide cannot do a good job if he is also a local radio aide. The district radio aide *nominally* is designated the radio aide of the licensee, which in this case is the d.w.c. city. As seen above, however, actually he is selected by agreement among the communities concerned.

c) Local radio aides then proceed to organize their own communities under the district plan, at the same time preparing the license

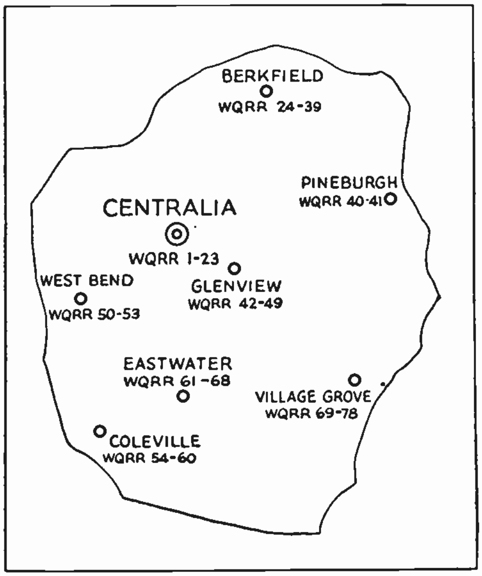


Fig. 1701 — **Fictitious air-raid warning district map, showing allocation of subnumbers based on the size and communication requirements of individual communities.**

data to be included in the license application. These data are forwarded to the district radio aide upon completion.

d) The district radio aide collects all license data from the local radio aides, executes the application, signs all operator permit applications and forwards the completed application to Washington.

It can readily be seen that licensing by districts is *not* practicable among communities which are at odds with one another. Unfortunately there are any number of reasons why district licensing might not be possible, such as: (a) the d.w.c. city may show no inclination to apply, or is in a very early stage of organization; (b) the d.w.c. city does not wish to have suburbs or adjacent communities included in its license application; (c) local defense officials insist upon an independent application, despite revelation of the advantages of the district system; (d) efforts on the part of outlying communities to help organize the d.w.c. city may have proved fruitless.

If, for one or more of the above or other reasons, it becomes necessary for communities of a district to apply independently, it is quite fitting and proper that they should do so rather than await a stalemated warning-district application. Individual communities thus applying should aim their organization at eventual consolidation with other communities of the district under the warning center. When the d.w.c. city becomes licensed, and is agreeable to a district license, the following steps then may be taken to effect a warning-area license:

a) Appointment of a district radio aide.

b) Preparation and submission of application for the entire district.

c) Surrender for cancellation of the independent licenses.

d) Request by independent licensees for authorization to continue operation under the old license until the district license is issued.

e) Re-issuance of operator permits to be signed by the district radio aide and good for operation anywhere in the district.

Since this process will require quite a bit of altruism on the part of local communities, and since it will probably mean temporary disruption and rehabilitation of the local organization, it is quite likely that many independent licensees will be unwilling to be a party to any such scheme, preferring to maintain their independent status. It is unthinkable, however, that independent licensees should want to crawl inside their shells and have nothing to do with any adjoining licensees. If an independently licensed community finds itself unwilling to consolidate itself into the district license, the next best thing to do is to arrange for regular contact by radio with the district warning center by having one of the independent licensee's station units represent its community in an intercommunicating network consisting of the control units of each community in the district, whether or not it be under the district license. The control unit of this "district net" should be located in the d.w.c. city, since the district warning center will be the only point in the district to receive shut-down or other orders from the Army Information Center as well as the point at which requests for aid will be received from and dispatched to adjoining communities.

Frequency allocations — It is one of the duties of the radio aide to coördinate the use of available frequencies by various communities so as to keep interference at a minimum. In setting up channels for employment by WERS stations, two factors must be considered: the frequency-stability requirements of FCC regulations, and the practical effect of possible interference between stations.

There are many schemes for frequency allocation in use among the hundreds of CD-WERS licensees in existence, and the one decided upon by any particular organization should be arrived at in consideration of the specific problems confronting that organization. Without doubt many of these problems will be peculiar to that organization, and therefore it is not possible for FCC, OCD or ARRL to prescribe an allocation scheme. The regulations give only the stability requirements and the band limits on which to base our plan.

OCD has proposed an allocation scheme which it has called its "Tri-Part Plan." This recommendation divides the 112–116-Mc. band into four segments, three of which are for use of CD-WERS while the other segment is to be used by SG- or CAP-WERS. Each segment is divided into a number of frequency channels 200 kc. apart, and each such band segment has a specific use. Nominally, they are the local-district (LD) band (112–112.8 Mc.) for the use of local control units in communicating with the district control unit, the local-fixed (LF) band (112.8–114 Mc.) for channels between local control units and fixed units of the local organization, the local-mobile (LM) band (114–115.2 Mc.) for contact with local control units and service headquarters by mobile units, and the Civil Air Patrol band (115.2–116 Mc.) for the use of the Civil Air Patrol and State Guard in areas where these categories of the services are in operation.[1]

A simpler and probably more practicable method of frequency allocation, modifications of which are in actual use in many districts, is the following: The radio aide, in conference with radio aides of nearby districts, chooses a suitable spot on the "high stability" (112–114 Mc.) section of the band for the warning-district net, and assigns at least one additional frequency to each community within the district for dispatch purposes or other use of its own choice. In the "low stability" section (114–116 Mc.) he should assign two frequencies to each community (including the d.w.c. city), one for a local net and the second for subnets

[1] For more details of this plan, see *QST*, Feb. 1943, p. 19.

of the local nets, constantly keeping in mind the geographical location of each community and the possibilities of mutual interference. Assuming that each community of our fictitious warning district intends to make full use of each each kind of service, the allocation for the first half of the band might be as follows:

Frequency	*Use*
112.1 Mc.	Guard Band
112.3	Warning District Network
112.5	Guard Band
112.7	Berkfield and Eastwater dispatch
112.9	Coleville and Pineburgh dispatch
113.1	Centralia dispatch
113.3	Glenview dispatch
113.5	West Bend dispatch
113.7	Village Grove dispatch
113.9	Guard Band

The second half of the band might be apportioned like this:

Frequency	*Sector Warden Nets*	*Lower Nets*
114.2	Centralia	Coleville
114.6	Village Grove, Berkfield	Glenview, Eastwater
115.0	Coleville	West Bend, Pineburgh
115.4	Glenview, Eastwater	Centralia
115.8	West Bend, Pineburgh	Village Grove, Berkfield

Actually the smaller communities may not have use for more than one channel, in which case more channels will be available for larger communities with a more complicated network scheme. An appreciable amount of duplicate and triplicate use of a single channel is quite practicable in lower nets of most warning districts. The type of receiver which will be in general use (the superregenerator) has the characteristic of featuring the loudest signal existent in its input circuit, completely annihilating any signal of appreciably less strength.

Note that 400-kc. channel separation is provided in the "low-stability" part of the band. This is to allow plenty of breathing room for each channel and provide for inevitable inaccuracy in frequency spotting by stations in the net. In the "high-stability" section of the band the channels need be only 200 kc. apart, since units capable of operating in this segment are expected to be more accurately calibrated and capable of better frequency stability.

The above plan makes no provision for State Guard or Civil Air Patrol operation in the area, but this should not imply that the possibility of their existence can be ignored. In the event that other categories of WERS *are* active, it is simply the duty of the radio aide to plan his frequency allocations accordingly, which will make the problem more complex because of resultant shortage of channels.

Because the "standard" vacuum tubes available at present do not perform efficiently at 224 Mc., that amateur band has not been included in the basic allocations above. For organizations whose amateurs already have suitable tubes and equipment for a successful 224-Mc. communications system, additional channels are available to simplify the problem.

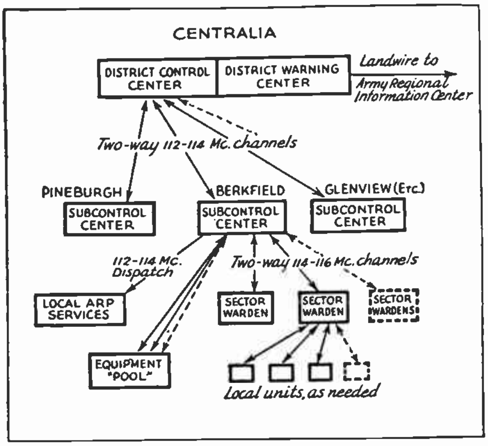


Fig. 1702 — A possible set-up of communication channels in an air-raid warning district, under one license. In this fictitious district, the city of Centralia houses the district warning center and the city itself is known as a main control center. The block "Local ARP Services" designates the fire, police, demolition-squad and other protection services to which a dispatching channel is desired. The block "Equipment Pool" designates a number of portable-mobile units available at a central point for immediate dispatch to an area suffering from unusual enemy bomb hits. For space reasons, only the Berkfield subcontrol center is shown subdivided here.

Operating personnel — Everyone who operates a WERS station unit must possess a WERS operator permit. *There are no exceptions to this rule.* To be eligible for such a permit, an individual must possess a radio license of any class issued by FCC. Thus a licensed amateur is already eligible, as is any commercial radio operator; he may secure an operator's permit simply by filing a completed FCC Form 457 with his radio aide, who forwards it to FCC.

With thousands of amateurs in military and government communication service, it has been found that very few communities have sufficient licensed operator personnel to carry out a satisfactory plan of CD-WERS communication. In most communities it has been found necessary to set up brief but intensive training courses for desirable personnel aimed either at securing an amateur license or a restricted radiotelephone permit, probably with emphasis on the latter because of the comparative ease in qualifying for it.

A course for the 'phone permit might be set up for a total of six hours, in two-hour periods. A preliminary period could well be spent on presenting a general background of radio and the need for regulation and licensing; then a period of brief general exposition of the high spots of commercial rules and regulations, followed by individual discussion of typical questions and answers; and lastly, several hours spent in simulated operation, practicing voice technique, phonetic word lists, signing the station off, repeating dummy message reports — all to make *operators* as well as licensees.

FCC commercial operator regulations provide to "employees of a division of local or

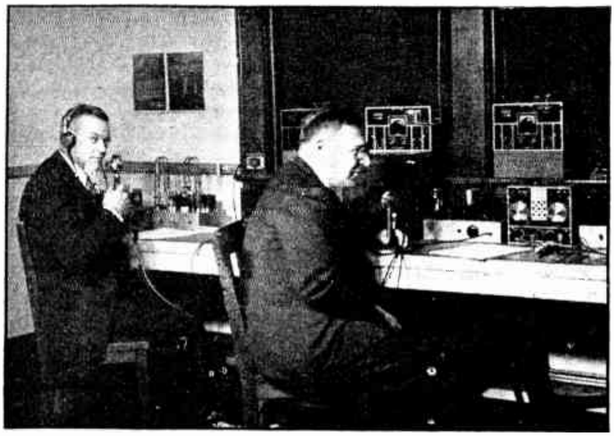


Fig. 1703 — A typical WERS district control center station — WODF-5, Akron, Ohio. Duplicate operating positions enable dual-channel contacts with the central control station and field groups simultaneously. Auxiliary transceivers are kept in reserve to supplement regular equipment in the event of failure and for use as field units.

state government" the convenience of "resident" examinations for the restricted 'phone permit. FCC has extended this privilege to personnel selected to operate WERS stations. In practice the municipality should communicate with the district inspector, submitting to him completed application forms with all other necessary attachments (fingerprint forms, proof of citizenship, photographs) for each applicant, along with the name of the person designated to supervise the examination (who might well be the radio aide) and the date and place it will be given. In the case of WERS stations, resident examinations will not be authorized prior to the issuance of the station license. Please note that this convenience applies to the restricted 'phone permit only — not to amateur licensing.

Operating procedure — One of the greatest faults with many CD-WERS organizations is that either they have failed to evolve any operating procedure or the procedure they have adopted is inadequate or unbecoming a government communications service. Realizing that the procedure used necessarily would vary according to the particular situation involved FCC did not limit WERS operation to any set form, except for a requirement that the procedure to be used be described as one of the supplementary statements in the application. From this it is not to be concluded that WERS units may say what they please on the air or conduct their operation in any way they see fit. "Rag-chewing," attempting to establish DX records, and experimenting on the air except for the sole purpose of establishing reliable communication between two or more licensed points, especially are to be discouraged. Operators should be trained to make their transmissions brief and to the point. Participating amateurs, in particular, should avoid discrepancy, since WERS is not an amateur service and should not be conducted like one.

It is difficult to foresee what actual communications need there will be during an air raid and therefore one can only plan as best he knows how, altering the specifications as it is found they can be improved. Thus all that can be discussed here are general principles.

Identification must be complete at the beginning and ending of each complete exchange of communication. The net control unit might call the roll and be answered by the net units in the following manner:

WXXX1: WXXX1 calling WXXX2, answer roll call, go ahead.
WXXX2: WXXX1 from WXXX2, answering roll call, go ahead.
WXXX1: OK 2. WXXX1 calling WXXX3, answer roll call, go ahead.
(Etc., until roll call completed)

WXXX1: 2 from control, ready for your traffic, go ahead.
WXXX2: Control from 2, here traffic. . . .

With enemy planes approaching, however, announcement of call letters might reveal location of units. In some sections the Army requires that call letters, as well as any other information that might be of value to the enemy, be omitted during actual air raids. At such times the Army will issue a *limited transmission order*, which will come through the district control center and must be immediately obeyed by CD-WERS networks. Identification at that time should be by unit number only, omitting all assigned call letters. *At all other times the FCC requirement that complete identification be given must be observed.*

To expedite the handling of incident reports, CD-WERS systems should use the same report forms as are used by the wardens in reporting incidents. These report forms usually consist of items numbered one to eleven. In transmitting messages using these forms the operator can merely designate the item number to follow, without repeating any other printed information on the form. Some systems of operating procedure, although following the order of the regular report form, nevertheless omit the item number altogether.

Since the urgency of messages will vary it is well to arrange priority ratings for messages, such as "regular" (a routine warden report not requiring action), "rush" (a warden report necessitating action to be taken by one of the protective services), or "duplicate" (indicating that no action had followed a previous report and the situation was becoming urgent).

Many licensees have devised systems for coding messages transmitted over CD-WERS networks, so that their meanings would not be apparent to the casual or enemy listener. Only information that might be useful to the enemy need be coded. The only effective way of putting such a code into practice is to agree on certain letters or numbers to represent certain words or phrases, and then to practise incessantly using the code. Code lists may consist of frequently used phrases, each containing its code number, which cover almost any conceivable need. To indicate locations, the city can be divided into zones, each street within that zone being numbered, so that a certain street could be indicated by saying "Zone 3–14," meaning the street numbered 14 in Zone 3, or "357 Zone 3–14" if it is desired to give the number of the house affected on that street. A simpler and perhaps more effective way of coding streets is to assign a number to each street in the city or district, starting, for example, with 0001 for Aaron Street and ending with 2500 for Zephyr Way; or, if desired, the numbers could correspond to the geographical location of the street instead of its alphabetical sequence; or there need be no relation between the street and the number assigned it. Each station unit should be equipped with a list of the number codes, arranged both numerically and alphabetically for ease of reference in coding and decoding.

Remember, however, that in handling traffic the important thing is to get the message through quickly. After efficient speed has been developed in handling messages, it is time enough to start considering coding devices. In no case should secrecy codes be developed at too great expense in speed.

In general, the following principles of operating procedure should be applied:

a) Comply with FCC and Army regulations.

b) Make transmissions short and to the point. Eliminate all words that can be eliminated without affecting the sense.

c) Adopt a procedure which will conform as closely as possible to that used by other branches of the ARP communications service.

d) Evolve your procedure around the slogan, "The enemy might be listening," and arrange it so that a listening enemy agent in no way may be aided by your transmissions.

e) Avoid unnecessary complications. The ideal operating procedure is one which will accomplish the objectives in view as simply, quickly and effectively as possible.

Application data — The person preparing a license application must make certain that it complies with the regulations, and he should be sure to prepare the application correctly *the first time* to avoid having it questioned by FCC — a time-wasting process, both for the applicant and for the Commission. On any doubtful issue it is wise to get the advice of someone who has "been through the mill." The comments to be made here, while necessarily of a general nature, are intended to cover points on which radio aides frequently go wrong.

The station-license application should be prepared on FCC Form 455 (available at any district FCC office). It may be executed only by municipal governments, such as cities, towns, counties, etc., and *not* by any subdivision of that government such as the police department, fire department or defense council. The application must be signed by the mayor, town manager or similar highest executive official of the government. For reasons of flexibility, many licensees have adopted the practice of listing all units as "portable" or "portable-mobile" except the fixed control units.

FCC will not issue a station license unless the application shows in detail the complete set-up of civilian defense communications. Extremely important is the map of operations, which should be included as a part of the supplementary data required in the regulations. A street map, available from the city clerk, should do the trick, with overmarking to show the communications plan. It should carry the locations of all station units, and if portable units are included their boundary lines of normal operation must be indicated. For warning-district organization, probably it will be more convenient to submit first a chart of the entire district, showing the channels between district and subcontrol centers, and second a series of smaller maps showing individual communities and details of unit locations.

The license application requires a number of attachments of supplementary statements and additional information, as follows:

a) Additional lists of equipment too long for inclusion on page two of the application.

b) A number of sheets identical to page three of the application, giving technical data on types of equipment not shown thereon.

c) Form 455 (a), certification of radio aide.

d) If the licensed community wishes to provide service to additional communities (as in the district plan of licensing), copies of intermunicipal agreements made between the licensee and the various other communities to be covered, properly signed in each case by the mayor or similar official, must be included.

(e) Forms 457, one for each operator authorization requested. Make certain that these are properly executed and that two full-face recent

COMBINED PHONETIC ALPHABET

Used by Armed Services of Great Britain and U. S. A.

A — Able (Affirm)*	J — Jig	R — Roger
B — Baker	K — King	S — Sugar
C — Charlie	L — Love	T — Tare
D — Dog	M — Mike	U — Uncle
E — Easy	N — Nan (Negat)*	V — Victor
F — Fox	O — Oboe (Option)*	W — William
G — George	P — Peter (Prep)*	X — X-ray
H — How	Q — Queen	Y — Yoke
I — Item (Interrogatory)*		Z — Zebra

* From U. S. Navy General Signal Book, used within the Navy for flag signaling.

0 — Zē-rō	3 — Thuh-ree′	7 — Se-ven
1 — Wun	4 — Fō-wer	8 — Ate
2 — Too	5 — Fi-yiv	9 — Niner
	6 — Six	

Example:
WQRR-49 — William Queen Roger Roger Fō-wer Niner

photographs, not over 2½ × 2½, signed on the back by applicant, are attached.

f) The general map of operations, as discussed above.

g) A statement concerning the scope of service to be rendered and type of messages handled. That is, whether a service for air-raid wardens alone, or for emergency equipment dispatch, or "incident officer" contact, or a combination of several.

h) Factual information on the exact area of operations to be included in the license.

i) A statement of the general operating procedure to be employed by all station units.

j) A list of all equipment procured, showing source (purchase, loan, gift, etc.) and distribution (usual area of operation of each unit).

k) A statement of methods to be used in supervising the operation of all station units, including data on monitoring, frequency-measurement methods, provision for frequent inspection by the radio aide, etc. If the latter has delegated controlling authority to deputies or assistants, this item should so state.

l) Methods used to ascertain the loyalty and integrity of operating personnel. This section should include a statement from the local chief of police giving the names of all operators for whom operating permits are requested and certifying to their character and loyalty to the United States. Also include data on plans for recruiting operators and whether they will serve on a paid or voluntary basis.

m) A positive statement of the applicant's ability to silence its units upon order of the regional defense commander. This entails establishing proof of close and continuous contact with the district control center and of arrangements for the immediate relaying of any such signals. The very nature of district organization permits an easy answer to this requirement, of course.

Carbon copies of all parts of the applications should be kept in the radio aide's files, so that he will be able to refer accurately to any part of the application which subsequently might be questioned by FCC.

Sample forms of agreements which may be adapted to the needs of the particular community are shown on this page. The first is the agreement executed between the amateur and the city when the former lends the city his equipment for the duration, and should be in duplicate. The second is the inter-municipal agreement for the license application under district organization, as adapted from the form used in the State of Massachusetts.

AGREEMENT

I, ..., residing at .... in the city of ...., ...; being the unconditional owner of the radio equipment described in detail below, do hereby convey all my right, title and interest to such equipment to the City of Berkfield, Faryland, for use solely in the War Emergency Radio Service; PROVIDED THAT it shall be returned to me by the City at the end of the present war.

List of Equipment:
....
....
....

Dated this .... day of ...., 194....

(Signed) ....

Witnessed by: ....

By its signed acceptance of this document the City of Berkfield acknowledges receipt of the equipment in good condition and pledges that at the end of the present war it will be returned to the above-named individual, his heirs or assigns, the City releasing all claims, right, title and interest thereto.

Accepted this .... day of ...., 194......

City of Berkfield, Faryland.

By ....

SEAL

WAR EMERGENCY RADIO SERVICE
INTER-MUNICIPAL AGREEMENT

IT IS AGREED THAT the City of ...., ... hereinafter known as the licensee, will apply to the Federal Communications Commission for permission to construct and operate radio stations in the War Emergency Radio Service in the area known as ....,

AND THAT the City of ...., hereinafter known as the sub-licensee, lies within said area and wishes to participate in a single War Emergency Radio Service network serving that area;

IT IS HEREBY AGREED by both parties:

THAT all radio equipment installed by the sub-licensee for the above purpose shall be under the direction and control of the licensee;

AND THAT the Radio Aide agreed upon by the licensee shall administer the operation of and be responsible to the licensee for all equipment in said network;

AND THAT during the existence of this agreement, the sub-licensee will not request individual authority for a War Emergency Radio Service station license;

AND THAT this agreement may be terminated at will by either of the parties concerned but that notification shall be given to the Federal Communications Commission sixty days prior to the termination of this agreement by either of the parties.

IN WITNESS WHEREOF, we hereunto set our hands this .... day of .... 19 .

The City of .... The City of ....
By .... By ....
Title .... Title ....
State of ....
County of ....

SUBSCRIBED and sworn to before me this .... day of .... 19 .

Notary Public ....

My commission expires ....

State Guard Stations

Reference to the regulations at the end of this chapter will indicate that much that has been said in the previous section concerning civilian defense stations applies also to state guard stations, but is presented from a civilian defense standpoint because specific information on state guard organization is lacking. This, of course, is because a state guard is a semi-military organization, and therefore many of its functions are not public information.

In state guard work the "communications officer" performs the same function for the state guard licensee that the radio aide performs for the civilian defense licensee, and applications are executed on the same three forms with practically the same set of supplementary statements necessary. One of the chief differences is that the license is issued to the state guard of an entire state, and must be signed by the commanding officer for that state. The original license contains authorization only for those units of the state guard which are prepared to take part in SG-WERS communication. Additional units can be included in the license later only by modification of the station license. The fact that an individual is a member of a state guard organization which is licensed for SG-WERS *does not* automatically authorize him to operate station units of that licensee simply at the command of his commanding officer. He must possess the necessary eligibility for a WERS operator's permit and must apply for such a permit on FCC Form 457, which form must be signed by the communications officer *of the licensee* and not by the communications officer of a local state guard unit; however, other authority to supervise the operation of local units may be delegated to local communications officers by the state communications officer.

All SG-WERS license applications must be forwarded through the Army Service Command of the area in which the state is included. After approval, the application is forwarded to FCC and the license issued if the application is found to be in order.

Note that the scope of service of state guard stations differs slightly from that allowed civilian defense stations, and that tests are allowed at any time of the day or night provided that such tests do not exceed four hours per week. Amateurs desiring more details of this service should communicate with the communications officer at the headquarters of their local state guard unit.

Civil Air Patrol Stations

Civil Air Patrol stations come under the same general rules and regulations as civilian defense and State Guard stations, and many of the specific rules for CAP-WERS stations are similar to those for CD- and SG-WERS.

CAP-WERS licenses are issued only to the duly appointed wing commanders of the Civil Air Patrol. A wing of the CAP usually covers an entire state, and under the wing commander is a wing communications officer similar to the radio aide in CD-WERS, who has the duties of preparing the license application and supervising the operation of all units of the wing. A wing is subdivided into "groups" and "squadrons," each of which has its commanding officer and its communications officer. Some wings, in particular those of the smaller states, have omitted the group organization and have only squadron units under the wing.

CAP-WERS organization goes from squadron to group to wing. A squadron of the CAP which wishes to participate in CAP-WERS should have application data prepared by its communications officer. These data then are approved by the squadron commander and forwarded to group headquarters. The group communications officer collects all such data, has them approved by the group commander, and forwards them to wing headquarters. The wing communications officer then prepares the license application much as the radio aide prepares a CD-WERS application. This application, when completed, must be signed by the wing commander, who is the licensee, and forwarded to the national headquarters of CAP. There it is perused by the national communications officer, sent back to the wing for correction if necessary, and finally forwarded to FCC for issuance of the license.

Note that, as in both other WERS categories, individuals who operate CAP-WERS units are in no way relieved from the necessity of possessing a WERS operator's permit. There is no limit to the amount of testing which can be conducted by CAP-WERS licensees, except that it shall not exceed the minimum necessary to ensure the availability of reliable communications. Amateurs interested in this service should get in touch with their local CAP squadron communications officer, or with the national CAP communications officer at CAP national headquarters, Washington, D. C.

Equipment

In the WERS regulations, power input to the transmitter r.f. output stage is limited to 25 watts or less, and the frequency bands available are each divided into two sections having differing frequency tolerances. The carrier frequency tolerance is set at 0.1 per cent in the lower half of the band and at 0.3 per cent in the upper half. With the possible exception of a few localities, where some 224-Mc. equipment may be available, all WERS communication is carried on in the 112-Mc. band because it is possible to operate with standard receiving tubes and ordinary circuit components at these frequencies. Only 112-Mc. equipment is discussed in this chapter, therefore.

Provision for frequency measurement is not necessary at every station, since it can be taken care of by the main station at each subcontrol center.

Measurements on representative types of amateur 112-Mc. equipment have shown that the stability requirement of 0.1 per cent can be readily met by any reasonably well-built oscillator-type transmitter, adjusted to frequency and thereafter left with its tuning controls untouched during an operating period. If the transmitter can be crystal-controlled, so much the better. For operating convenience crystal control is desirable at control stations, since two or more frequencies can be made available in the same transmitter simply by providing the necessary crystals and a switch. It should be possible to use any of the frequencies, provided the maximum frequency separation is not more than 2 Mc. or so, without touching the tuning controls. If a crystal-controlled transmitter is not available, it is desirable to provide separate transmitters for each frequency to be used, to avoid the important practical difficulty of returning to exactly the same frequency each time a shift is made.

Each subcontrol center should have one receiver which can be used for continuous monitoring of the district warning center control station. This is a necessity as a means of securing immediate radio silence when so ordered by the Army. A second receiver should be available for communication within the local network. Although it is desirable, as a rule, to have separate transmitters and receivers rather than transceivers at the control station, the second receiver in the control station may be in a transceiver which is used, in the "transmit" position, as the transmitter for communication with the sector wardens.

One transmitter and one receiver (the two may be combined in a transceiver) should suffice for the sector wardens' stations (field stations). If a separate frequency channel is maintained for portable-mobile apparatus it will be necessary for the operators of the sector stations to monitor both this frequency and that on which they are to communicate with the subcontrol center. A transceiver, with its automatic tuning of both transmitter and receiver to the same frequency, may be simpler to operate than separate transmitter and receiver units, unless two transmitters are available to take care of both frequencies. Where a transceiver is used, the two tuning spots should be plainly marked on the dial. In many networks both functions can be carried out on one frequency, thus simplifying operation.

Mobile apparatus, which is required to work only on one frequency, can be of any type.

Transceivers, particularly the dry-battery-operated units, cannot be depended upon to maintain carrier stability within the 0.1 per cent necessary to meet the requirements for operation in the low-frequency half of the band. The more powerful transceivers, for operation on a.c. or from storage-battery supply, are fairly stable as transmitters, but if the receiver has to be tuned to two or more frequencies the unavoidable inaccuracies in resetting the dial preclude the possibility that the frequency can be reliably maintained within 0.1 per cent. It is better to confine the use of transceivers to the high-frequency half of the band; the average one is readily capable of meeting the requirement of 0.3 per cent.

All equipment must be capable of operating from sources of power supply independent of the a.c. mains, and must in fact have such power supply available. So that emergency power can be conserved for the time when it is needed, it is helpful to equip fixed stations with alternative power supplies.

The apparatus designs in this chapter should be considered as a source of circuit and constructional ideas rather than as items to be duplicated. Fortunately 112-Mc. equipment suitable for WERS work is quite simple, and it is readily possible to adapt an idea used in one piece of equipment to a different circuit layout. A few general principles:

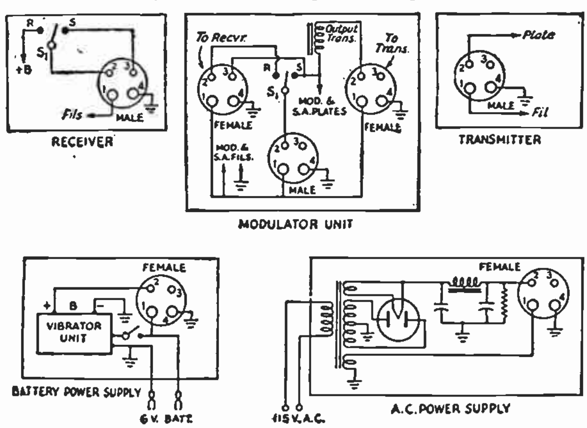
1) The apparatus must be simple to operate. Much of the actual operation must be done by those who have never built radio equipment or operated anything more complicated than a broadcast receiver. If the apparatus can be enclosed so that only the on-off and change-over controls are accessible, there is that much less chance that the equipment will be misadjusted, accidentally or otherwise.

2) Keep in mind the possibility that replacement of some components may be necessary. If, for example, a choice of tubes exists, take the type which has the best chance of being available at short notice, even though it may not actually be the best.

3) Build the equipment, including the power supply, so that it can be moved to a new location set up for operation again with a minimum of delay. Construction must withstand the knocks and jars of transportation. Enclose the equipment so that wiring cannot be damaged or projecting components knocked loose.

4) Use uniform control methods for all installations, so that an operator filling in at a station to which he is not normally assigned can work with minimum special instruction.

5) Standardize on a system of mak-



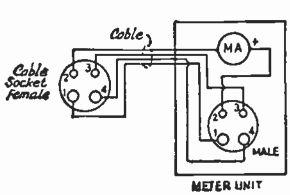


Fig. 1704 — Interconnection and switching system for various units of the emergency station. Connections are made by means of three-foot cables provided with a plug at one end and a socket at the other. Four conductors are required in the cables; to minimize filament voltage drop, the type of cable having two heavy conductors should be used. Alternatively, pairs of wires in an ordinary six-wire cable can be connected in parallel to lower the resistance.

ing connections, particularly between the power supply and the r.f. and audio parts of the assembly, so that interchangeable units can be substituted in case of failure. Do not neglect such considerations as connections for microphones and headsets. To avoid the possibility of having two or three different kinds of plugs and jacks, or even pin jacks, standardize on one type at the outset, so that any microphone or headset will fit any transmitter or receiver.

A standardized connection scheme is suggested in Fig. 1704. The system is based on the use of universal four-conductor cables for quick and positive interconnection. On each unit a socket (female) is used for *outgoing* power and a plug (male) for *incoming* power; thus there is no danger of shock nor possibility of making wrong connections.

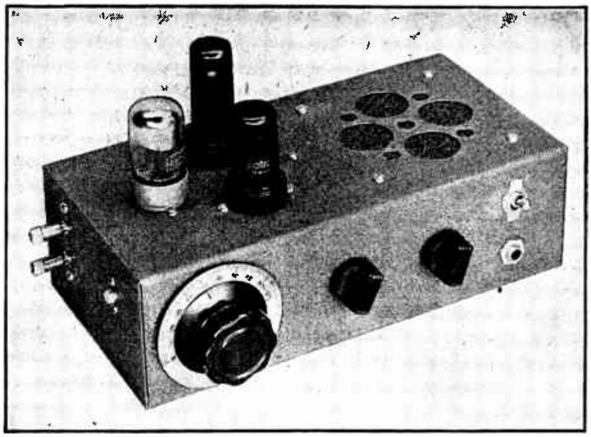


Fig. 1705 — A superregenerative receiver with built-in speaker, constructed on a standard chassis base. The detector trimming condenser is mounted on the side. The audio gain control is mounted next to the tuning control (extreme left). The regeneration control is between the volume control and the 'phone jack and on-off switch.

The unit plan illustrated by Fig. 1704 has a number of advantages. Should a particular unit develop trouble in operation, a spare can be substituted with negligible loss of time and the defective unit may be serviced without interrupting communication. Extra units can be built in anticipation of such a contingency with a probable saving in time, effort and components as compared to the alternative of providing a spare transmitter-modulator or complete transmitter-receiver, since with this system a few spare units should take care of the replacement requirements of a fair-sized network; it is unlikely that all parts of a station would fail simultaneously while it is somewhat more inconvenient to set up a multi-unit system than to place one integral station in operation, the standardized connection idea nevertheless can be applied with distinct benefits.

Control Station Receivers

By far the simplest type of receiver for very high-frequencies is the superregenerator, long a favorite in amateur work. It provides good sensitivity with a small number of tubes and very elementary circuits. Its disadvantages are lack of selectivity and the fact that, since the detector is an oscillator coupled to an antenna, it will radiate a signal which may cause interference to other receivers. To some extent the lack of selectivity is advantageous, since it increases the chances of hearing a call even though the transmitter and receiver may have drifted somewhat in frequency since the last contact was established. The radiation ques-

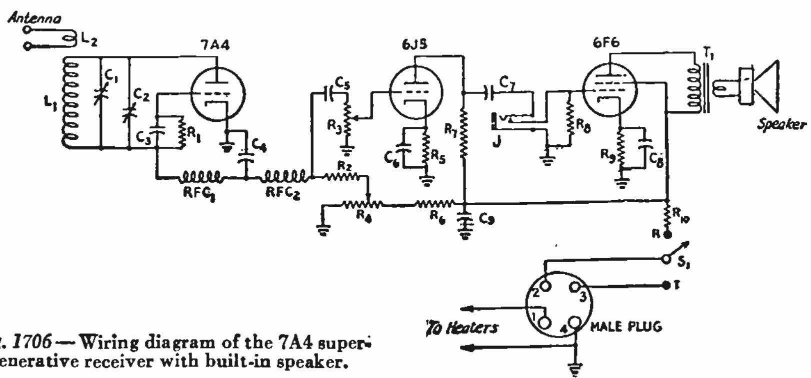


Fig. 1706 — Wiring diagram of the 7A4 superregenerative receiver with built-in speaker.

C_1 — 25-$\mu\mu$fd. air trimmer (Hammarlund APC-25).
C_2 — 5-$\mu\mu$fd. tuning condenser (National UM-15 with 2 stator and 2 rotor plates removed).
C_3 — 50-$\mu\mu$fd. midget mica.
C_4 — 0.006-μfd. mica.
C_5, C_7 — 0.01-μfd. 600-volt paper.
C_6, C_8 — 10-μfd. 25-volt electrolytic.
C_9 — 8-μfd. 450-volt electrolytic.
R_1 — 5 megohms, ½-watt.
R_2 — 25,000 ohms, 1-watt.
R_3 — 0.5-megohm volume control.
R_4 — 50,000-ohm wire-wound pot.
R_5 — 1500 ohms, ½-watt.
R_6, R_7 — 50,000 ohms, 1-watt.
R_8 — 0.1 megohm, ½-watt.
R_9 — 500 ohms, 1-watt.
R_{10} — 2000 ohms, 10-watt wire-wound or higher. See text.
J — Closed-circuit jack.
RFC_1 — V.h.f. choke (Ohmite Z-1).
RFC_2 — 80-mh. r.f. choke (Meissner 19-2709).
S_1 — S.p.d.t. toggle switch.
T_1 — Output matching transformer.
Speaker — 4-inch p.m. type.
L_1 — 1¾ turns No. 14 e., ½-inch inside diameter, spaced diameter of wire.
L_2 — ⅞ turns No. 14 e., ½-inch inside diameter.

tion is more serious. The avenue of approach in the reduction of radiation is to use every means possible, in the way of building as good an oscillator circuit as the circumstances will permit, to operate the detector at the lowest plate voltage that will give satisfactory operation. A means of providing regeneration control is essential in any supergenerative receiver. Should existing equipment not have such a control, it is urged that one be installed.

Superregenerative receiver with built-in speaker — The receiver shown in Figs. 1705, 1706 and 1707 is built in a 10 × 5 × 3-inch metal chassis, with the tubes and speaker mounted on the 5 × 10-inch face. One side is used for a panel and the opposite side is left clear in case one wishes to operate with the receiver resting on this side. The antenna terminals and the detector padding condenser are mounted on the left-hand side, and the four-prong power plug is mounted on the right-hand side. The only care necessary in laying out the chassis is to mount the tuning condenser and the padding condenser so that their respective terminals come close together, to make the leads as short as possible. The tuning condenser, C_2, is supported back of the panel on long (1¾-inch) 6–32 screws, and the padding condenser is mounted directly on the side of the chassis. A bakelite shaft extension is fastened to the tuning condenser shaft and brought out through a panel bearing. The quench-frequency filter choke, RFC_2, is supported between the two audio tube sockets on a ½-inch pillar.

Each socket has a soldering lug placed under one screw, and all of the grounds for that particular stage are made to the lug. Most of the resistors and condensers can be mounted directly on the tube-socket or variable resistor terminals.

The coil, L_1, can be trimmed slightly by squeezing the turns together or pulling them apart until the desired amount of bandspread is obtained. The antenna adjustment is made by moving the antenna coil, L_2, closer to L_1 until the regeneration control must be set at about ⅔ full for "supering" to start. This adjustment is made with the antenna connected.

A simple superheterodyne receiver — A superheterodyne control station receiver of medium selectivity is shown in Figs. 1708 and 1710. The circuit arrangement is shown in Fig. 1709. The 955 mixer tunes from 112 to 116 Mc., while the h.f. oscillator, in which a second 955 is used, tunes from 91 to 95 Mc. The tuning condensers of these two circuits are ganged. The 1852 i.f amplifier and the 6J5 superregenerative detector are tuned to 21 Mc. Transformer coupling is used between the detector and the 6J5 first audio stage. The output tube which feeds the speaker is resistance coupled to the preceding stage. The power supply is a simple choke-input affair with a VR105-30 regulator tube controlling the plate voltage of the h.f. oscillator and mixer stages. R_9 is the detector superregeneration control, and R_{11} is the audio volume control.

Most of the constructional details are apparent from Figs. 1708 and 1710. The chassis measures 3 × 7 × 15 inches. All components of the v.h.f. circuits, including the tubes, are mounted underneath the chassis. In Fig. 1710 the double-section tuning condenser, C_1C_2, is mounted near the top. Each section of the original condenser consisted of two rotor plates and one stator, giving a maximum capacity of 10 μμfd. per section. By removing one rotor plate in each section and double-spacing the stator, sufficient reduction in tuning rate was obtained to spread the 112- to 116-Mc. band over 110 degrees on the dial. If this much bandspread is not desired, the stator plates need not be double-spaced.

Immediately above the tuning condensers are the two acorn tubes, with the oscillator tube nearer the panel. The self-supporting mixer and oscillator coils, L_2 and L_3, are mounted at right angles to each other and soldered to their respective condenser terminals.

The 1852 i.f. tube is mounted on top of the chassis in the rear right-hand corner. The first

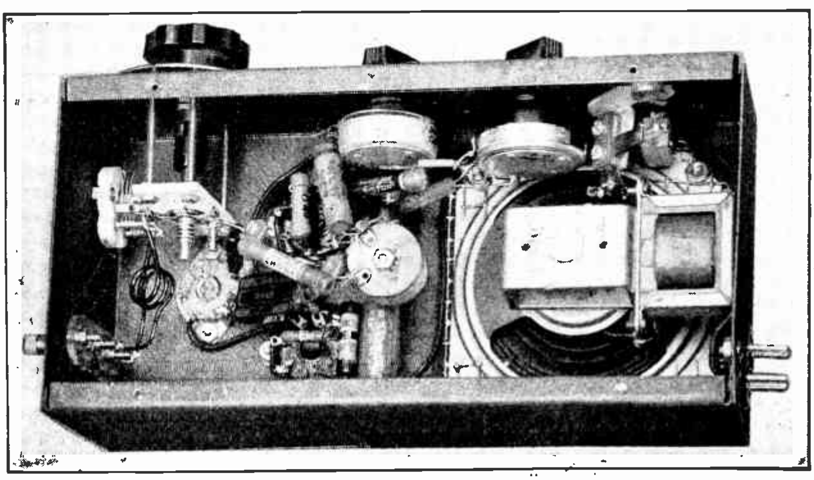


Fig. 1707 — This underside view of the 7A4-detector superregenerative receiver shows the loudspeaker mounted at one end of the 5 × 10 × 3-inch chassis. The detector tuning condenser is mounted on long machine screws from the front "panel." The r.f. chokes may be seen in the clear at the left center. The male plug for the power-supply cable is to be seen mounted on one end at the corner near the loudspeaker.

i.f. transformer is composed of two coils, L_4L_5 in Fig. 1709, wound on a polystyrene form ¾ inch in diameter. It is placed underneath the chassis as close as possible to the submounted 1852 socket and at right angles to L_2 and L_3. No shielding of these windings, other than that provided by the chassis, is necessary.

The second i.f. transformer, L_6L_7, is made in a similar manner and is mounted in the shield can on top of the chassis to the left of the 1852. The i.f. tuning condenser, C_{15}, also is mounted inside the shield in such a position that it can be adjusted by means of a screwdriver inserted in a hole in the top of the can. The plate lead of the 1852 should be as short as possible and well shielded to prevent regeneration in this stage. All r.f. ground connections for the i.f. amplifier are brought to a single point on the metal ring by which the socket is fastened to the chassis. Particular care also should be exercised in grounding the can shielding the second i.f. transformer.

The 6J5 superregenerative detector is placed to the left of the second i.f. transformer. The

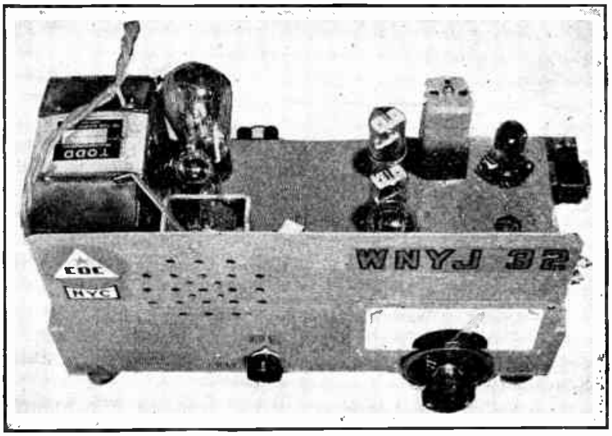


Fig. 1708 — Top view of the superheterodyne receiver. The i.f. amplifier tube and output transformer are in the rear right-hand corner. The detector and audio tubes are in line to the right. Power-supply components and loudspeaker are at the left-hand end of the chassis. This unit was constructed by Frank Heubner, N. Y. C. WERS licensee.

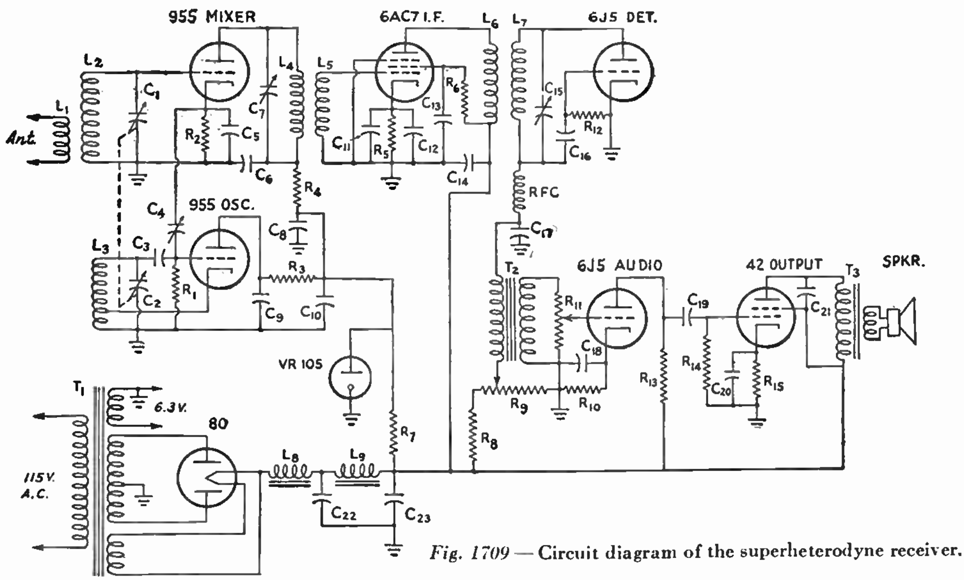


Fig. 1709 — Circuit diagram of the superheterodyne receiver.

C_1, C_2 — 10-μμfd. modified midget variable (see text).
C_3, C_6, C_{16} — 100-μμfd. midget mica.
C_4, C_7, C_{15} — 3–30-μμfd. trimmer.
C_5 — 50-μμfd. midget mica.
C_8, C_{10}, C_{17} — 0.001-μfd. midget mica.
C_9 — 500-μμfd. midget mica.
C_{11} — 0.002-μfd. midget mica.
C_{12}, C_{13}, C_{14} — 0.01-μfd. 400-volt paper.
C_{18}, C_{20} — 25-μfd. 25-volt electrolytic.
C_{19} — 0.05-μfd. 400-volt paper.
C_{21} — 0.002-μfd. 400-volt paper.
C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
R_1 — 20,000 ohms, ½-watt.
R_2, R_3, R_4 — 10,000 ohms, ½-watt.
R_5 — 200 ohms, ½-watt.
R_6 — 60,000 ohms, ½-watt.
R_7 — 7000 ohms, 10-watt.
R_8 — 250,000 ohms, 1-watt.
R_9 — 75,000-ohm wire-wound potentiometer.
R_{10} — 2000 ohms, 1-watt.
R_{11} — 0.5-megohm volume control.
R_{12} — 2 megohms, ½-watt.
R_{13} — 50,000 ohms, 1-watt.
R_{14} — 0.5 megohm, ½-watt.
R_{15} — 500 ohms, 1-watt.
L_1 — 4 turns No. 20 hook-up wire, ½-inch diameter.
L_2 — 8 turns No. 12, ½-inch diameter, 1¼ inches long.
L_3 — 5 turns No. 12, ½-inch diameter, ⅝ inch long, tapped 2 turns above ground.
L_4 — 12 turns No. 18, ¾-inch diameter, close-wound.
L_5 — Same as L_4, spaced 3/16 inch from L_4 on same form.
L_6 — 10 turns No. 18, ¾-inch diameter, close-wound.
L_7 — 15 turns, No. 18, ¾-inch diameter, close-wound, spaced ½ inch away from L_6 on same polystyrene form.
L_8, L_9 — 20-henry filter choke.
T_1 — Power transformer, 700 volts, c.t., at 60 ma., with 5-volt rectifier-filament and 6.3-volt heater windings.
T_2 — Interstage audio transformer.
T_3 — Pentode output-matching transformer.
RFC — 2.5-mh. r.f. choke.

two audio tubes are in line in front of the detector tube. The audio transformer, T_2, had to be mounted outside the chassis at the right-hand end because it picked up hum in any other position. With a shielded transformer it is probable that this trouble would not be encountered.

The regeneration control, R_9, is fastened underneath the chassis, since it does not require attention once it has been adjusted for proper operation of the detector.

Power-supply components are placed at the left-hand end of the chassis, both above deck and below. The speaker is mounted at the left-hand end of the panel where a screen of ⅜-inch holes has been drilled. A pair of feed-through insulators at the right-hand end of the chassis serves as the antenna terminals. Half-inch rubber feet fastened to the bottom edges of the chassis provide clearance for the tuning knob.

When the receiver was first placed in operation the grid of the 1852 was tapped down four turns on L_5, which was tuned with a trimmer condenser. L_4 and L_5 were spaced ¾ inch apart. While crystal-controlled signals came through perfectly with this arrangement, the selectivity was too great for satisfactory reception from modulated-oscillator transmitters. As a result, the tuning condenser across L_5 was removed, the grid of the 1852 was moved to the top of the coil, and the distance between L_4 and L_5 was reduced. These changes provided increased gain as well as the desired reduction in selectivity.

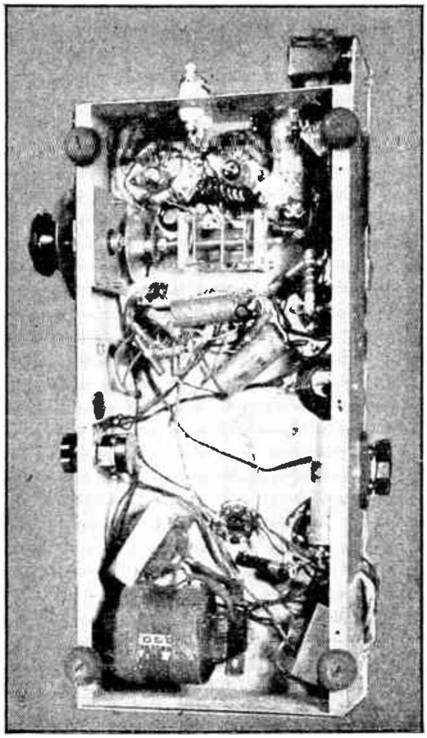


Fig. 1710 — Bottom view of the WERS control-station receiver. The two acorn tubes in the circuits are visible above the tuning condenser, near the top of the chassis. The power-supply occupies the lower portion.

Alignment — The receiver may be lined up with the aid of an all-wave receiver or any other source which will serve as a signal generator at the required frequencies.

In tuning up the i.f. amplifier and adjusting the superregenerative detector, the 955 oscillator tube should be removed from its socket and a two-foot length of wire attached to the plate lead of the mixer tube where it connects to the top of L_4. The regeneration control, R_9, should then be advanced until the 6J5 detector goes into superregeneration. If an all-wave receiver is used as the test-signal source, it should be tuned slowly between 20 and 30 Mc. with its antenna attached. At some point between these limits the signal from the oscillator in the all-wave receiver should block the detector. The all-wave receiver should then be tuned to approximately 21 Mc. and the detector adjusted to this frequency by listening for the dead spot as C_{15} is turned through its range. After this initial adjustment, move the all-wave receiver some distance away, disconnect its antenna and readjust C_{15} on the weaker signal. At the same time the input circuit of the i.f. amplifier may be tuned up by adjusting C_7. R_9 should then be readjusted for maximum blocking and greatest reduction in hiss when the test signal is tuned in.

With the detector and i.f. amplifier lined up, the next step is to put the 955 oscillator tube back in its socket and remove the antenna from L_4. If frequency-measuring apparatus for putting the oscillator on 91 to 95 Mc. is not available, either a very low-power 112-Mc. oscillator or a harmonic from the oscillator in the all-wave receiver may be used to produce a test signal at the operating frequency. If the signal cannot be heard at some point on the dial as C_2 is tuned, adjust the inductance of the oscillator coil by squeezing together or spreading the turns of L_3. The coupling condenser, C_4, should be set at about three-quarters of its maximum capacity and then adjusted for maximum mixer response as indicated by the degree of silencing when the test signal is tuned in. C_2 must be readjusted each time the capacity of C_4 is changed. When the right amount of injection has been determined, the spacing of the turns of L_3 should be adjusted so that a 114-Mc. signal is heard when C_2 is set at half its maximum capacity. Then C_7 and R_9 should again be readjusted slightly for maximum signal response consistent with good quality. Optimum operation will be obtained with a detector plate voltage of about 20.

As a final adjustment, the mixer should be checked for resonance. By squeezing or spreading the turns of L_2 very slightly while tuning in signals at 112 and 116 Mc. alternately, determine if more or less capacity is required to peak the signal. By bending one end of the rotor plate of C_1, the mixer can be adjusted to resonate over the entire band.

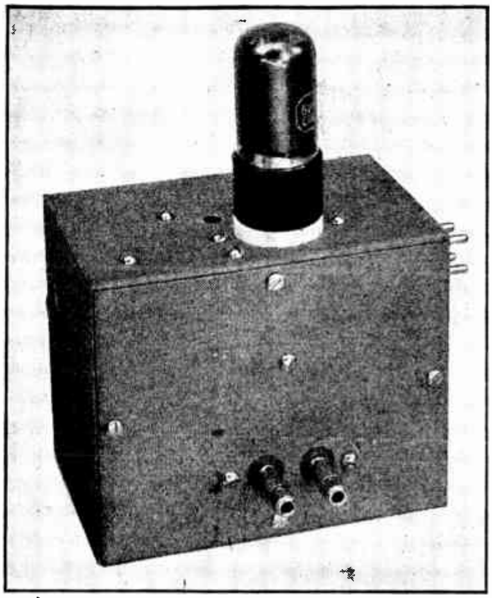


Fig. 1711 — A stable 112-Mc. transmitter. The oscillator is built in a small metal box, with only the tube, power plug and antenna posts on the outside. The small hole on the top beside the tube is for adjustment of the excitation condenser. The grommeted hole on the left side is for screw-driver adjustment of the tank condenser.

Control Station Transmitters

The 112-Mc. oscillator shown in Fig. 1711 is designed to minimize frequency modulation, and to that end is constructed around a high-*C* tank circuit of somewhat unconventional design. Reduction of frequency modulation is a step toward minimizing interference, since frequency modulation broadens the signal. The circuit requires a minimum of parts, although the tank condenser, C_1, may be difficult to obtain except from salvage stock.

The tank circuit consists of the balanced condenser, C_1, and the U-shaped metal piece whose dimensions are given in Fig. 1715. This "coil" is designed to have as much surface area as possible, thereby reducing resistance and losses, and also to provide the lowest possible contact resistance where it connects to the condenser. The ends of the U-shaped inductance fit under the stator-plate assemblies, which in the types of condensers specified are provided with flat holding plates to which the individual condenser plates are soldered. The slots in the ends of the U allow the inductance to be slid in and out to adjust the *L/C* ratio over a small range. To assemble the tank circuit the condenser must be dismounted from the base, and washers about the same thickness as the metal of the tank coil must be inserted between the base and the rotor supports. This raises the rotor to correspond to the increased height of the stators. It is not difficult to replace the stators so that the plate spacing is uniform. If the inductance is made exactly as specified, the slotted ends should come within about 1/16th inch of the far side of the base to give the proper frequency range.

The inductance shown in the photographs was cut from a small piece of scrap sheet copper somewhat less than 1/16th inch thick. Aluminum also works well. The metal should have low resistance, although its thickness is of no importance except for mechanical stiffness.

The oscillator is assembled in a 3 × 4 × 5-inch metal box as shown in Figs. 1713 and 1714. The various views should make the construction obvious. Chief considerations are to keep the grid and plate leads short, to which end the tube socket is mounted directly above the plate section of the tank condenser, with the latter just far enough below the plate prong to allow room for soldering a connection; and to keep the tank inductance as near the center

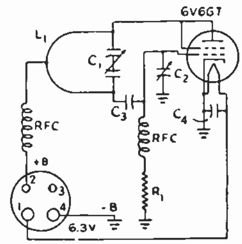


Fig. 1712 — Stable 112-Mc. transmitter circuit diagram.

C_1 — 100 μμfd. per section (Hammarlund MCD-100-S or Millen 24100).
C_2 — 3-30-μμfd. padder (National M-30, Millen 28030, Hammarlund MEX, etc.).
C_3 — 50-μμfd. midget mica.
C_4 — 250-μμfd. midget mica.
R_1 — 15,000 ohms, ½-watt.
L_1 — See Fig. 1608.
RFC — 1¼-inch winding of No. 28 d.s.c. on ¼-inch polystyrene rod, no spacing between turns.

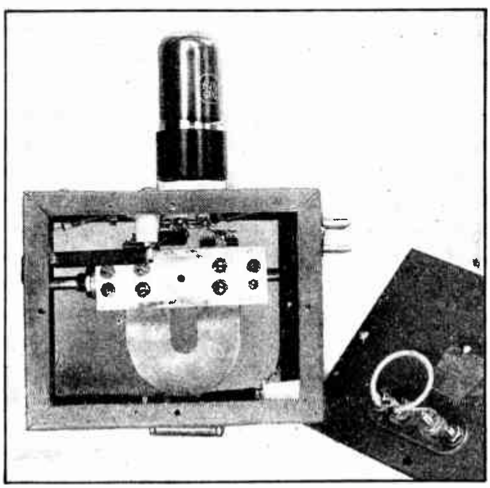


Fig. 1713 — Looking into the 112-Mc. transmitter from the antenna-terminal side. The grid choke is in the upper left corner, with its "hot" end supported by a small ceramic stand-off insulator. The plate choke is partly visible in the lower right corner; it is mounted endwise, also on a ceramic stand-off insulator. The one-turn antenna coil may be seen mounted on the antenna terminals on the side plate, shown here removed from the box.

of the box as possible so its flat sides will be well spaced from the steel side plates of the box. This spacing is accomplished by mounting the condenser on a 1-inch ceramic pillar fastened by a machine screw at the center hole in the base. The other end of the pillar is fastened to the side of the case. On the same side directly below is the r.f. output terminal assembly. The antenna pick-up coil is a 1-inch diameter single turn of No. 14 wire covered with spaghetti tubing. The antenna coupling is adjusted by bending the supporting leads for the pick-up coil to bring the turn closer to or farther away from the tank inductance. The coupling is ordinarily rather close, physically, because of the peculiar shape of the field about a tank inductance of this construction.

The tank condenser is screwdriver-adjusted, a slot being sawed in the end of the shaft. The rotor shaft of the condenser cannot be grounded since the circuit is not actually balanced; grounding the rotor changes the excitation and reduces the output to negligible proportions. The capacity between the rotor and the case also should be kept as low as possible.

The plate voltage is fed to the tank circuit near the center of the U. The lead from the cathode to ground should be as short as possible and made of heavy wire, likewise the lead from the grounded filament pin. The same connection may be used for both, and also for the No. 1 pin. The excitation condenser, C_2, should be mounted so as to keep it as far as possible from the plate section of the tank condenser.

Oscillator adjustment — The adjustments to be made are to determine whether the frequency range is correct and to set the output coupling and excitation for maximum stability

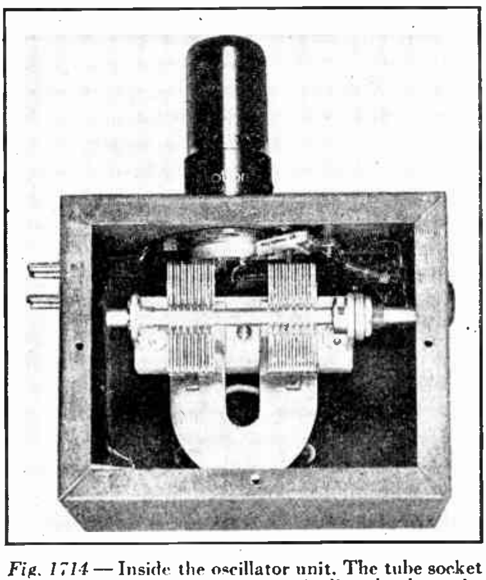


Fig. 1714 — Inside the oscillator unit. The tube socket is placed so that the plate prong is directly above the left-hand tank condenser stator terminal, making an extremely short plate connection. The grid condenser connects the grid prong and right-hand stator terminal.

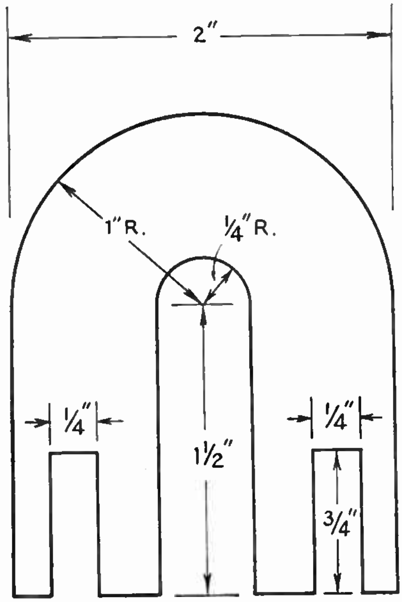


Fig. 1715 — Tank-inductance construction. This layout drawing is full-size and may be used as a template.

and output. The tank inductance will be properly adjusted when it is set (by sliding the ends in and out under the stator-plate assemblies) so that with the condenser at maximum capacity the frequency is between 111 and 112 megacycles. The frequency may be measured by using Lecher wires. The output may be judged by connecting a dial light (150-ma. size or larger) to the output terminals, whereupon varying the coupling and adjusting C_2 will readily show the optimum settings. The stability is more difficult to check unless a 112-Mc. superheterodyne receiver is available. However, the maximum stability is obtained when the capacity of C_2 is set at the largest value which will give good output, and it is advisable to adjust C_2 by first increasing its capacity to the point where the output drops off and then decreasing it just to the point where the output comes back to normal. As the capacity is decreased still more the output should decrease somewhat.

With normal operation the plate current, with load, should be between 50 and 60 ma. The exact value will vary somewhat with individual tubes, and if it tends to be outside these limits it may be regulated by using a slightly different value of grid leak, larger values giving less plate current and vice versa. The plate current will drop a few milliamperes when the load is removed.

To adjust the coupling for working into a 600-ohm line, a 1-watt resistor of 500 or 600 ohms may be used as a load. To indicate current through the resistor a 60-ma. dial light may be connected in series with it. A 150-ma. lamp also may be used, but is a less convenient indicator since it glows only dimly. The coupling should be adjusted for maximum current.

A master oscillator power amplifier — The 6V6GT unit shown in Fig. 1712 can be used as the oscillator with an 815 amplifier. A suitable m.o.p.a. is shown in Figs. 1716 and 1718. Not only does it provide for all the input power the regulations will allow, but it permits removal of modulation from the frequency-determining stage. From the practical operating standpoint, perhaps an even more important feature of such an arrangement is that it provides an effective measure of isolation between the antenna and the oscillator so that changes in position or length of the antenna do not drag the frequency all over the band.

The circuit diagram is shown in Fig. 1717. Excitation can be adjusted for best output and stability by means of C_2. An 815 dual pentode is used as a push-pull amplifier. The two stages are coupled inductively, since experience has proved this method to be superior to link coupling. The output tank is of the linear type. Jacks are provided so that milliammeters may be plugged in to read oscillator plate current and amplifier grid and plate currents.

The unit is built on the panel of a standard steel carrying case for portability. The outside dimensions of the case are 12 × 7 × 6 inches, the panel measuring 12 × 7 inches. The chassis is cut from a piece of aluminum sheet 15 inches long and 5 inches wide. A ¾-inch edge is bent down at each end of the sheet, the sheet then being bent into the form of a step at one end with a 3-inch "riser" and a 2½-inch "tread."

A separate inverted L-shaped mounting, also made of sheet aluminum, is provided for the oscillator components, as shown in Fig. 1718. It consists of a 4 × 5-inch piece of sheet metal bent at right angles at the center of the 5-inch dimension, each half then measuring 4 × 2½ inches.

The oscillator tank inductance, L_1, is a piece of 1/16-inch aluminum sheet cut to the shape of a U, 2 inches wide and 2½ inches long overall. A half-inch slot 1¾ inches long in the center makes the conductor ¾-inch wide at all points. To provide good surface contact between the inductance and the stators of the tank condenser, C_1, slots ¼-inch wide and ¾-inch long are cut in the ends of the legs of the U, spaced so that they will slide under the stator sections when the mounting screws of the latter are loosened up. Washers may be used to raise the rotor sections a corresponding amount. This assembly is then fastened to the under side of the horizontal part of the inverted-L mounting.

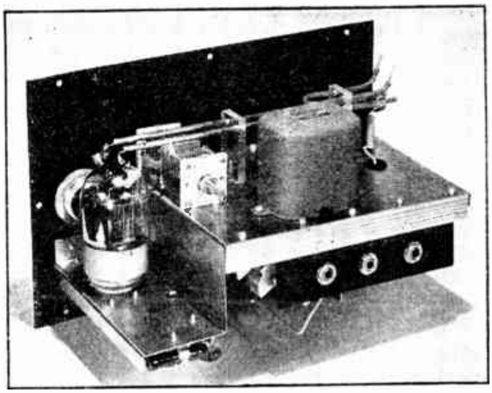


Fig. 1716 — Top view of the 112-Mc. m.o.p.a., showing the 815 amplifier tube and its linear plate tank circuit. The modulation transformer is mounted toward the rear of the chassis and the meter jacks are beneath on the bakelite strip. This unit was constructed by W2MYH.

All tank condensers are mounted with their control shafts facing toward the rear, from where they may be adjusted with a screwdriver. The oscillator-tube socket is mounted on the vertical part of the L mounting, so that its terminals are close to the condenser stators, with the tube in a horizontal position. After the oscillator unit is assembled and wired, it is fastened to the panel by means of aluminum angle pieces and self-tapping screws.

The amplifier grid-circuit components are mounted underneath the "tread" portion of the step at one end of the chassis. The grid tuning condenser, C_3, is placed at the inner edge of the "tread," leaving sufficient space alongside it for the 815 socket. The grid inductance, L_2, is a U-shaped piece of 3/16-inch copper tubing approximately 2 inches long and 1½ inches wide. The ends of the tubing are soldered to the stator terminals of C_3. The grid tuning condenser should be mounted so that L_2 is centered over L_1, with a separation of about 1 inch between the two inductances.

At the rear of the chassis an 8-inch length of angle stock provides a mounting for a bakelite strip which carries the three metering jacks.

The components of the linear output circuit for the 815 and the modulation transformer are

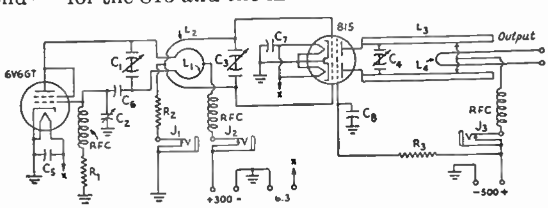


Fig. 1717 — Circuit of the 112-Mc. m.o.p.a.

C_1 — 100 μμfd. per section variable (Hammarlund MCD-100-S).
C_2 — 3-30-μμfd. mica trimmer (National M-30).
C_3 — 25 μμfd. per section variable (Hammarlund HFD-30-X).
C_4 — 15 μμfd. per section variable (Cardwell ET-15-AD).
C_5, C_7, C_8 — 250-μμfd. midget mica.
C_6 — 50-μμfd. midget mica.
R_1 — 15,000 ohms, ½-watt.
R_2 — 15,000 ohms, 1-watt.
R_3 — 10,000-ohm wirewound for plate voltage of 300; 15,000-ohm 10-watt wirewound for plate voltage of 400.
RFC — 1¼-inch winding of No. 28 d.s.c. wire on rod, ¼-inch close-wound, or Ohmite Z-1 v.h.f. r.f. choke.
L_1, L_2, L_3, L_4 — See text.
J_1, J_2, J_3 — Closed-circuit jack.

placed on the top side of the chassis. The line consists of a pair of ¼-inch copper tubes 9 inches long, spaced 1 inch center to center. They are supported on two pieces of polystyrene ½-inch thick, each of which has been tapped at the bottom edge so that it may be fastened to the chassis by a machine screw from underneath. The sections of tubing are passed through holes drilled in the polystyrene at such a height that the tubing sections almost touch the condenser stator terminals as they pass over C_4 to the plate terminals of the 815. The piece supporting the outer ends of the tubes is made somewhat longer than the other so that an extra pair of holes may be provided at the top for mounting the "hair-pin" antenna-coupling loop above the tubing sections. The loop is a piece of No. 12 wire covered with "spaghetti" and then bent to shape. The high-voltage lead to the amplifier passes up through a rubber grommet in the chassis to the r.f. choke, *RFC*, which is soldered to the sliding shorting bar which tunes the line.

The amplifier chassis is fastened to the panel by sections of angle stock. Plate and heater voltage are fed to the unit through a plug and receptacle near one end of the panel.

Tuning — In tuning the transmitter the oscillator is first set to the correct frequency by adjusting L_1. This is done by sliding it backward or forward under the stators of C_1 until it is possible to tune to the center of the 112-Mc. band with C_1 set near maximum capacity. With the oscillator running the grid circuit of the amplifier should be tuned for resonance, which will be indicated by maximum amplifier grid current. C_2 should now be turned toward maximum capacity until the grid current starts to fall off and then backed off slightly to restore normal grid current. Maximum stability will be obtained when C_2 is set as close to its maximum capacity setting as it can be without impairing the output of the oscillator.

The oscillator does a good job of driving the 815 amplifier. When the oscillator is adjusted correctly and the grid circuit of the amplifier is tuned to resonance, there should be no difficulty in obtaining a grid-current reading of 6 or 7 ma. It may be necessary to vary the coupling between the two circuits to obtain a maximum grid-current reading. When the oscillator is loaded, its plate current should run about 50 ma.

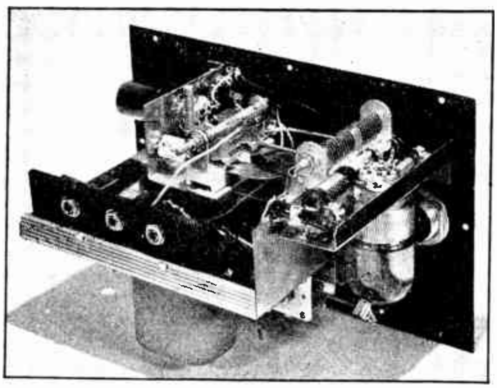


Fig. 1718 — Bottom view of the 112-Mc. m.o.p.a. transmitter. The amplifier grid-circuit components are grouped near the 815 tube socket at the right, while the oscillator section is built up as a separate unit on the inverted L-shaped aluminum bracket at the upper left.

Amplifier plate and screen voltages may now be applied and the plate circuit tuned to resonance. Adjustment of the shorting bar may be necessary to bring the band within the range of C_4. Plate voltage should be applied for only short intervals until the resonance point has been found.

With the antenna attached, the amplifier may be loaded by increasing the coupling until the plate current at resonance increases to about 150 ma. The plate voltage should be limited to 400 if the amplifier is to be plate modulated. Under load conditions, the grid current will drop to 3 or 4 ma. For WERS work, the unit should be operated at reduced input to comply with the regulations, which limit the permissible power input to 25 watts. With a 300-volt vibrator-pack supply the plate current to the amplifier should be adjusted to a maximum of about 80 ma.

The entire transmitter shown, as well as a modulator of sufficient power, may be operated economically from a dual vibrator-type power pack in emergency service with a carrier output of 15 watts.

A linear-tank circuit m.o.p.a. — The successful construction of an m.o.p.a. transmitter employing the more common type of triode tubes in the output stage is difficult with the type of oscillator described because of the increased driving power required. To solve this problem without sacrificing stability a high-frequency triode with a linear tank circuit may be used as the oscillator.

Such a unit is shown in Figs. 1719 and 1721. The circuit diagram appears in Fig. 1720. The oscillator circuit is of the simple shunt-fed ultraudion variety with a linear tank, connected between grid and plate, replacing the usual *L/C* tank. High voltage is fed in at the center of the line through the r.f. choke. The frequency of the oscillator may be set to the desired value by adjustment of the position of the line shorting bar. Excitation may be adjusted by means of the mica trimmer condenser, C_2. Bias is obtained from the grid leak, R_1.

The amplifier is coupled to the oscillator through the trimmer, C_3. A linear tank is also used in the plate circuit of the amplifier. The circuit is tuned by means of the small variable air condenser, C_5. Out-of-phase voltage for neutralizing is obtained by virtue of the grid-filament and plate-filament capacities of the tube which serve to split the tank circuit. C_4 is the neutralizing condenser connected between grid and the "bottom" of the tank circuit. The antenna is coupled by means of the well-known "hairpin" loop adjacent to the amplifier plate line. Bias is obtained from the grid leak, R_3.

The modulator is quite conventional with a 6L6 operating under conditions approximately Class A. The microphone is fed directly to the grid of the 6L6 through the mike transformer, T_1. The output of the modulator is coupled to the plate circuit of the r.f. amplifier by means of a 10-watt multimatch audio transformer, T_2.

The transmitter is contructed on a 7 × 9 × 3-inch chassis. The two r.f. tube sockets are submounted about the center of the chassis. Both oscillator and amplifier lines are made of copper tubing 5⁄16 inch in diameter. Each line consists of two sections of tubing, 15 inches long, bent into gooseneck shape, so that the upper end is close to one of the caps on top of the tube when the lower end is near the chassis. The lower end of each section is flattened out for about an inch so that the pair may be mounted, flat portions overlapping, on a single ceramic button or feed-through insulator set in the chassis. The sections are kept at a uniform spacing of one inch, center to center, by means of small strips of polystyrene drilled to fit the tubing. The oscillator line has additional support against vibration from a polystyrene strip extending from a metal angle piece fastened to the chassis to a similar angle piece fastened to the spacing strip on the line near the tube end. The shorting bar for adjusting frequency consists of a pair of metal strips about ½-inch wide and 1½ inches long, curved at the ends to fit the tubing, which clamp either side of the tubing by means of a machine screw at the center of the shorting strips. The oscillator grid condenser, C_1, is connected directly between one end of the line and the tube grid cap, while the other end of the line connects to the plate terminal. The r.f. choke and grid leak in series are connected between the grid cap and a 'phone-tip jack, set in the chassis, which serves as a feed-through insulator to the oscillator grid-meter jack. The excitation control condenser, C_2, connects between the grid terminal and the chassis.

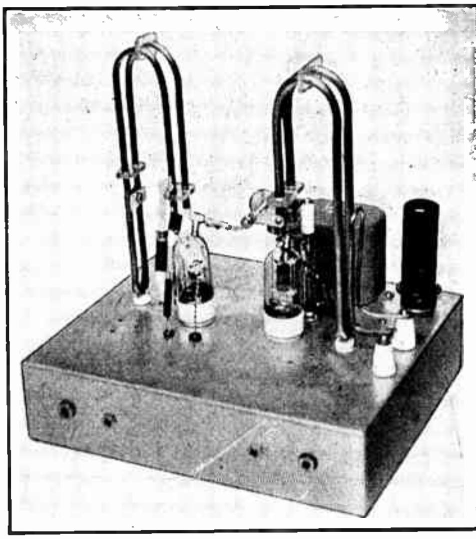


Fig. 1719 — The linear-tank circuit m.o.p.a. showing the mounting of the amplifier tuning and neutralizing condensers. The modulator is in the foreground. The jack in front is the mike connection and the socket for power-supply input. The lines are made of 5⁄16-inch copper tubing. This unit was constructed by W8UJB.

The coupling condenser, C_3, is suspended on heavy leads between the plate cap of the oscillator tube and the grid cap of the amplifier tube. The amplifier line is made similar to the oscillator line with the omission of the shorting bar. C_5 is a small variable-gap air condenser of the type sometimes used in neutralizing beam tubes. The plates are one inch in diameter. The supporting strips for the plates are fastened directly to the ends of the tubing with machine screws through the tubing. The neutralizing condenser, C_4, which is of the same type as C_5, is mounted on top of a 2½-inch standoff insulator in a position close to the grid cap of the amplifier tube. Very short leads are then required to connect one side of the condenser to the grid cap and the other side to

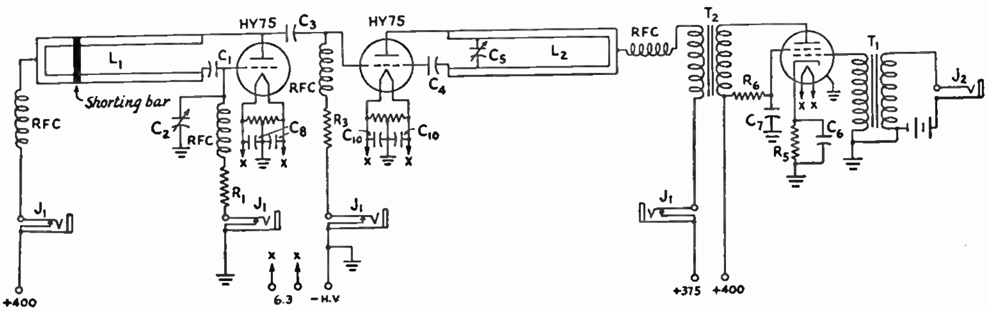


Fig. 1720 — Circuit diagram of the parallel-line linear-tank circuit m.o.p.a.

C_1 — 50-μμfd. silvered-mica fixed.
C_2, C_3 — 3-30-μμfd. ceramic mica trimmer (Hammarlund type MEX).
C_4, C_5 — Adjustable-gap-type air neutralizing condenser, 1-inch plates (Bud NC890).
C_6 — 8-μfd. electrolytic midget.
C_7 — 0.01 μfd.
C_8, C_9, C_{10}, C_{11} — 0.005-μfd. mica.
R_1 — 5000 ohms, 1-watt.
R_2, R_4 — 50-ohm center-tapped.
R_3 — 10,000 ohms, 1-watt.
R_5 — 750 ohms, 10-watt.
R_6 — 25,000 ohms, 1-watt.
L_1, L_2 — See text.
T_1 — Microphone transformer, single-button to grid.
T_2 — 10-watt modulation trans. (UTC S18 or S19).

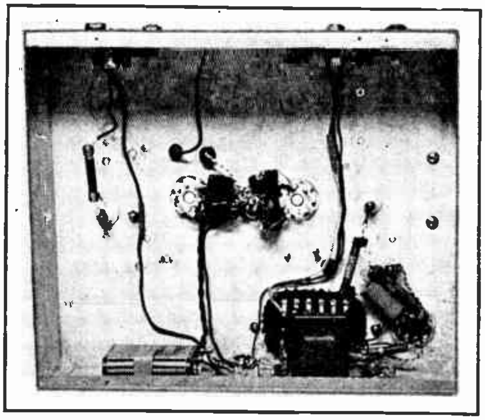


Fig. 1721 — Bottom view of the linear-tank circuit m.o.p.a., showing r.f. filament by-passing and microphone transformer and battery. Sockets are submounted.

the proper end of the line. The amplifier grid r.f. choke and grid leak in series are mounted in a manner similar to corresponding units on the oscillator. The antenna coupling loop is made of heavy wire so as to be self-supporting and is mounted on a pair of small stand-off insulators opposite the amplifier plate line.

The 6L6 socket and the output transformer are mounted at the back of the chassis, while the mike transformer is fastened to the rear edge underneath. Meter jacks are placed along the front edge of the chassis, while the mike jack and power-cable connector are at the rear. The microphone jack and the two plate-current jacks must be insulated from the chassis, while the two for grid current may be mounted without insulation. Care should be exercised in using the plate-current jacks, since full plate voltage will appear between the exposed portion of the jack and the chassis.

The entire transmitter may be operated from a single 400-volt, 175-ma. supply. An additional section of filter is used in the high-voltage lead supplying the modulator. This consists of a choke and an 8-μfd. condenser. The purpose of this extra section of filter is both to drop the voltage slightly to 375 for the 6L6 and also to prevent changes in modulator plate current at an audio rate from producing corresponding changes in oscillator plate voltage because of insufficient voltage regulation of the power supply.

Plate voltage should not be applied to the amplifier until it is neutralized. The frequency of the oscillator should be first set to the center of the band by adjustment of the shorting bar, moving it toward the tube to increase frequency and away from the tube to decrease frequency. If the dimensions given are followed closely, the circuit should tune to 113 Mc. with the shorting bar 2½ inches from the chassis end of the line. C_2 and C_3 should then be adjusted for maximum amplifier grid current consistent with reasonable oscillator input. It should be possible to obtain a grid current of at least 15 ma. without difficulty. Since the adjustment of both C_2 and C_3 will have an effect upon the frequency of the oscillator, the frequency should again be set to the center of the band before attempting to neutralize the amplifier.

With the oscillator running, there will be a noticeable change in amplifier grid current when C_5 is tuned through resonance. Keeping C_5 adjusted so that the amplifier is near resonance, the neutralizing condenser, C_4, should be adjusted in small steps until the grid current remains steady as the plate current is tuned through resonance, indicating that the amplifier is neutralized. The oscillator may now be set to the desired frequency and C_2 and C_3 readjusted for optimum performance. An indication of excitation may be obtained by checking the grid current of the oscillator. With the amplifier coupled, the oscillator grid current should run between 10 and 12 ma., while the plate current will be approximately 60 ma.

With no antenna attached to the output terminals, and the oscillator running, plate voltage may now be applied to the amplifier and the plate circuit tuned accurately to resonance as indicated by the dip in plate current. If the circuit is working properly, the plate current should dip to a minimum of about 18 ma. Since no fixed bias is provided, plate voltage should never be applied to the amplifier until after the oscillator has been turned on and is supplying excitation. When loaded, the amplifier plate current should not fall below 10 to 12 ma.

In coupling the modulator to the plate circuit of the r.f. amplifier, the output-transformer taps are adjusted for about 5000 ohms primary and 6500 ohms secondary. The use of a screen resistor of 25,000 ohms and a cathode resistor of 750 ohms results in a static plate current of about 50 ma. to the 6L6.

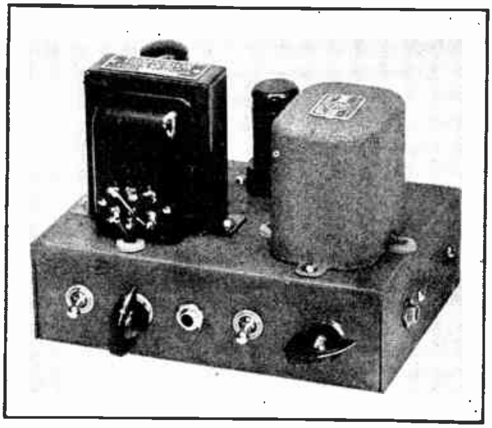


Fig. 1722 — The Class-B modulator unit. The output transformer is at the left, the driver transformer at the right. Controls along the front are send-receive switch, 'phone-c.w. switch, key jack, microphone battery switch, and gain control. The microphone jack is on the right-hand edge, around the corner from the gain control.

Class-B modulator — Modulators suitable for use with these control station transmitters are required to deliver between 6 and 12 watts of audio output depending on the input power employed. For a 25-watt input, the modulator may be a conventional Class-B arrangement, using a 6N7 driven by a 6J5. Class-B is used because of its higher plate efficiency and relatively low idling plate current. The oscillator load will be between 5000 and 6000 ohms, depending upon the plate current, and it will be sufficient to take the nearest value furnished by the output transformer, using a plate-to-plate load of 8000 ohms for the 6N7. There is ample gain with the single speech amplifier stage for ordinary single-button microphones operated from a 3-volt battery.

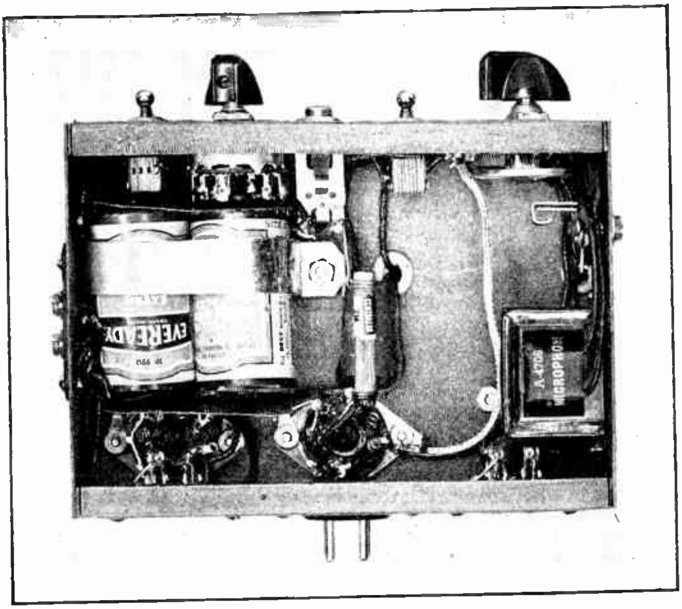


Fig. 1723 — Underneath the modulator chassis. The microphone transformer is mounted on the chassis edge alongside the microphone jack. The power plug and the two outgoing power sockets for the transmitter and receiver are mounted on the rear edge of the chassis (bottom edge in this view). A terminal strip for connecting an external microphone battery is located on the left-hand edge. The two flashlight cells of the microphone battery are held in place by a metal strip; the cells are protected from accidental short-circuit by a piece of thin fiber or cardboard which is bent in the shape of a U to cover the terminal.

In the units shown in Figs. 1722–1724, power input and output connections conform to the standards previously described. To give tone modulation for code transmission, the speech amplifier tube is made to oscillate by connecting the primary of the microphone transformer as a tickler in series with the plate circuit. A four-pole double-throw switch is necessary to change from 'phone to c.w., two poles being used to transfer the primary of T_1, a third to close the plate circuit for 'phone, and the fourth to disconnect the cathode condenser for tone modulation. This last is essential for good keying (the speech amplifier tube is keyed in the cathode circuit). The c.w. tone pitch depends upon the value of the cathode resistor and the volume control setting, but with several microphone transformers tested falls in the optimum region (500–1000 cycles) with a 2000-ohm cathode resistor. When modulated c.w. operation is not required, it may be omitted by substituting the alternative input circuit of Fig. 1727.

A separate switch is provided to open the microphone battery circuit whenever desired. The battery would normally be left on while receiving when communication is being carried on, but during stand-by periods it would be desirable to switch off the microphone current to prolong battery life. The same effect can be secured by pulling the microphone plug out of the jack, but the switch is more convenient. A battery of two flashlight cells connected in series is made a permanent part of the unit, since there is sufficient mounting room under-

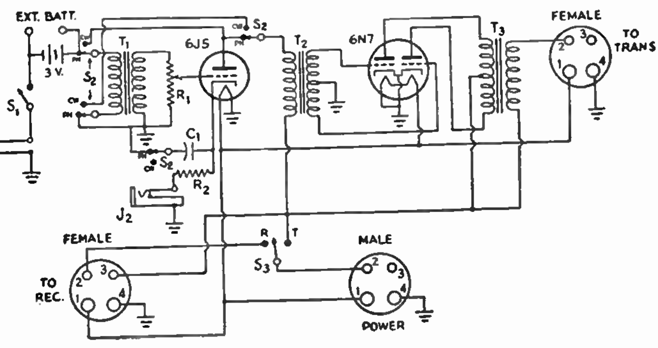


Fig. 1724 — Circuit diagram of the Class-B modulator unit.
C_1 — 10-μfd. 50-volt electrolytic.
R_1 — 0.5-meg. volume control.
R_2 — 2000 ohms, 1-watt.
T_1 — Single-button microphone-to-grid transformer (Stancor A-4706 or equivalent).
T_2 — Class-B driver transformer (UTC S8 or equivalent).
T_3 — Class-B output transformer, 6N7 to 5000–6000 ohms (Thordarson T-19M13 or equivalent).
J_1 — Open-circuit jack.
J_2 — Closed-circuit jack.
S_1 — S.p.s.t. toggle switch.
S_2 — 4-p.d.t. rotary switch (Yaxley 3242J or equivalent).
S_3 — S.p.d.t. toggle Switch.

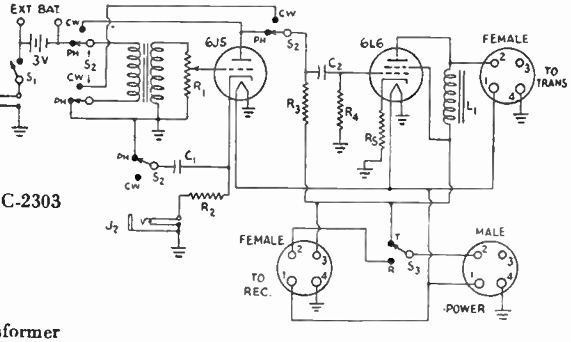


Fig. 1725 — Circuit diagram of the alternative choke-coupled 6L6 Class-A modulator.

C_1 — 10-μfd. 50-volt electrolytic.
C_2 — 0.01-μfd. paper.
R_1 — 0.5-megohm volume control.
R_2 — 1500 ohms, 1-watt.
R_3 — 50,000 ohms, 1-watt.
R_4 — 0.25 megohm, ½ watt.
R_5 — 500 ohms, 1-watt.
L_1 — 10-15-henry 100-ma. filter choke (Stancor C-2303 or equivalent).
J_1 — Open-circuit jack.
J_2 — Closed-circuit jack.
S_1 — S.p.s.t. toggle switch.
S_2 — 4-pole double-throw rotary switch.
S_3 — S.p.d.t. toggle switch.
T_1 — Single-button microphone-to-grid transformer (Stancor A-4706 or equivalent).

neath the chassis, but additional terminals are provided for an external battery should the internal one wear out during an emergency. To use an external battery it is necessary to disconnect one of the leads to the self-contained battery unit.

The microphone jack is mounted on the side of the chassis, so the microphone plug and cord will be out of the way of the controls on the front. The key jack is on the front. Since the modulator unit is small (the chassis is 5 × 7 × 2 inches) the send-receive switch is placed at the end, where it is easiest to handle.

The plate current taken by the modulator and speech amplifier tubes is in the vicinity of 35 ma. with no excitation. When the r.f. oscillator is added, the current drain is just under 100 ma. With 100 per cent voice modulation, the maximum current is 110 to 115 ma.

Class-A modulator — While the Class-B type of modulator is to be preferred because of its higher audio-frequency power output for a given plate power input, Class-B transformers are practically unobtainable at the present time. Therefore, unless suitable transformers can be salvaged from old equipment, the probability is that a Class-A modulator will have to be used. Such a modulator, using a 6L6 with a preceding 6J5 as a speech amplifier, is shown in Fig. 1726. Provision for tone modulation also is incorporated in this unit, and again may be omitted by using the alternative speech amplifier connections of Fig. 1727.

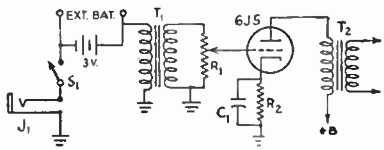


Fig. 1727 — Alternative microphone input circuit for eliminating the tone-modulation feature incorporated in the modulator circuits of Figs. 1724 and 1725. Circuit values correspond to those given in Figs. 1724 and 1725.

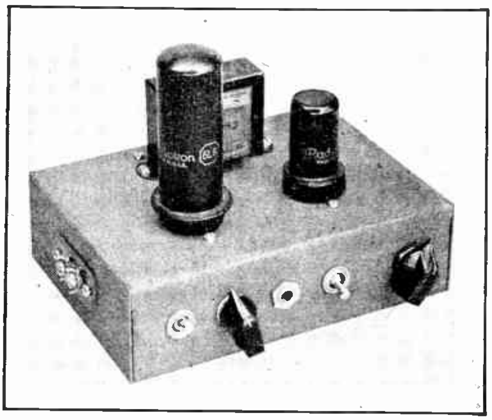


Fig. 1726 — A Class-A 6L6 choke-coupled modulator, using a minimum of transformers. The controls along the front are, left to right, the send-receive switch, 'phone-c.w. switch, key jack, microphone battery switch, and gain control. Terminals for external microphone battery are on the left edge. The microphone jack, not visible, is mounted on the right-hand edge of the chassis.

Any filter choke capable of maintaining an inductance of 10 henrys or more with 100 ma. d.c. through its winding will serve as a coupling choke for the modulator. The higher the inductance the better the low-frequency response, but since "quality" is not a consideration so long as completely understandable speech is transmitted it is unnecessary to use higher inductance than is found in the ordinary 100-ma.choke.

To keep the total modulator plate current down to 40 or 45 ma. and thus avoid overloading the vibrator power supply (which delivers 100 ma. at 300 volts) when the modulator and r.f. oscillator are operated simultaneously, the cathode resistor is higher than is normal for a Class-A 6L6 at this plate voltage.

The plate-voltage switching and the input and output socket and plug arrangement are identical with the Class-B unit already described.

Controls and power-supply outlets are arranged similarly to the controls on the Class-B modulator. The chassis is the same size, 5 × 7 × 2 inches. There are no especially critical points involved in wiring, and practically any parts layout will be satisfactory. The two-cell microphone battery is held in place by a bracket fitting around the battery, fastened to the chassis by the machine screws which mount the choke.

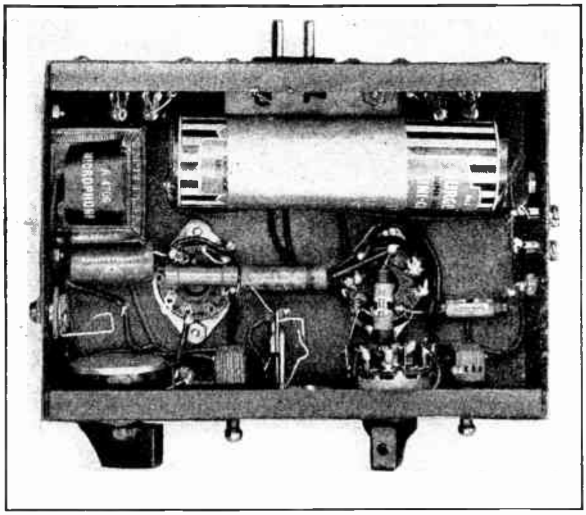


Fig. 1728 — Underneath view of the 6J5-6L6 Class-A modulator unit. Parts are placed wherever found most convenient in wiring.

Transmitter-Receivers

It frequently is convenient to build the transmitting and receiving equipment in one assembly, arranged with a switch to shift both the antenna and plate supply from "receive" to "transmit" in a single operation. A station of this type is shown in Figs. 1729–1732.

In the circuit diagram of the transmitter-receiver, in Fig. 1731, a 955 acorn tube is used as the self-quenching superregenerative detector, while the transmitter section is built around an HY75. Each of these circuits is coupled to the output terminals by a low-impedance link (L_1 and L_4). The acorn as a detector not only is more efficient than a conventional tube such as the 6J5, but will operate satisfactorily at a much lower plate voltage, thereby reducing receiver radiation. The padder condenser, C_3, is for the purpose of adjusting the tuning range of C_1 to cover the band. R_{11} is the regeneration control.

The audio section, which consists of a 6J5 resistance-coupled stage and a 6V6 Class-A output stage, is used both as an audio amplifier, feeding the 3-inch p.m. dynamic speaker, and as a modulator for the transmitter. The use of the transceiver transformer at the input of the audio section makes it unnecessary to switch the input back and forth between the microphone and the detector output; the separate windings for each permit both to be permanently coupled. When transmitting, the output of the detector is cut off by opening the cathode circuit, while the microphone circuit is opened up during the receiving periods. Separate gain controls are provided for the audio amplifier in the receiving and transmitting positions, so that speaker volume and microphone gain may be adjusted independently to the desired levels.

A 4-p.d.t. switch, S_1, takes care of all switching operations in changing over between transmitting and receiving positions. In the receiving position, the antenna input is connected to the detector, the detector cathode circuit is closed, the grid of the 6V6 is connected to the receiver volume control and the output of the 6V6 is connected to the speaker input transformer. In changing over for transmission, the antenna input is connected to the transmitter, the microphone circuit closed, the grid of the 6V6 connected to the microphone gain control and the output of the 6V6 connected in series with the plate-voltage lead to the HY75.

The chassis, which is of standard 7 × 9 × 2-inch size, has adequate space for all components without crowding. Viewed from the rear, the detector occupies the right-hand side of the chassis while the transmitter oscillator is at the left. The audio equipment makes use of the space at the center between the two r.f. units. The mounting for the 955 acorn tube in the detector is quite unique. The ceramic acorn socket is mounted on an inverted Johnson No. 20 stand-off insulator. The holes in the

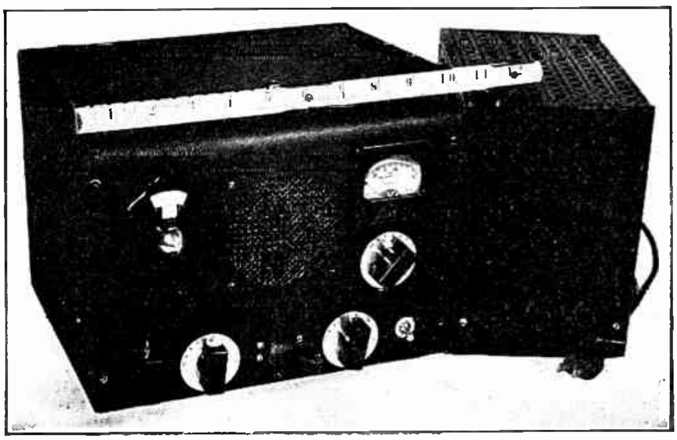


Fig. 1729 — A 2½-meter transmitter-receiver in its cabinet and with accompanying power-supply unit. Power may be obtained from any power pack, a.c. or vibrator pack, delivering about 300 volts, 100 ma. The dial at the left on the panel is for tuning the receiver, while the transmitter tuning control is immediately under the plate milliammeter. Along the bottom, from left to right, are the detector regeneration control, the receiver audio gain control, the change-over switch, and the transmitter audio gain control. The toggle switch shown at the right was later replaced by a microphone jack. This unit was constructed for combination portable and portable mobile work by W2DKJ.

Fig. 1730 — Plan view of the transmitter-receiver. The acorn detector assembly is at the right, the transmitter oscillator at the left, and the audio equipment at the center.

mounting feet of this insulator are spaced exactly right to match the mounting holes of the National type XCA acorn socket. The recess in the stand-off provides room for the projecting glass bottom of the tube and protects it from injury. The screw at the terminal end of the insulator is used for mounting the assembly on the chassis. This brings the tube socket 1¾ inches above the chassis, where there is just room to squeeze the grid condenser, C_3, between the grid terminal of the tube socket and the stator terminal of the tuning condenser, C_1, which is mounted on the panel with spacers at three of the four corners. It is placed on the panel so that the shaft is 2¾ inches above the chassis and 2 inches from the edge of the panel. A couple of short pieces of wire serve to connect the padder, C_2, in parallel with C_1 and to mount it with its adjusting screw toward the rear so that it may be reached with a screw-driver after the unit is placed in the cabinet. The rotor of C_1 is grounded to the panel through a soldering lug under one of the condenser mounting screws. The other upper mounting screw holds a bracket for the dial light. The coil, L_2, is soldered directly across the rotor and stator terminals of C_1. The two-turn antenna-coupling link is fastened in place above L_1 by soldering one end to the grounding lug mentioned above and the other end to one of the mounting screws of the acorn-tube socket which serves as an insulated anchorage. A wire then goes from this anchorage down through a small polystyrene feed-through insulator in the chassis to the proper change-over switch terminal underneath.

The two u.h.f. r.f. chokes in the detector circuit are mounted vertically from the plate and cathode terminals of the tube socket, their lower ends being supported on small fibre washers, while connections to the lower ends of the choke windings are made with push-back wire passing through small holes in the chassis under the washers. The grid leak, R_1, and the by-pass condenser, C_4, are soldered directly between the socket terminals and the

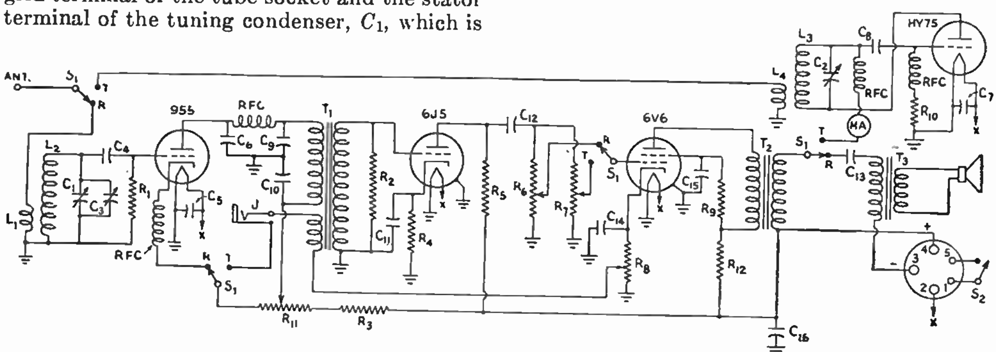


Fig. 1731 — Circuit diagram of the W2DKJ transmitter-receiver.

C_1, C_2 — National UM15 midget variable, with all but 1 stator and 1 rotor plate removed.
C_3 — 30-μμfd. mica trimmer (National M-30).
C_4 — 50-μμfd. mica.
C_5, C_6, C_7, C_8 — 100-μμfd. mica.
C_9 — 0.006 μfd.
C_{10} — 0.5-μfd., 200-volt paper.
C_{11} — 10-μfd., 25-volt electrolytic.
C_{12}, C_{13} — 0.1-μfd., 400-volt paper.
C_{14} — 25-μfd., 25-volt electrolytic.
C_{15} — 0.25-μfd., 400-volt paper.
C_{16} — 8μfd., 450-volt electrolytic.
R_1 — 2 megohms, ¼-watt.
R_2, R_3 — 0.25 megohm, ¼-watt.
R_4 — 2000 ohms, ½-watt.
R_5 — 0.1 megohm, ¼-watt.
R_6, R_7 — 0.5-megohm potentiometer (Centralab 103).
R_8 — 500 ohms, 10-watt, with variable tap.
R_9 — 30,000 ohms, 1-watt.
R_{10} — 5000 ohms, 10-watt.
R_{11} — 50,000-ohm potentiometer.
R_{12} — 2000 ohms, 25-watt.
L_1, L_4 — 2 turns No. 16 insulated wire, ⅜-inch diameter.
L_2 — 4 turns No. 16 wire, ½-inch diameter, turns spaced approximately the diameter of the wire.
L_3 — 5 turns No. 16 wire, ½-inch diameter, turns spaced approximately the diameter of the wire.
MA — Milliammeter, scale 0–100 ma.
RFC — U.h.f. r.f. choke (Ohmite Z-1).
S_1 — Poles of 4 p.d.t. switch (Federal anti-capacity).
S_2 — Volume-control power switch on R_6
T_1 — Transceiver transformer (Thordarson T72A59).
T_2 — Modulation, approximately 10,000 ohms primary to 5000 ohms secondary, 10-watt (Thordarson T17M59).
T_3 — Universal speaker transformer.

chassis, while C_5 is soldered across the filament terminals of the socket.

Since both sides of the transmitter tuning condenser, C_2, are "hot," the components comprising the transmitter section are grouped back from the panel to reduce hand-capacity effects when tuning. The shaft of the condenser is extended to the panel by using an Isolantite flexible shaft coupling and a section of ¼-inch bakelite rod. The condenser is mounted on the chassis by means of a small piece of angle brass. The panel bearing for the shaft extension is 1⅛ inches above the chassis and 2 inches from the edge of the panel in line with the center of the milliammeter immediately above. L_3 is mounted by soldering it across the terminals of C_2. The two r.f. chokes shown in the transmitter circuit are mounted either side of the tuning condenser. The HY75 is placed close behind this assembly, with its socket submounted on angle pieces to lower slightly the grid and plate terminals at the top.

The modulation transformer, T_2, is mounted at the center of the chassis immediately behind the chassis, while T_3 is fastened underneath the speaker. Sockets for the two audio tubes are submounted along the back edge of the chassis.

The microphone jack must be insulated from the chassis with fibre washers. The transceiver transformer, T_1, is fastened to the left-hand edge of the chassis. A jack terminal strip (National FWJ) is located at the rear edge for making convenient connections to a low-impedance line to the antenna.

Exact placement of resistors and most of the fixed condensers underneath the chassis is unimportant. However, C_7 should be connected across the tube-socket terminals and C_9 should be placed close to the r.f. choke.

Mobile transmitter-receiver — An alternative is to construct the transmitter and receiver as a unit and install it in some convenient spot, such as the glove compartment of the car. Such an installation is shown in Figs. 1733 to 1735. Normally the glove compartment will be too small to contain the necessary apparatus, but, if the fiber box is removed, the space behind the dash will, in most cases, be ample. In the set illustrated, the fiber box is replaced by a metal box of the same dimensions, backed by the chassis on which the transmitter and receiver are mounted. The box and chassis are made of No. 24 gauge sheet metal. If patterns are prepared beforehand, it will be possible to have the metal work done by a tinsmith at comparatively small cost. The actual dimensions naturally will be determined by the shape and size of the space available in the particular model of car in which the station is to be installed. The chassis shown in the photographs measures 12 inches wide and 10½ inches deep.

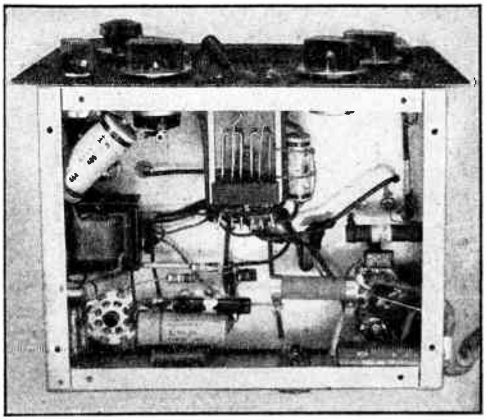


Fig. 1732 — Bottom view of the W2DKJ transmitter-receiver. The transformer at the left is the transceiver transformer, T_1. The change-over switch is at the center.

The circuit diagram of the mobile unit is given in Fig. 1735. Electrically, it is practically equivalent to the transmitter-receiver circuit of Fig. 1738. The transmit-receive switch is a Federal Radio anti-capacity type, with "outboard" antenna change-over contacts added. Below the change-over switch is the 1½-inch permanent-magnet dynamic speaker. To the left side of the speaker are the volume control and microphone jack. C_5, C_6, C_7, C_{16}, RFC_3 RFC_4, R_2, and the tank circuit, C_5L_4, are clustered around the base of the 6V6 oscillator tube. Trial showed that C_7 is a worth-while addition; while the 6V6 will oscillate without it, the output is considerably less.

To the right side of the speaker are the regeneration control and filament switch, the pilot light and the detector circuit. C_1, C_3, RFC_1, R_1, and the tank circuit, C_2L_1, are bunched about the base of the 6J5G. RFC_2 and C_4 are mounted on the underside of the chassis pan. RFC_2 is a low-frequency r.f. coil which, together with the associated condensers, forms a filter for the quench frequency.

The high-frequency chokes, RFC_1, RFC_3 and RFC_4, consist of 55 turns of No. 30 enameled wire wound on a drill shank, then sprung loose and fastened with Duco cement. The chokes

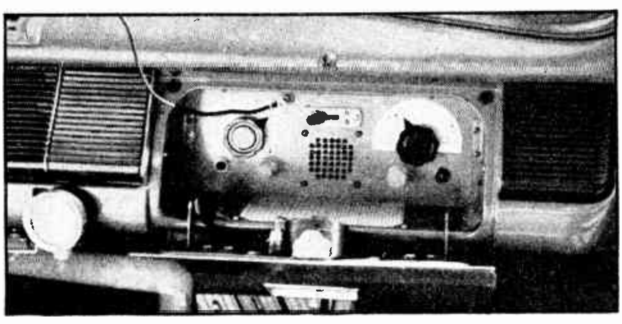


Fig. 1733 — A transmitter-receiver unit for mobile use, installed in the glove compartment of a car. This installation was made by W2DVG.

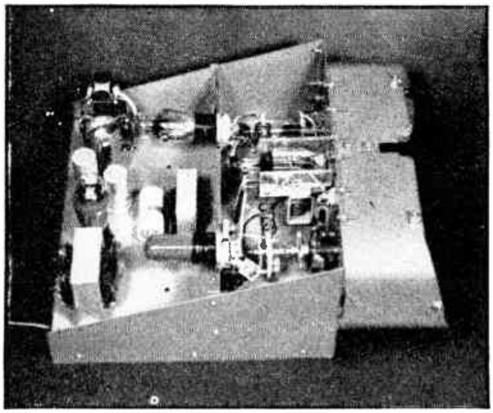


Fig. 1734 — Side view of the mobile unit, showing the chassis and mounting-hood arrangement. The metal chassis and hood are made of No. 24 gauge "black metal." Plan view of the mobile station, showing location of components. The oscillator and detector tubes are mounted horizontally from the partition, with associated tank-circuit components arranged close to the tube sockets. The audio equipment is at the rear. Most of the power wiring is concealed underneath the chassis.

are very light in weight and are safely mounted by the wire with which they are wound.

Directly behind the tube-mounting partition and between the detector and oscillator tubes is the filter choke, L_5. (This outfit obtained its plate power from the car broadcast receiver, which had a resistance-capacity filter. When this power was used for the transmitter-receiver, the filter resistor was shorted out, leaving only the condensers; hence the necessity for L_5 and C_{15}.)

Behind L_5 is the audio coupling condenser, C_{12}, which is too large to fit below deck. Left to right, across the back of the chassis, are the output transformer, the modulator tube, condensers C_9, C_{13}, C_{14} and C_{15}, the first audio tube and the audio input transformer. The latter had enough room between the original windings and the core for the installation of approximately 50 turns of No. 30 enameled wire for microphone input. The new winding is covered with cellophane tape to keep it in place and to protect it from moisture.

Beneath the chassis are RFC_2, C_4, R_5, R_6, R_7, R_8, R_9, C_{10}, C_{11}, R_{11}, and most of the audio and power wiring. Only two leads, each fused, leave the chassis. These are marked "A+" and "B+" in Fig. 1735. A third connection for negative returns is made through the frame of the car.

The numbered dials available were either too large to fit on the panel or too small for comfortable gripping. Pointer knobs were used, therefore, together with home-made scales. The scales were drawn in India ink on the backs of filing cards cut to size and fastened to the panel with rubber cement. The white plastic-knob pointers did not provide sufficient contrast with the white scale cards, and so the pointers later were roughed up with fine sandpaper and coated with black India ink.

The volume- and regeneration-control knobs are made of ¾-inch Lucite rod, drilled for shaft size and drilled and tapped for set screws. They project out from the panel an inch and one-half for easy handling.

The regular car-radio antenna is used for the transmitter-receiver. It is of the telescoping type, passing through the roof. The antenna is connected to the switch terminal by an 18-inch lead and the system extended to ¾ wavelength. A vernier adjustment for antenna tuning is obtained by sliding the antenna up or down inside the car. This does not change its physical length but alters the portion which closely parallels the windshield dividing strip.

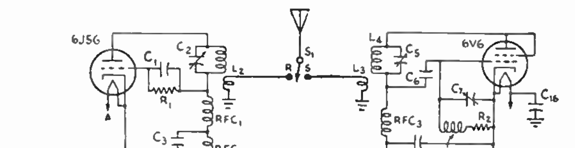


Fig. 1735 — Circuit of the mobile station.
C_1 — 20-μμfd. mica.
C_2, C_5 — 3-plate variable (Cardwell Trim-Air).
C_3 — 0.006-μfd. mica.
C_4, C_6, C_8 — 0.001-μfd. mica.
C_7 — 3–30-μμfd. trimmer.
C_9 — 20-μfd. electrolytic.
C_{10}, C_{11} — 5-μfd. electrolytic.
C_{12} — 0.5-μfd. paper.
C_{13}, C_{14}, C_{15} — 20-μfd. electrolytic.
C_{16} — 100-μμfd. mica.
R_1 — 5 megohms, ½-watt.
R_2 — 15,000 ohms, 1-watt.
R_3 — 0.5-megohm pot.
R_4 — 2250 ohms, 1-watt.
R_5 — 0.1 megohm, ½-watt.
R_6 — 20,000 ohms, 1-watt.
R_7 — 0.5 megohm, ½-watt.
R_8 — 500 ohms, 1-watt.
R_9 — 200 ohms, 1-watt.
R_{10} — 50,000-ohm pot.
R_{11} — 50,000 ohms, 2-watt.
L_1, L_4 — 3 turns No. 14, ½-inch diameter, turns spaced so tanks tune to WERS band.
L_2, L_3 — 1 turn No. 14, ½-inch diameter.
L_5 — Replacement-type filter choke.
B — 6-volt pilot lamp.
RFC_1, RFC_3, RFC_4 — 55 turns No. 30, self-supporting.
RFC_2 — Low-frequency choke coil.
S_1 — Anti-capacity change-over switch (see text).
S_2 — Filament switch (on R_{10}).
Spkr — 1½-inch p.m. speaker (Cinaudagraph).
T_1 — Transceiver transformer (WE213-D with microphone winding of 50 turns No. 30 e. See text).
T_2 — Universal output transformer, push-pull type.

Transceivers

Although differing widely in constructional details, most 112-Mc. transceivers for WERS employ the fundamental circuit arrangement shown in Fig. 1736-A.

The basic transceiver circuit is a conventional parallel-fed ultraudion. Any of the more numerous types of receiving-type triodes (or pentodes with screen and suppressor grids connected to plate) will work satisfactorily. For operation at 200 volts or less a 6J5GT or HY615 is preferable as the oscillator, since they work more efficiently than some of the other types. However, a metal 6J5 or 6J5G, 6C5, 6F5, 6K5G, 6P5G or 7A4 could be used as the detector-oscillator with minor adjustments in coil dimensions and grid-leak resistance. Similarly, a 6A4, 6V6G, 6L6, 6L6G, 6K6G, 6G6G or 6F6G will serve in the output stage.

The r.f. tube takes 20 or 25 ma. at 200 volts when transmitting, and has an r.f. output of a watt or so. Including the audio system, the total current drain in the transmit position is in the neighborhood of 60 ma. at 200 volts. In reception the plate current of the r.f. tube is negligible, and the total current at 200 volts is only about 35 ma. An easily constructed vibrator-type power supply for 200-volt operation is described in a later section. Where a 300-volt supply is used, a 6V6GT or similar detector-oscillator tube is recommended.

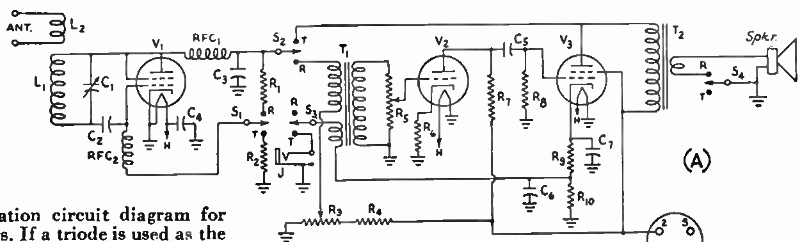
The "transceiver transformer," T_1, as used in Fig. 1736-A is an ordinary interstage audio (about 3:1 ratio) with a microphone primary added. There is usually enough space between the core and the windings to get in at least one layer of fairly fine wire, such as No. 30. It is necessary to take the core apart and possibly to remove some of the paper already around the windings. In the unit shown, the microphone primary is one layer of No. 30 s.c.c. (about 50 turns) wound over the existing windings. It was given a coat of shellac to hold it in place, and covered with paper to prevent short circuits to the core.

If a single switch wafer of the desired number of poles and circuits cannot be obtained, any 4-pole double-throw switch may be used. Usually some sort of wafer switch can be salvaged from old equipment; if it is necessary to use more than one gang, the only result is that the switch is more bulky.

The regeneration control divider, consisting of R_3 and R_4 in series, permits operating the detector at the lowest plate voltage consistent with good superregeneration, and thus holds receiver radiation to a minimum. The fixed resistor makes the setting of the control less critical, and also keeps the voltage across the variable resistor to a safe value.

The microphone current is obtained from the cathode circuit of the modulator tube, by tapping the microphone across part of the cathode resistor. The single by-pass condenser from cathode to ground is sufficient to prevent feed-back between the modulator and microphone circuits. In reception the microphone circuit is opened by the switch, with the result that the bias on the output tube rises and the plate current is reduced.

The use of a common audio volume control for transmitting and receiving means continual



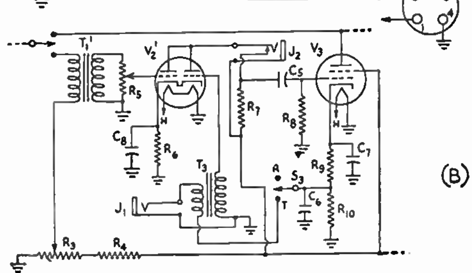


Fig. 1736 — **Combination circuit diagram for low-power transceivers. If a triode is used as the oscillator (V_1), omit the screen-grid connection.**

C_1 — 10-15 μμfd. midget variable.
C_2 — 50-μμfd. mica.
C_3 — 0.005-μfd. mica.
C_4 — 250-μμfd. mica.
C_5 — 0.1-μfd. paper, 400 volts.
C_6 — 25 to 50 μfd. electrolytic, 50 volts.
R_1 — 5 megohms, ½-watt.
R_2 — 5000 ohms, 1-watt (for 6J5, 6C5); 10,000 ohms, 1-watt (for 6V6, etc.).
R_3 — 0.5-megohm volume control.
R_4 — 1000 ohms, ½-watt.
R_5 — 0.1 megohm, 1-watt.
R_6 — 0.5 megohm, ½-watt.
R_7 — 250 ohms, 1-watt.
R_8 — 200 ohms, 1-watt.
R_9 — 50,000-ohm volume control.
R_{10} — 50,000 ohms, 1-watt.
L_1 — 3 turns No. 12, 9/16-inch diameter, ½-inch long.
L_2 — 1 turn No. 12 or No. 14.
RFC_1, RFC_2 — 55 turns No. 30 d.c.c., close-wound, ¼-inch diameter.
T_1 — Transceiver transformer (see text).
$T_1{}^1$ — Audio interstage transformer.
T_2 — Output transformer, pentode to voice coil.
T_3 — Microphone transformer.
S_{1-4} — 4-pole double-throw switch.
J_1 — Open-circuit jack.
J_2 — Closed-circuit jack.
Spkr — 3-inch permanent-magnet speaker.
V_1 — 6J5, 6C5, 6V6, 6F6, etc.
V_2 — 6J5, 6C5. $V_2{}^1$ — 6N7.
V_3 — 6V6, 6F6, etc. (GT types preferred).

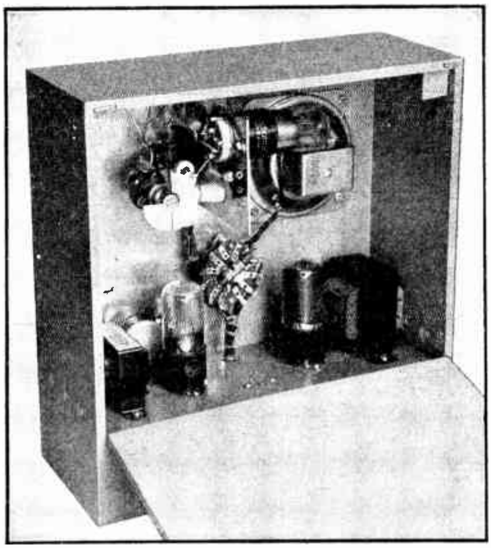


Fig. 1737 — Rear view of a transceiver installed in its case. The oscillator-detector is constructed as a unit.

setting and resetting of the volume control between transmissions if maximum transmitter performance is to be obtained. Aside from the nuisance angle, most volume controls deteriorate rapidly under constant use. In Fig. 1736-B, this problem is solved by employing a dual triode as an audio mixer. The detector output is coupled through T_1 to one grid of the 6N7 while the microphone transformer feeds the second grid, thereby providing a separate and independent audio channel for each and making a transceiver transformer unnecessary. The plates of the audio mixer tube are in parallel, so that signals on either grid will be reproduced in the common plate circuit.

In Fig. 1736-A, the audio system consists of a triode first stage (6J5 or 6C5) followed by a 6V6 (or equivalent, in any of the varieties of glass or metal) used as a modulator in transmitting and to drive the loudspeaker in reception. There is no provision for headphone reception in this unit, but if it is wanted a jack can easily be connected in the 6J5 plate circuit as in Fig. 1736-B. If this is done an additional switch section should be provided to cut out the headphone circuit when transmitting. As a matter of economy when operating from emergency power, the "B" drain can be cut to a very low value during reception periods if headphones only are used, since the change-over switch could be arranged to cut the "B" lead to the plate and screen of the pentode audio power tube. If speaker is not wanted or if a suitable unit is not available, this is a worth-while modification.

A simple transceiver — The transceiver shown in Figs. 1737 and 1738 is constructed from parts which in most cases can be readily salvaged from old equipment. The panel is a 10 × 10-inch piece of ¼-inch tempered Presdwood, while the shelf which holds the audio circuits is a 3½ × 10-inch piece of the same material. The shelf is mounted 1½ inches above the bottom of the panel, leaving room for the resistors and condensers underneath. The box in which the transceiver is housed is made of ¼-inch plywood, the inside dimensions being 10 × 10 × 3½ inches. At the back a door, hinged at the bottom, gives access to the tubes and r.f. section.

The oscillator is all in one unit, built on a 3 × 4-inch piece of scrap aluminum with ½-inch bent over at one end to form a mounting lip. The metal base projects 3½ inches behind the panel, the same depth as the shelf for the audio section. In general, the oscillator circuit has been arranged to make the leads between the tube and tuned circuit as short as possible. The mechanical layout may have to be varied for tuning condensers of different construction. A condenser having a maximum capacity of 10 to 15 μμfd. is required for C_1. The one used in this unit is a Hammarlund MC-20-S (originally having a maximum of 20 μμfd.) with one plate removed. To reduce capacity to ground, the rear bearing assembly should be taken off by sawing the rotor shaft and the side rods holding the stator plate. Removing this excess material noticeably increases the efficiency.

The tuned-circuit coil, L_1, is mounted under the condenser panel-mounting nut, the other end being soldered to the side rod holding the stator plate. Since both sides of the condenser must be insulated from ground, the condenser is mounted on a midget stand-off insulator. An insulated coupling and extension shaft connect the rotor to the tuning dial.

The plate and grid chokes are mounted from insulated lugs at the "cold" ends, the hot ends being placed as close as possible to the points in the circuit where they connect. The power leads from the r.f. section are cabled and brought down to the switch.

To protect the speaker cone from damage, the grille holes are backed by a piece of window-screen material which is held in place by the

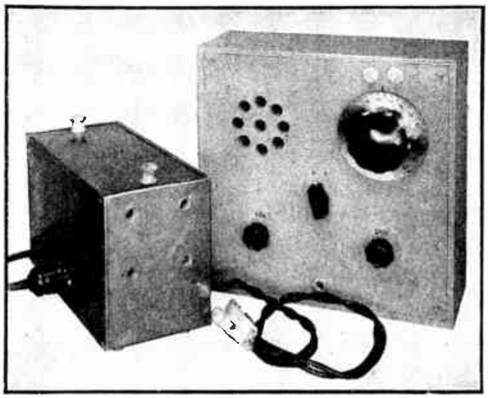


Fig. 1738 — This low-power transceiver and vibrator power supply can be built from receiver components which nearly every amateur can salvage from old equipment. The addition of an antenna, microphone, and storage battery makes it a complete emergency station.

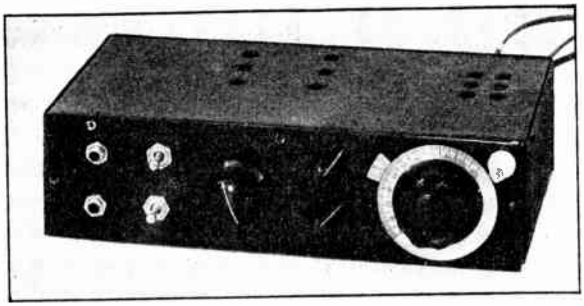


Fig. 1739 — Front view of another transceiver, which fits into the glove compartment of a car. The toggle switches are for remote control of filament and plate power supply in the rear compartment.

bolts which are used to fasten the speaker to the panel.

A metal strip running from top to bottom of the panel serves as a shield to prevent body capacity and also as a low-inductance ground connection between the oscillator and the audio section. It makes direct contact with the oscillator support, the rotor of R_5, and the metal frames of the switch and microphone jack.

In the rear view the transformer at the left is T_1, the revamped audio transformer. The audio gain control, R_5, is on the panel between R_1 and the 6J5 first audio. The modulator tube and speaker transformer are at the right, with the regeneration control, R_3, behind them on the panel. All leads from the switch are cabled and pass through a hole in the shelf near the panel. The two grid leaks, R_1 and R_2, are mounted directly on the switch contacts, but all other resistors are below the shelf. The below-shelf arrangement is of no particular consequence, since there are no r.f. circuits underneath, but the grid leads to both tubes should be kept short, so that hum pick-up will be minimized. The dropping resistor, R_4, for the regeneration control circuit is mounted on the lug strip at the rear; the other two resistors which connect together at this strip are the two sections of the modulator cathode resistor.

A compact transceiver — Another constructional arrangement for the transceiver, also based on the circuit of Fig. 1736, is shown in Figs. 1739 and 1740. The front view shows the use of a metal chassis, with dimensions of approximately 9 × 6 × 4 inches, as a cabinet. This allows ample space on the panel for the gain and regeneration controls which are shown with pointer knobs, and for the adjacent send-receive switch which has a knob. The two toggle switches at the left control the filament and plate power supplies, which are mounted in the rear of the car for mobile work. The speaker, which is a separate unit, is plugged into one of the jacks at the left and the single-button microphone into the other. Cables to the power supply from the panel of the main transmitter are shown in the rear. Several half-inch holes drilled in the top of the case provide ventilation for the tubes inside.

The tubes are mounted horizontally. This arrangement places the terminals of the sockets right at the points of connection to component parts which are mounted within the front section of the unit. Exceptions are the microphone and interstage transformers, T_1 and T_2, which are mounted at the opposite ends of the top section and as far apart as space will permit. All of the r.f. components, including the HY615 detector-oscillator, are confined to one compartment, as shown in Fig. 1740.

General — In alternative layouts the proper r.f. choke specifications may differ from those given. The grid choke is the more critical. In both cases the number of turns should be adjusted so that the cold end can be touched with the finger without disturbing the operation of the oscillator. Effective superregeneration depends considerably on the grid choke and also on the capacity of the plate by-pass condenser, C_3. The circuit may not superregenerate at all with less than 0.002 μfd. at C_3, while values higher than 0.005 tend to cut down the audio output because of the by-pass effect across the primary of the audio transformer, T_1. Two or more condensers may be connected in parallel or series if the exact capacity required is not obtainable in one unit.

The coil inductance is adjusted by spreading or squeezing the turns until the proper frequency range is secured. It is best to adjust the inductance to bring 112 Mc. near the maximum-capacity end of the tuning range.

The size of the antenna coupling coil, which is usually mounted on feed-through insulators on the panel, will depend upon the antenna system used. Ordinarily a turn or two of wire is sufficient, the coupling being adjusted by bending the leads so that the position of the antenna coil is changed with respect to the tank coil.

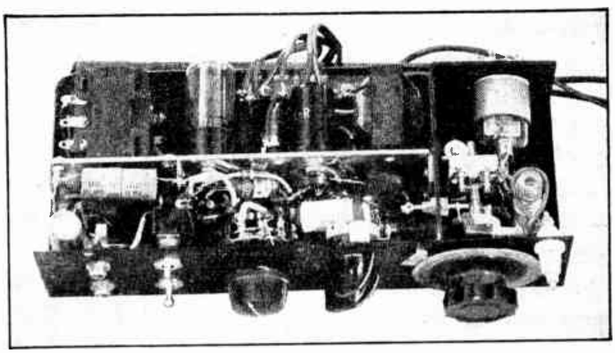


Fig. 1740 — Top view of the transceiver showing the horizontal mounting of the tubes for short leads. The oscillator-detector and associated r.f. components are in the shielded compartment to the right.

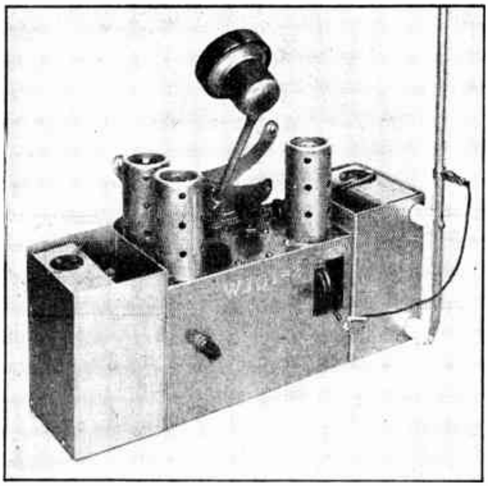


Fig. 1741 — The entire transreceiver assembly complete with battery power supply and antenna is supported from a Western Electric microphone breastplate. The 1G4G driver and 1G6G modulator tubes are at the right, with the change-over switch on the front panel below. The 1G6G oscillator tube is at the left, with the detector tuning condenser and regeneration control knob behind it. The knob at left front on the top panel is on the volume control. The unity-coupled tank coil can be seen through the oblong opening on the front panel.

Portable Equipment

Completely portable equipment is essential in many WERS applications. A unit of the walkie-talkie type is shown in Figs. 1741–1743.

The circuit, shown in Fig. 1742, includes a self-quenched superregenerative detector of the "Minute Man" type, covering a range of from 110 to 120 Mc., using an HY114-B triode, a unity-coupled modulated oscillator built around a 1G6G dual triode, and a common audio system consisting of a 1G4G transformer-coupled to another 1G6G which is operated as a Class-B modulator and second audio amplifier.

Change-over from receiving to transmitting is effected by a 3-p.d.t. anti-capacity switch with an "off" position, salvaged from a discarded transceiver. When the receiver is in use the filament circuit of the 1G6G oscillator is opened, while in transmission the detector filament is cut out of the circuit. The microphone battery circuit is opened during receiving periods. Two points on the switch are used for disconnecting the antenna from the receiver circuit when the transmitter is on the air.

Of two available means of providing necessary bias for the 1G4G first audio tube, a stack of four 1.25-volt Mallory bias cells was chosen for its light weight and compactness. Properly, five of the cells should have been used in order to comply with the tube manufacturer's recommendation of 6 volts bias for operation with a 90-volt plate supply. The manufacturer's holder requires the application of insulating strips to prevent the possibility of the cells being shorted out by the shielding used in this rig. As an alternative, four penlite cells in series could be employed. Automatic bias is not desirable in this circuit, since the difference in total "B"-battery current drain between the receiving and transmitting conditions is too great and the same audio circuit conditions must be maintained because of the tubes' dual function as receiver amplifier and transmitter modulator.

The transmitter tank circuit, shown as a single turn in the circuit diagram to simplify the drawing, actually is 2 turns of 3/16-inch copper tubing, 1 inch in diameter. It is soldered directly to the plate prongs on the oscillator tube socket, the coil plane being at right angles to the plane of the bottom of the socket. The grid coil, consisting of two turns of No. 18 wire, is threaded through the tank coil, from which it is insulated by varnished cambric tubing.

The entire equipment is designed to be supported from the breastplate and webbed shoulder-harness of an old-style Signal Corps type Western Electric microphone. The only external wiring is in the headphone cord, which may be made very short. An adjustable aluminum tubing antenna, cut down from an earlier 5-meter telescoping antenna, is supported by stand-offs.

The oblong opening over the oscillator coil permits a wide range of adjustment of the antenna tap, as well as access to the tank condenser, C_5, for frequency setting.

The trimmer condenser seen in the antenna

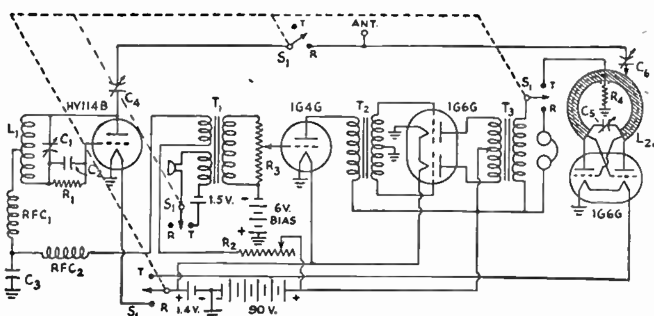


Fig. 1742 — Circuit diagram of the walkie-talkie type transmitter-receiver shown in Figs. 1741 and 1743.

C_1 — 10-$\mu\mu$fd. midget variable.
C_2 — 50-$\mu\mu$fd. mica.
C_3 — 0.005-μfd. mica.
C_4, C_5, C_6 — 3–30-$\mu\mu$fd. mica trimmer.
R_1 — 20 megohm, ½-watt.
R_2 — 50,000-ohm midget variable.
R_3 — 500,000-ohm midget variable.
R_4 — 500-ohm wire-wound, 10-watt.
L_1 — 7 turns No. 14 enameled, ½-inch diameter, 1-inch long.
L_2 — 2 turns 3/16-inch copper tubing, 1-inch diameter.
L_3 — 2 turns No. 18 enameled (wound inside L_2 and insulated from L_2 with spaghetti).
RFC_1 — Ohmite Z-0.
RFC_2 — 80-μh. r.f. choke (preferably shielded).
S_1 — 3-p.d.t. anti-capacity switch with neutral position.
T_1 — 3-winding transceiver transformer.
T_2 — Midget Class-B input transformer.
T_3 — Midget Class-B output transformer.

lead serves as a blocking condenser to isolate d.c. from the antenna and also provides some degree of adjustment of the electrical length of the antenna. The clip which connects the antenna lead to the tank circuit is set by trial. Generally it will be placed fairly close to the center of the tank coil.

The batteries are contained in the metal pockets at the ends of the unit. The two 45-volt units in series and two 1.4-volt units in parallel are connected by a cable, conducted through grommets in the walls of the partitions, to the distributing points within the case. The arrangement is designed to distribute the weight of the batteries evenly and at the same time make them accessible for replacement. Covers are provided for the tops of the battery compartments.

The main case is a salvaged shield can measuring 4 × 7½ × 5 inches, with 4 × 3½ × 5-inch pockets added at each end for the batteries.

All components except the change-over switch and the microphone are mounted on the top deck of the case and upon the vertical shield partition suspended from it. Thus, when servicing is required, it is necessary only to remove the knob and mounting nut from the change-over switch assembly and the mounting screws from the microphone breastplate, and then lift the entire assembly from the shield-case.

Fig. 1743 shows the r.f. end with its separate circuits for receiver and transmitter. Some rather careful work is required in the construction of the unity-coupled coil for the oscillator. Soft copper tubing, 3⁄16-inch diameter, is cut to a 4-inch length and the exact center marked. At this point a ⅛-inch hole is drilled through one wall of the tubing. This hole is enlarged longitudinally to a width of about ¼-inch and the edges smoothed with a fine file and emery cloth. On the same side, saw cuts are made ⅜-inch from each end, halfway through the tubing, and a longitudinal half-section removed. The edges left here are also buffed smooth.

The tubing is then coiled about a 1-inch diameter dowel or piece of pipe, with the hole and the cuts on the inside of the coil. Considerable care must be exercised to make the coil symmetrical on each side of the center hole and to avoid a sharp bend or a break at the point where the wall has been weakened by the drilling. A 2-turn coil is formed — or rather, a 1⅔-turn coil, as the halved ends are allowed to project at right angles to the coil, parallel to each other and separated by the distance between the No. 3 and No. 6 prongs (plate prongs) on the octal socket to be used for the 1G6G oscillator. This socket should be of the best quality obtainable, both with respect to quality of insulation (preferably Isolantite) and strength of prongs, which must support the weight of the coil assembly soldered to them.

Two pieces of high-grade varnished insulating tubing are threaded from the open ends of the coil until they are seen to meet at the center hole. Then two pieces of No. 18 enameled copper wire are threaded through the insulating tubing in a similar manner. The inside ends of wire and tubing are carefully fished out through the center hole, just enough of the insulation being pulled through to insulate the wire from the edges of the hole in the copper tubing. The ends of the wire are cleaned and soldered to form the center tap of the grid coil. Enough wire is left at the outer ends to allow them to be crossed over and connected to the grid prong on the 1G6G socket which is opposite to the plate prong to which that end of the plate coil is soldered.

The soldering is the next step. The least possible heat should be applied consistent with a rigid job of mounting; too much heat will impair the usefulness of the insulation. The same caution must be observed when soldering the center tap to the plate coil, which is done at a point directly behind the center hole through which the grid tap is brought out, and in making the soldered connection between the grid tap and the grid resistor. Finally, the 3–30-μμfd. trimmer condenser, C_5, is soldered across the outer ends of the plate coil with equal care.

The mountings for the detector tuned circuit and the HY114-B tube were made up separately and wiring completed as far as possible before placing them on the chassis. The arrangement is such as to provide for the shortest possible leads between the tuned circuit

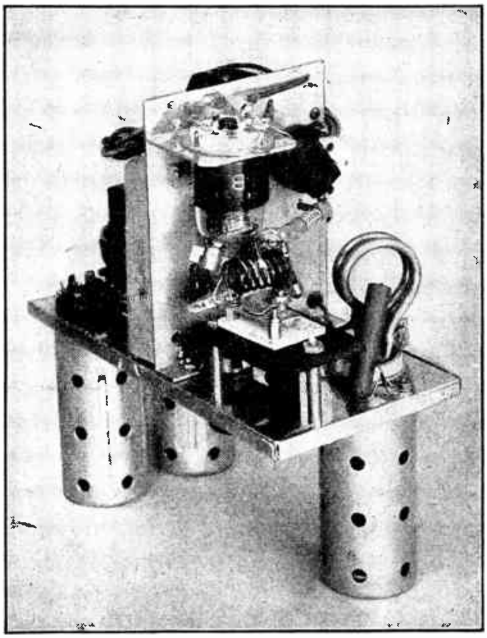
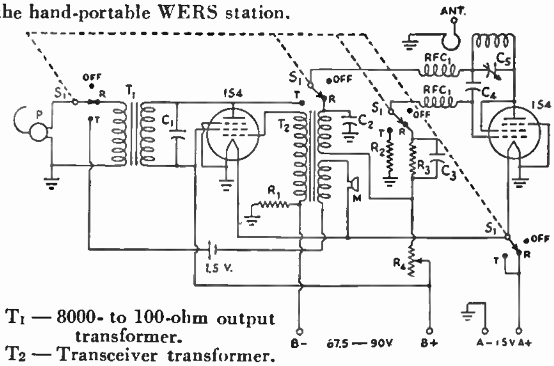


Fig. 1743 — R.f. end of the subpanel assembly, with the detector circuit at the left and the oscillator tank coil at the right. The large resistor in the foreground is the grid resistor for the unity-coupled coil. The 80-μh. interruption-frequency choke, RFC_2, is at the right of the HY114-B, behind the fixed condenser, C_3. An insulated shaft coupling was used on the receiver tuning condenser to avoid body capacity effects. The audio equipment is mounted on the other side of the shield partition.

Fig. 1744 — Circuit diagram of the hand-portable WERS station.



C_1 — 0.001-μfd. midget mica.
C_2 — 0.003-μfd. midget mica.
C_3 — 0.005-μfd. midget mica.
C_4 — 50-$\mu\mu$fd. midget mica.
C_5 — 2-plate midget variable.
R_1 — 750 ohms, ½-watt carbon.
R_2 — 25,000 ohms, ½-watt carbon.
R_3 — 10 megohms, ½-watt carbon.
R_4 — 100,000-ohm potentiometer (regeneration control).
L_1 — 1 turn No. 12, ½-inch inside diameter.
L_2 — 4 turns No. 12, ½-inch inside diameter.
RFC_1 — V.h.f. r.f. choke (Obmite Z-0 or home-made equivalent).
M — Single-button carbon microphone (see text).
P — Single headphone, 100 ohms (see text).
S_1 — Four-pole three-position rotary switch.
T_1 — 8000- to 100-ohm output transformer.
T_2 — Transceiver transformer.

and the plate and grid caps of the tube. The 80-mh. r.f. choke, RFC_2, would better be shielded.

Wherever leads pass through the baffle, rubber grommets are provided. As a final precaution, it would be well to cement an insulating layer over the inner surfaces of the shield case to avoid accidental shorts.

Handie-talkie — A portable unit of the handie-talkie type, similar to those used by the armed forces, is shown in Figs. 1745–1746. The complete transceiver and its battery power supply, enclosed in an aluminum case as shown, weighs but 3⅞ pounds. The outside dimensions of the case are only 2⅝ × 2¾ × 11 inches.

The transceiver circuit, shown in Fig. 1744, is more or less conventional in most respects. It was drawn up around the type 1S4 tube, which is the secret of the set's compactness. Two tubes of this type are used in the unit. One, with its screen tied to the plate, is used as

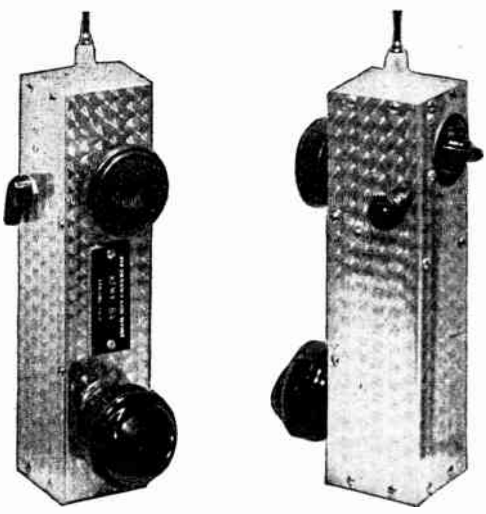


Fig. 1745 — These two views of the hand-portable transceiver show the arrangement of controls. The left-side view shows the headphone, microphone, and, on the side, the change-over switch. The right-side view shows the regeneration control on the side and the tuning knob and scale on the rear. Constructed by W6TWL.

a triode oscillator and detector. The other 1S4 is used as an audio amplifier and modulator.

Many have encountered trouble in getting the 1S4 to superregenerate on frequencies as high as 112 Mc. In this circuit this difficulty was solved by placing a 0.005-μfd. condenser, C_3, across the receiving grid lead, R_3.

The microphone is insulated from the aluminum case so that it may be connected in series with the single headphone, P, for side-tone transmitting. The headphone, which has an impedance of about 100 ohms, was provided with an output transformer, T_1, for proper impedance matching to the output of the 1S4. When transmitting, the secondary is opened and the primary is used as a Heising modulation choke.

Bias for the audio amplifier or modulator is provided by operating the "B" — below ground potential and using the voltage drop across R_1. Regeneration is controlled by a simple variable series resistance, R_4, in the "B" + lead to the detector so that an additional switch is not necessary to eliminate battery drain when the set is turned off.

The change-over switch, S_1, is a four-pole triple-throw rotary switch. In addition to the two usual positions, a third position, labeled Off in Fig. 1744, is provided where all battery circuits are open. The filament circuit is completed when the switch is placed in either the receiving or the transmitting position.

The filament battery consists of two No. 3 flashlight cells connected in parallel for longer life. They are held in the case by spring clips. The plate battery is a 67½-volt Minimax or two 45-volt hearing-aid batteries in series. These batteries are not too difficult to buy now, since stores are selling them on the open market after the nominal installation date has passed. One penlite cell provides adequate voltage for the microphone.

The case for the unit is made of pieces of sheet aluminum fastened together with self-tapping screws. The metal was given a "swirl" finish by applying a spinning cork, held in the chuck of a drill press, after the aluminum sheet had been smeared with a thin coating of a mixture of valve-grinding compound and oil.

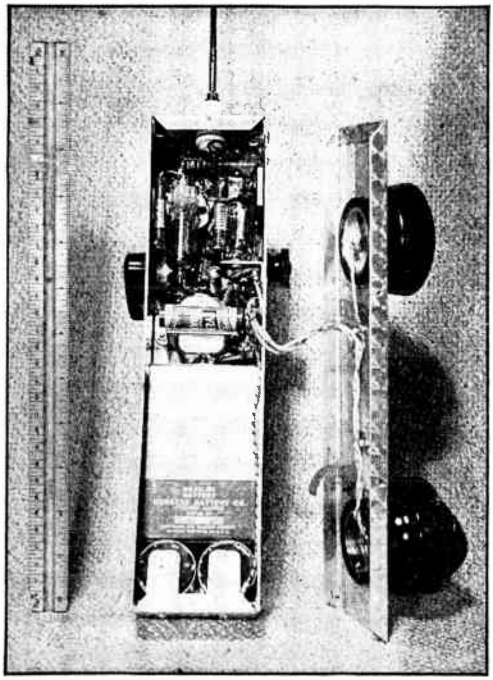


Fig. 1746 — The handie-talkie unit with the front cover removed. The miniature plate and filament batteries occupy the lower portion of the case, while the c.f. and audio components are fitted into the top section.

The microphone, mounted in the lower end of the panel, and the single headphone at the top are from a Western Electric handset in which the brass threads moulded in the handle were cut off with a hacksaw to provide mountings. The regeneration control is on the right side and the change-over switch on the left side in a position convenient for one-hand operation.

The batteries occupy the lower section of the case, the "A" battery cells being mounted in clips at the bottom and the "B" battery immediately above. The penlite microphone battery is mounted in a clip on top of the headphone transformer. The life of the batteries is very good despite their size, since the current drain is small.

The transceiver components occupy the remaining space at the top. The r.f. tube is mounted in an inverted position in a polystyrene socket, while the audio tube socket is fastened to a shelf at the left in the inside view of the unit. The tank coil, L_2, is soldered directly to the terminals of the tuning condenser, C_5, which is mounted immediately below the antenna terminal.

The antenna is a quarter-wave rod, 24 inches long, which plugs into a receptacle on top of the case. It is coupled to the tank coil by a single turn of wire, L_1, one end of which is grounded to the case. With this arrangement transmission and reception are good even with the antenna at an angle or in a horizontal position.

Construction of Power Supplies

With the possible exception of control stations, which in many instances will be installed at locations where emergency power already is provided for, WERS stations will have to be furnished with a type of power supply independent of the a.c. mains. The 6-volt automobile-type storage battery is by far the best choice for primary power, and while dynamotors, genemotors, converters, and complete vibrator-type "B" supplies cannot now be purchased new, it is not difficult to remodel an old receiver-type power transformer for use in a vibrator supply. Furthermore, the power supply in a car radio receiver can be pressed into service if necessary. Aside from the vibrator and transformer (and rectifier tube, if one is used), most of the components can be found in old b.c. receivers or existing amateur equipment.

It must be emphasized again that no dependence should be placed on the continuance of power from the ordinary 115-volt lines during an emergency. An a.c. supply can be used for routine testing, of course, and also during an actual emergency so long as the line power

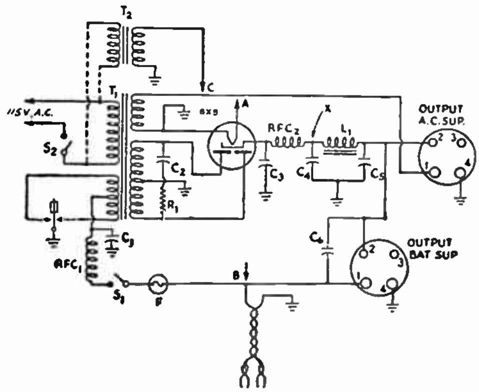


Fig. 1747 — Combination 115-volt and battery supply.
C_1 — 0.5-μfd. paper, 50-volt rating or higher.
C_2 — 0.005 to 0.01 μfd., 1600 volts (see text).
C_3 — 0.01-μfd. 600-volt paper.
C_4 — 8-μfd. 450-volt electrolytic.
C_5 — 32-μfd. 450-volt electrolytic.
C_6 — 100-μμfd. mica.
L_1 — 10–12 henry 100 ma. filter choke, not over 100 ohms (Stancor C-2303 or equivalent).
R_1 — 5000 ohms, ½- or 1-watt.
RFC_1 — 55 turns No. 12 on 1-inch form, close-wound.
RFC_2 — 2.5-mh. r.f. choke.
S_1 — S.p.s.t. toggle switch, heavy duty (10–12 amp.).
S_2 — S.p.s.t. toggle switch.
F — 15-ampere fuse.
VIB — Mallory 500P, 294, etc.
T_1 — Special vibrator transformer with 115-volt and 6-volt primaries, to give approximately 300 volts at 100 ma. d.c. (Stancor P-6166 or equivalent).
T_2 — 6.3-volt filament transformer, to be used when 6.3-volt filament winding on T_1 will not supply all heaters in both transmitter and receiver.
X — Insert a series resistor of suitable value to drop the output voltage to 300 at 100-ma. load, if necessary. If transformer gives over 300 volts d.c., a second filter choke may be used to give additional voltage drop as well as more smoothing.
NOTE — All ground connections should be made to a single point on the chassis.

lasts, but some independent source of power *must* be available. In addition to the power supplies described here, other supplies suitable for WERS work may be found in Chapter Eight.

Combination supplies — In a vibrator supply built from individual components it is necessary to filter out hash and to adjust the waveform to minimize sparking at the vibrator contacts. When such a supply is built around a manufactured transformer it is advisable to use the type which has both 115-volt and 6-volt primaries, thereby making an a.c.-d.c. supply which uses the minimum of parts for both purposes. Such transformers have been made in various ratings. A suitable circuit diagram is given in Fig. 1747.

R.f. filters for reducing hash are incorporated in both primary and secondary circuits. The secondary filter consists of a 0.01-μfd. paper condenser directly across the rectifier output, with a 2.5-mh. r.f. choke in series ahead of the smoothing filter. In the primary circuit a low-inductance choke and high-capacity condenser are needed because of the low impedance of the circuit. A choke of the specifications given should be adequate, but if there is trouble with hash it may be beneficial to experiment with other sizes. The wire should be large — No. 12, preferably, and No. 14 as a minimum. Manufactured chokes such as the Mallory RF583 are more compact and give higher inductance for a given resistance because they are bank-wound, and may be substituted if obtainable. C_1 should be at least 0.5 μfd.; even more capacity may help in bad cases of hash.

The power supply should be built on a metal chassis, with all unshielded parts underneath. A

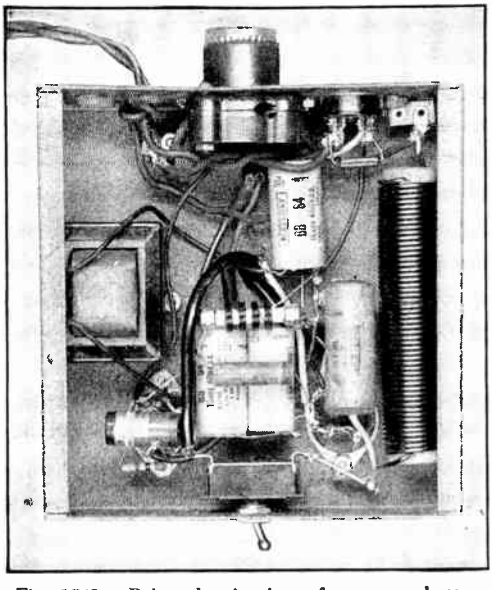


Fig. 1748 — Below-chassis view of a storage-battery power supply using a rewound transformer. The circuit is given in Fig. 1749. The various components can be easily recognized in this view. The transformer, vibrator and rectifier tube are mounted on top of the chassis.

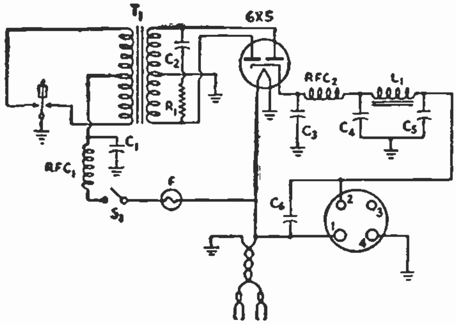


Fig. 1749 — Vibrator power-supply circuit diagram. Except for T_1 and T_2, all components are identical with those of Fig. 1747. T_1 may be either a regular 6-volt input vibrator-type power transformer or a home-altered 115-volt receiver transformer as described in the text.

bottom plate to complete the shielding is advisable. The transformer case, vibrator case and metal shell of the tube all should be grounded to the chassis. If a glass tube is used it should be enclosed in a tube shield. The battery leads should be evenly twisted, since these leads are more likely to radiate hash than any other part of a reasonably well-shielded supply. A little care in this respect usually is more productive than experimenting with different values in the hash filters. Such experimenting should come *after* it has been found that radiation from the leads has been reduced to an absolute minimum. Shielding the leads is not particularly helpful.

The 100-$\mu\mu$fd. mica condenser, C_6, connected from the positive output lead to the "hot" side of the "A" battery, may be helpful in reducing hash in certain power supplies. A trial is necessary to see whether or not it is required. It should be mounted right on the output socket.

Testing for methods of eliminating hash should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiver antenna leads by radiation from the supply itself and the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the supply and leads.

The smoothing filter for battery operation can be a single-section affair, but there will be some hum (readily distinguishable from hash because of its deeper pitch) unless the filter output capacity is fairly large — 16 to 32 μfd.

Rewinding transformers — Those who cannot get either complete vibrator assemblies or special transformers, or who want to assemble a vibrator supply at the least possible expense, can find many of the necessary parts in old broadcast receivers. A power transformer with a 100-milliampere secondary is needed; the voltage rating should be 350 or so with any transformer of this type, but the exact value does not matter too much. The high-voltage secondary must be in good shape. Pick out a transformer with a case — one of the "fully shielded" type — but not one immersed in pitch. The receiver usually will have a filter choke or two as well as filter condensers which may be usable.

Before dismantling the transformer, measure the output voltages of the windings if these are not already known. This will require a multi-range a.c. voltmeter. If the builder does not have such an instrument, the measurements can be made by a radio repairman or at the local parts store.

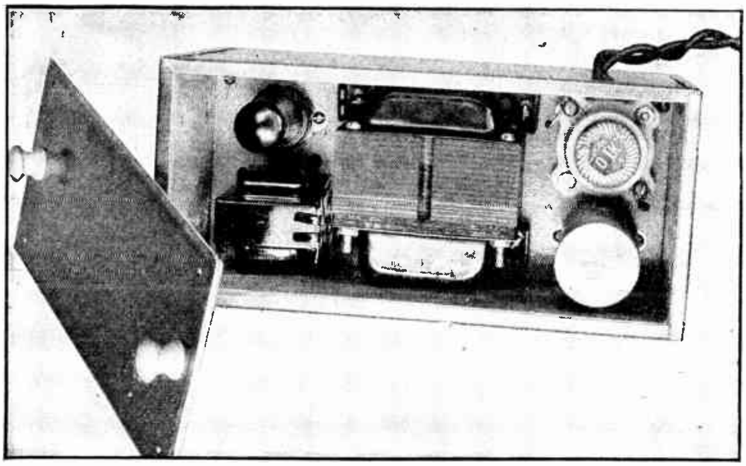


Fig. 1750 — A view inside the vibrator-type power-supply unit shown assembled in Fig. 1736. The rectifier tube is at the upper left with the filter choke just below. The primary fuse socket and vibrator are at the right. A synchronous-type vibrator may be substituted for the interrupter-type if it is desired to eliminate the rectifier tube.

Next take the transformer apart, being careful to avoid damaging the windings or bending the core pieces. The filament secondaries are nearly always on the outside of the coil assembly, so remove the outer layers of paper to expose the uppermost filament winding. Count the number of turns and divide this figure by the output voltage of the winding to find the number of turns per volt. Most small transformers have about three turns per volt. Make a note of the exact figure and then remove the remaining filament secondaries, leaving only the primary and high-voltage secondary.

When this has been done, slide one of the core pieces inside the coil and see how much space has been made available by removing the low-voltage secondaries. The battery primary to be put on will not have many turns, but the wire should be large to keep the losses low, so generally two layers will be required. The current to be carried will be in the vicinity of 8 amperes at full load, but since the primary is to be center-tapped each half of the winding carries current only half the time. Thus the heating effect is equivalent to 4 amperes. No. 12 or No. 14 wire is suitable. It would not be advisable, however, to use smaller wire than No. 16, and that size only when a larger size will not fit the space. If the space still is too small, remove the 115-volt primary.

If the normal transformer output was about 300 volts at 100 milliamperes through an ordinary filter (this should be ascertained before taking the transformer apart, by hooking up a power supply and making a d.c. measurement) it is useful to save the old primary if possible, since such a transformer can be used for a combination a.c.-battery supply. However, it does not pay to save the old 115-volt primary at the expense of using too-small wire on the 6-volt primary; the efficiency and regulation will be better with larger wire sizes.

Whether the old primary is inside or outside the high-voltage secondary is a matter of chance. If the old primary is on the inside and it is necessary to remove it, the job can be done by pulling the outermost layer through the side of the assembly, after which the rest can easily be unwound. One half of the new primary should be wound directly on the insulating sleeve into which the core fits, then the high-voltage secondary slipped over it, and finally the second half of the new primary wound on top. Both halves should be wound in the same direction so that the end of the first half can be connected to the beginning of the second to give a center tap with the proper polarities. If separate leads are brought out from each half (this is usually the most convenient method) it is easy to check the polarities after the transformer is reassembled. Connect two leads together for trial, then apply 115 volts across the high-voltage winding. If the voltage across the outer ends of the new winding is twice the voltage across each half, the polarity is correct. A filament voltmeter should be used for this check, since the voltage is low.

To obtain 300 volts at the rated current of 100 ma. from the supply, using a 6X5 rectifier and a filter having a choke with a resistance of about 100 ohms, the secondary/primary turns ratio should be 70:1, assuming an even 6 volts from the storage battery. Multiply the original a.c. output voltage of the high-voltage secondary by the number of turns per volt to find the total number of turns, then divide the product by 70 to find the proper number of turns for the primary. For example, if the output voltage was known or measured to be 750 volts a.c. (375 each side of center-tap) and the transformer had three turns per volt, the total number of secondary turns is 750 × 3, or 2250. Dividing 2250 by 70 gives 32 (dropping the fraction) as the total primary turns, or 16 each side of the center-tap.

The new windings should be sufficiently well insulated so that there is no possibility of a short-circuit to the core or secondary, but otherwise no special precautions are necessary since the voltage is low. Reassemble the transformer, interleaving the laminations. It is advisable to use no more than two laminations on a side before interleaving from the other side, but it is not necessary to interleave them singly. With careful packing it should be possible to get back all of the core pieces.

Once the transformer is rebuilt, the remainder of the supply is constructed and adjusted as previously described. If the job has been done properly the efficiency should be about normal for vibrator supplies. Individual transformers have been found to vary somewhat, in that for an output of 100 ma. at 300 volts the battery current ranges from 7.5 to 9 amperes with the different units. This does not include the current taken by the rectifier heater. Because of this current and the power loss in the plate-cathode circuit of the rectifier tube, the over-all efficiency of the tube rectifier type of supply is not quite as high as with the synchronous vibrator. With no load on the supply the battery current should be about 1.5 amperes.

Low-voltage supply — A vibrator supply for operation at lower voltage (in the vicinity of 200 volts d.c.) is shown in Figs. 1750 to 1752. This is especially suitable for use with the transceiver previously described when a 6J5 oscillator tube is used, or for commercial dry-battery transceivers modified to permit connecting an external power supply.

The transformer is a universal replacement-type unit having a d.c. output, when operated from 115 volts a.c., of about 70 ma. at 250 to 300 volts, and provided with 6.3-, 5- and 2.5-volt filament windings. The circuit is much the same as in the case of the homemade units just described. As shown in the circuit diagram, Fig. 1752, the filament windings on the transformer are used in the battery circuit; the 6.3-volt winding provides one side of the battery primary, and the other side consists of the 5-volt winding in series with half the 2.5-volt winding. This method gives lower output voltage than can be obtained with a properly proportioned primary, but avoids the inconvenience of rewinding the transformer. The output voltage is about 200 with a load of 60 ma.

Before the battery primary is permanently connected, the proper polarities of the filament windings must be determined. Apply line voltage to the regular 115-volt primary and connect the 6.3- and 5-volt windings in series. Measure the total voltage across the two. If it is something over 11 volts the polarity is correct, but if the voltage is very low the connections to one of the windings should be reversed. Then add half the 2.5-volt winding to the 5-volt winding and measure the voltage across these two in series. It will be between 6 and 7 volts when the polarities are correct. The connection between the 6.3- and 5-volt windings becomes the center-tap of the battery primary, as shown in the circuit diagram, Fig. 1752.

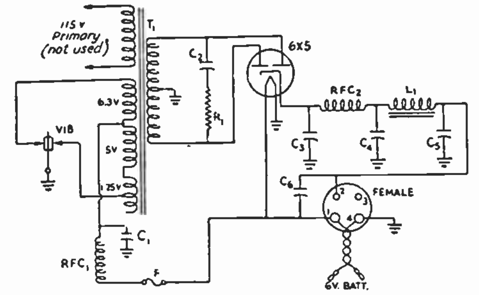


Fig. 1752 — **Circuit diagram of the low-voltage vibrator-type power-supply unit.**
C_1 — 0.5-μfd. paper, 200 volts.
C_2 — 0.008-μfd. paper, 1600 volts.
C_3 — 0.01-μfd. paper, 600 volts.
C_4 — 8-μfd. electrolytic, 300 volts.
C_5 — 16- to 32-μfd. electrolytic, 300 volts.
C_6 — 100-μμfd. mica.
R_1 — 5000 ohms, 1 watt.
L_1 — 10-henry 60-ma filter choke.
RFC_1 — 52 turns of No. 12, close-wound on 1-inch diameter form.
RFC_2 — 2.5-mh. r.f. choke.
T_1 — Power transformer, 300 volts each side of c.t., 60 to 70 ma.; with 6.3-, 5- and 2.5-volt windings. 115-volt primary is unused.
F — 10 ampere fuse.
VIB — Vibrator (Mallory 294, etc.).

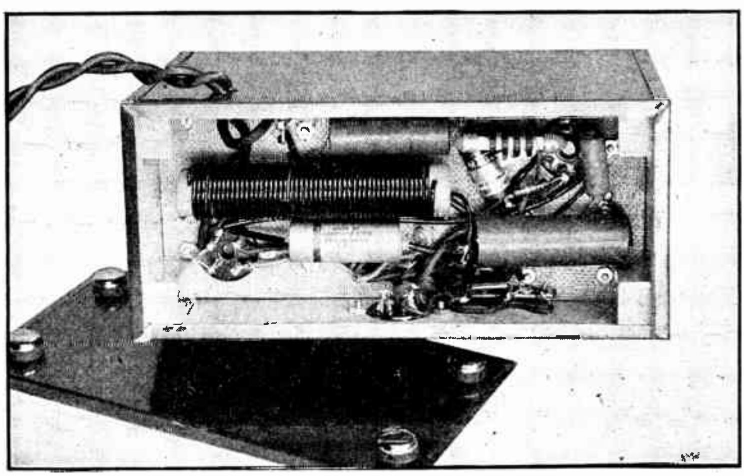


Fig. 1751 — **Hash and smoothing filter components are mounted in the bottom of the low-voltage vibrator power supply. The 4-prong outlet socket is mounted on the side.**

All the components in the supply with the exception of the four-prong outlet socket are mounted on a piece of quarter-inch tempered Masonite measuring 3¾ × 9 inches. This fits into a plywood box having inside dimensions (3¾ × 9 × 5½ inches) just large enough to contain the equipment. The Masonite shelf rests on ¾-inch square blocks, 1¼ inches long, glued to the corners of the box at the bottom. The top and bottom of the box are removable. To provide shielding and thus reduce hash troubles, the box is covered with thin iron salvaged from 5-quart oil cans. Where the edges bend around the box to make a joint, the lacquer is rubbed off with steel wool so the pieces make electrical contact, and the metal is tacked to the plywood with escutcheon pins.

To make sure that the shielding will be complete, the top and bottom of the box slide into place from the side, with the metal covering extending out so that it fits tightly under a lip bent over from the metal on the sides. These lips also are cleaned of lacquer to permit good electrical contact. The general construction should be quite apparent from the photographs. The bottom is provided with rubber feet, and the top has a small knob at each end so that it can be pushed out. This is essential, since the fit is good and there is no way to get either the top or bottom off, once on, without having some sort of handle to grip.

Antenna Systems

In many cases, particularly at control stations, it will be necessary to use non-directive antennas because of the necessity for working field stations at random points of the compass. At field stations which normally work with only a single control station, however, it may be advantageous to use a simple form of directive array. The power gain will be worth while in bettering the signals in both directions, and in addition will minimize interference to and from other networks. The simpler forms of antennas described in Chapters Ten and Eighteen are quite suitable for WERS work.

More important, perhaps, than the antenna itself is its location. Every effort should be made to get the antenna well above its surroundings and to provide, whenever possible, a clear path between the control station and the network stations with which it must communicate. Having a line of sight between antennas will ensure successful communication even though the power is very low and the antenna itself is nothing more than a simple half-wave wire. Where there are intervening obstructions, it will be helpful to use as much height as possible.

Vertical polarization is to be preferred to horizontal, since vertical polarization is better suited to mobile operation. A simple vertical antenna has practically no horizontal directivity, therefore it will work equally well in all directions except for effects attributable to its surroundings and to the terrain over which the signal must travel. The signal strength will be poor if a horizontally polarized antenna is used to receive a vertically polarized signal.

A half-wave antenna, two half waves fed in phase stacked vertically, or an extended double Zepp, all will be satisfactory in WERS, and are very simple types to construct. Design details will be found in Chapter Ten. If the station is to be operated on a fixed frequency, the antenna length should be adjusted for that frequency. If the same antenna is to work on several frequencies, the length had best be chosen midway between the two extremes.

Transmission lines — At nearly all fixed locations it will be necessary to use a transmission line between the antenna and the radio equipment, since the latter will be indoors where it is easily accessible while the former will be placed on the roof of the building to secure adequate height. Low-loss concentric line is ideal for working into the center of a half-wave antenna, but there is little likelihood it can be obtained except in isolated instances. The alternative is an open-wire line having an impedance of 500 to 600 ohms. It is advisable to keep the spacing between wires small, to prevent radiation loss; 2-inch spacing is about right, provided the line can be installed fairly rigidly so that it will not swing in a breeze and cause the transmitter frequency to change. This close separation also requires a fairly large number of spacers — at intervals of perhaps three to four feet. Lacking more suitable materials, the spacers may be made of two-inch lengths of quarter- or half-inch wooden dowel or cut pieces of square section (preferably of maple), boiled in paraffin to make them waterproof. In paraffining the wood, take care that the temperature does not get high enough to scorch it. Such spacers will provide adequate insulation at the power levels permitted for WERS transmitters. Spacers may also be cut from scrap bakelite panels.

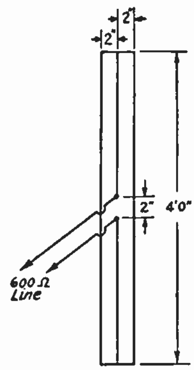


Fig. 1753 — Three-wire folded-doublet antenna for matching a 600-ohm line. The three conductors are connected together at the ends, as indicated. They may be made of wire, rod or tubing, and can be mounted on stand-off insulators on a wooden support.

To make such a line nonresonant it will be necessary to install a matching stub at the antenna. The design and adjustment of such stubs also is covered in Chapter Ten. As an alternative, a multi-wire doublet antenna may be used to couple directly to a line having an impedance of the order of 500 to 600 ohms without special matching provisions. Such an antenna is shown schematically in Fig. 1753. It gives a 9-to-1 impedance step up at the line terminals, hence practically automatic matching to a 600-ohm line, assuming the normal doublet impedance of 70 ohms. In addition, it has a broad resonance characteristic and therefore is well suited to working anywhere in the band.

To avoid the necessity for impedance matching, two-wire lines may be operated as tuned lines if desired. Such operation has been successful with lines up to at least 100 feet long. Since in most cases the coupling device at the transmitter or receiver is a single-turn coil, the simplest method of tuning the line is to adjust the feeder length until the current in the line is maximum when the transmitter is operating on the chosen frequency. A small dial light or flashlight bulb, connected in series with one side of the line right at the transmitter terminals, may be used as a current indicator. The transmission line should be made about four feet longer than necessary, its length being adjusted by cutting off an inch or two at a time until maximum bulb brilliancy is obtained.

From a constructional standpoint it is desirable to use the same antenna for both transmitting and receiving. The change-over switch for this purpose should have low capacity, and preferably should have low-loss insulation. The ordinary type of wafer switch is satisfactory, particularly if it is ceramic insulated. A small porcelain-base d.p.d.t. knife switch also may be used for this purpose. If possible, the antenna switch should be combined mechanically with the power-supply change-over switches for the transmitter and receiver so that all the necessary switching from transmission to reception can be done in one simple operation.

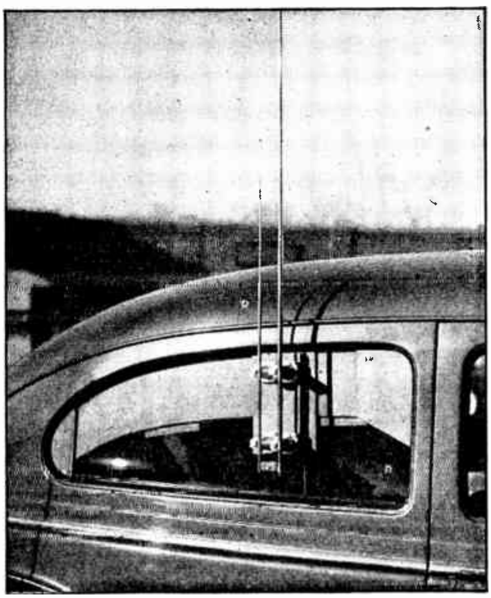


Fig. 1754—A J-type antenna for 112-Mc. mobile operation can be mounted easily in the window of a car, allowing the radiator proper to be placed above the roof of the vehicle. The dimensions are given in Fig. 1755.

Mobile antennas—It is probable that most WERS networks will have one or more stations installed in cars, for dispatching to points which may be in urgent need of communication. The equipment previously described is readily adaptable to car installations; the transceiver, in particular, can be set up with little difficulty, and can get its power from the car broadcast receiver, if there is one. This would require only the installation of a suitable power socket in the car receiver, together with a switch to cut the power from the receiver when the transceiver is in use. Antennas suitable for such mobile WERS installations are described in Chapter Eighteen.

For a solid but easily detachable mounting for a mobile antenna, the arrangement shown in Fig. 1754 is suggested. It is held in place by a panel of wood, cut to the shape of the window, on which the antenna is mounted. By running up the window the panel is held firmly in place. The antenna is of the "J" type, shown in Fig. 1755. This type of installation places the radiator proper above the roof of the car, and has the advantage that it can be readily removed from the car when not in use or when needed elsewhere.

The unit shown is built of ¼-inch plywood, since the usual thickness of the window glass in cars is ¼ inch. Run down the window of the car about half way, or enough to leave at least a 6-inch opening, and make a pattern of cardboard using the top edge of the window glass for the guide. Trim the cardboard to this shape, and then push it up in the window and use the edge of the glass to mark the bottom edge of the pattern. From the pattern, mark the piece of plywood and cut it out with a saw. Additional small pieces to form stops in the corners are fastened to the main piece with glue and brads. A piece of plywood about 6 × 8½ inches should be fastened to the large piece at the point where the antenna is to be supported, using glue and brads, and the four stand-off insulators which support the antenna bolted to this piece. If the insulators are not long enough for the antenna to clear the side of the car, they can be raised by wood strips.

Two small strips should be nailed along the inside of the main piece so that they extend down below the edge a few inches and form, with the outside pieces, a yoke to keep the assembly in the proper position on the window.

The feeder can be made of flexible rubber-covered wire (obtained by splitting a length of parallel lamp cord) separated by small plastic or dry wood spacers. The antenna ends of the wires are soldered to the heads of the large bolts in the upper stand-off insulators, and the wire is run out through holes in the wood.

The antenna and matching-section rods are regular automobile whip antennas and are supported on the stand-off insulators by small loop-shaped metal clamps. The shorting bar is made along the same lines, with bars of heavy metal on both sides of the clamp loops.

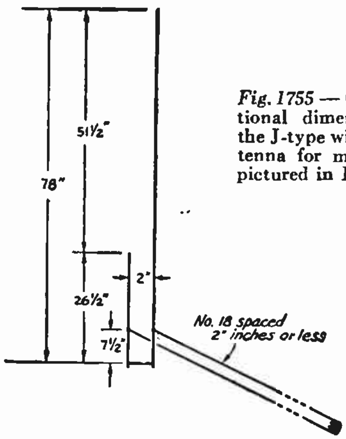


Fig. 1755 — Constructional dimensions of the J-type window antenna for mobile use pictured in Fig. 1754.

Frequency Measurement

Under the WERS regulations provision must be made for measurement of frequency of the transmitters in the network, and for checking the carrier stability to make sure that the frequency deviation does not exceed that permitted in the section of the band in which the transmitter operates.

Probably the simplest means of measuring frequency is the Lecher wire system, which is a pair of parallel bare wires to which the transmitter or receiver can be coupled. The parallel wires form a transmission line along which standing waves appear, and the distance between consecutive current loops along the line gives the wavelength directly.

Even under good conditions, frequency determination by the Lecher wire method (see Chapter Twenty) is subject to inaccuracy of the order of 0.1 per cent by the limitations of the means available for measuring length, as well as other small but avoidable errors. More accurate measurements require more elaborate equipment, although not necessarily equipment which is not already available or which cannot be constructed readily.

At lower frequencies it is customary to employ an oscillator whose fundamental frequency is such that harmonics appear at intervals of some multiple of 100 kc., the harmonics being used to provide calibration points for a receiver or heterodyne frequency meter. Methods of construction and calibration are fully described in Chapter Nineteen. If a regular communications receiver is so calibrated it can readily be used for checking the frequency of 112-Mc. transmitters. A simple method, suggested by W1EAO, is shown in block-diagram form in Fig. 1756. An auxiliary oscillator capable of tuning over the range 14–14.5 Mc. (not necessarily bandspread tuning) is required. Any simple oscillator circuit may be used, and it may be operated at any convenient plate voltage from 100 volts upward.

The method of measurement is as follows: On the regular 112-Mc. receiver, tune in the signal to be measured. Set the auxiliary oscillator frequency so that its 8th harmonic is heard beating with the 112-Mc. signal. Adjust to zero beat. Then tune the communications receiver to the fundamental frequency of the auxiliary oscillator. Adjust the receiver to zero beat and read the frequency as accurately as possible from the calibration curve. Multiplying this figure by 8 will give the 112-Mc. transmitter frequency.

Two initial precautions must be observed in using this method. First, it must be determined that the auxiliary oscillator is tuning over the 14–14.5-Mc. range. The chief cause of error here is the possibility of a spurious response (such as an image) in the communications receiver, which would result in a misleading frequency indication. For this reason the signal must not be too strong. Only enough antenna should be used to obtain a signal of moderate strength. Second, ascertain that the 8th harmonic is the one actually being used by giving the 112-Mc. receiver or transmitter an initial check with Lecher wires. The 14-Mc. frequency is used so that there will be no possibility of getting the wrong harmonic after the 112-Mc. band is known even roughly. The auxiliary oscillator dial should be marked with the 14- and 14.5-Mc. limits, so there will be no chance of tuning far off frequency.

In zero-beating the oscillator harmonic to a 112-Mc. signal in a superregenerative receiver a series of beat notes will be heard as the auxiliary oscillator frequency is varied. However, one of these will be much stronger than the others; it represents the true beat.

The accuracy of measurement depends upon the precision with which the auxiliary oscillator and communications receiver are set to zero beat, and on the accuracy of the receiver calibration. An accuracy well within 0.01 per cent is possible with reasonable care.

Frequency-checking procedure — For the regular frequency-checking procedure, the best plan would seem to be to calibrate a bandspread receiver at the control station, using whatever frequency-measurement means is available, and then by means of the receiver to measure the frequency of each transmitter on the air as it checks into the network in test periods. With relatively little initial cut-and-try each station can be set on its proper frequency, after which only minor adjustments

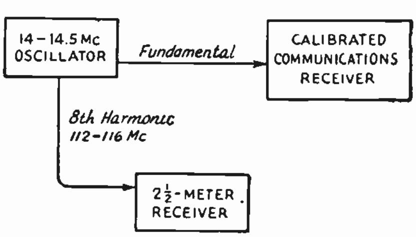


Fig. 1756 — Block diagram showing use of a calibrated communications receiver and auxiliary 14-Mc. oscillator for checking frequency of 1.12-Mc. WERS stations.

should be necessary even over quite long periods of time. A frequency deviation in any transmitter reporting to the control station will instantly be observed because it will appear at a different setting of the receiver dial. The receiver calibration should be checked at regular intervals. This check should be made with the antenna connected to the receiver.

References

The *QST* articles listed in the following bibliography deal with the technical aspects of WERS, and provide information supplementary to that contained in this chapter. Those marked with an asterisk (*) describe in more detail apparatus treated in this chapter.

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2) "Receivers for 112-Mc. Emergency Work," Goodman, Jan., 1942.
3) "An Experimental 112-Mc. Receiver," Brannin, Dec., 1941.
4) "Off the Ultrahighs" (W6ANN's 112-Mc. superhet), Tilton, July, 1942.
5) "More Selectivity in WERS Reception," Grammer, Sept., 1943.
6) "A.C.-D.C. Gear for 112 Mc.," H & K, Aug., 1943.
7) *"A WERS Control Station Receiver," Heubner, July, 1944.
8) "Compact Gear for 224-Mc. WERS," Semel, Nov., 1944.

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9) "Defense Network Control Station," Stiles, Feb., 1942.
10) *"A 25-Watt 2½-Meter M.O.P.A.," Bailey, Dec., 1942.
11) "A Crystal-Controlled Transmitter for WERS," Brooks, April, 1942.
12) "A Stabilized 2½-Meter Oscillator," Goodman, Nov., 1940.
13) *"A 112-Mc. Emergency Transmitter," Grammer, Dec., 1941
14) "WERS Gear, 1942 Style," Hieronymous, Nov., 1942.
15) "Mica Trimmer Tank Condensers in WERS Gear," H & K, June, 1943.
16) "A.C.-D.C. Gear for 112 Mc.," H & K, Aug., 1943.
17) "A V.H.F. Transmitter for Emergency Service," Hay and Harpster, Sept., 1943.
18) "A Battery-Powered Camper's Combination," French, May. 1944.
19) *"A Simple M.O.P.A. for WERS Service," Pattison, July, 1944.
20) "A Receiving-Tube 112-Mc. M.O.P.A.," Espy, Sept., 1944.
21) "Compact Gear for 224-Mc. WERS," Semel, Nov., 1944.
22) "A Versatile WERS Mobile Station," Rand, Nov., 1944.

Transmitter-Receivers:

23) "Low-Power 112-Mc. Transmitter-Receiver," May, 1941.
24) "A 112-Mc. Transmitter-Receiver Combination," Brannin, May, 1942.
25) "Building WERS Gear from Salvaged B.C. Sets," Mix, Sept., 1942.
26) "A.C.-D.C. Transmitter-Receiver for 2½," H & K, Sept., 1942.
27) "A Simple Transmitter-Receiver for War Emergency Work," Rand, Nov., 1942.
28) *"A 112-Mc. Transmitter-Receiver," Lynch, Jan., 1943.
29) "An Economical Transmitter-Receiver for WERS," Magee, June, 1943.
30) "Rebuilding TR-4s for Non-Priority Tubes," Mix, July, 1943.
31) *"Constructional Aspects of WERS Mobile Installations," Forster, Aug., 1943.
32) "Notes Covering the WERS Transmitter-Receivers for Allegany County, Maryland," H & K, Aug., 1943.
33) "CD-WERS, 1944 Style," Long, Nov., 1943.
34) "Plug-In Headphone Adapter for TR-4s," H & K, Nov., 1943.
35) "Simplified Transmitter-Receiver Switching Arrangement," H & K, Nov., 1943.

Transceivers:

36) "A Battery Transceiver for 112 Mc.," Chambers, April, 1940.
37) "A Simple Transceiver for Two and One-Half," H & K, April, 1942.
38) *"A Transceiver for WERS," Grammer, Oct., 1942.
39) "Revamping 5-Meter Transceivers for 2½," H & K, Oct., 1942.
40) "Boosting Transceiver Performance," H & K, Dec., 1942.
41) "A Transceiver for Mobile WERS Work," Bradley, Dec., 1943.
42) "A Simple WERS Transceiver with Transformerless Power Supply," Roth, Jan., 1944.
43) "WKXM-8 — A Novel WERS Transceiver," Mitchell, April, 1944.
44) "Building WERS Transceivers in the School Shop," Metzger, May, 1944.
45) "A Single-Tube WERS Transceiver," Abell, Oct., 1944.
46) *"A Versatile WERS Mobile Station," Rand, Nov., 1944.

Walkie-Talkies and Handie-Talkies:

47) "A Pack Set for 112-Mc. Defense Work," Chambers, April, 1942.
48) "A Talkie-Walkie for Civilian Defense," Kopetzky, June, 1942.
49) "Handy Andy," Palmer, Oct., 1943.
50) "On the Spot with a Walkie-Talkie," Burkle, Nov., 1943.
51) "A WERS Handie-Talkie for $1538.77," Long, Feb., 1944.
52) *"A Walking WERS Station," French, March, 1944.
53) *"A Self-Contained Handie-Talkie," Haist, June, 1944.

Power Supplies:

See 17, 31, 38 and 42 above, and the following:

54) *"Power Supply for Emergency Equipment," Grammer, Jan., 1942.
55) "A Gas-Driven Generator for Emergency Power Supply," Landes, Feb., 1943.
56) "A Portable Power Supply for WERS," Long, May, 1944.

Antennas:

See 8, 33 and 45 above, and the following:

57) *"112-Mc. Emergency Gear" (three-wire folded doublet), Grammer, Dec., 1941.
58) *"Antennas for 112-Mc. Mobile Work," Goodman, Feb., 1942.
59) "A Simple Collapsible Rotary Antenna for 2½-Meter Mobile Work," H & K, March, 1942.
60) "Feeding the Coaxial Dipole with an Open-Wire Line," H & K, May, 1942.
61) "A Four-Element Continuously-Rotatable Antenna for 112 Mc.," H & K, Sept., 1942.
62) "Folding Car-Roof V.H.F. Antenna," H & K, Aug., 1943.
63) "Three-Element Directional Antenna for Portable 112-Mc. Work," H & K, Aug., 1943.
64) "Simple Method for Investigating Performance of 112-Mc. Antennas," H & K, Nov., 1943.

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65) "A Lecher-Wire System for U.H. Frequency Measurement," Grammer, Oct., 1941.
66) "A Simple Method of Frequency Measurement for WERS," Woodward, Sept., 1942.
67) "CD-WERS, 1944 Style" (WERS absorption-type frequency meter), Long, Nov., 1943.
68) "A Junk-Box Frequency Meter for 112 Mc.," Adams, Nov., 1943.
69) "Frequency Measurement in WERS," Bliss, Dec., 1943.
70) "Sensitive Battery-Operated Test Rig for WERS," H & K, Sept., 1944.

Chapter Eighteen

Antenna Construction

The use of good materials in the antenna system is important since the antenna is exposed to wind and weather. To keep electrical losses low, the wires in the antenna and feeder system must have good conductivity and the insulators must have low dielectric loss and surface leakage, particularly when wet.

For short antennas, No. 14 gauge hard-drawn enameled copper wire is a satisfactory conductor. For long antennas and directive arrays, No. 14 or No. 12 enameled copper-clad steel wire should be used. It is best to make feeders of ordinary soft-drawn No. 14 or No. 12 enameled copper wire, since hard-drawn or copper-clad steel wire is difficult to handle unless it is under considerable tension at all times. The wires should be all in one piece; where a joint cannot be avoided, it should be carefully soldered.

In building a resonant two-wire feeder, the spacer insulation should be of as good quality as in the antenna insulators proper. For this reason, good ceramic spacers are advisable. Wooden dowels boiled in paraffin may be used with untuned lines, but their use is not recommended for tuned lines. The wooden dowels can be attached to the feeder wires by drilling small holes and binding them to the feeders with wire.

The ends of tuned feeders or the ends of the antenna are points of maximum voltage. It is at these points that the insulation is most important, and Pyrex glass, Isolantite or steatite insulators with long leakage paths are recommended. Glazed porcelain also is satisfactory. Insulators should be cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung from one chimney to another or from a chimney to a tree. Small trees usually are not satisfactory as points of suspension for the antenna because of their movement in windy weather. If the antenna is strung from a point near the center of the trunk of a large tree, this difficulty is not so serious. Where the antenna wire must be strung from one of the smaller branches, it is best to tie a pulley firmly to the branch and run a rope through the pulley to the antenna, with the other end of the rope attached to a counterweight near the ground. The counterweight will keep the tension on the antenna wire reasonably constant even when the branches sway or the rope tightens and stretches with varying climatic conditions.

"A"-Frame Mast

The simple and inexpensive mast shown in Fig. 1801 is satisfactory for heights up to 35 or 40 feet. The materials required are the 2 × 2-inch lumber, five ¼-inch carriage bolts 5½ inches long (with washers), a few spikes, about 300 feet of No. 12 galvanized iron wire, and several small strain insulators. The latter are used every 10 to 12 feet to break the guy wires into sections. Clear, sound lumber should be selected. The completed mast may be protected by two or three coats of house paint.

If the mast is to be erected on the ground, a couple of stakes should be driven to keep the bottom from slipping and it may then be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation — lifting the mast, carrying it to its permanent berth and fastening the guys — with the mast vertical all the while. It is entirely practicable, therefore, to erect this type of mast on any small, flat area of roof.

By using 2 × 3s or 2 × 4s, the height may be extended up to about 50 feet. The 2 × 2 is too flexible to be satisfactory at such heights.

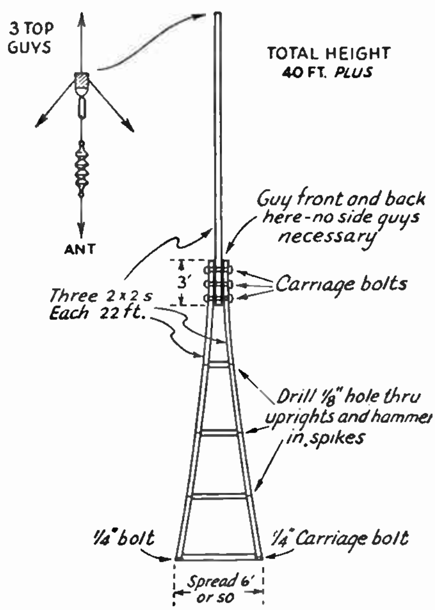


Fig. 1801 — Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

Simple 40-Foot Mast

The mast shown in Fig. 1802 is relatively strong, easy to construct, readily dismantled, and costs very little. Like the "A" frame, it is suitable for heights of the order of 40 feet.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about two feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the bottom the two legs are bolted to a length of 2 × 4 which is set in the ground. A short length of 2 × 3 is placed between the two legs about half way up the bottom section, to maintain the spacing.

The two back guys at the top pull against the antenna, while the three lower guys prevent buckling at the center of the pole. The two sets of back guys may be anchored at the same point. The guy anchors should be 15 feet or more from the bottom of the pole.

The 2 × 4 section should be set in the ground so that it faces the proper direction, and then made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. With the lower section laid on the ground, bolt *A* should be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section may be bolted in place and the mast pushed up, using a ladder or another 20-foot 2 × 3 for the job. As the mast goes up, the slack in the guys can be taken up so that the whole structure is in some measure continually supported. When the mast is vertical, bolt *B* should be slipped in place and both *A* and *B* tightened. The lower guys can then be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they should be adjusted to pull the top section into line.

The 2 × 4 should extend at least 3 feet into the ground. Rocks in the hole will provide bracing. The mast will stand alone without guying when the two bottom bolts are in place.

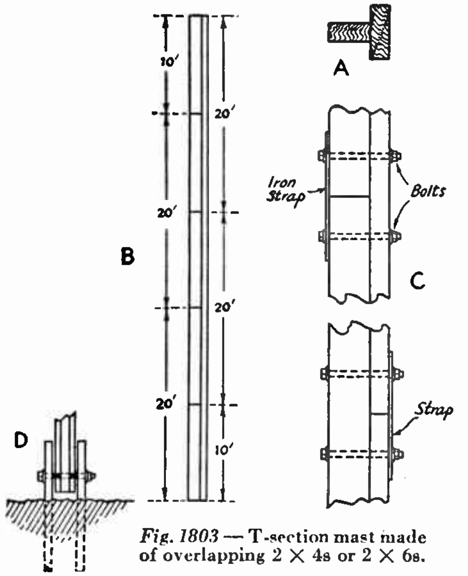


Fig. 1803 — T-section mast made of overlapping 2 × 4s or 2 × 6s.

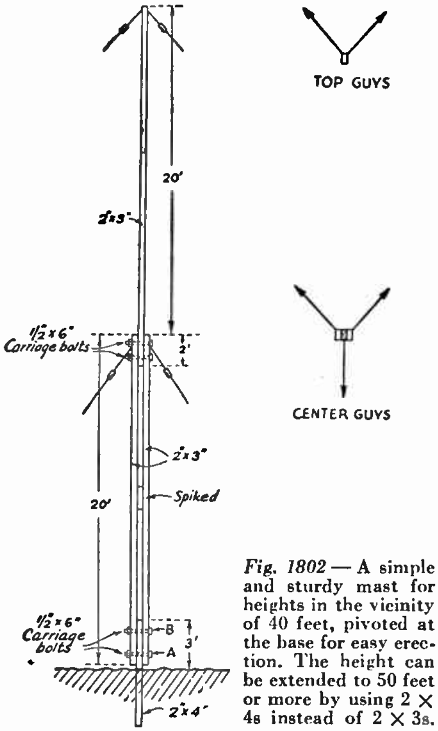


Fig. 1802 — A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.

T-Section Mast

A type of mast suitable for heights up to about 80 feet is shown in Fig. 1803. The mast is built up by butting 2 × 4 or 2 × 6 timbers edgewise against a second 2 × 4, as shown at A, with alternating joints in the edgewise and flatwise sections. The construction can be carried out to greater lengths simply by continuing the 20-foot sections. Longer or shorter sections may be used, if more convenient.

The method of making the joints is shown at C. Quarter-inch or $\frac{3}{16}$-inch iron, $1\frac{1}{2}$ to 2 inches wide, is recommended for the straps, with $\frac{1}{2}$-inch bolts to hold the pieces together. One bolt should be run through the pieces midway between joints, to provide additional rigidity.

Although there are many ways in which such a mast can be secured at the base, the "cradle" illustrated at D has many advantages. Heavy timbers set firmly in the ground, spaced far enough apart so the base of the mast will pass between them, hold a large carriage bolt or steel bar which serves as a bearing. This bolt goes through a hole in the mast so that it is pivoted at the bottom. As the mast is swung upward in an arc while being raised, the bottom will be free to pivot on the bearing.

The job of raising the mast can be simplified, when a bottom bearing of this nature is used, because half of the guys can be put in place and

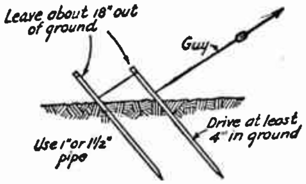


Fig. 1804 — Pipe-guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.

tightened up before the mast leaves the ground. Four sets of guys should be used, one in front, one directly in the rear, and two on each side at right angles to the direction in which the mast will face. Since the base position is fixed by the bearing, all the side guys can be put in place, anchored and tightened while the mast is lying on the ground. A set of guys should be used at each of the joints in the edgewise sections, the guy wires being wrapped around the pole rather than fastened to bolts or passed through holes in the pole; either of the latter methods tends to weaken the joints.

For heights up to 50 feet, 2 × 4-inch members may be used throughout. For greater heights, use 2 × 6s for the edgewise sections, 2 × 4-inch pieces will do for the flat sections.

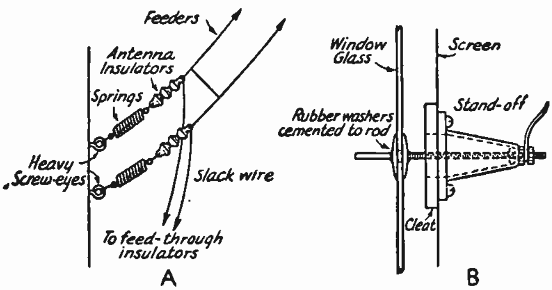


Fig. 1805 — (A) Anchoring feeders takes the strain from feed-through insulators or window glass. (B) Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

Guys and Guy Anchors

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material (No. 12 iron wire is considerably heavier than copper). Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

Guy wires should be broken up by strain insulators, to avoid the possibility of resonance at the transmitting frequency. Common practice is to insert an insulator near the top of each guy, within a few feet of the pole, and then cut each section of wire between the insulators to a length which will not be resonant either on the fundamental or harmonics. An insulator every 25 feet will be satisfactory for frequencies up to 30 Mc. The insulators should be of the "egg" type with the insulating material under compression, so that the guy will not come down if the insulator breaks.

Guy wires may be anchored to a tree or building when they happen to be in convenient spots. For small poles, a 6-foot length of 1-inch pipe driven into the ground at an angle will suffice. Additional bracing will be provided by using two pipes, as shown in Fig. 1804.

Halyards and Pulleys

A free-running pulley and a long-lived halyard are definite assets to an antenna system. Common clothesline is strong enough for small antennas, but does not stand the weather well and should be renewed frequently. Sash cord is better, but still not weather resistant. A satisfactory halyard is ⅜- or ½-inch waterproofed manila rope, the larger size being needed only to hold long stretches of wire. Ordinary rope or cord can be waterproofed by soaking it a day or two in automobile top dressing.

A good grade of galvanized iron pulley will be satisfactory in locations where the atmosphere is free from salt, but at seashore locations a pulley intended for marine use should be used. One of the best types is a hardwood block with a bronze roller-bearing shaft, which will resist corrosion under adverse conditions.

Bringing the Antenna or Transmission Line into the Station

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 1805, to remove strain from the lead-in insulators. Holes cut through the walls of the building and fitted with feed-through insulators are undoubtedly the best means of bringing the line into the station. The holes should have plenty of air clearance about the conducting rod, especially when using tuned lines which develop high voltages. Probably the best place to go through the walls is the trimming board at the top or bottom of a window frame which provides flat surfaces for lead-in insulators. Cement or rubber gaskets may be used to waterproof the exposed joints.

Where such a procedure is not permissible, the window itself usually offers the best opportunity. One satisfactory method is to drill holes in the glass near the top of the upper sash. If the glass is replaced by plate glass, a stronger job will result. Plate glass may be obtained from automobile junk yards and drilled before placing in the frame. The glass itself provides insulation and the transmission line may be fastened to bolts fitting the holes.

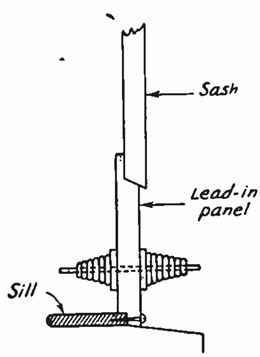


Fig. 1806 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Sealing the overlapping joint will help make it weatherproof.

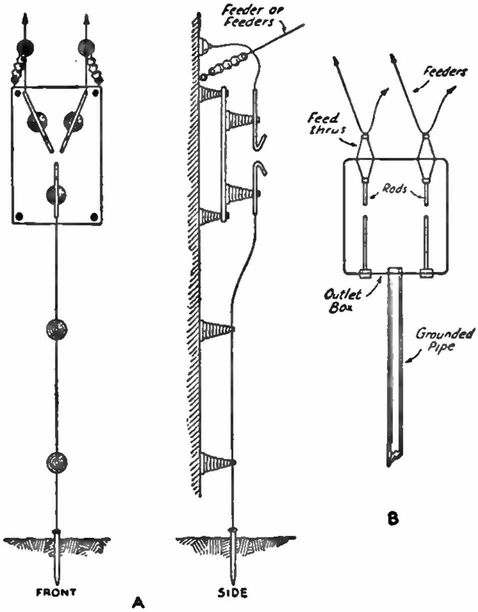


Fig. 1807 — **Low-loss lightning arresters for transmitters.**

Rubber gaskets cut from inner tube will render the holes waterproof. The lower sash should be provided with stops at a suitable height to prevent damage when it is raised. If the window has a full length screen, the scheme shown in Fig. 1805-B may be used.

As a less permanent method, the window may be raised from the bottom or lowered from the top to permit insertion of a board which carries the feed-through insulators. This lead-in arrangement can be made weatherproof by making an overlapping joint between the board and window sash, as shown in Fig. 1806, and covering the opening between sashes with a sheet of soft rubber from a discarded inner tube.

Lightning Protection

An ungrounded radio antenna, particularly if large and well elevated, is a lightning hazard. When grounded, it provides a measure of protection. Therefore, grounding switches or lightning arresters should be provided. Examples of construction of low-loss arresters are shown in Fig. 1807. At A, the arrester electrodes are mounted by means of stand-off insulators on a fireproof asbestos board. At B, the electrodes are enclosed in a standard steel outlet box. The gaps should be made as small as possible without danger of breakdown during operation. Lightning-arrester systems require the best ground connection obtainable.

The most positive protection is to ground the antenna system when it is not in use; grounded flexible wires provided with clips for connection to the feeder wires may be used. The ground lead should be short and run, if possible, directly to a driven pipe or water pipe where it enters the ground outside the building.

Antenna Switching

It is often desirable, particularly in DX work, to use the same antenna for transmitting and receiving. This requires switching of antenna from transmitter to receiver. One of two general systems may be employed. In the first, the transmitter and receiver each are provided with an antenna tuner, and the antenna transmission line is switched from one to the other. In the second system, one antenna tuner is provided for each antenna and the switch is in the low-impedance coupling line. Several typical arrangements are shown in Fig. 1808.

The high voltages which develop on tuned lines require switches and wiring having good insulation. Frequently relays with low-capacity contacts are substituted for manually operated switches. Either way is satisfactory.

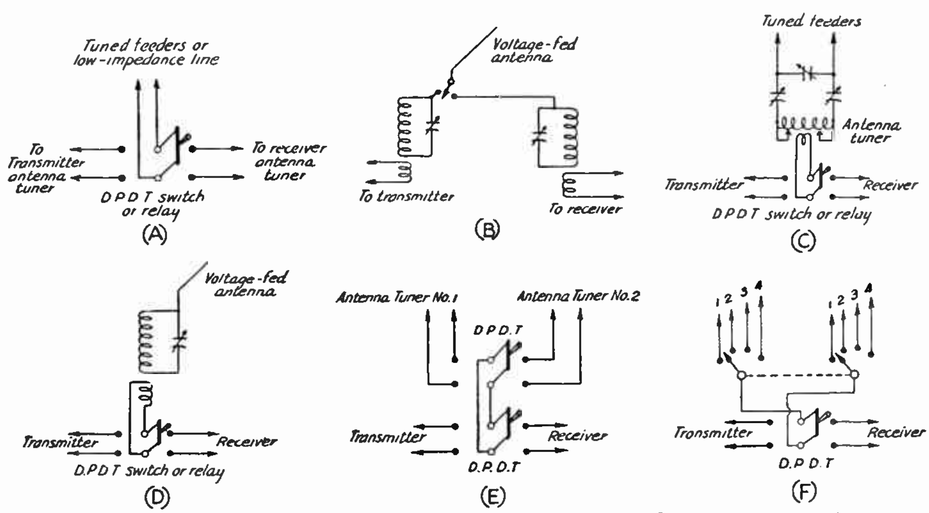


Fig. 1808 — **Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. E — For two tuned-line antennas with a tuner for each antenna or for two low-impedance lines. F — For combinations of several two-wire lines.**

Rotary Beam Construction

Many amateurs mount the simpler types of directive antennas in such a way that the antenna can be rotated to shift the direction of the beam at will. Obviously the use of such rotary antennas is limited to the higher frequencies, if the structure is to be of practicable size. For this reason the majority of rotary-beam antennas are constructed for use on 14 Mc. and higher frequencies. The problems in rotary-beam construction are those of providing a suitable mechanical support for the antenna elements, furnishing a means of rotation, and attaching the transmission line so that it does not interfere with the rotation of the system. The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the rotating structure. The large diameter of the conductor is beneficial also in reducing resistance, which becomes an important consideration when close-spaced elements are used.

When the elements are horizontal a supporting structure is necessary, made usually of light but strong wood. Dural tubes often are used for the elements, and thin-walled corrugated steel tubes with copper coating also are available for this purpose. The elements frequently are constructed of sections of telescoping tubing, making length adjustments quite easy. Electricians' thin-walled conduit also is suitable for rotary beam elements.

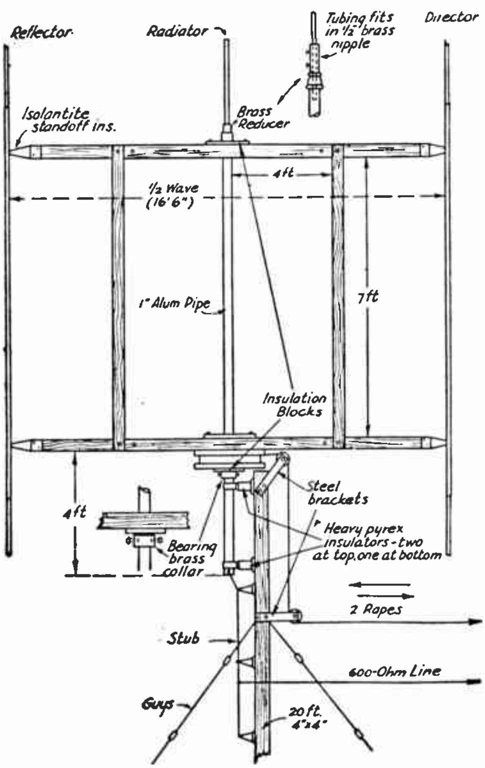


Fig. 1809 — A practical vertical-element rotatable array for 28 Mc. The driven antenna is fixed and the reflector and director elements, parasitically excited, rotate around it. Close-spaced elements may be used if desired.

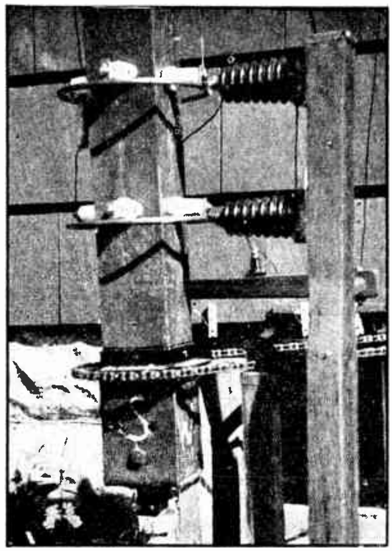


Fig. 1810 — One form of rotating mechanism. A bicycle sprocket and chain turn the pole which supports the beam antenna. Feeder connections from the antenna are brought to the metal rings, which slide against spring contacts mounted on the large stand-off insulators.

Various means of rotation and of making contact to the transmission line have been devised. Fig. 1809 shows a mechanical arrangement suitable for use with vertical elements. The antenna, which is a vertical section of metal tubing, is mounted in a fixed position and is provided with a director and reflector which rotate about it. The advantage of this arrangement is that no provision need be made for special contacts between the antenna and the feeder system, since the position of the antenna is fixed. A rope-and-pulley arrangement provides rotation from the operating room, so that, when a signal is picked up, the antenna can be rotated rapidly to the position which gives maximum response. It is then also pointing in the proper direction for transmission. The system can be varied in dimensions and details; for instance, close element spacing might be used to give greater gain.

Parts from junked automobiles often provide gear trains and bearings for rotating the antenna. Rear axles, in particular, can readily be adapted to the purpose. Some amateurs use motor-driven rotating mechanisms which, although complicating the construction, simplify remote control of the antenna. More or less elaborate indicating devices, which show the direction in which the antenna is pointed, often are used with motor-driven beams.

One method is shown in Fig. 1810. In this case the pole is rotated by a chain-and-sprocket arrangement, with the base resting on a bearing. Feeders are brought down the pole from the antenna to a pair of wire rings, against which sliding contacts press.

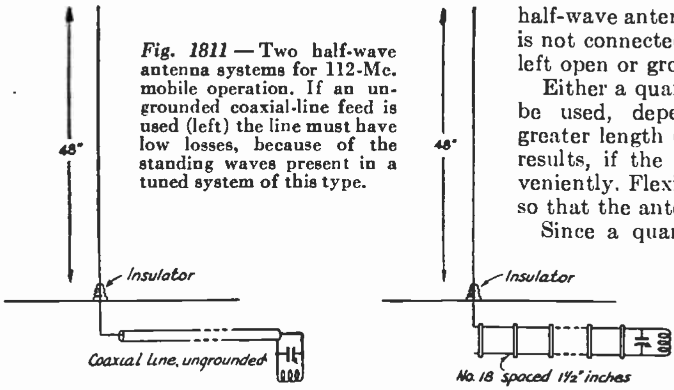


Fig. 1811 — Two half-wave antenna systems for 112-Mc. mobile operation. If an ungrounded coaxial-line feed is used (left) the line must have low losses, because of the standing waves present in a tuned system of this type.

Mobile Antennas

For mobile work on the very-high frequencies, a flexible rod or "whip" antenna is commonly used, mounted vertically on stand-off or feed-through insulators attached to the car body. Where possible the antenna should be a half-wavelength long, since this length will give the best two-angle radiation. A quarter-wave antenna, working against the metal car body as a counterpoise or "ground," may be used but it is not so efficient a radiator as the half-wave antenna.

As in the case of antennas for fixed stations, it is important that the car antenna be mounted as high as possible, to avoid screening effects of the car and to give maximum range. The best location for mounting the antenna is in the middle of the roof in the case of a car with a metal top. If the antenna cannot be mounted so that it is entirely above the top of the car, it should still be made to have a major portion of its effective radiating length above the roof. The top forms a "ground" of good conductivity and improves the performance. Convertibles and coupés have a convenient spot for mounting the antenna on the deck in back of the rear window. The lead-in can be brought into either the luggage compartment or the driver's seat, depending upon the location of the radio gear. Sedans lend themselves more readily to mounting the antenna alongside the hood or on the rear bumper. An antenna mounted alongside the car body but projecting above the metal top will transmit best in the direction across the top of the car.

It is advantageous to mount the antenna as near the transmitter as possible, in order to simplify the feed system. Special feeder systems, such as low-loss coaxial lines, are necessary if the antenna is located at one end of the car and the transmitter at the other. A quarter-wave tuned line is a suitable system, using appropriate tuning methods. When used with an end-fed half-wave antenna, the end of the feeder which is not connected to the antenna may either be left open or grounded to the car body.

Either a quarter- or half-wave antenna may be used, depending upon conditions. The greater length of the latter will lead to better results, if the installation can be made conveniently. Flexible metal rod is generally used, so that the antenna will be self-supporting.

Since a quarter-wave antenna normally is supported at a low-voltage point, hard rubber insulators are satisfactory. However, a half-wave antenna will usually be supported at a high-voltage point and thus requires good insulation for best efficiency. Ceramic insulators usually can be obtained to fit any case. It is wise not to skimp on size because of the greater chance of breakage with the smaller units. The feed through types and the stand-off types with metal base rings are least likely to break.

The two methods of feeding the half-wave antenna shown in Fig. 1811 are probably the most convenient. Both systems use tuned feed lines, and thus require a tuning-system at the transmitter end.

If a quarter-wave antenna is to be mounted permanently on the car it should be located on the roof, otherwise it is likely that the radiation pattern will be quite irregular. The resulting directional effects will be a help on some occasions but a definite hindrance on others. The antenna can be fed by a tuned line or by a coaxial line, as shown in Fig. 1812. The coaxial line can be of the 70- or 100-ohm type.

The coaxial line feed can be checked by observing its detuning effect on the transmitter; a good match will have been obtained when the detuning is a minimum. The antenna length should be about 22 to 24 inches, and this length and the capacity of the condenser should be varied until connecting the other end of the line to the transmitter causes a minimum of frequency change. Loading is controlled at the transmitter by adjusting the coupling coil, *not* by varying the condenser at the antenna. The antenna is made longer in small steps and the condenser adjusted until the concentric line introduces a minimum of reactance at the transmitter (shows the least detuning effect on the tank circuit). The method is simply to vary the length of the radiator until it shows an impedance near that of the line and then to cancel the reactance by the series condenser.

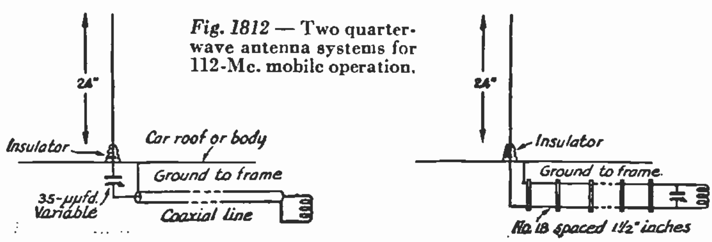


Fig. 1812 — Two quarter-wave antenna systems for 112-Mc. mobile operation.

℄ V.H.F. Antennas

Although antennas for the very-high frequencies are constructed on the same principles as those for lower frequencies, the smaller dimensions permit structural arrangements which would be unwieldy, if not impossible, on lower frequencies. The extended double Zepp, used vertically, is particularly easy to mount, the elements being made of ¼-inch copper or dural rod or tubing and fastened to the side of a pole by stand-off insulators. Open-wire feeder systems are better if the line is long, the losses being lower than with twisted pair.

A simple, practical application of the end-fire principle (§ 10-12) is the use of two lengths of copper tubing, bent to form a "pitchfork" one half-wavelength long (down to the bend) and with a quarter- to an eighth-wavelength separation. If the pole can be made to rotate 180°, full advantage may be taken of the directivity of the system. Tuned feeders may be used if the length is not more than one or two wavelengths; for greater lengths, an untuned line and a matching stub are desirable.

Combination collinear and broadside arrays as described in Chapter Ten give good gain and are not difficult to construct. The elements can be made of wire or tubing. The assembly may consist simply of wires hung from a rope stretched between two supports.

Close spacing and balance are important factors in v.h.f. feeder operation to minimize radiation from the line. For this reason the coaxial line is the best type of feed for the v.h.f. antenna, but the open wire line is quite effective if care is taken in its construction. If a matching section is used, it should be symmetrical and loaded on both sides, to maintain current balance in the matching section.

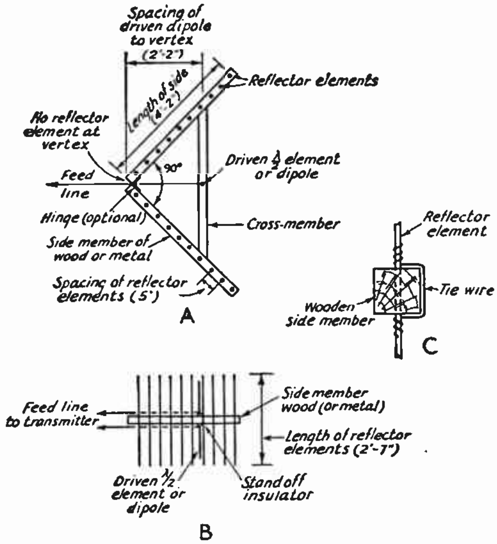


Fig. 1813 — A corner reflector antenna system with a spine- or grid-type reflector. The reflector elements are made of stiff wire or tubing. The dimensions shown are for the 224 Mc. band and must be doubled for 112 Mc.

Corner reflector antenna — A type of highly directive antenna system for the v.h.f. and u.h.f. ranges above 56 Mc. which is comparatively easy to construct is the "corner" reflector, shown in Fig. 1813. It consists of two plane reflecting surfaces set at an angle of 90°, with the antenna set on a line bisecting this angle. The distance of the antenna from the vertex should be 0.5 wavelength. The reflector surfaces are made of spines spaced about 0.1 wavelength apart.

The antenna used may be a center-fed full-wave affair with a two-wire line. Since the radiation resistance of the antenna is raised when the deflector is used, an impedance matching system will be required if ordinary types of lines are used. For this reason a tuned line is advisable. Alternatively, a folded dipole (§ 10-14) may be used directly with a 500-ohm line (No. 12 wire spaced 2 inches).

The transmission line should be run out at the rear of the reflector, to keep the system symmetrical and thus avoid any unbalance.

The corner reflector antenna will give a gain of approximately 10 db. over a simple half-wave dipole. The front-to-back and front-to-side ratios are of the order of 35 and 25 db., respectively, in a typical case, and the directional pattern is relatively free from secondary lobes of appreciable amplitude.

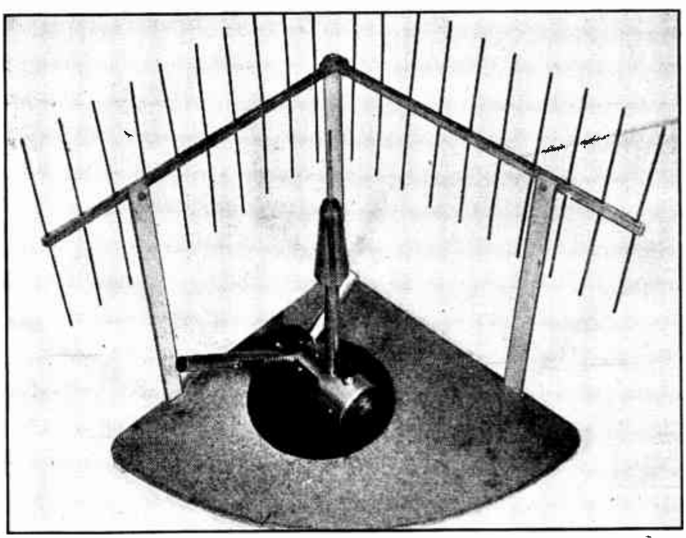


Fig. 1814 — A 750-Mc. transmitter using an acorn triode as an oscillator. The concentric antenna is an integral part of the oscillator unit. The square-corner reflector concentrates the radiation in the desired direction. (W6IOJ.)

Chapter Nineteen

Carrier-Current Communication

WARTIME restrictions on radio communication have led many amateurs into the experimental investigation of non-radio methods of communication. A number of such methods have been explored — including induction-field, ground-current, light-beam and supersonic transmission — but the only one to achieve widespread use is carrier-current (wired-wireless) transmission. Hundreds of individuals in various sections of the country now are engaged in experimental communication using this method.

Although carrier-current communication systems have been in use for many years by telephone and power companies, the subject is a relatively new one to radio amateurs. For this reason, it seems appropriate to preface this chapter with a brief résumé of the principles involved, the results which may be reasonably expected, and the difficulties which may be encountered.

Carrier-Current Fundamentals

Essentially, carrier-current communication is similar to radio communication. The process is one of impressing modulation, either in the form of voice or the telegraph code, on a radio-frequency carrier (§ 4-1; § 5-1) and then demodulating the carrier at the receiving end. The only difference is in the medium by which the r.f. carrier wave is transported.

Since most amateurs are familiar with the feeders or transmission lines commonly used to feed antennas, the basic principles of the carrier-current system can be explained by saying simply that it is communication by means of feeder wires or transmission lines. R.f. energy is fed into the transmission line by a transmitter at the sending station, and is delivered to a receiver (instead of an antenna) coupled to the distant end of the line. For communication by this method the transmission line is, therefore, a prerequisite.

While telephone companies, which use the carrier system to carry several conversations simultaneously over a single conductor for long-distance circuits, employ lines especially designed for the purpose, the electric power companies have very successful systems operating over the same high-tension lines by which power is distributed from central generating plants. This leads to the thought of using the same lines which supply electric power to our homes as the transmission lines required for carrier-current communication, since all homes within a wide area usually are coupled to the same power system in one way or another.

Frequencies — A balanced feeder system radiates very little energy compared with the radiation from an antenna at the high frequencies normally assigned for amateur use in peacetime. Nevertheless, even a well-balanced two-wire line will radiate more energy at these frequencies than is permitted under wartime restrictions. Since radiation from a given conductor decreases as the frequency is lowered, most carrier-current systems operate at frequencies lower than 200 kc.

From the private experimenter's point of view, there is an even more important reason for using low frequencies. Power lines feeding the average home ordinarily are extremely poor transmission lines for high-frequency currents. Not only are these lines shunted by very low impedances in the house itself — lamps, heaters, b.c. receivers, motors and other appliances which consume r.f. energy — but the lines, once outside the house, are interrupted by transformers whose high capacities shunt much of the remaining energy to ground. For these reasons it may be said that, in general, the lower the frequency the better the performance for carrier-current work.

There are other considerations which limit the advisable extent to which the frequency may be lowered for purposes of private communication, however. It happens that the public utilities operating carrier circuits make use of the frequencies below 160 kc. Since individuals will not want to run the risk of creating interference with vital services, the very low frequencies are to be avoided. Another reason for giving preference to somewhat higher frequencies is that their use reduces, to a certain extent, the very real problem of supplying the large values of inductance and capacity required for oscillator tank circuits. Frequencies of 160 to 200 kc. are sufficiently low for reasonably successful work and avoid the range commonly used by public-utilities systems.

Operating restrictions — Although no Federal license is required for the operation of carrier-current equipment, there are two restrictions which must be observed. The first of these is the FCC regulation (Sec. 2.102) which limits the *radiation* field strength to a value of 15 microvolts per meter at a distance in feet of $157{,}000/f_{kc.}$, where $f_{kc.}$ is the frequency in kilocycles. At a frequency of 150 kc., for example, the radiation field strength should not exceed 15 microvolts at a distance of 157,000/150 or 1046 feet from any power line which may be carrying r.f. from the transmitter.

The second restriction is one imposed in certain regions by military authorities, in the form of public proclamations prohibiting the use of *any* equipment capable of being employed for communication within specified restricted zones. These zones are designated by notices posted at every local selective service board office, post office, court house or town hall within each restricted zone. Where such a military order is in effect, it means that carrier-current communication (as well as any other kind) is specifically prohibited.

As might be expected, the noise level of most domestic power lines is rather high. To overcome this difficulty, the use of high transmitter power might seem desirable. However, because of the legal limitation on radiation field strength previously mentioned and the fact that harmonic output must be kept low to prevent interference with broadcast reception, the use of transmitter power inputs exceeding 50 watts or so is seldom advisable.

Further restrictions involving the rights of the utility companies in controlling the use of their wire lines may be encountered. Since in many instances their own carrier-current transmissions are carried over the lines, care must be taken to avoid interference on the frequencies used by them, or by any other established channel of communication.

Although there would be a reduction of r.f. losses if the transmitter could be coupled to the line on the street side of the meter, such connections are illegal, as are any connections made to telephone lines.

Ranges — Since performance depends so largely upon line conditions, it is impossible to predict with any degree of accuracy the distance range which may be expected. In general, greater distances can be covered in rural districts, where open-wire lines are more often employed, distribution transformers are less frequent, and loading is less along the line. In the cities ranges usually will be less, because much of the wiring is carried in grounded conduit and the distances between loading points are short. However, the city dweller has an advantage in that usually he does not need to cover as great a distance to find someone with whom to communicate. With transmitter power inputs of 25 watts or less, distances up to five miles often are reported in metropolitan areas. Rural stations frequently are able to increase their ranges to ten miles or more. It should be remembered that, to cover an air-line distance of three or four miles between transmitting and receiving stations, the signal may have to travel a considerably greater distance in following the power lines.

The fact that two stations may receive power from different distribution systems does not necessarily mean that communication is impossible, since there is evidence that the signal may be transferred from one line to another by induction provided the two lines run close together at some point.

The useful range usually is greater in the daytime than at night because line loading is less during the day. For the same reason, the noise level is lower in the daytime and during the late hours at night than it is in the evening.

While both telegraphic and voice communication have been carried on successfully, telegraphy will carry better through noise and more advantage may be taken of the noise-reducing properties of a selective receiver. A high percentage of modulation is advisable for 'phone work.

Getting started — The best first step for anyone interested in getting started in carrier-current work is to find someone not too distant from his location to work with him. The pages of *QST's* "Experimenter's Section" frequently list the names of other interested persons in many communities. Having located another enthusiast, one person can build the transmitter while the other takes on the job of making the receiver or converter. If the receiver is so designed that it may be operated in a car, so much the better, since it will then be possible to form a good idea of where and how far the signal is traveling by following

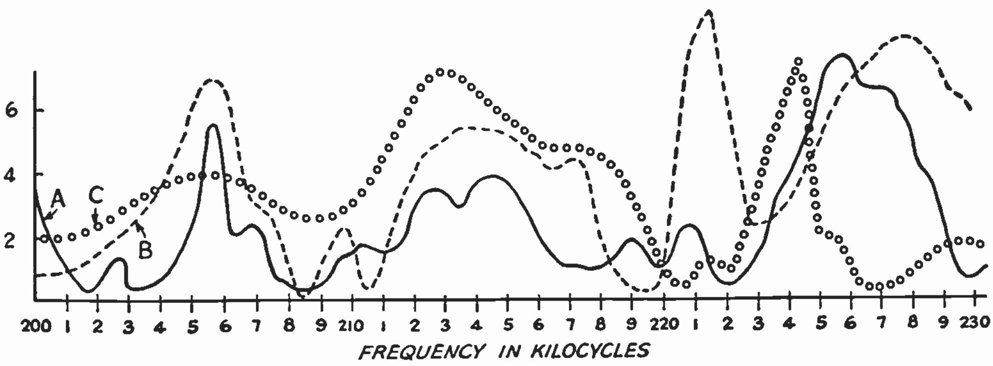


Fig. 1901 — A set of transmission curves run by three carrier-current stations to determine optimum operating frequencies for intercommunication. Such peaks as those at 205.6 kc. on the A curve, 221.3 kc. on the B curve, and 224 kc. on the C curve, are suited only for the sharply tuned carrier of a c.w. signal. From these curves it would appear that the optimum frequency for these stations to use in common would be 224.5 kc. The units of the vertical axis of the curve are in a.f. volts as indicated by an output meter.

power lines. Tests also can be made at a distance to determine the effects of experimental adjustments at the transmitter. Such tests eliminate much of the guesswork ordinarily connected with an experimental carrier-current system.

Selecting operating frequencies — In addition to the general considerations outlined under "Frequencies" on page 404, a critical survey of transmitting conditions over a wide range of frequencies should be made between each carrier-current station and every other such station with which it desires to maintain communications.

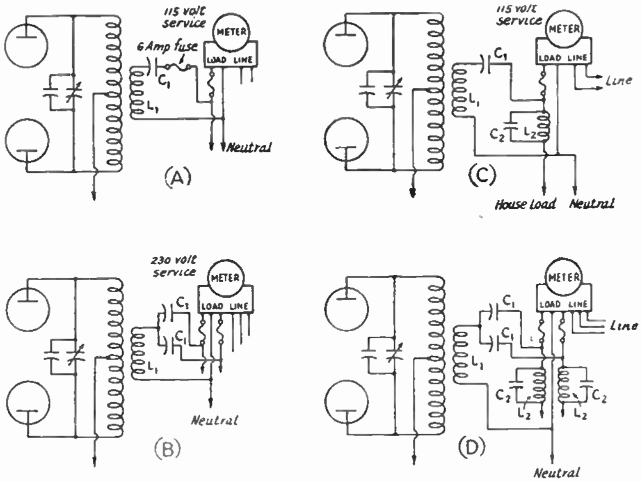


Fig. 1903 — **Methods of coupling the transmitter to the power line.**

Highly varied load conditions and by-passing effects prevail over the complicated networks of power lines in cities. Even in the case of a simple power line between two stations, without branches, transformer banks or substations, the choice of a frequency will often be found to be critical.

Transmission curves should be run between all stations in a network, in every possible combination of receive and transmit, and using the same variable frequency oscillator and receiver in each instance. The oscillator should be calibrated and cover a range of from 25 to 300 kc. or more. Tone modulation should be provided to enable the receiver to follow the signal. If the oscillator is not calibrated, its tuning control should be provided with as good a dial as can be obtained. Dial readings are logged so that good operating points may be located on the transmission curves. A satisfactory receiver for the tests is a simple non-regenerative detector with a stage of audio and an output meter, or a vacuum-tube voltmeter.

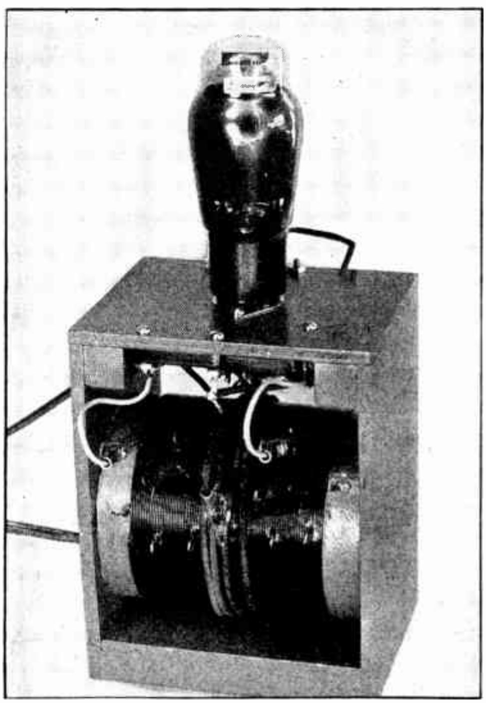


Fig. 1902 — **A 25-watt 6L6 transmitter for wired wireless. Since variable condensers large enough to give sufficient frequency change are difficult to secure unless paralleled b.c. receiver gangs are used, a tapped coil is used for tuning. Circuit diagram is shown in Fig. 1904.**

As many spot readings as possible should be taken. It is well to cover the entire range in 1-kc. steps, plotting the output meter of v.t.v.m. reading against the frequency setting or dial reading. The reception should be monitored aurally in order to identify false peaks caused by broadcast band harmonics, commercial transmitters, other carrier-current transmitters, etc.

Examination of the completed curve may reveal steep-sided peaks of less than 1 kc. in width. If the transmitter is set precisely on such a point, it will be excellent for c.w., but poor for 'phone, since most of the side-bands would be cut off.

Satisfactory peaks for 'phone operation will be four or five kc. in width, though they need not be flat-topped for the entire width.

A typical set of transmission curves run by three stations to determine optimum operating frequencies for intercommunication is shown in Fig. 1901. The separate voltage curves are identified by different types of lines in the chart, and differently colored inks or pencils will serve the same purpose in the actual test curves. A synthesis of the curves readily reveals the best average peak of all curves at all stations. Some stations will be obliged to sacrifice something in signal strength in order to join others on a mutually advantageous frequency. When a compromise frequency cannot be worked out satisfactorily, a net may be subdivided into sections which can share a relatively good frequency.

Other advantages secured in the running of transmission curves are the noting of particularly noisy spots which may thus be avoided as well as the locations of frequencies already occupied by established communications systems.

ℭ Station Equipment

Except for the antenna, the apparatus required for carrier current consists of the same units as used in radio communication — transmitter, receiver, power supplies, modulator (if used) and microphone or key. The apparatus may consist of anything from a simple self-excited oscillator for the transmitter and a regenerative receiver (if c.w. alone is used) to a modulated m.o.p.a.-type rig with a superheterodyne receiver or a converter working into a communications or broadcast receiver which is used as the i.f. and a.f. amplifier. Representative examples of the kinds of equipment commonly used are shown in the photographs and diagrams which appear throughout this chapter.

Checking frequency — The frequency of the transmitter may be checked by picking up harmonics on a near-by broadcast receiver. For example, when the transmitter is tuned to 150 kc. the fourth harmonic will be heard at 600 kc., the fifth harmonic at 750 kc., the sixth harmonic at 900 kc., etc. The number of kilocycles between any consecutive pair of harmonics will give the transmitter frequency. If harmonics are separated more than 200 kc. the transmitter frequency is too high, while a separation of less than 150 kc. will indicate that the transmitter frequency is too low. Frequencies of 150 kc. and 200 kc. are most suitable for checking in this manner, since their harmonics fall in broadcasting-station channels where the beats with the broadcast signals are easily spotted.

It is advisable to move the transmitter frequency to a setting such that the harmonics fall between broadcast channels and, in particular, well away from the frequencies used by local stations, to avoid interference with neighboring listeners.

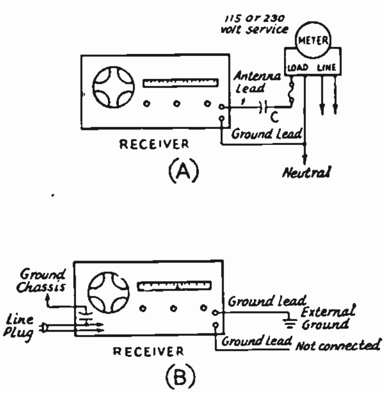


Fig. 1904 — Methods of coupling the receiver to the line.

Coupling to line — Various methods of coupling the output of the transmitter to the power line may be used, as shown in Fig. 1903.

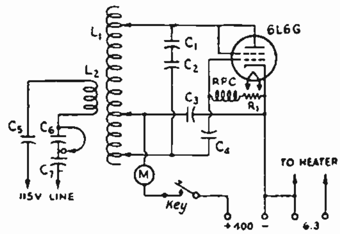


Fig. 1905— Circuit of the 25-watt 6L6 transmitter.

C_1, C_2 — 0.006-μfd. 2500-volt mica.
C_3, C_5 — 0.1-μfd. 600-volt paper.
C_4 — 100-μμfd. mica.
C_6, C_7 — 0.05-μfd. 600-volt paper.
R_1 — 50,000-ohm wire-wound, 10-watt.
RFC — 80-mh. r.f. choke (Meissner 19-2709).
L_1 — 80 turns No. 18 e., close-wound on 3½-inch diameter form, tapped every 5th turn.
L_2 — 4 turns No. 18 rubber-covered wire, wound over the center of L_1.

In these circuits, C_1 serves a dual purpose as both blocking and tuning condenser for the line circuit. The value to be used depends to a considerable extent on the line constants, and should be determined experimentally. In practice, it has been found that the capacity required varies from about 0.01 μfd. to as much as 0.05 μfd. The condenser used must be capable of withstanding the line voltage. The coupling coil, L_1, should be of sufficient size to provide the necessary coupling to the final tank circuit. It is advisable to start out with a fairly large coil, wound over the final tank coil and tapped every few turns so that the loading can be adjusted.

While it is possible to work without the use of load-isolating filters if the necessary material is not available for their construction, a considerable improvement in over-all efficiency can be obtained by their use, since a large percentage of the total power loss may be attributed to the house load. It can be seen from diagrams C and D, Fig. 1903, that the purpose of the filters is to prevent r.f. power from being expended in the shunt load normally connected to the house side of the meter.

It should be borne in mind that the inductance coils in the isolating filters must have sufficient current-carrying capacity for the total connected load without causing any serious loss of voltage at the line frequency. C_2 does not have to withstand any considerable voltage but must have the proper capacity to tune the coil, L_2, to the operating frequency. As in the case of the tank circuit, a wide range of values may be used. If a 0.005-μfd. mica condenser is used for C_2, a coil which will resonate at approximately 150 kc. may be wound with 70 turns of No. 14 enameled wire on a 3½-inch form. The voltage drop in such a coil should be

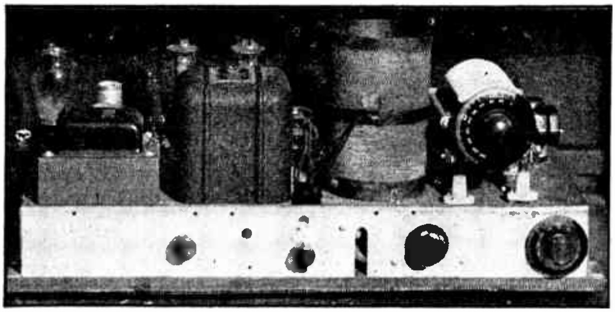


Fig. 1906 — M.o.p.a.-type carrier-current transmitter. Transmitter, modulator and power supply are on one chassis, arranged for remote control from the operating position. The final tank coil and tuning condenser are at right. Circuit diagrams are shown in Figs. 1907 and 1903.

negligible under normal house-wiring loads.

Suggested methods for coupling the receiver to the power line are shown in Fig. 1904.

Transmitter Construction

Hartley 6L6 transmitter — A carrier-current transmitter of the simplest type is pictured in Fig. 1902. The circuit, shown in Fig. 1905, is the conventional series-fed Hartley. The tank condenser, C_1C_2, consists of two 0.006-μfd. mica condensers connected in series (to decrease the possibility of breakdown). Frequency and excitation are adjusted by selection of the proper taps on the coil. The output or "antenna" coupling is adjusted by the proper selection of the condensers in series with the coupling coil, L_2.

The inductance is wound on 3½-inch diameter cardboard tubing, cut to a length of about 4¾ inches. The cardboard should be given a coat or two of shellac. Then 80 turns of No. 18 enameled wire must be wound on, as tightly as possible. Taps are made at every 5th turn, making a 1-inch loop of wire at each tap and twisting it tightly for several turns so that the loop will not pull apart as the rest of the coil is wound. When the coil is finished the loops should be scraped bare of insulation. As a finishing touch, spots of Duco cement may be applied to secure the twisted portions in place.

The framework used to support the coil and other components is made of ¼-inch plywood, except for the two corner strips at the top, which are of ½-inch square stock, and the two 1 × 2-inch bottom strips. The whole assembly is held together by brads and glue. The box is made just wide enough to allow the coil to be forced in, the pressure of the sides then serving to hold the coil firmly in place. The box measures 6⅞ inches high, 5¼ inches long, and 4¼ inches deep.

The tube socket is submounted in a hole in the center of the top of the framework. The tank condensers, C_1 and C_2, are fastened underneath by screws. The grid choke, *RFC*, is supported on a ½-inch pillar attached to the rear screw holding the socket. The coupling condensers, C_5, C_6 and C_7, are supported on lugs under the heads of screws which serve also for coupling taps at the rear of the box. Flexible leads from the coil are fastened to the terminals of the tank condensers.

When completed, the transmitter should be hooked up to a power supply delivering 350 or 400 volts at 100 ma. A meter and key should be connected in the positive high-voltage lead, as shown in Fig. 1905. While there may be some objection to placing the key in the positive lead, there is a measure of safety in the fact that, as long as the operator's hand is off the key (as it would be when making adjustments) there is no chance for shock when adjusting the coil taps. Care must be exercised, of course, and it is recommended that all tuning adjustments be made with one hand in the pocket.

Initial tuning-up should be done without the

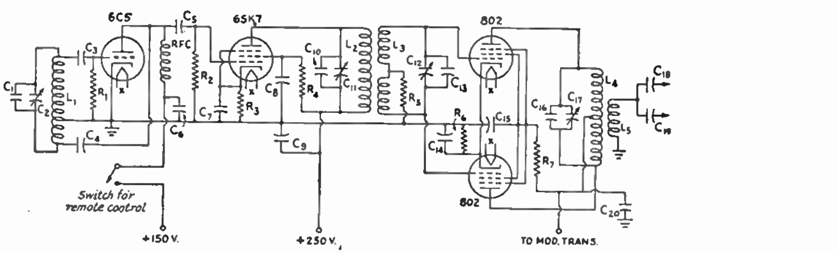


Fig. 1907 — Circuit diagram of the master-oscillator power-amplifier wired-wireless transmitter.

C_1 — 0.002-μμfd. mica.
C_2 — 350-μμfd. variable.
C_3 — 500-μμfd. mica.
C_4 to C_9, inc. — 0.1-μfd. 600-volt paper.
C_{10} — 0.002-μfd. 2500-volt mica.
C_{11}, C_{12} — 350-μμfd. variable.
C_{13} — 0.002-μfd. 2500-volt mica.
C_{14} — 0.1-μfd. 600-volt paper.
C_{15} — 0.002-μfd. 1000-volt mica.
C_{16} — 0.002-μfd. 5000-volt mica.
C_{17} — 350-μμfd. variable.
C_{18}, C_{19} — These condensers correspond to C_1 in Fig. 1903.
C_{20} — 0.002-μfd. mica, 1000 volts.
R_1 — 0.1 megohm, 1-watt.
R_2 — 50,000 ohms, 1-watt.
R_3 — 300 ohms, 1-watt.
R_4 — 50,000 ohms, 1-watt.
R_5 — 20,000 ohms, 1-watt.
R_6 — 300 ohms, 10-watt.
R_7 — 15,000 ohms, 10-watt.
L_1 — 160 turns No. 28 enam. on 1½-inch form, tapped 50th turn from bottom.
L_2 — 90 turns No. 22 enam. on 3-inch form.
L_3 — 90 turns No. 22 enam. on same form as L_2, half each side of L_2.
L_4 — 80 turns No. 18 enam. on 3½-inch form, tapped at center.
L_5 — See section on coupling.
RFC — 30-mh. choke.

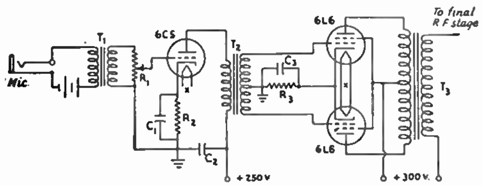


Fig. 1908 — **Circuit diagram of the modulator used with the m.o.p.a. carrier-current transmitter.**

C_1 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 10-μfd. 50-volt electrolytic.
R_1 — 0.5-megohm volume control.
R_2 — 1000 ohms, 1-watt.
R_3 — 125 ohms, 10-watt.
T_1 — Single-button microphone-to-grid transformer.
T_2 — Interstage audio transformer.
T_3 — Modulation transformer, 5000 ohms plate-to-plate to 3000-ohm load.

oscillator coupled to the line. Set the clips so that there are 60 turns between grid and plate, and attach the cathode tap at 25 turns from the grid end. Press the key and read the plate current; then try again with the cathode tap on either side of the first position. The setting of the cathode tap which gives the lowest plate-current reading is the one to use. With a 350-volt supply, the no-load plate current should run around 25 or 30 ma.

The next step is to connect the output circuit to the line. Set the coupling clip so that neither C_6 nor C_7 is being shorted (the condition of loosest coupling). The plate current under load should increase to 30 or 40 ma., depending upon the frequency of the transmitter.

The quality of the note can be checked by listening to a harmonic with the communications receiver set to its lowest frequency range. As the coupling is increased the note will become rough or "yoopy," indicating that the coupling is too tight or that the cathode tap needs some adjustment. The note will roughen up before it chirps; the roughness can be tolerated, but the chirp makes copying rather difficult.

Be careful when making adjustments — you will have deliberately hooked onto the 115-volt line, and you can get a severe shock from it!

M.o.p.a. transmitter — The photograph of Fig. 1906 illustrates a typical transmitter of the m.o.p.a. type. The circuit diagram appears in Fig. 1907. The tube line-up consists of a 6C5 Hartley oscillator, 6SK7 buffer, and push-pull 802 final amplifier. In addition to the r.f. circuits, the chassis also includes a plate-and-screen modulator for 'phone work, and a power supply. The wiring diagram of the modulator unit is shown in Fig. 1908.

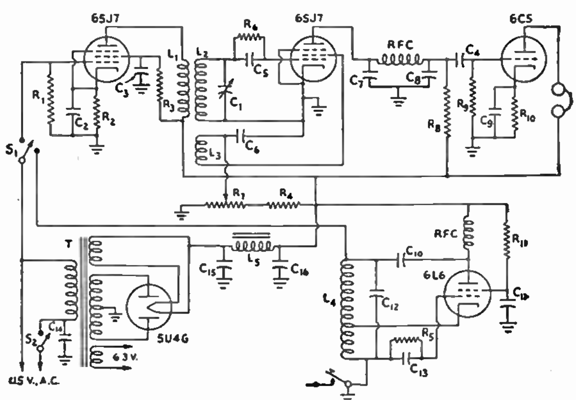
This transmitter operates at an input of about 20 watts, with a final plate current of 70 ma. and screen current of 30 ma. It is, however, capable of handling higher power with increased final-amplifier plate and screen voltages.

Almost any combination of tubes could be used in a similar arrangement. Triodes will require neutralizing circuits, of course.

Combination transmitter-receiver — In Fig. 1909 is shown the circuit diagram of a c.c. transmitter-receiver. The receiver consists of a 6SJ7 regenerative detector and a single-stage audio amplifier with a 6C5, preceded by an untuned stage of r.f. using a second 6SJ7 with the hot side of the power line tied directly to the grid. The detector is quite conventional. R_7 is the regeneration control. The receiver coils, L_1, L_2 and L_3, are scramble-wound with No. 32 enameled wire, on a piece of cardboard tubing about one and one-half inches in diameter. The secondary, L_2, should be wound on the cardboard form first and covered with a layer of friction tape. The primary, L_1, is then wound close to the grid end of L_2, and the tickler, L_3,

Fig. 1909 — **Circuit diagram for a combined transmitter-receiver for carrier-current communication.**

C_1 — 700-μμfd. variable.
C_2, C_3, C_4 — 0.01 μfd.
C_5 — 250-μμfd. mica.
C_6 — 0.25 μfd.
C_7, C_8 — 500-μμfd. mica.
C_9 — 0.1 μfd.
C_{10}, C_{11} — 0.002 μfd.
C_{12} — 0.006 μfd.
C_{13} — 250-μμfd. mica.
C_{14} — 0.1 μfd.
C_{15}, C_{16} — 8 μfd., 450 volts.
R_1 — 75,000 ohms, 1-watt.
R_2 — 1000 ohms, 1-watt.
R_3, R_4, — 50,000 ohms, 1-watt.
R_5 — 50,000 ohms, 2-watt.
R_6 — 2 megohms, ½-watt.
R_7 — 50,000-ohm variable.
R_8 — 0.25 megohm, ½-watt.
R_9 — 0.1 megohm, ½-watt.
R_{10} — 2500 ohms, 1-watt.
R_{11} — 15,000 ohms, 1-watt.
RFC — 25 mh. to 80 mh.
S_1 — S.p.d.t. toggle switch.
S_2 — S.p.s.t. toggle switch.
T — Power transformer: 350–0–350, 5 and 6.3 volts.



L_1 — 100 turns No. 32 e., 1½-inch diameter; see text.
L_2 — 300 turns No. 32 e., 1½-inch diameter; see text.
L_3 — 75 turns No. 32 e., 1½-inch diameter; see text.
L_4 — 150 turns No. 18 e., 2¼-inch diameter, tapped at 50 turns from grid end.
L_5 — 15- to 30-henry filter choke.

The ground connections which are shown in the diagram indicate connections to the chassis and not to actual earth.

close to the ground end. Finally, the entire coil should be given another covering of tape.

The transmitter is a regulation Hartley oscillator using a 6L6G. With the plate voltage available the input runs about 12 watts. The note should be crystal d.c.

The power supply is common to both the receiver and transmitter, the receiver acting as a bleeder for the supply.

All condensers are of the tubular type except the transmitter tank condenser, the grid condensers, and the variable used for detector tuning. The latter is a two-section gang condenser of 365 $\mu\mu$fd. capacity per section, with both sections in parallel. If the system shown in the circuit diagram is used, all ground connections shown must be made to the chassis and not to actual ground; otherwise, the key will short-circuit the line. If it is desired to ground the chassis for safety, place C_{14} between the arm of S_1 and the ungrounded side of the line. The chassis may then be grounded and the danger of shock or short-circuit removed. A 25-mh. to 80-mh. r.f. choke inserted between the key and L_4 will improve the keying. A 0.1-μfd. condenser should then be placed across the key.

Receiver Construction

Almost any superheterodyne receiver can be converted to operate on 150 kc. by making changes in the oscillator and tuned circuits.

While the circuits used in superheterodynes vary considerably from one model to another, a typical arrangement is that shown in Fig. 1910. The oscillator circuit, before necessary changes are made, is shown at A, while C is the revised circuit. C_2 is the series tracking condenser in the oscillator circuit; this condenser is removed and connected in parallel with C_1, the oscillator tuning condenser. Fig. 1910-B shows the radio-frequency circuit. L_1 and L_2 are removed and the coils shown in D are connected in place of those removed. The dimensions of the coils are given in the drawing. Coil L_2 has

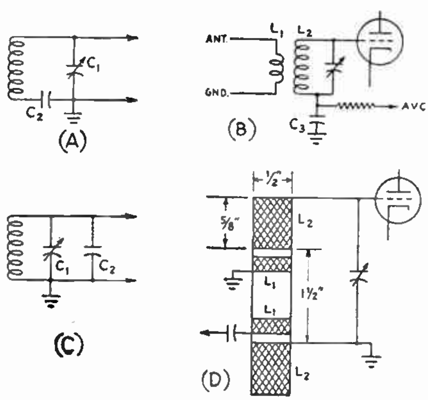


Fig. 1910 — Circuit changes required for converting a b.c. superheterodyne receiver for operation in the 150–160 kc. region. A, normal oscillator circuit; B, revised oscillator circuit; C, normal r.f. circuit (mixer input); D, revised r.f. circuit, showing cross-section of coil for tuning to 150 kc. See text for further description.

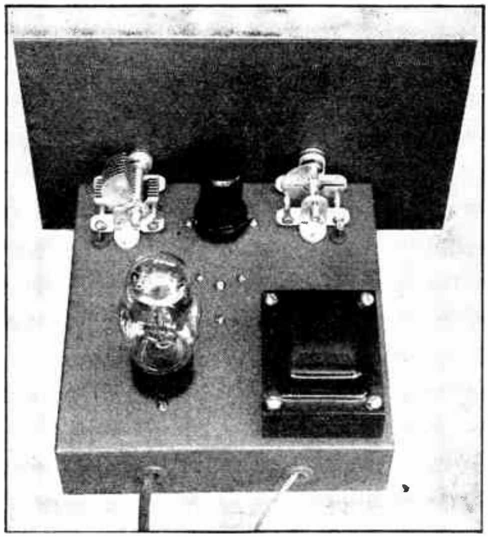


Fig. 1911 — Top view of the superheterodyne carrier-current converter. Note the adjusting screw for the output tank condenser, C_5, in front of the 6SA7 tube.

approximately 300 turns and L_1 approximately 25 turns. Both coils may be boiled in wax. After cooling, they should be wrapped with cotton tape so they will hold their form.

The simpler regenerative-detector receivers also have been used successfully for carrier-current work. The basic circuit of the simple regenerative receiver described in Chapter Twelve (Figs. 1201–1207) may be used with the provision of suitable coils and tuning condensers. Using a broadcast-type 365-$\mu\mu$fd. variable condenser for C_3, L_1 may be 120 turns of No. 32 enameled wire wound on a 3-inch diameter form and L_2 about 20 turns on the same form as L_1. In this instance, the band-spread condenser, C_2, should be omitted, and the antenna coupling condenser, C_1, will not be used.

The grid coil, L_1, of this receiver will be coupled to the 115-volt lines most efficiently if a link-coupling coil similar to L_2 in the transmitter diagram, Fig. 1905, using the same number of turns and size wire, is wound approximately at the center of L_1 of the receiver. Coupling condensers similar to C_5, C_6, and C_7 also must be used.

The three-tube wide-range general-coverage and bandspread superheterodyne receiver described in the same chapter (Figs. 1213–1220) also is suitable for carrier-current use when the low-frequency coils (Range A) are plugged in.

A superhet converter — A simple superheterodyne converter for carrier-current work is shown in Figs. 1911 and 1913.

The circuit of the converter, shown in Fig. 1912, is quite conventional. It consists of a 6SA7 mixer tube with the output on 1950 kc., so that it can be hooked into any communications receiver which will tune to 1950 kc. The grid circuit tunes the range 150 to 200 kc. and, in order to give the output frequency of 1950 kc., the oscillator tunes from 1800 to 1750 kc. The oscillator could also be made to tune from 2100 to 2150 kc., but by using the former range it can be checked on a communications receiver which covers only the amateur bands.

The converter is built on a 7 × 7 × 2-inch chassis. The tuning condensers, C_1 and C_3, are bolted to the chassis in a position such as to allow the panel to be supported by their panel bushings. The toggle switch and the screw holding the oscillator coil also help to hold the panel to the chassis. The mixer and oscillator padding condensers, C_2 and C_4, are fastened to the sides of the chassis, under their respective tuning condensers, and C_5, the output-circuit tuning condenser, is mounted on the chassis directly behind the 6SA7. The output coil, L_3, is fastened to the chassis near its tuning condenser. Both L_2 and L_3 are wound on one-inch bakelite forms. L_1 is a winding taken from a 175-kc. i.f. transformer.

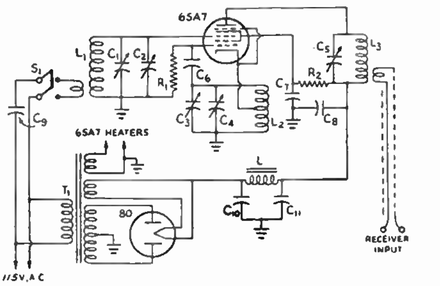


Fig. 1912 — Wiring diagram of the c.c. converter.
C_1 — 100-μμfd. variable (Hammarlund MC-100-S).
C_2, C_5 — 260-μμfd. trimmer (Hammarlund CTS-160).
C_3 — 20-μμfd. variable (Hammarlund MC-20-S).
C_4 — 350-μμfd. mica trimmer (Hammarlund CTS-230).
C_6 — 50-μμfd. mica.
C_7 — 0.1 μfd. 400-volt paper.
C_8 — 0.01 μfd. 600-volt paper.
C_9 — 0.1-μfd. 600-volt paper.
C_{10}, C_{11} — 8-μfd. 450-volt electrolytic.
R_1 — 20,000 ohms, ½-watt.
R_2 — 20,000 ohms, 1-watt.
S_1 — D.p.s.t. toggle switch.
T_1 — Power transformer, 240-0-240 volts each side of center-tap, 6.3- and 5-volt filament windings (Thordarson T-13R19).
L — 8 henrys, 40 ma. (Thordarson T-13C26).
L_1 — 175-kc. i.f. transformer replacement winding (Carron S735). Antenna winding is 11 turns of No. 32 d.s.c. wire wound over L_1.
L_2 — 43 turns No. 32 d.s.c., close-wound on 1-inch diameter form. Cathode tap at 5th turn from ground end.
L_3 — 50 turns No. 32 d.s.c., close-wound on 1-inch diameter form. Output coil is 14 turns of No. 32 d.s.c., close-wound, ⅛-inch from L_3.

The primary winding for L_1 is put on after two layers of cellophane tape have been wound over L_1 to serve as insulation. The primary then is wound on and cemented with coil dope or Duco cement.

The converter is put into service by connecting its output to the input of a communications receiver. While both the converter and receiver are warming up, set the receiver to 1950 kc. Then adjust the output trimmer, C_5, for maximum noise from the receiver. Next, check the converter oscillator range by setting the oscillator tuning condenser, C_3, to minimum capacity and the receiver to 1800 kc. It should then be possible to tune in the converter oscillator signal by adjusting C_4 until the signal is heard. Check the range of the oscillator by setting C_3 at maximum; if it can be tuned in at 1750 kc. on the receiver, the range is right on the nose. If the range is too great (oscillator frequency lower than 1750 kc. at maximum capacity), it indicates that fewer turns are required on L_2, and vice versa. The range is not critical, of course, since the converter is not ganged.

The receiver can now be reset to 1950 kc. and C_1 and C_3 set to the middle of their ranges. After adjusting C_2 for maximum noise (with S_1 closed), the converter is lined up for action. It will be found that the mixer tuning control is not too sharp and will need attention only after the signal has been tuned in with the main tuning control, C_3. Remember that C_3 tunes backward to the usual way: i.e., the converter is tuned to 150 kc. when it is at minimum capacity and to 200 kc. when it is at maximum capacity, the reverse of the mixer condenser action.

The coupling switch, S_1, is included so that the converter will not receive radio-frequency current during transmission periods. It should be used to disconnect the converter from the line whenever the transmitter is being keyed.

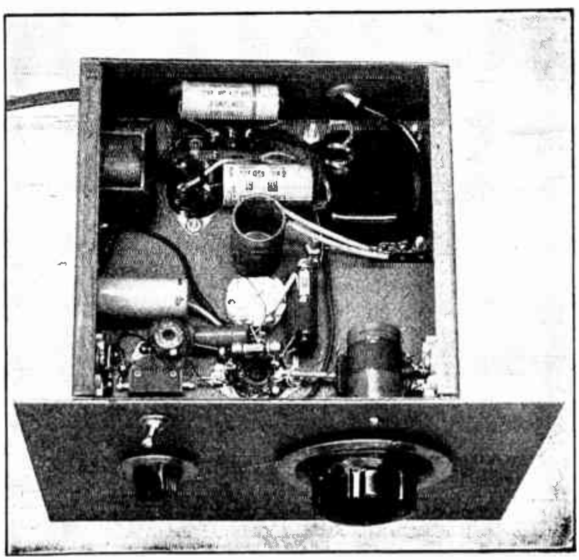


Fig. 1913 — This bottom view of the converter also shows the panel layout. The output tank coil, L_3, can be seen in the center of the chassis; L_2 is directly under the oscillator tuning dial, and the mixer grid coil, L_1, can be seen at the right, next to the toggle switch. Padding condensers C_4 and C_2 are mounted on the left- and right-hand sides of the chassis, beneath their respective tuning condensers.

Chapter Twenty

Measurements and Measuring Equipment

THE PROPER OPERATION of all but the very simplest of transmitters and receivers calls for the use of measuring instruments of various types. While the amateur station can be operated successfully with nothing more than a means for checking transmitter frequency and power input — and modulation, in the case of a 'phone transmitter — the progressive amateur is interested in instruments and measurements as an aid to better performance. The measure of the perfection of an amateur station, once a satisfactory transmitter and receiver have been provided, is the extent and utility of the auxiliary measuring and checking apparatus available.

Fundamentally, the process of measurement is that of comparing a quantity with a reference standard. Measuring equipment divides into two types: (1) fixed *standards* giving a reference point of known accuracy, used with associated equipment for making comparisons between the known and unknown quantities, and (2) direct-reading instruments or *meters* which have previously been calibrated in terms of the quantity being measured.

WWV SCHEDULES

All U. S. frequency calibration is based on the standard frequency transmissions from the National Bureau of Standards standard-frequency station, WWV. This station is on the air continuously, day and night, its radio frequencies of 5, 10 and 15 Mc. (and 2.5 Mc. from 7 P.M. to 9 A.M. with 440-cycle modulation only) modulated by standard audio frequencies of 440 and 4000 cycles per second, the former corresponding to A above middle C. In addition, there is a 0.005-second pulse every second, heard as a faint tick, which provides an accurate time interval for purposes of physical measurements.

The audio frequencies are interrupted on the hour and every five minutes thereafter for one minute to give the station announcement and to provide an interval for checking r.f. measurements. The station announcement is the call, WWV, sent in code, except on the hour and half hour when it is given by voice.

The accuracy of all frequencies is better than a part in 10,000,000. The 1-minute, 4-minute, and 5-minute intervals marked by the beginning and ending of the announcement periods are accurate to a part in 10,000,000. The beginnings of the periods when the audio frequencies are interrupted mark accurately the hour and the successive 5-minute periods.

Methods of making the measurements required in the amateur station will be discussed in this chapter, and design and construction of representative types of the instruments used in making these measurements will be described.

Frequency Measurement

Dependable frequency-measuring apparatus is desirable in the amateur station for several closely related purposes:

To insure that the transmitter is operated in the desired frequency band.
To set the transmitter to a desired frequency (if a self-controlled oscillator is used).
To determine the frequency of a received station, or to calibrate the receiver.
To determine the harmonic at which a frequency multiplier stage operates.
To determine the harmonic output of the transmitter.

Sec. 12.135 of the FCC Regulations states:

> The licensee of an amateur station shall provide for measurement of the transmitter frequency and establish procedure for checking it regularly. The measurement of the transmitter frequency shall be made by means independent of the frequency control of the transmitter and shall be of sufficient accuracy to assure operation within the frequency band used.

Frequency (§ 2-7) is measured by counting the number of cycles or oscillations per second. Since this cannot be done directly, except at very low frequencies, in practice the measurement is made (a) by noting the response of a selective resonant device, such as a tuned circuit (absorption frequency meter, Wien bridge, etc.) or mechanical resonator (tuning fork, vibrating reed, etc.) previously calibrated in terms of frequency, or (b) comparing the unknown with a known frequency from a separate source, either matching it directly with a variable calibrated source (heterodyne frequency meter), or measuring the difference between it and a fixed source (frequency standard) the frequency of which is known with high precision, by interpolation.

Calibrated receiver — In the absence of more elaborate frequency-measuring equipment, a calibrated receiver may be used to indicate the approximate frequency of an oscillator. If the receiver is well-made and has good inherent stability, a bandspread dial calibration can be relied on to within perhaps 0.2 per cent. Some manufactured models having factory calibration may be used to even closer limits. For most accurate measurement the oscillator should be unmodulated and maximum response in the receiver indicated by a carrier-operated tuning indicator (§ 7-13), the receiver beat-oscillator being turned off.

When checking transmitting frequency the receiving antenna should be disconnected. If the signal blocks the receiver, the transmitter frequency can be checked by listening to the oscillator, with the power amplifier turned off.

Absorption frequency meters — The simplest type of frequency meter consists of a coil and condenser, tunable over the frequency range desired (Fig. 2001). A frequency meter of this type, when tuned to the frequency of the transmitter and loosely coupled to the tank coil, will extract a small amount of energy from the tank. This energy can be used to light a small flashlight lamp. Maximum current will flow when the frequency meter is tuned exactly to the transmitter frequency; hence, the brightness of the lamp indicates resonance. A more accurate indication may be obtained by use of a thermogalvanometer or vacuum-tube voltmeter. A crystal detector and d.c. milliammeters may also be used (Figs. 2002–2003).

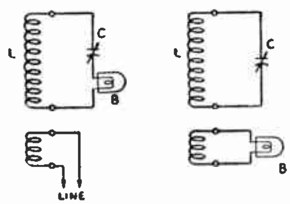


Fig. 2001 — **The simple absorption frequency meter circuit at left is used chiefly in transmitter checking, with link line coupling to the circuit being checked. Circuit at right uses a flashlight-bulb indicator loosely coupled to the tuned circuit, giving a sharper resonance point.**

Ordinary coil-and-condenser combinations cannot be used successfully at frequencies much above 200 Mc. Instead, special low-capacity circuits (Fig. 2004) or transmission lines (Figs. 2005–2007) are used.

Although this type of frequency meter is not well adapted to precise measurement of frequency, it is useful for checking (1) the fundamental frequency of an oscillating circuit, (2) presence and order of amplitude of harmonics, (3) frequency of parasitic oscillations, (4) neutralization of an amplifier, (5) field strength on a qualitative basis, (6) presence of r.f. in undesired places such as power wiring, or any other application where the detection of a small amount of r.f. and measurement of its frequency may provide useful information.

Calibration of the absorption frequency meter is most easily accomplished with a receiver of the regenerative type to which the coil in the meter can be coupled. With the detector oscillating weakly, the frequency meter

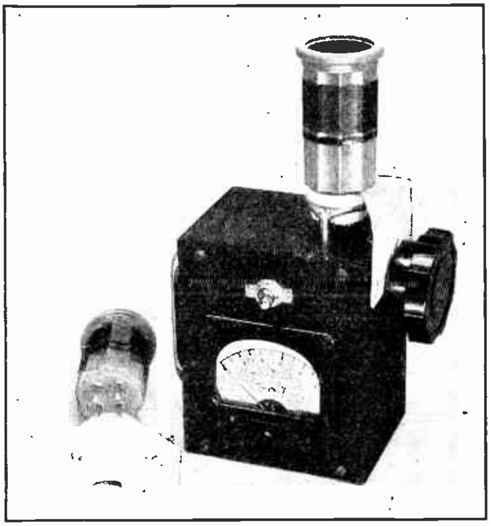


Fig. 2002 — **A sensitive absorption-type frequency meter with a crystal-detector rectifier and d.c. milliammeter indicating circuit. Individual calibration charts mounted directly on each coil form make the meter direct-reading. The toggle switch places a 10-ma. shunt across the 0–1 ma. meter; this range is used for preliminary readings, to avoid burning out meter or crystal. The meter gives indications at several feet from a low-power oscillator.**

should be brought near the detector coil and tuned over its range until a setting is found which causes the detector to stop oscillating. The coupling between meter and receiver should then be loosened until the stoppage of oscillation occurs at only one spot on the meter tuning dial. The meter is then tuned to the frequency at which the receiver is set. If the receiver is set on several stations of known frequency, a number of points for a calibration curve can be obtained for each coil.

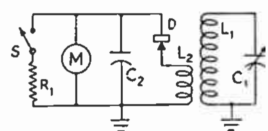


Fig. 2003 — **Indicating frequency-meter circuit diagram.**

C_1 — 140-μμfd. variable (Hammarlund HFA-140-A).
C_2 — 0.001-μfd. mica.
D — Fixed crystal detector (Philmore).
M — 0.1 d.c. milliammeter (Triplett Model 321).
R_1 — 3-ohm shunt; see general data on meter shunts.
S — S.p.s.t. toggle switch.
L_1, L_2 — Plug-in coils wound on 1½-inch diameter forms:

Frequency Range	*Wire Size*	L_1	*Length*	L_2 [1,3]
1.1–3.5 Mc.[2]	No. 28 e.	81¾	1⅞″	17 turns
2.5–8.0 Mc.[2]	No. 24 t.	37¾	1⅝″	11 "
4.5–14 Mc.[2]	No. 20 t.	17¾	1½″	6 "
7.5–25 Mc.[2]	No. 16 t.	8¾	1¼″	4 "
22–70 Mc.	No. 16 e.	2¾	1″	2 "
40–120 Mc.	No. 16 e.	¾	—	¾ "

[1] Closewound, No. 30 d.s.c., ¼-inch from primary.
[2] Available commercially (Hammarlund SWK-4).
[3] Because the impedance of individual crystal detectors varies considerably, experiment with the number of turns on L_2 is necessary for maximum current indication. If meter reads backwards, reverse crystal connections.

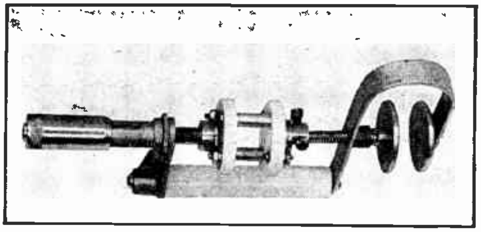


Fig. 2004 — A "micrometer" type-absorption u.h.f. frequency meter covering a range of approximately 200 to 800 Mc. The main assembly consists of the adjustment head and large pillar insulator from a National NC800 neutralizing condenser. Silver-plated brass strip, ½ inch wide, bent into a U 1½ inches wide and 2 inches long (excluding the supporting extension to the insulator) constitutes the inductance, while the variable condenser is a pair of 1⅛-inch aluminum plates. The exponential capacity curve of this type of condenser gives maximum band-spread at the high-frequency end of the range and a nearly linear frequency scale near the center.

The same method may be used with a superheterodyne receiver, but it must be remembered that the oscillator frequency differs from the signal frequency by the intermediate frequency. For instance, if the receiver dial reads 6500 kc. and the receiver i.f. is 456, the oscillator frequency will be 6956 kc., which is the frequency which should be marked on the meter calibration scale. It is necessary to know whether the oscillator is on the high or low side of the incoming signal; in most receivers it is on the high side throughout, but in some it is shifted to the low side on the high-frequency ranges.

If the oscillator coils in the receiver are not accessible, the frequency meter may be capacity coupled through a few turns of insulated wire wrapped around the frequency-meter coil, one end of the wire being placed near the stator plates of the oscillator condenser.

For transmitter frequency checking a flashlight lamp or other indicator is not entirely necessary, since resonance will be indicated by a change in the plate current of the stage being checked as the meter is tuned through resonance. However, for locating parasitic oscillations, determining amplitude of harmonics, checking neutralization, locating stray r.f. fields, etc., a sensitive indicator is indispensable.

The inherent errors in the absorption-type frequency meter ordinarily limit its useful accuracy to about 1 per cent.

Lecher wires — At very-high and ultra-high frequencies it is possible to determine frequency by actually measuring the length of the waves generated. The measurement is made by observing standing waves on a two-wire parallel transmission line or Lecher frame. Such a line shows pronounced resonance effects, and it is possible to determine quite accurately the current loops (points of maximum current). The distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read directly in meters (inches × 39.37 if a yardstick is used), or in centimeters for the very short wavelengths.

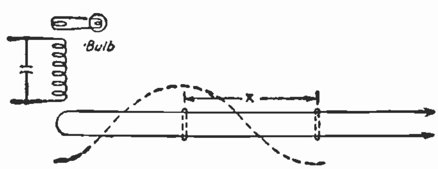


Fig. 2005 — Coupling a Lecher-wire system to a transmitter tank coil. Typical standing-wave distribution is shown by the dashed line. The positions of the shorting bar at the current loops equals one-half wavelength.

The Lecher wire line should be at least a wavelength long — that is, 9 feet or more on 112 Mc. — and should be entirely air-insulated except where it is supported at the ends. It may be made of copper tubing or of wires stretched tightly between any two convenient supports. The spacing between wires should be approximately one to one and one-half inches. The positions of the current loops are found by means of a "shorting bar," which is simply a metal strip or knife edge which can be slid along the line to vary its effective length. The system can be used more conveniently and with greater accuracy if it is built up in permanent fashion and provided with a shorting bar maintained at right angles to the wires (Figs. 2006 and 2007). The support may consist of two 12-foot pieces of "1 by 2" (actually about ¾ × 1⅝ inch) pine fastened together with wood screws to form a T girder, this arrangement being used to minimize bending of the wood when the wires are tightened.

A slider holds the shorting bar and acts as a guide to keep the wire spacing constant. A piece of wood held in the hand can be used; it is an easy matter to regulate the pressure so that free movement is secured. A spring device may be arranged for the same purpose.

For convenience in measuring lengths directly in the metric system used for wavelength rather than in inches, the supporting beam may be marked off in decimeter (10-centimeter) units. A 10-centimeter transparent scale (obtainable at 5 & 10 cent stores) may be cemented to the slider, extending out from the front, so that readings can be taken to the nearest millimeter. The difference between any two readings gives the half wavelength directly.

Making measurements — Resonance indications can be obtained in several different ways. Let us suppose the frequency of a transmitter is to be measured. A convenient and fairly sensitive indicator can be made by soldering the ends of a one-turn loop of wire, of about the same diameter as the transmitter tank coil, to a low-current flashlight bulb, then coupling the loop to the tank coil to give a moderately-bright glow. A similar coupling loop should be connected to the ends of the Lecher wires and brought near the tank coil, as shown in Fig. 2005. Then the shorting bar should be slid along the wires outward from the transmitter until the lamp gives a sharp dip in brightness. This point should be marked (a piece of string can be tied on one of the wires) and the shorting bar moved out until a second dip is obtained. Marking the second spot, the distance between the two points can be measured and will be equal to half the wavelength. If the measurement is made in inches, the frequency will be

$$F_{Mc.} = \frac{5906}{\text{length (inches)}}$$

If the length is measured in meters,

$$F_{Mc.} = \frac{150}{\text{length (meters)}}$$

A frequency of 112 Mc. corresponds to a length of slightly less than 52¾ inches (1.34 meters) and a frequency of 116 Mc. to 50²⁹⁄₃₂ inches (1.29 meters).

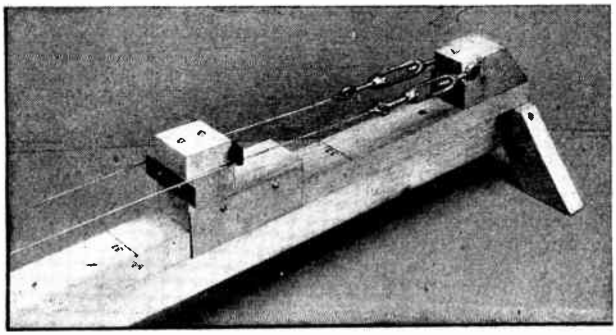


Fig. 2006 — One end of a typical Lecher-wire system. The feet at each end keep the assembly from tipping over when in use. The wires terminate in airplane-type strain insulators at one end, and at the other in small turnbuckles for maintaining tension. The wire is No. 16 bare solid copper antenna wire (hard-drawn). The turnbuckles are held in place by a 3⁄16 × 2-inch bolt through the anchor block. This end of the line is thus short-circuited; it does not matter whether it is open or shorted, since the other end is the one connected to the pick-up loop.

In checking a superregenerative receiver, the Lecher wires may be similarly coupled to the receiver coil. In this case the resonance indication may be obtained by setting the receiver just to the point where the hiss is obtained, then as the bar is slid along the wires a spot will be found where the receiver goes out of oscillation. The distance between two such spots is equal to a half wavelength.

In either case, the most accurate readings result only when the loosest possible coupling is used between the line and the tank coil. After taking a preliminary reading to find the regions along the line in which resonance occurs, loosen the coupling until the indications are just discernible and repeat the measurement. Unless this is done the tuning of the line will affect the frequency of the oscillator and inaccurate indications will be obtained. As the coupling is loosened the resonance points will become sharper, which is a further aid to accurate determination of the wavelength.

The shorting bar must be kept at right angles to the two wires. A sharp edge on the bar is desirable, since it not only helps make good contact but also definitely locates the *point* of contact.

The accuracy with which frequency can be measured by such a system depends principally upon the technique of measurement. The necessity for using very loose coupling to the transmitter or receiver has already been mentioned. In addition, careful measurement of the exact distance between two current loops also is essential. Even if all other sources of error are eliminated, measurements within 0.1 per cent require an accuracy within 1 part in 1000, or 1 millimeter in one meter, in measuring the distance along the wires. This means that an accurate standard of length is necessary — a good steel tape, for instance — and that care must be used in determining the length exactly.

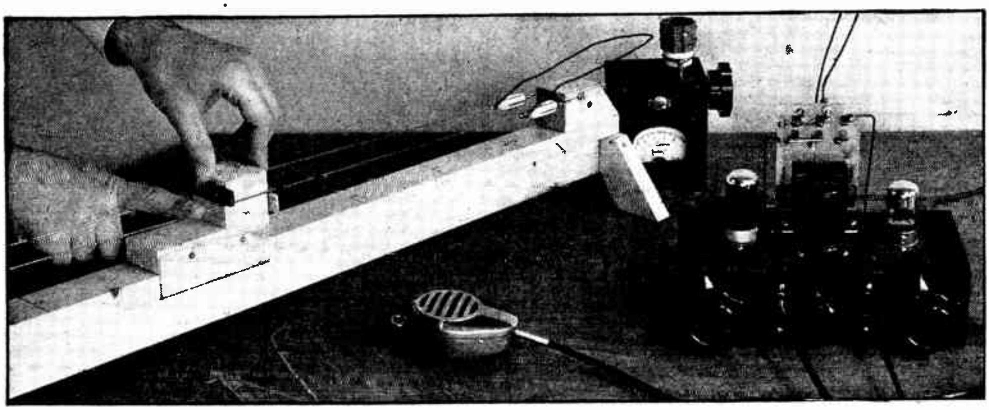


Fig. 2007 — A Lecher-wire system set up for frequency measurement, using a crystal-detector absorption frequency meter, loosely coupled to the oscillator tank, as a resonance indicator. Because only very loose coupling to the oscillator is required, this system will give more accurate results than coupling the wires directly to the transmitter tank. The shorting bar is of brass with a sharp edge for better contact and more precise indication; the wooden slider keeps it at right angles to the wires. A horizontal strip of bakelite at the back of the slider keeps the wires tight against the shorting bar. Sheet metal pieces screwed to the sides of the sliding block are bent under the horizontal member of the T to keep the block in place. At the back a horizontal strip of bakelite keeps the wires pressed close to but not actually touching the shorting bar, being pressed down on the bar only when a reading is taken.

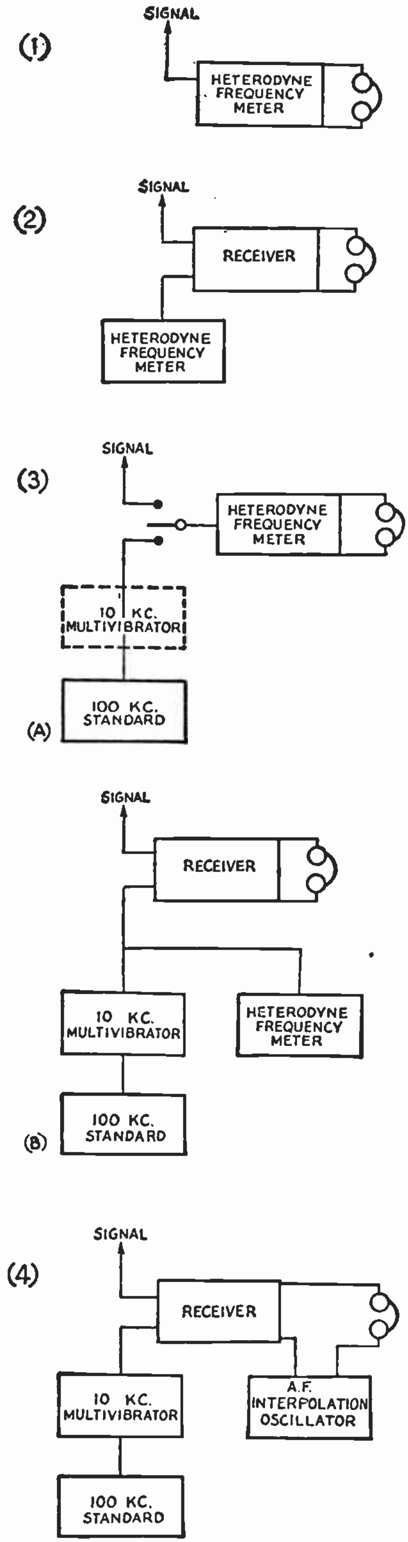


Fig. 2008 — Frequency measurement methods.

Comparison methods — For more accurate frequency measurements, beat-frequency comparison methods must be used. Typical combinations capable of varying degrees of accuracy are illustrated in the block diagram of Fig. 2008.

1) The simplest method for checking the frequency of local oscillators is the use of a calibrated heterodyne frequency meter (with a built-in detector or mixer) and with a pair of headphones as a zero-beat indicator.

2) By using a receiver in conjunction with the heterodyne frequency meter, incoming received signals can be measured.

3) For greater accuracy, the heterodyne frequency meter may be used as a linear r.f. interpolation oscillator in conjunction with a 100-kc. standard (with or without a 10-kc. multivibrator as a frequency divider). The fixed harmonics from the standard provide accurate check or reference points on the calibrated frequency meter scale.

A receiver (if adequately calibrated) with an internal b.f.o. may be substituted for the frequency meter as an interpolation oscillator. Alternatively, it may be used (non-oscillating) in conjunction with the heterodyne meter for measuring received signals (3-B).

4) Although ordinary amateur practice does not require greater accuracy than is possible with method (3). even more precise measurements can be made by the use of a calibrated audio oscillator used in the same manner as an r.f. interpolation oscillator; it is set at zero beat with the beat-note resulting from the combination of the unknown signal frequency with the nearest 10-kc. multivibrator harmonic.

Using careful design and construction, accuracies of 0.01 per cent (100 parts in a million) can be attained with methods (3) A and B. Method (4) is accurate to 10 parts in a million with ordinary equipment; when precision laboratory apparatus is used it is reliable to better than 1 part in a million.

For greatest accuracy the 100-kc. reference standard should be calibrated on the WWV transmissions (page 408). These transmissions may be tuned in on a receiver (beat oscillator off) and the standard adjusted until its harmonic is exactly at zero beat with WWV. The calibration should be rechecked each time a precision measurement is to be made.

Heterodyne frequency meters — For more accurate measurement of transmitter frequency, a heterodyne frequency meter is used. This is a small, completely shielded oscillator with a precise frequency calibration covering the lowest frequency band in use. It must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

The signal from this oscillator (or a harmonic thereof) is fed into a receiver or simple detector together with the signal to be measured, and the two frequencies are heterodyned. When the frequency-meter oscillator is tuned to zero beat with the signal, its frequency or the harmonic multiple is the same as the unknown frequency, and the latter therefore can be read directly from the frequency-meter dial.

The oscillator used in the frequency meter must be very stable. Mechanical considerations are most important in its construction. No matter how good the instrument may be electrically, its accuracy cannot be depended upon if it is flimsily built. Inherent frequency stability can be improved by avoiding the use of phenolic compounds and thermoplastics (bakelite, polystyrene, etc.) in the oscillator circuit, and employing only high-grade ceramics for insulation. Plug-in coils or switches ordinarily are not acceptable; instead, a solidly built and firmly mounted tuned circuit should be permanently installed and the oscillator panel and chassis reinforced for rigidity.

To obtain high accuracy, the frequency meter must have a dial which can be read precisely to at least one part in 500; ordinary dials such as are used for transmitters and inexpensive receivers are not capable of such precision without the addition of vernier scales. Select a dial which has fine lines for division marks and an indicator close to the dial scale, so that the readings will not appear different because of parallax when viewed from different angles.

A stable oscillator circuit suitable for use in a heterodyne frequency meter is the electron-coupled circuit (§ 4-2). The oscillation frequency is practically independent of moderate variations in supply voltages, provided the plate and screen voltages are properly proportioned, and it is possible to take output from the plate with but negligible effect on the frequency of the oscillator. A third feature is that strong harmonics are generated in the plate circuit. A typical electron-coupled frequency meter is shown in Figs. 2009–2010.

When the frequency meter is first turned on some little time is required for the tube to reach its final operating temperature; during this period the frequency of oscillation will drift slightly. Although the drift will not amount to more than two or three kilocycles on the 3500-kc. band and proportionate amounts on the other bands, it is desirable to allow the frequency meter to "warm up" for about a half hour before calibrating or before making measurements in which utmost accuracy is desired. Better still, it may be left on permanently. The power consumption is negligible, and the long-time stability will be vastly improved.

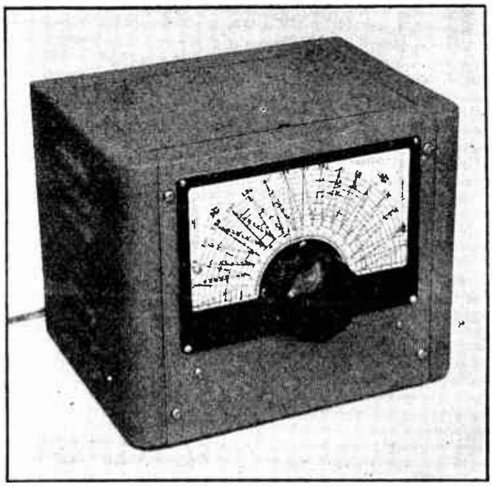
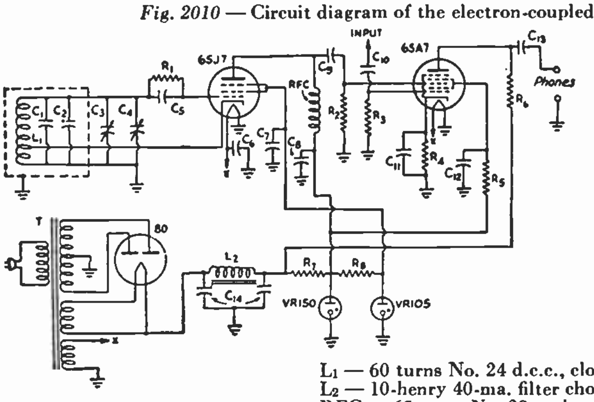


Fig. 2009 — Electron-coupled heterodyne frequency meter with harmonic amplifier and voltage regulator. The direct-reading dial is calibrated for every 10-kc. point from 1750 to 1900 kc. Axial lines passing through these points are intersected by ten semi-circular subdivision lines. Diagonal lines connecting the ends of adjacent 10-kc. lines, in conjunction with the subdivisions, enable reading the scale accurately to 1 kc. or better.

Although some frequency drift is unavoidable, it can be minimized by the use of voltage-regulator tubes in the power supply and low-drift silvered-mica or zero temperature-coefficient fixed condensers in the tuned circuit. A small negative temperature-coefficient capacity may be included to compensate for residual drift.

Calibration of the frequency meter is readily accomplished if a low-frequency standard (discussed later in this chapter) is available, the required calibration points being supplied by

Fig. 2010 — Circuit diagram of the electron-coupled heterodyne frequency meter.



C_1 — 350-μμfd. zero-drift.
C_2 — 40-μμfd. negative-temperature.
C_3 — 100-μμfd. midget.
C_4 — 50-μμfd. trimmer.
C_5 — 100-μμfd. silver-mica.
C_6, C_8 — 0.005-μfd. mica.
C_7, C_9, C_{10} — 0.002-μfd. mica.
C_{11} — 25-μfd. 25-volt electrolytic.
C_{12} — 0.01-μfd. 400-volt paper.
C_{13} — 0.5-μfd. 400-volt paper.
C_{14} — Dual 8-μfd. 450-volt electrolytic.
R_1, R_2 — 1 megohm, ½-watt.
R_3 — 200,000 ohms, ½-watt.
R_4 — 150 ohms, ½-watt.
R_5 — 5000 ohms, 1-watt.
R_6 — 50,000 ohms, 2-watt.
R_7 — 3500-ohm, 10-watt wire-wound.
R_8 — 2500-ohm, 5-watt wire-wound.
L_1 — 60 turns No. 24 d.c.c., close-wound, 1¼-inch diameter, tap at 12 turns.
L_2 — 10-henry 40-ma. filter choke.
RFC — 65 turns No. 28 e. close-wound, 9/16-inch diameter.
T — 300-volt, 50-ma. power transformer.

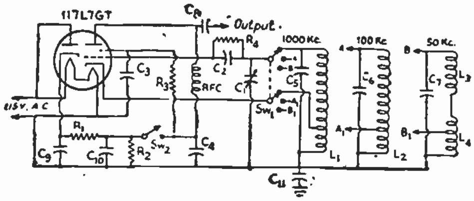


Fig. 2011 — Wiring diagram of a 50-, 100- and 1000-kc. electron-coupled frequency standard.

C_1 — 100-$\mu\mu$fd. midget variable (Hammarlund HF-100).
C_2 — 250-$\mu\mu$fd. mica.
C_3, C_4, C_{11} — 0.02-μfd. 400-volt paper.
C_5 — 500-$\mu\mu$fd. mica.
C_6 — 0.001-μfd. mica.
C_7 — 0.002-μfd. mica.
C_8 — 50-$\mu\mu$fd. mica.
C_9, C_{10} — 8-μfd. 450-volt electrolytic.
R_1, R_2 — 50,000 ohms, 1-watt.
R_3, R_4 — 0.1 megohm, 1-watt.
RFC — 2.5-mh. r.f. choke.
S_1 — 3-position 2-circuit rotary switch (Mallory 3223J).
S_2 — S.p.s.t. toggle switch.
L_1 — 100 turns No. 34 d.c.c. close-wound on 9/16-inch form. Cathode tap at 30th turn from ground.
L_2 — 2.5-mh. r.f. choke. Cathode tap is between first and second pies from ground.
L_3 — 2.5-mh. r.f. choke, at right angles to L_4.
L_4 — 2.5-mh. r.f. choke. Cathode tap between second and third pies from ground.

NOTE — Because of manufacturing tolerances in r.f. chokes and condensers, additional capacity may be required on the 50- and 100-kc. ranges. If C_1 does not tune to desired frequency, add 100 $\mu\mu$fd. to C_6 or C_7 as required. For additional output, decrease R_1 to 25,000 ohms. All d.c. connections must be isolated from the chassis.

harmonics from the standard. The frequency meter is tuned to zero beat with these harmonics, using either a built-in detector or the station receiver to combine the two signals to provide an audible beat. When a sufficient number of points have been established they may be marked on graph paper, and a calibration curve drawn. For maximum convenience, a direct-reading dial scale can be constructed.

If no frequency standard is available, calibration points may be obtained from other sources of known frequency, such as the transmitter crystal oscillator, harmonics of local broadcasting stations, etc. As many such points as possible should be secured, so that individual inaccuracies will average out.

In use, the signal from the frequency meter is fed into the receiver through a lead from the plate of the oscillator through a very small capacity to the input of the receiver. The signal to be measured is then tuned in as usual and the frequency meter adjusted to zero beat.

For convenience in checking the frequency of the transmitter or other local oscillators which generate sufficiently strong signals, it is desirable to incorporate a detector in the frequency meter which will combine the signals and deliver the audio beat-note output to headphones or to a visual zero-beat indicator. A frequency-converter tube such as the 6L7 or 6SA7 is especially suited for this purpose.

With a stable oscillator, a precision dial and frequent and careful calibration, an over-all accuracy of 0.05 to 0.1 per cent may be expected of the heterodyne frequency meter. The principal limiting factors are the precision with which the calibrated dial can be read and the "reset" stability of the tuned circuit.

Frequency standards — To make more precise frequency measurements, particularly of amateur-band limits, a secondary frequency standard is required. This is a highly stable low-frequency oscillator, usually operated at 50 or 100 kc., the harmonics of which are used to provide reference points every 50 or 100 kc. throughout the spectrum. Since all amateur band edges fall at multiples of these frequencies, it is possible to establish band limits with extreme accuracy. A 1000-kc. frequency is often added to facilitate preliminary identification of frequency ranges, especially on v.h.f.

An electron-coupled oscillator built according to the principles previously outlined for frequency meters, with a tuned circuit for 50 or 100 kc., will serve as a simple and inexpensive standard. A typical circuit for such a unit is shown in Fig. 2011. A standard of this type is inherently more accurate than a heterodyne frequency meter because (a) the lower-frequency oscillator has better inherent stability and (b) the frequency setting once made is not thereafter changed, eliminating the reset and calibration errors.

Even better long-time stability can be obtained from a crystal-controlled oscillator of the type shown in Fig. 2012. Special X-cut crystals are available which oscillate at either of two frequencies, determined by the thickness (1000 kc.) and length (100 kc.). Either may be selected by a switch which connects in a tuned circuit resonant at the desired frequency. A parallel trimmer across the crystal will permit adjusting the frequency precisely to 100 kc. (provided natural crystal frequency is on the high side).

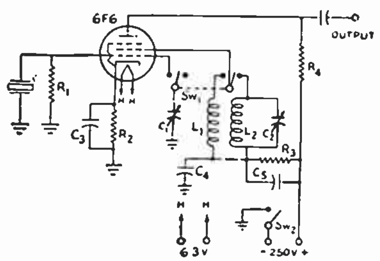


Fig. 2012 — Circuit diagram of a dual-frequency 1000–100-kc. crystal-controlled secondary frequency standard.

C_1 — 35-$\mu\mu$fd. midget variable (Hammarlund HF-35).
C_2 — 100-$\mu\mu$fd. mica trimmer (Hammarlund CTS-85).
C_3, C_4, C_5 — 0.1-μfd. 400-volt paper.
C_6 — 0.001-μfd. midget mica.
R_1 — 5 megohms, ½-watt.
R_2 — 500 ohms, ½-watt.
R_3 — 25,000 ohms, 1-watt.
R_4 — 0.25 megohm, ½-watt.
L_1 — 8-mh. r.f. choke (Meissner 1920-78).
L_2 — 2.5-mh. r.f. choke (all but one pie removed).
S_1 — D.p.d.t. toggle switch.
S_2 — S.p.s.t. toggle switch.
Crystal — Bliley SMC-100.

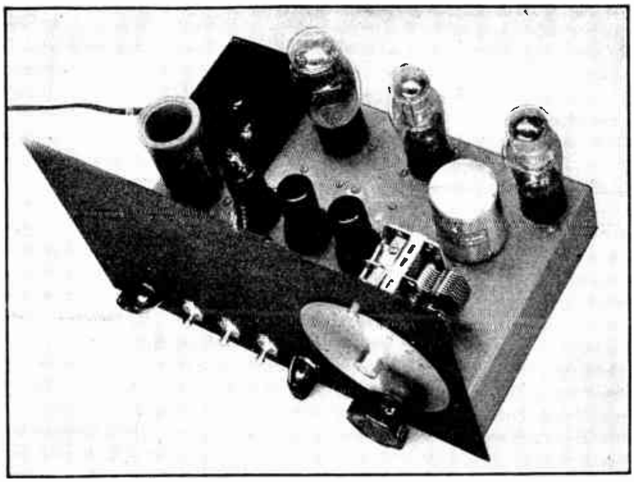


Fig. 2013 — A secondary frequency standard, incorporating a 100-kc. low-drift crystal oscillator, a 10-kc. multivibrator, and a harmonic amplifier-modulator. The vernier dial is used for precise setting of crystal frequency. Controls along the bottom are, left to right: output tuning, C_{14}; on-off switch, S_1; "B" switch, S_2; multivibrator switch, S_3, and multivibrator control, R_8. Power transformer, rectifier and regulator tubes are along the rear edge of the 7 × 12-inch chassis. The crystal oscillator is at the right, multivibrator tube in the center, and output circuit at the left. The output circuit is tuned to the band in use, with output taken either through C_{17} or a link winding. Output coupling is adjusted to give desired signal strength in the receiver. The crystal frequency can be adjusted to precisely 100 kc. by the vernier dial controlling C_1. Switching the multivibrator section on or off will cause a frequency change of less than 1 part in a million.

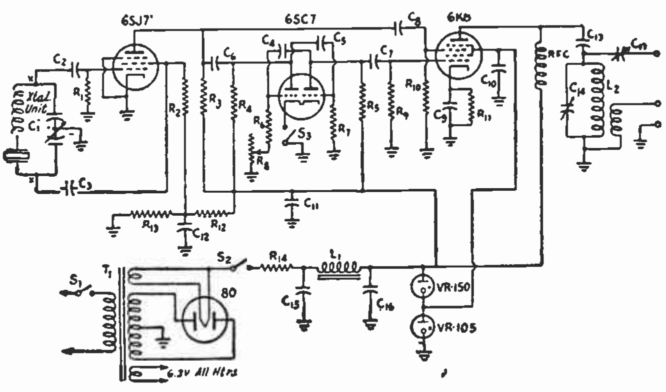
For highest accuracy in frequency measurement and calibration, the most suitable instrument is a precision crystal-controlled secondary standard, provided with a multivibrator (§ 3-7) for frequency division (Figs. 2013–2014) to mark 10-kc. intervals throughout the communications spectrum. The frequency of a signal can then be checked by noting its location with respect to two adjacent 10-kc. points on the dial of a calibrated receiver or heterodyne frequency meter and estimating the exact frequency by interpolation.

Before adjusting a frequency standard, at least a 15-minute warm-up period should be allowed. For initial adjustment, couple its output into a broadcast receiver and adjust the oscillator to zero beat with a b.c. station operating on a multiple of 100 kc. If the oscillator is self-excited a second station 100 kc. away should be checked, to make sure it is working on 50 or 100 kc. rather than another frequency which gives an odd harmonic. Since broadcasting stations are required to stay within 20 cycles of assigned frequency, the maximum error will be less than 30 parts in one million.

In adjusting the multivibrator, two adjacent 100-kc. points are first noted on the dial of a calibrated receiver. The multivibrator is then turned on, and its frequency control (R_8 in Fig. 2014) set at half-scale. The number of separate audio beats *between* the two marked 100-kc. points is then counted. If it is a number other than nine (indicating 10-kc. intervals), readjust the frequency control until nine beats are observed. Mark this point. Note also the points on the scale where 8 and 10 beats occur, indicating approximately 11- and 9-kc. separation. The odd frequencies are occasionally useful in checking frequencies very close to the 10-kc. harmonics where the low beat-frequency makes it difficult to secure zero-beat, particularly when an interpolation oscillator is used.

The 100-kc. points usually can be identified because they are louder than the 10-kc. harmonics. This identification can be facilitated by applying audio modulation to the 100-kc.

Fig. 2014 — Circuit diagram of the precision low-drift crystal-controlled 100-kc. secondary frequency standard.



C_1 — Dual 365-μμfd. variable.
C_2, C_3 — 0.01-μfd. 400-volt paper.
C_4, C_5 — 0.001-μfd. midget mica.
C_6, C_7 — 10-μμfd. midget mica.
C_8 — 50-μμfd. midget mica.
C_9, C_{10}, C_{11}, C_{12} — 0.1-μfd. 400-volt.
C_{13} — 0.002-μfd. midget mica.
C_{14} — 140-μμfd. variable.
C_{15}, C_{16} — 8-μfd. 450-volt electrolytic.
C_{17} — 3–30-μμfd. mica trimmer.
R_1 — 1 megohm, ½-watt.
R_2, R_3 — 0.5 megohm, 1-watt.
R_4, R_5 — 50,000 ohms, 1-watt.
R_6, R_7 — 20,000 ohms, ½-watt.
R_8 — 15,000-ohm potentiometer.
R_9 — 0.3 megohm, ½-watt.
R_{10} — 0.1 megohm, ½-watt.
R_{11} — 800 ohms, ½-watt.
R_{12} — 25,000 ohms, 1-watt.
R_{13} — 50,000 ohms, 1-watt.
R_{14} — 1500 ohms, 10-watt.
RFC — 2.5 mh. r.f. choke.
S_1, S_2, S_3 — S.p.s.t. toggle.
T_1 — Power transformer, 250 volts, 40 ma.
L_1 — 7-henry, 40-ma. filter choke.
L_2 — Plug-in coil for band in use.

The crystal is a Bliley SOC-100 (supplied complete with oscillator coil in same mounting). For checking 1000-kc. points, a 150 microhenry coil (75 turns of No. 30 d.c.c. on a 1½-inch form) may be substituted for the crystal unit, connected between points X-X in the diagram. With C_1 near maximum capacity, the oscillator circuit will tune to 1000 kc.

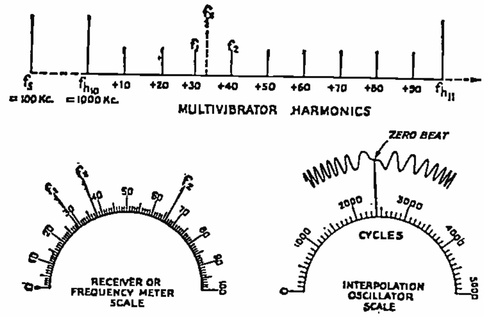


Fig. 2015 — Use of interpolation methods in measuring frequencies between standard harmonics. At the top is shown the relative location of the frequency-standard fundamental and harmonics in the spectrum, together with the multivibrator harmonics, as related to the unknown frequency under measurement (f_x). At the left is shown a small segment of this spectrum as it appears on the dial of a calibrated receiver or heterodyne frequency meter, and at right the appearance of the audio oscillator dial when using the comparison audio beat-note method.

signal only, causing the modulated points to stand out because of the distinctive tone.

At v.h.f. and u.h.f., the 100 kc. harmonics may be difficult to identify. In this region, therefore, a higher standard oscillator frequency — 1 Mc. or 10 Mc. — is followed by a chain of intermediate frequency dividers. For amateur use, a low-drift 7-Mc. crystal and an assymetrical 1-Mc. multivibrator constitute a useful combination.

Interpolation — When measuring exact frequencies with the aid of a frequency standard and multivibrator providing equi-spaced harmonic points, it is necessary to determine the exact location of the unknown frequency by interpolation between adjacent standard harmonics. This can be done (a) by use of a calibrated receiver or heterodyne frequency meter with a scale which is linear with frequency (Fig. 2008–3), or (b) by comparison of the audio beat frequency with a calibrated audio oscillator (Fig. 2008–4).

In method (a), the points at which the unknown frequency and the nearest lower and higher harmonics appear on the dial of the receiver or frequency meter are noted, as shown in Fig. 2515. Knowing the exact frequencies of the harmonic points, f_1 and f_2, the unknown frequency, f_x, can be determined as follows:

$$f_x = f_1 + \frac{S_x - S_1}{S_2 - S_1}(f_2 - f_1)$$

where S_1 is the dial setting for f_1, S_2 for f_2 and S_x for f_x.

Method (b) consists of combining the standard and unknown frequencies in a detector and measuring the resulting audio beat frequency by zero-beating with a calibrated audio oscillator having a linear frequency range covering half the difference between adjacent harmonics (0–5000 cycles with a 10-kc. multivibrator), as shown in Fig. 2015. The measured frequency is then equal to the reading of the audio oscillator, added to or subtracted from the nearest standard harmonic. To determine whether to add or subtract this audio difference, it is necessary only that the frequency under measurement be known within 5 kc. from the receiver (or auxiliary oscillator) calibration.

In addition to the beat note resulting from the nearest adjacent harmonic, f_1, there will also be another higher beat from f_2. However, by tuning the receiver midway between f_1 and f_x, its adjacent-channel selectivity will discriminate against f_2 and reduce the higher beat note to a negligible level.

The interpolation audio oscillator should have a scale which reads linearly with frequency (as opposed to the logarithmic scale commonly found in laboratory oscillators). A beat-frequency oscillator with a variable tuning condenser of suitable plate shape in series with the correct value of fixed capacity will have such a scale. A well-made resistance-capacity oscillator also has a nearly linear scale.

A suitable detector-mixer is a pentagrid frequency tube converter (§ 7-9) with some form of zero-beat indicator in the plate circuit. The interpolation-oscillator output is connected to the oscillator grid and the audio output from the receiver to the signal grid.

Zero-beat indicators — Use of the heterodyne method of frequency comparison requires a means for determining when the known and unknown frequencies are synchronized; i.e., when they are at zero beat. The point at which zero beat occurs can be determined approximately by listening to the output of the receiver or detector in the headphones or loudspeaker. For greatest accuracy, however, some form of auxiliary visual zero-beat indicator is desirable. This may be a rectifier-type a.f. voltmeter with a copper-oxide or diode rectifier (§ 2-3), a neon-tube "flasher," or an electron-ray tube (§ 7-13) with its triode grid connected to the receiver output. Headphones usually are required for preliminary adjustments, since most visual indicators respond only to frequencies of less than about 25 cycles.

A simple beat indicator of the "magic eye" tube type, which not only affords an exact zero-beat indication but gives a visual comparison of any two frequencies nearly zero beat, may be obtained by combining the two inputs in the triode grid of the electron-ray tube (Fig. 2016). The input voltages should be adjusted so that the "eye" of the tube normally closes to about 45°.

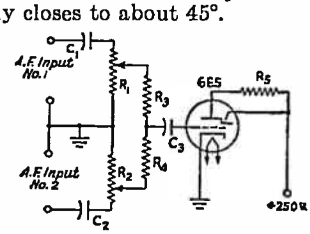


Fig. 2016 — A visual zero-beat indicator circuit.
C_1, C_2 — 0.1-μfd. paper. C_3 — 0.01-μfd. paper.
R_1, R_2 — 0.5-megohm volume control.
R_3, R_4 — 0.25 megohm. R_5 — 1 megohm.

Audio-frequency measurement — The measurement of unknown audio frequencies can be accomplished by a number of direct comparison, bridge and counting methods. Laboratory measurements commonly are made with the calibrated a.c. bridge (§ 2–11).

The Wien bridge is a simple and satisfactory instrument for measuring audio frequencies with good accuracy because it contains no inductance to pick up stray magnetic fields and can be used over a wide frequency range. In the circuit of Fig. 2017, R_1 and R_2 are identical variable resistors ganged on a common shaft. The dial can be calibrated directly in frequency. Decimal multiplying factors are obtained merely by changing the capacities in the circuit by a three-position range switch. Since a frequency range of 10 to 1 can be covered with a single value of capacity, the use of three pairs of condensers enables coverage of the complete audio range from 20 to 20,000 cycles.

In practice it is impossible to maintain equality between the resistances of R_1 and R_2 with the degree of accuracy required for perfect balance. The compensating potentiometer, R_5, assists in sharpening the balance with but little effect upon the frequency calibration.

In the middle frequency range (200 to 2000 cycles) balance is most readily attained by the use of headphones. On the other frequency ranges an indicating meter, such as a sensitive vacuum-tube voltmeter, is desirable.

The unknown frequency is preferably applied through an electrostatically shielded isolating transformer. Accuracy will be impaired if harmonics are present. Since the bridge is unbalanced for these harmonics they appear prominently in the output, even though representing only a small percentage of the input. With headphones, in the middle audio range it is usually possible to balance the bridge for the fundamental despite the presence of harmonics. When an indicating instrument is employed, however, a low-pass filter may be required in the output of the neutral arm.

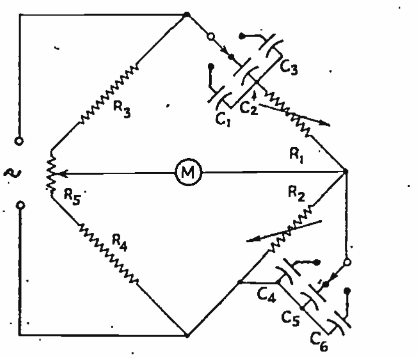


Fig. 2017 — Wien bridge circuit for a.f. measurements.
C_1, C_4 — 1.0-μfd. 200-volt paper.
C_2, C_5 — 0.1-μfd. 400-volt paper.
C_3, C_6 — 0.01-μfd. mica.
R_1, R_2 — 10,000-ohm wire-wound variable (logarithmic taper).
R_3, R_4 — 5000-ohm ½-watt wire-wound.
R_5 — 200-ohm wirewound variable.
M — Vacuum-tube voltmeter or equivalent (see text).

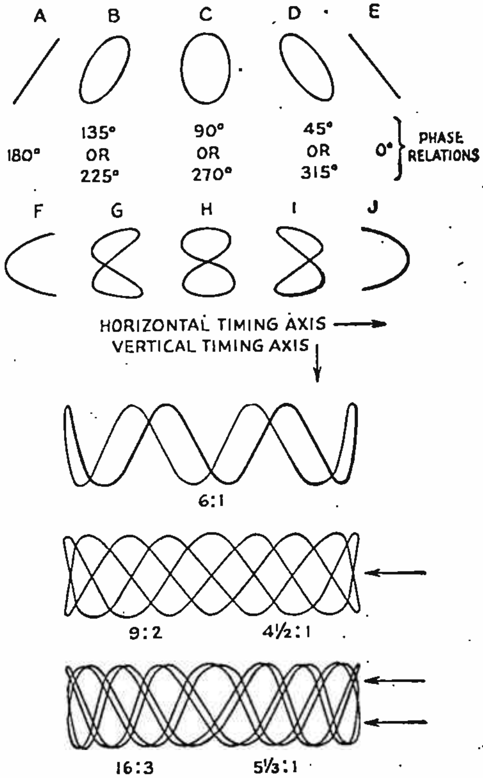


Fig. 2018 — Lissajou's figures as used in measuring audio frequencies by comparison with a known source on a cathode-ray oscilloscope. Figures A through E illustrate the pattern produced by different phase relationships when the two voltages have a 1:1 frequency ratio. Figures F through J show the same phase relationships with a 2:1 frequency ratio, the higher frequency being applied to the vertical plates. The next figure shows a ratio of 6:1, determined by counting the peaks of the waves in the horizontal plane (in this instance the higher frequency is applied to the horizontal plates). Complex ratios are identified by one or more cross-overs, as indicated by the arrows opposite the 9:2 and 16:3 figures. In principle, frequency ratios are determined by counting both horizontal and vertical peaks (number of cross-overs plus 1). Care must be taken not to confuse the back lines (return trace shown by light line in 6:1 figure) in counting cross-overs. This can be done by counting only those peaks which travel in the same direction across the screen when the frequency ratio is so adjusted that the pattern rotates slowly.

The cathode-ray oscilloscope (§ 3-9) is extremely useful in measuring frequencies by the comparison method when a reliable standard source is available. Applying the unknown and standard frequencies to opposite pairs of cathode-ray tube deflecting plates will result in images termed Lissajou's figures. By proper interpretation of these figures, forms of which are shown in Fig. 2018, frequency ratios up to 10 to 1 can be identified conveniently. Thus, with a 1000-cycle oscillator, calibration points between 100 and 10,000 cycles are available. A 60-cycle a.c. source similarly can be used up to 600 cycles or higher.

When the frequency ratio is greater than about 10 to 1, the pattern becomes so complex that it is difficult to count the cross-overs accurately. A method which yields clearer figures and thus is easier to interpret at high ratios is to apply the standard frequency, f_S, to both pairs of plates through a phase-splitting circuit consisting of capacity and resistance in series, while the unknown frequency, f_X, is connected in series with the anode No. 2 supply, as shown in Fig. 2019. A modified pattern is then obtained, the lower frequency being caused to produce an elliptical or circular trace by the phase-splitting circuit, C_1 and R_4, while the higher frequency modulates this circular low-frequency pattern by its introduction into the anode No. 2 circuit. The result is a "gear-wheel"-shaped trace (provided the one frequency is an even integer of the other). The ratio of the two frequencies is then determined by counting the number of teeth. If the frequency ratio is odd, two gear-wheel patterns will appear superimposed; in this case the teeth of both must be counted and the result divided by 2.

An alternative method is to apply the higher frequency in series with the control-grid circuit of the cathode-ray tube. This produces a "spot-wheel" pattern by modulating the light intensity; frequency ratios are determined by counting the spots in the same manner as above.

In either case the patterns appear stationary only when the two frequencies have an exact harmonic ratio. At other than integral ratios the figure will appear to rotate slowly in one direction or the other.

Where a calibrated audio oscillator is available, it may be used as a comparison standard as described above. If no electrical frequency standard is available, the audio frequency can be converted into sound through a power amplifier and loudspeaker and then measured by aural comparison with a properly tuned piano, remembering that middle C is 256 cycles and each octave above or below doubles or halves the frequency. Intermediate points can be obtained by multiplying each successive half-note above C in any octave by 1.05946 (e.g., if C is 1, C# equals 1.05946, D equals 1.1225, etc.).

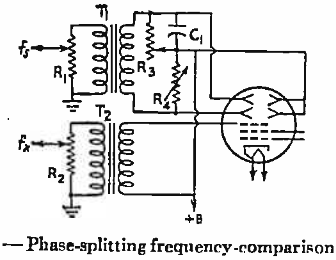


Fig. 2019 — Phase-splitting frequency-comparison circuit.
C_1 — 0.5-μfd. 400-volt paper.
R_1, R_2 — 0.5-megohm variable.
R_3 — 0.25 megohm, ½-watt.
R_4 — 25,000-ohm variable.
T_1, T_2 — Interstage audio transformers, 6:1 ratio.

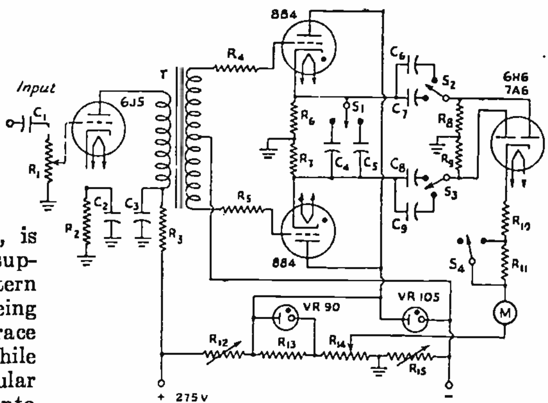


Fig. 2020 — A direct-reading audio-frequency meter.
C_1 — 0.25 μfd. 200-volt paper.
C_2 — 50-μfd. 50-volt electrolytic.
C_3 — 16-μfd. 300-volt electrolytic.
C_4 — 0.002-μfd. mica.
C_5 — 0.05-μfd. mica or paper.
C_6, C_8 — 500-μμfd. mica.
C_7, C_9 — 0.02-μfd. mica.
R_1 — 0.5 megohm, ½-watt.
R_2, R_{10} — 1000 ohms, ½-watt.
R_3 — 5000 ohms, ½-watt.
R_4, R_5 — 0.15 megohm, ½-watt.
R_6, R_7, R_8, R_9 — 3000 ohms, ½-watt.
R_{11} — 10,000 ohms, ½-watt.
R_{12} — 10,000-ohm 10-watt semi-variable.
R_{13} — 20,000 ohms, 2-watt.
R_{14} — 500-ohm wire-wound.
R_{15} — 2500-ohm wire-wound variable.

Direct-reading a.f. meter — A number of circuits have been devised by means of which thermionic or gas-discharge tubes are used in trigger- or counter-type circuits (§ 3-10) to measure the frequency of a.c. Instruments of this type have been constructed for use at frequencies as high as 100 kc. and are capable of accuracies of 3 per cent or better.

The circuit of such an instrument is shown in Fig. 2020. It consists essentially of an electronic counter circuit and an indicator. When an alternating voltage is applied to the grids of the 884 gaseous discharge tubes through the 6J5 input amplifier, each tube becomes alternately conducting and nonconducting. At each transition of the current from one tube to the other a current pulse traverses the indicator circuit, charging a condenser on one half of each cycle and discharging it on the next. The average (d.c.) value of current flowing into the condenser depends upon the number of these pulses per second. This charging current flows through a 6H6 or 7A6 (the latter type is preferred) diode rectifier and the milliammeter, M. The push-pull diode circuit prevents reverse current flowing through the meter.

Since the average d.c. current in the indicating circuit is directly proportional to the number of pulses received per second, the reading of the indicating meter, if the range limits are properly calibrated, is directly proportional to the frequency of the input voltage. The upper frequency limit of such an instrument, using gaseous tubes, is about 50 kc.

Audio-frequency signal generator — In Fig. 2021, a Wien bridge frequency-selective feed-back circuit is applied to a resistance-capacity audio oscillator of relatively simple design which features stability of frequency and amplitude, immunity from the effects of variations in supply voltage and tube characteristics, rapid warm-up time, simplicity in circuit design and construction, low cost, and a reasonably linear tuning scale.

In principle, the oscillator consists of a resistance-coupled amplifier back-coupled through the frequency-selective Wien bridge. Both positive and negative feed-back are obtained by connecting one output terminal to the grid of the 6SK7 and the other terminal to the cathode. One of the fixed arms comprises the cathode-bias resistor of this tube. One input terminal is grounded, while the other is connected through a blocking condenser to the plate of the 6ST7. At all frequencies except that for which the bridge is balanced, the negative feed-back exceeds the positive; thus, by increasing the positive feed-back slightly, the circuit can be made to oscillate. Oscillation cannot occur at any frequency other than the balance frequency, however.

The frame of the four-gang variable condenser, C_2C_3 and C_4C_5, is at grid potential, the resistance to ground being of the order of megohms, and so it must be carefully insulated from ground by ceramic insulators with long leakage paths. Otherwise the calibration will be affected by subsequent changes in the leakage resistance of this mounting. For maximum linearity at all frequencies, leakage resistance and stray capacities must be kept to a minimum. A ceramic-type insulated coupling should be used between the variable condenser and the control dial. All circuit elements associated with the first two tubes, including the gang condenser, should be in a shielded compartment.

Since the frame of the variable condenser is at a.f. potential, it is necessary to balance its comparatively large capacity to ground by connecting a condenser, C_1, across the upper section (C_2C_3 in Fig. 2021). The effective capacities of the two sections may be made approximately equal over the tuning range by adjustment of this balancing condenser.

Provided the pairs of resistors for each range (R_1R_6, R_2R_7, etc.) are accurately matched, adjustment of the small trimmers built into each variable-condenser section will complete the balance sufficiently accurately for satisfactory waveform and amplitude over the full extent of each range. The setting of the feed-back control, R_{11}, will be found somewhat critical; the optimum setting for good waveform and constant amplitude throughout each tuning range should be determined with the aid of an oscilloscope. The balancing condenser, C_1, can be set at various values while the other adjustments are being made until the performance on all ranges is found to be satisfactory.

The oscillator proper delivers sufficient voltage output to realize the full power capabilities of the 6V6. The 6SJ7 serves only as an isolating or buffer amplifier.

Automatic amplitude control is obtained by rectifying the output in the diode portion of the 6ST7 and applying the resulting d.c. bias to the 6SK7. The output amplitude is held constant within 1 db. throughout the audio-frequency range. On the highest range the shunting effect of stray capacities causes appreciable falling off.

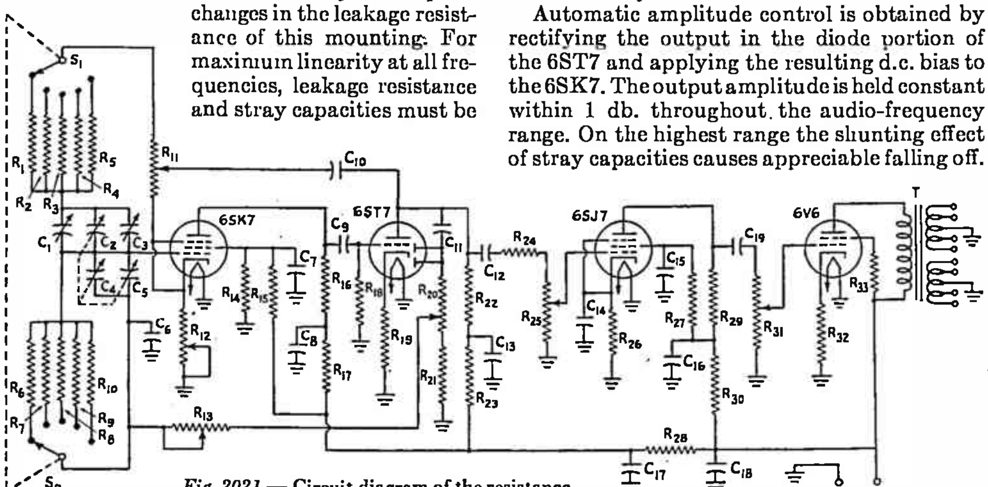


Fig. 2021 — Circuit diagram of the resistance-capacity audio-frequency signal generator.

- C_1 — 100-μμfd. air trimmer.
- C_2–C_5 — 410-μμfd. per section four-gang variable.
- C_6, C_8, C_{10}, C_{13}, C_{16} — 1.0-μfd. 400-volt paper.
- C_7, C_9, C_{11}, C_{12}, C_{15}, C_{19} — 0.1-μfd. 400-volt paper.
- C_{14} — 50-μfd. 50-volt electrolytic.
- C_{17}, C_{18} — 16-μfd. 450-volt electrolytic.
- R_1, R_6 — 15 megohms, ½-watt.
- R_2, R_7 — 2 megohms, ½-watt.
- R_3, R_8 — 0.25 megohm, ½-watt.
- R_4, R_9 — 40,000 ohms, ½-watt.
- R_5, R_{10} — 7500 ohms, ½-watt.
- R_{11} — 10,000-ohm variable.
- R_{12} — 2500-ohm semi-variable.
- R_{13} — 2-megohm variable.
- R_{14}, R_{21}, R_{23}, R_{30} — 50,000 ohms, ½-watt.
- R_{15}, R_{16}, R_{17}, R_{29} — 25,000 ohms, ½-watt.
- R_{18}, R_{24}, R_{27} — 0.5 megohm, ½-watt.
- R_{19} — 1000 ohms, ½-watt.
- R_{20} — 50,000-ohm variable.
- R_{22} — 15,000 ohms, ½-watt.
- R_{25} — 0.1-megohm variable.
- R_{26} — 600 ohms, ½-watt.
- R_{28} — 10,000 ohms, 2-watt.
- R_{31} — 0.5-megohm variable.
- R_{32} — 500 ohms, 2-watt.
- R_{33} — 25,000 ohms, 2-watt.
- T — Universal-type impedance-matching output transformer (UTC VM-0, LS-56, etc.).

Measurement of Current, Voltage and Power

D.c. instruments — Instruments for measuring direct current (§ 2-6) are based on the d'Arsonval moving-coil principle, comprising an indicating pointer moving across a calibrated scale, actuated by the flow of current through a coil located in a constant magnetic field.

Ammeters and voltmeters are basically identical instruments, the difference being in the method of connection. An ammeter is connected in series with the circuit and measures the current flow. A voltmeter is a milliammeter (ammeter reading one-thousandth of an ampere) which measures the current through a high resistance connected across the source to be measured; its calibration is in terms of the voltage drop in the resistance or *multiplier*.

The ranges of both voltmeters and ammeters can be extended by the use of external resistors,

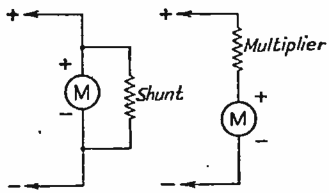


Fig. 2022 — How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.

connected in series with the instrument in the case of a voltmeter or in shunt in the case of an ammeter. Fig. 2022 shows at the left the manner in which a shunt is connected to extend the range of an ammeter and at the right the connection of a voltmeter multiplier.

To calculate the value of a shunt or multiplier it is necessary to know the resistance of the meter. If it is desired to extend the range of a voltmeter, the value of resistance which must be added in series is given by the formula:

$$R = R_m (n - 1)$$

where R is the multiplier resistance, R_m the resistance of the voltmeter, and n the scale multiplication factor. For example, if the range of a 10-volt meter is to be extended to 1000 volts, n is equal to 1000/10 or 100.

If a milliammeter is to be used as a voltmeter, the value of series resistance can be found by Ohm's law (§ 2-6):

$$R = \frac{1000\ E}{I}$$

where E is the desired full-scale voltage and I the full-scale reading of the instrument in ma.

To increase the current range of a milliammeter, the resistance of the shunt is

$$R = \frac{R_m}{n - 1}$$

where R_m is the meter resistance, as above.

Homemade milliammeter shunts can be constructed from any of the various special kinds of resistance wire, or from ordinary copper magnet wire if no resistance wire is available. The Copper Wire Table on page 461 gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size which will carry the full-scale current (at 250 circular mils per ampere). Measure off enough wire (pulled tight but not stretched) to provide the required resistance. Accuracy can be checked by causing a current to flow through the meter which makes it read full-scale without the shunt; connecting the shunt should then give the correct reading on the new full-scale range.

Copper has an appreciable temperature coefficient of resistivity (actually, 0.00393/° C.), and therefore a change in temperature of only 5° F. will change the effective resistance of a copper-wire shunt about 1 per cent. The heating effect of the current through the shunt must be taken into account if high accuracy is desired, as must temperature rise in enclosed cabinets and even room-temperature variations.

The following table shows the minimum wire sizes which should be used for shunts required to carry typical values of current with reasonable temperature rise, together with the resistance per foot at temperatures of 20° C. (68° F.) and 25° C. (77° F.).

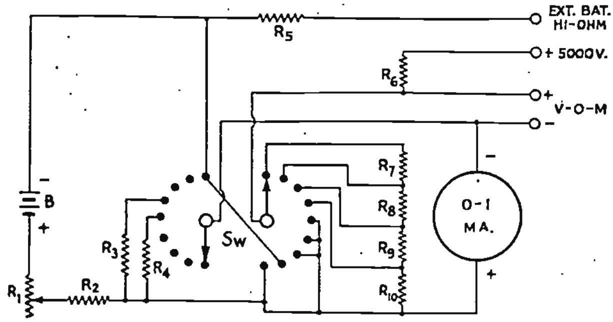
Current	*Wire Size*	*Resistance in Ohms Per Foot* 20° C.	25° C.
60 ma.	No. 38	0.6596	0.6726
100 "	" 36	0.4148	0.4295
150 "	" 34	0.2609	0.2660
250 "	" 32	0.1641	0.1673
400 "	" 30	0.1032	0.1052
600 "	" 28	0.06490	0.06617
1000 "	" 26	0.04082	0.04162

Compensation for temperature variations may be obtained by adding resistance in series, using wire having a negative or zero temperature coefficient. For example, with a 500-microampere movement enough series resistance may be added to make the useful sensitivity of the instrument 1 millivolt.

Shunts for current measurement may be wound with copper wire for ranges up to 50 or 100 ma. to match the coefficient of the copper coil in the movement. At higher current values manganin or similar low-temperature coefficient wire is desirable.

Precision wire-wound resistors used as voltmeter multipliers cannot readily be made by the amateur because of the much higher resistance required (as high as several megohms). As an economical substitute, standard metallized fixed resistors may be used. Such resistors are supplied in tolerances of 5, 10 or 15 per cent ± the marked values. By obtaining matched pairs from the dealer's stock, one of which is, for example, 4 per cent low while the other is 4 per cent high, and using the pairs in parallel or series to obtain the required value of resistance, good accuracy can be obtained at small cost. High-voltage multipliers are preferably made up of several resistors in series; this not only raises the breakdown voltage but tends to average out errors in the individual resistors due to manufacturing tolerances.

Fig. 2023 — Circuit of a low-cost V–O–M.
R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½-watt.
R_3 — 100-ma. shunt, 0.33 ohms (see text).
R_4 — 10-ma. shunt, 3.6 ohms (see text).
R_5 — 40,000 ohms, ½-watt.
R_6 — 4 megohms, 4-watt (four 1-megohm 1-watt resistors in series).
R_7 — 0.75 megohm, 1-watt (0.5 megohm and 0.25 megohm ½-watt in series).
R_8 — 0.2 megohm, ½-watt.
R_9 — 40,000 ohms, ½-watt.
R_{10} — 10,000 ohms, ½-watt.
SW — 9-point 2-pole switch (Mallory-Yaxley 3109).
B — 4.5 volts (Burgess 5360).



A portable combination milliammeter-voltmeter-ohmmeter (often called a V–O–M, or multimeter) having several ranges is extremely useful for experimental purposes and for trouble-shooting in receivers and transmitters. As a voltmeter such an instrument should have high resistance, so that very little current will be drawn in making voltage measurements. A low-resistance voltmeter will give inaccurate readings when connected across a high-resistance circuit. A resistance of 1000 ohms per volt is satisfactory for most uses; a 0-1 milliammeter or 0-500 microammeter (0-0.5 ma.) is the basis of most multi-range meters of this type. Microammeters having a sensitivity of 0-50 μa., giving a voltmeter resistance of 20,000 ohms per volt, are found in units available at reasonable cost. Multipliers for the various ranges are selected by switches.

For maximum convenience in use, multiplier ranges should read in similar basic units, preferably in decimal multiples such as 10×, 5× or 2.5×. This minimizes likelihood of confusion in making readings and simplifies the mental arithmetic involved. On the other hand, the highest accuracy with any given movement occurs in the vicinity of 75 per cent of the maximum range. Thus it is desirable to provide as many ranges as practicable, to facilitate keeping the more commonly encountered important values (such as 6.3 or 250–300 volts) near this region on one scale or another.

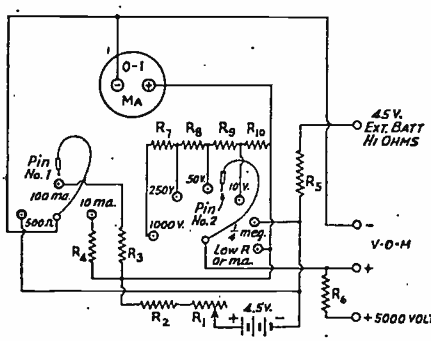


Fig. 2024 — Alternative V–O–M circuit employing pin-jack connectors. Except for the rotary switch, *SW*, all values are the same as in Fig. 2023. Voltage and high-resistance measurements are made with Pin No. 1 disconnected and Pin No. 2 in the appropriate jack. For measuring low resistances by the shunt method, Pin No. 1 is plugged into the 500-ohm jack and Pin No. 2 into the low-ohm-ma. jack. Leaving Pin No. 2 in this same position and removing Pin No. 1 to neutral, the 1-ma. range can be covered. To cover additional current ranges merely requires plugging Pin No. 1 into the 10- or 100-ma. jacks, while Pin No. 2 remains in the low-ohm-ma. jack.

The various current ranges on a multi-range instrument are obtained by using a number of shunts individually switched in parallel with the meter. Particular care must be taken to minimize contact resistance in the switch.

A variety of mechanical and electrical arrangements may be used in a meter of this type. One simple and inexpensive version is shown in Fig. 2023. Using a 0–1 ma. meter, this unit provides five voltage ranges at 1000 ohms per volt: 0–10, 50, 250, 1000 and 5000 volts. The current ranges are 0–1, 10 and 100 ma. There are two ohmmeter ranges: a series range of 0–250,000 ohms and a shunt range of 0–500 ohms. The high-ohms scale is multiplied by 10 if the positive terminal of a 45-volt "B" battery is connected to the "Hi-Ohm" terminal as indicated (the unknown resistance being connected between the negative battery terminal and the ohmmeter negative).

The circuit in Fig. 2023 is based on the use of a multi-position rotary selector switch. An alternative circuit for use where such a switch is not available, employing 'phone-tip and jack connectors instead, is shown in Fig. 2024.

When d.c. voltage and current are known, the power in a d.c. circuit can be stated by simple application of Ohm's law: $P = EI$ (§ 2-6). Thus the voltmeter and ammeter are also the instruments used in measuring d.c. power.

A.c. instruments — Since d.c. meters will not function on alternating current, therefore it is necessary either to rectify the a.c. and measure the resulting d.c. or to use special instruments that will indicate on a.c. (§ 2-8).

In a.c. ammeters and voltmeters utilizing the moving iron-vane principle, the maximum sensitivity usually is 15 to 25 ma. (40 to 67 ohms per volt), so that iron-vane voltmeters consume substantial power. Thus while suitable for measuring filament and line voltages, they cannot be used with circuits which are unable to sustain an appreciable measuring load.

For a.c. measurements at frequencies above a few hundred cycles, or where the power taken by iron-vane meters makes them unsuitable, special devices enabling the use of d.c. movements are employed. The most common of these is the full-wave copper-oxide rectifier, which converts a low-resistance 0-1 ma. d.c. milliammeter into a high-resistance 0-0.909 ma. a.c. milliammeter, making possible an a.c. voltmeter having a sensitivity of 1000 ohms per volt. The design of multipliers for such voltmeters must allow for the fact that the rectifier resistance varies with current. Two scales are usually provided, one for use below 50 volts and the other above. The frequency error averages 0.5 per cent per 1000 cycles.

In a.c. power measurements, the simple multiplication of current and voltage is in error unless the load is purely resistive. If the a.c. current and impedance are known, the power is I^2Z. For ordinary power calculations, such as the input to a transformer, the product of a.c. voltage and current is sufficiently accurate.

R.f. instruments — The measurement of high-frequency a.c. or r.f. quantities involves special problems. Practical instruments read in terms of d.c. from a conversion device.

R.f. current usually is measured by means of a thermoammeter. This is a sensitive d.c. microammeter connected to a thermocouple associated with a heater made of a short piece of resistance wire. Thermoammeters have been made with an r.f. sensitivity of 1 ma., but the ranges used by amateurs for measuring antenna current, etc., are from 0–0.5 ampere up.

The most suitable r.f. voltmeter is a peak-reading vacuum-tube voltmeter (Fig. 2025). When properly designed, its accuracy at r.f. is limited only by the variation of input resistance with frequency. The peak diode voltmeter has little error even at 60 Mc. The same is true of the self-biased and slide-back types, if tubes having low input capacity are used. The oscilloscope also can be used as an r.f. voltmeter for potentials of several volts or more.

R.f. power measurements may be made by measuring the current through a resistor or reactance of known value. Approximate measurements can be made by using ordinary 6- or 115-volt light bulbs as a substitution or "dummy" load, connected either singly or in series-parallel to provide the required resistance and power rating. The approximate resistance of the bulb can be computed from its wattage rating at 60 cycles. Special noninductive resistance units, enclosed in vacuum bulbs mounted on tube bases, are available for this purpose. For higher-power work the units can be connected in series-parallel (§ 4-9).

Where the substitution load method is impractical, r.f. power can be measured by multiplying the current through the circuit (using a thermoammeter) by the r.f. voltage across the circuit, as read on an r.m.s. meter (or 70.7 per cent of the reading on a peak voltmeter).

R.f. power may also be measured by the photometric method, in which a calibrated light-sensitive cell (a photographer's exposure meter will serve) is used to compare the relative brilliance of a light bulb as a substitution load and on the 115-volt 60-cycle supply.

Vacuum-tube voltmeters — The most generally useful instrument for the measurement of d.c., a.c. and r.f. voltages is the vacuum-tube voltmeter. Its chief advantages are (a) high input resistance (i.e., negligible power taken from the circuit under measurement), and (b) good accuracy over a wide range of frequencies extending into the u.h.f. region.

The vacuum-tube voltmeter operates by virtue of the change in plate current in a vacuum

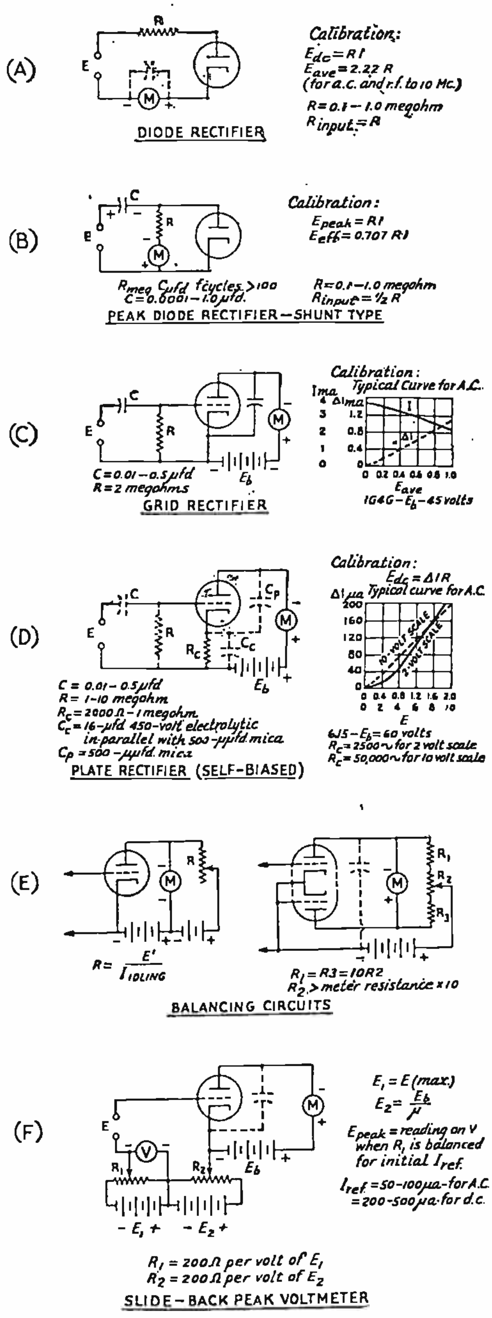


Fig. 2025—Fundamental vacuum-tube voltmeter circuits.

tube caused by a change in the voltage applied to the grid. Thus, by introducing the voltage which is to be measured into the grid circuit, the resulting plate-current change constitutes a measure of the applied voltage. In the case of a.c. the tube acts as a rectifier, and the measurement is in terms of rectified d.c.

Representative vacuum-tube voltmeter circuits are shown in Fig. 2025. The simple diode rectifier (A) can be almost any type of tube; in a triode or multi-grid type, all electrodes except the control grid are connected to cathode (or negative filament). A Type 30 or 1G4G tube with a flashlight cell for filament supply makes a convenient portable unit. Tubes with low input capacity (1N5G, 6T7G, 954, etc.) should be used for high frequencies. The frequency range is limited by the tube input capacity shunting the load resistance. The calibration will be linear above 2 or 3 volts, provided the load resistance exceeds 0.1 megohm. The meter, M, usually is a sensitive microammeter (0–100 or 200 μa.); however, a 0-1 ma. meter may be used, with reduced sensitivity.

The peak diode voltmeter at B, shunt-connected to eliminate the necessity for a d.c. return in the measured circuit, reads peak a.c. voltage. The input resistance is comparable to that of the simple diode for equivalent sensitivity, but the high-frequency error is less. The time constant of the RC circuit should be at least 100 for the lowest frequency to be measured ($RCF < 100$). Typical values are 0.5 megohm and 0.5 μfd. for audio frequencies and 0.1 megohm and 0.05 μfd. (mica) for r.f. and i.f.

The grid-rectification circuit at C is equivalent to the diode rectifier of B followed by a triode amplifier, with sensitivity greatly increased over the ordinary diode. The input resistance is low (0.1 to 1.0 megohm) with small inputs because of grid current. The plate current is maximum when idling and decreases with signal. This circuit is useful chiefly because it can be used with inexpensive meters. The instrument may be calibrated from a known 60-cycle source; the scale is square-law for small signals, becoming linear with increasing input. The value of R is non-critical. C should have a reactance small compared with R at the operating frequency; i.e., 0.01-μfd. mica from 1 kc. up, 0.1-μfd. paper for low a.f. For d.c., C is, of course, omitted. A high-μ tube will have lower idling (no-signal) plate current.

The self-biased plate-rectification or reflex voltmeter at D has a very high input resistance and fair sensitivity. It is normally connected directly across the circuit to be measured; if no d.c. return is available, a coupling circuit must be added, as shown by the dotted lines (C — 0.01 μfd; R — 10 megohms or more). A low-μ tube is preferable, to minimize contact potential and grid current. The cathode resistance, R_2, controls the sensitivity; the higher its resistance, the more linear and stable will be the calibration. A range switch can be provided, connecting in various values of cathode resistance from 2000 ohms to 0.5 megohm to give full-scale ranges from about 2.5 to 250 volts. The plate and cathode by-passes may be 0.001-μfd. mica condensers, the cathode by-pass being shunted by a 10-μfd. electrolytic for 60-cycle calibration and for low audio-frequency measurements.

The no-signal plate current present in the circuits of C and D can be balanced out by bridge or bucking circuits, typical forms of which are shown at E. An auxiliary battery (or a section of the voltage divider in an a.c. power supply) is connected back to the meter through a variable resistor, providing a controllable opposite current flow which can be made to equal exactly the residual plate current of the tube. When used with C this balancing circuit allows the meter terminals to be reversed, thereby making the meter read forward instead of backward. The resistor, R, should have a value not less than ten times the internal resistance of the meter.

At the right in E an automatic self-balancing circuit is shown wherein a duplicate triode (usually the second section of a twin-triode tube) takes the place of the adjustable resistor, R. The current flowing through R_1 and R_3 being equal and opposite with no signal, there is no potential across the meter and consequently no reading. When a voltage is applied to the grid of the voltmeter triode this balance is disturbed, however, and the meter registers current flow. A small zero-setting resistor, R_2, is provided to correct for any discrepancies in the tubes or resistors. The values for R_1 and R_3 depend on the plate-supply voltage available; the higher their resistance, the better the sensitivity and stability. The minimum value is several times the meter resistance.

The "slide-back" voltmeter at F is a comparison instrument in which the peak value of an a.c. or r.f. voltage is read in terms of a d.c. substitution voltage; the voltmeter tube and the milliammeter, M, merely indicate when the two voltages are equal. With the input terminals shorted and R_1 set so that V reads zero, the tube is biased nearly to cut-off by adjustment of R_2. The residual plate current becomes the reference current ($I_{ref.}$) or "false zero." When an a.c. voltage, E, is applied across the input terminals, plate rectification of the positive peaks causes the plate current to rise. By adjustment of R_1, additional bias voltage is introduced to balance out the a.c. voltage. The additional bias required to bring the plate current back to the reference value ($I_{ref.}$) is equal to the peak value of the voltage being measured. In operation, R_1 should be adjusted (after setting $I_{ref.}$) so that all of E_1 is in the circuit, to avoid burning out the milliammeter when the signal is applied. After the unknown voltage has been connected, the bias is reduced by R_1 until the reference current is reached. The slide-back voltmeter is capable of high accuracy and, since it requires no a.c. calibration, is particularly useful for a temporary set-up.

R, Z, C, L and Q Measurements

Resistance — The volt-ammeter, ohmmeter and Wheatstone bridge methods are commonly used in measuring resistance. In the volt-ammeter method, the resistance is determined from Ohm's Law by measuring the current through the resistor when a known d.c. voltage is applied. The resistance can be determined with a voltmeter alone, when

$$R = \frac{eR_m}{E} - Rm$$

where R is the resistance under measurement, E is the voltage read on the meter, e is the series voltage applied, and R_m is the internal resistance of the meter (full-scale reading × ohms-per-volt).

The ohmmeter is a practical application of this method, with a low-current d.c. voltmeter and a source of voltage (usually dry cells) connected in series with the unknown resistance. If the meter reads full-scale with the connecting leads shorted, insertion of the resistance under measurement will cause the reading to decrease in proportion to the resistance inserted. The scale thus can be calibrated in ohms.

In Fig. 2026-A, the series resistance is adjusted until the milliammeter reads full-scale when the test leads are shorted. When the meter reading changes as the battery ages the series resistance is reduced to compensate for the change. In B, the series resistance is kept constant but the sensitivity of the meter is varied to compensate for the changing voltage. The circuit of C is useful for measuring resistances below a few hundred ohms. The unknown resistance is connected as a shunt across the meter, reducing the current reading. Values of a fraction of an ohm can be read in this way.

The ratio of resistance valves which can be measured on a single ohmmeter range averages about 100 to 1, or from one-tenth to ten times the center-scale value.

Only approximate measurements can be made with an ohmmeter. For greater accuracy, the unknown resistor may be compared with a standard resistance of known accuracy by means of a Wheatstone bridge (§ 2-11). If resistance measurements only are to be made, the bridge can be powered from a battery and a milliammeter used for the balance indicator. If reactances also are to be measured, an a.c. source is required (Fig. 2028).

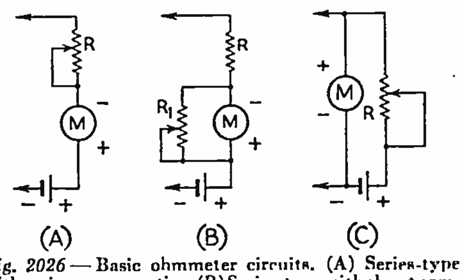


Fig. 2026 — Basic ohmmeter circuits. (A) Series-type with series compensation. (B) Series-type with shunt compensation. (C) Shunt-type for measuring low resistances.

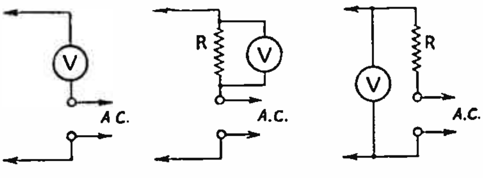


Fig. 2027 — Reactance-meter circuits for checking C and L.

Capacity and inductance — The capacity of condensers and the inductance of coils can be measured (a) in terms of their reactance, (b) by comparison with a standard, or (c) by substitution methods.

The reactance method is simplest but least accurate. The method is similar to the d.c. ohmmeter, except that impedance is measured instead of resistance. In Fig. 2027-A, the unknown reactance is placed in series with an a.c. rectifier-type voltmeter across the 115-volt a.c. line. With a 1000-ohms-per-volt meter, capacities can be identified from approximately 0.001-μfd. to 0.1-μfd. At B the reactance is connected in series with a 1000-ohm resistance; the proportionate voltage drop across this resistance indicates the reactance of condensers from 0.1 μfd. to 10 μfd. and of inductances from 0.5 henry to 50 henries, when Q is greater than 10. Because the lower end of the scale of a rectifier-type meter is somewhat crowded, a better reading can be had by using the connection at C for large reactances. Approximate calibrations for each connection may be made by checking typical condensers and coils of known values and drawing calibration curves for the voltmeter in use.

The reactance method at best gives only approximate indications of inductance and capacity. For accurate measurements, an a.c. bridge must be used.

A simple bridge for the measurement of R, C and L is shown in Fig. 2028. Its accuracy will depend on the precision of the standards, the sensitivity of the detector or balance indicator, the voltage and frequency of the a.c. source, and the ratio of the unknown value to the standard. The signal source may be a 1000-cycle audio oscillator with low harmonic content and the detector a pair of headphones. A "magic-eye" tube can be used as a detector.

For maximum accuracy the ratio of the unknown to the standard should be kept small, so that R is read near the center of its scale. The ratio can be as high as 10 to 1 in either direction with good accuracy, and an indication can be had even at 100 to 1. Additional standards may be included for other ranges if desired.

The potentiometer, R, must be calibrated as accurately as possible in terms of the *ratio* of resistance on either side of its mid-point, which may be arbitrarily marked 10. If the potentiometer is next set at 500 ohms, the ratio of resistances is 1 to 10 and the scale may be marked 1. The corresponding point on the other end of the scale is marked 100. Intermediate points are similarly marked according to the resistance ratios. These ratios will then correspond with the ratio of the unknown resistance, inductance or capacity to the standard in use, when the bridge has been balanced for a null indication on the detector.

Since direct current flowing through a coil changes its inductance, allowance must be made for this effect when measuring choke coils and transformers carrying d.c.

Condensers should be checked for leakage as well as for capacity. This check must be made with the rated d.c. voltage applied, a microammeter being connected in series with the high voltage source. The resistance of good paper condensers should be above 50 megohms per microfarad, that of mica condensers above 100.

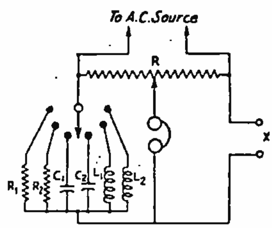


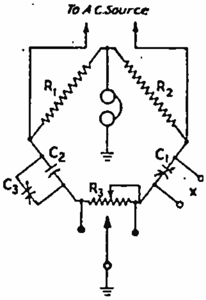
Fig. 2028 — Simple a.c. bridge for measuring *R*, *C* and *L*.
C_1 — 0.01-μfd. mica.
C_2 — 1.0-μfd. paper.
R — 10,000-ohm wire-wound.
R_1 — 100-ohm wire-wound.
R_2 — 10,000-ohm wire-wound.
L_1 — 125-mh. r.f. choke.
L_2 — 12-henry filter choke.

The condition of electrolytic condensers can be checked roughly with an ohmmeter. With the positive terminal of the condenser connected to the positive of the ohmmeter battery, high-voltage electrolytics should show a resistance of 0.5-megohm or so; low-voltage cathode by-pass condensers should be over 0.1 megohm. Electrolytics can also be checked by measuring the leakage current when the rated d.c. polarizing voltage is applied. It should read about 0.1 ma. per μfd. The maximum for a useful unit is about 0.5 ma. per μfd. Low leakage current also indicates a faulty unit. An electrolytic condenser which has lain idle on the shelf will show leakage current as high as 2 ma. per μfd. per 100 volts. After "aging" for a few minutes with rated d.c. voltage applied it should return to normal, however.

The measurement of small capacities under 0.001 μfd. is not possible with a bridge of the type previously described because stray reactances affect the accuracy. A more accurate bridge for measurement of small capacities is shown in Fig. 2029. It is of the substitution type with a calibrated air condenser, C_1, for the variable arm. C_2 is a fixed reference capacity. C_3 is used to balance out stray capacity including that of the leads to C_x. The bridge is first balanced by adjusting C_3, with C_1 at maximum capacity and the leads to C_x in place. C_x is then connected and the bridge again balanced by adjusting the capacity of C_1 to compensate for C_x. The difference in capacity (ΔC) of C_1 between its new setting and maximum capacity is the capacity of C_x.

It is impossible to get a zero null indication from the detector unless the resistance as well as the capacity of the two condensers being

Fig. 2029 — A substitution-type capacity bridge circuit.
C_1 — 100-μμfd. straight-line-capacity condenser (may be dual 500-μμfd. with sections in parallel).
C_2 — 900-μμfd. silver-mica.
C_3 — 100-μμfd. variable trimmer.
R_1, R_2 — 500-ohm wire-wound (1 per cent accuracy).
R_3 — 1000-ohm wire-wound potentiometer.



compared are equal. R_3 is therefore included to aid in achieving a resistive balance. Generally speaking, R_3 will be in the C_2 leg when measuring a mica condenser and in the C_1 leg for an air condenser. The bridge is brought into balance by alternately varying the standard capacity, C_1, and equalizing the power factor by means of R_3 until zero indication is obtained.

The bridge can be made direct-reading in μμfd. by using a dial with 100 divisions and a 10-division vernier (such as the National Type N), installed so that 0 on the dial corresponds to maximum capacity on C_1. Then, as the capacity of C_1 is decreased to compensate for the addition of C_x, $C\Delta$ is numerically equal to the dial reading times 10. The true capacity of C_1 will depart from linearity with the dial setting as it nears zero, but the percentage error remains small up to at least 90 on the dial ($C_x < 900$ μμfd). The over-all accuracy can be made better than 1 per cent.

L, C and Q measurements at r.f. — The low-frequency a.c. bridge method of measuring inductance is of value only for the high-inductance coils used at power and audio frequencies. I.f. and r.f. coils must be measured at the frequencies at which they are used.

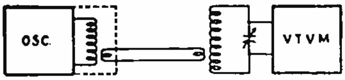
The method commonly employed is to determine the frequency at which the coil resonates when connected across a capacity of known value. This may be done (1) by connecting the coil-condenser combination in a two-terminal oscillator (§ 3-7) and measuring the resulting oscillation frequency on a calibrated receiver, or (2) by connecting the coil to a calibrated condenser as in Fig. 2030, supplying the circuit with r.f. power from a suitable oscillator, and tuning the condenser until resonance is indicated by maximum indication on a vacuum-tube voltmeter. With the capacity known in μμfd. and the resonant frequency in kc., the apparent inductance of the coil in microhenries can be computed:

$$L = \left(\frac{159,160}{f}\right)^2 \frac{1}{C}$$

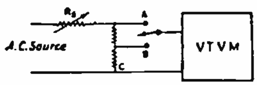
The apparent inductance thus computed is in error, however, in that it also includes the distributed capacity of the coil. This will be discovered if a similar measurement is made at another frequency (for example, the harmonic of f_1), for it will be found that a different value of inductance results. However, by combining the two measurements the true inductance can be determined:

$$L = \frac{10^{12}}{13.15 f_2^{\,2}} \times \frac{1}{C_1 - C_2}$$

when f_2 is the second harmonic of f_1, C_1 the capacity required to tune to f_1, and C_2 the capacity required to tune to f_2.



A



B

Fig. 2030 — (A) Circuit used for measuring inductance, capacity and Q at r.f. The calibrated variable-frequency oscillator should have a tuning range in excess of 2-to-1. (B) Circuit for calibrating the v.t.v.m. for Q measurements from 60-cycle a.c. R_{BC} is 70.7 per cent of R_{AC}. With the switch in position A, R_S is adjusted to give a voltmeter deflection near the upper part of its scale; this is the peak-deflection reference point. The switch is then turned to position B, and the new reading noted. By making a number of measurements with different initial input levels, a graph can be plotted showing both peak and 70.7 per cent readings for a wide range of inputs.

A convenient source of r.f. for the two-frequency method of measurement is the transmitter exciter unit, provided it has good second-harmonic output. The oscillator output and link circuit (shown inside the dashed lines in Fig. 2030) should be shielded or at a distance sufficiently remote from the measuring circuit so that the v.t.v.m. shows no indication when there is no coil in the circuit. The calibrated condenser must have sufficient capacity to tune over a 2-to-1 frequency range; it may be calibrated by means of a bridge such as the substitution-type capacity bridge of Fig. 2029.

The resonance method can also be used for accurate measurement of capacity. A standard coil of suitable inductance must be provided; the exact value is not important. The standard condenser, C_1, is first tuned to resonance at the oscillator frequency. The unknown capacity, C_x, is then added in parallel, whereupon the capacity of C_1 is reduced until the circuit again resonates at the oscillator frequency. The difference between the two settings (ΔC) represents the capacity of C_x.

The arrangement of Fig. 2030 is additionally useful in that it constitutes a Q meter, and thus can be used for measuring r.f. resistance and impedance. Since the resistance in a tuned circuit broadens the resonance curve (§ 2-10), measuring the frequency difference between the two points at which the output voltage equals 70.7 per cent of the peak voltage (i.e., where the resistance in the circuit equals its reactance), will give the Q of the circuit.

One method of determining these points involves the use of a calibrated variable-frequency oscillator to determine the band-width. Another employs a calibrated variable condenser to measure the capacity change.

When the calibrated variable oscillator and v.t.v.m. are used, the frequency and r.f. voltage at resonance must first be noted. The oscillator frequency is then varied either side of resonance until the v.t.v.m. reads 70.7 per cent of its initial value. Then Q is equal to the frequency divided by the band-width, or

$$Q = \frac{f_r}{\Delta f}$$

where Δf is the difference between f_1 and f_2.

When the frequency of the oscillator is fixed and a calibrated variable condenser is used, the capacity at resonance (C_r) is noted, as well as that on either side at which the meter reads 70.7 per cent of maximum. Then

$$Q = \frac{2\,C_r}{C_2 - C_1}$$

The foregoing applies to measuring the Q of coils. Actually, the value of Q thus derived is that of the tuned circuit as a whole, including the condenser. The Q of the condenser used as a standard must, therefore, be high. An efficient air condenser with steatite or mycalex insulation is required; it should be operated near maximum capacity, using short, heavy leads and with stray capacities kept as low as possible.

The Q of other air condensers and of mica condensers can be determined by first measuring the Q of the circuit with a standard coil in place, and then connecting C_x in parallel with C and again measuring the Q. The Q of the unknown condenser is

$$Q_x = \frac{(C_1 - C_2)\,Q_1\,Q_2}{C_1\,(Q_1 - Q_2)}$$

Low-Q mica and paper condensers ($Q <$ 1000) can be measured by inserting the unknown in series with L and C. Q_1 is measured with a shorting bar across the unknown; the bar is then removed and Q_2 determined. Then

$$Q_x = \frac{(C_2 - C_1)\,Q_1\,Q_2}{C_1\,Q_1 - C_2\,Q_2}$$

If C_2 is larger than C_1, the reactance is inductive rather than capacitive; i.e., the "condenser" is actually an inductance at the measurement frequency.

The r.f. resistance, reactance and impedance of other components can be measured by the same methods. If an external r.f. impedance (such as a transmission line or an r.f. choke) is inserted in a coil-condenser circuit, it will both detune the circuit and broaden its resonance curve. Using a standard coil and condenser suitable for the operating frequency, connect the unknown across C_1 (for high impedances) or in series with L and C (for low impedances), and proceed as above. If C_1 must be increased to restore resonance, the reactance of the unknown is inductive; if it must be decreased, the reactance is capacitive.

Oscilloscopes

Uses — The most versatile of all measuring devices is the cathode-ray oscilloscope (§ 3-8, 3-9). Although relatively expensive, it replaces a number of less satisfactory types of measuring equipment. It is useful on d.c., a.c. and r.f., and is particularly suited to a.f. and r.f. measurements because of the high input resistance and small frequency error.

An oscilloscope is, in effect, a complex voltmeter capable of measuring any two voltages simultaneously by the deflection of a weightless electron-beam pointer. Moreover, because this beam projects an image on a retentive luminous screen, the measurements include the additional factor of time. It is possible, therefore, to see the actual form of a transient pulse or one or more repetitive cycles of an a.c. voltage and to measure not only amplitude but also frequency and waveform (§ 3-9).

Constructional considerations — The oscilloscope should be housed in a metal cabinet, both to shield the tube from stray fields which might deflect the beam and to protect the operator from the high voltages employed. It is good practice to provide an interlock switch which automatically disconnects the high-voltage supply when the cabinet is opened for servicing or other reasons.

In building the unit, the cathode-ray tube must be placed so that the alternating magnetic field from the transformer has no effect on the electron beam. The transformer should be mounted directly behind the base of the tube, with the axes of the transformer windings and of the tube on a common line.

It is important that provision be included either for switching off the electron beam or reducing the spot intensity, or for swinging the beam to one side of the screen with d.c. bias, when no signal voltage is being applied. A thin, bright line or a spot of high intensity will "burn" the screen of the cathode-ray tube.

If trouble is experienced in obtaining a clean pattern from a high-power transmitter because

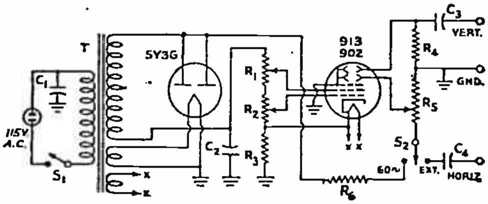


Fig. 2031 — An oscilloscope for modulation monitoring.
C_1 — 0.01-μfd. 400-volt paper.
C_2 — 0.5-μfd. 800-volt paper or oil-filled.
C_3 — 0.005-μfd. mica.
C_4 — 0.1-μfd. 600-volt paper.
R_1 — 50,000-ohm variable.
R_2, R_5 — 0.5-megohm variable.
R_3 — 1 megohm, 1-watt.
R_4, R_6 — 0.5 megohm, 1-watt.
S_1 — S.p.s.t. toggle switch.
S_2 — S.p.d.t. toggle switch.
T — Replacement-type transformer; 350 volts, 40 ma.; 5 volts, 3 amperes; 6.3 volts, 2 amperes.

of r.f. voltage introduced by the 115-volt line, by-pass condensers (0.01 or 0.1 μfd.) should be connected across the primary of the power transformer in the common connection being grounded to the case.

A simple oscilloscope — The circuit of a simple cathode-ray oscilloscope is shown in Fig. 2031. Either a 1-inch 913 or a 2-inch 902 tube can be used. The cathode-ray tube may be mounted, together with the associated rectifier tube and other components, in a cabinet made of a standard 3 × 5 × 10-inch steel chassis with bottom plate.

This circuit is useful primarily for modulation checking in radiotelephone transmitters (§ 5-10). Horizontal sweep voltage may be obtained either from an audio-frequency source (such as the modulator stage of the transmitter) or from the 60-cycle a.c. line, as selected by S_2. Using a sinusoidal a.c. sweep, the pattern appearing on the screen will be in the form of a trapezoid or triangle (depending on the percentage of modulation).

R_5 controls the amplitude of the applied horizontal sweep. R_1 is the intensity control and R_2 the focusing control. If needed, a 2.5-mh. 125-ma. r.f. choke may be connected in series with the lead to the rotor of R_5 to correct leaning of patterns caused by r.f. coupling.

Amplifiers — The usefulness of the oscilloscope is enhanced by providing amplifiers for both the horizontal and vertical sweep voltages, thereby insuring that sufficient voltage will be available at the deflection plates to give a pattern of suitable size. With small oscilloscope tubes (3-inch and smaller screens) the voltage required for a beam deflection of one inch varies from about 30 to 100 volts, depending upon the anode voltages, so that an amplifier tube capable of an undistorted peak output voltage of 100 or so is necessary. (With such an amplifier, the voltage difference, or total voltage "swing," between the positive and negative peaks is 200 volts.) A resistance-coupled pentode-tube voltage amplifier (§ 3-3) is ordinarily used because of the high gain obtainable. The amplifier should be designed to have flat frequency response over a wide range of frequencies (§ 3-3, 3-6, 5-9). Voltage gains of 100 to 150 or more per stage are readily obtainable, so that full deflection of the beam can be secured with an input of one volt or less.

Time base — For actual studies of waveform, the use of a sweep circuit having a linear time base is necessary (§ 3-9). The sweep circuit proper usually employs a grid-controlled gaseous discharge tube (the 884 and 885 are especially designed for this purpose), operating as a synchronized relaxation oscillator.

The voltage output of this type of sweep circuit is limited, because the charging rate of the condenser is linear only on that portion of the logarithmic charging curve (§2-6, 3-9) which is practically a straight line. A linear charging rate over a longer period of time can be secured by substituting a current-limiting device, such

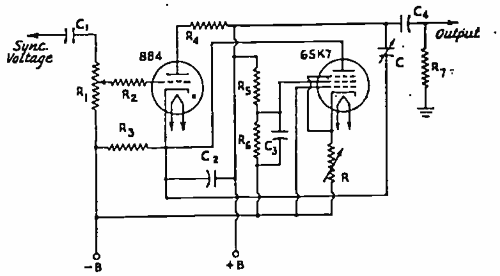


Fig. 2032 — Sweep oscillator with current-limiting tube.

C — 0.001–0.25 μfd.
C_1, C_4 — 0.1-μfd. paper.
C_2 — 100-μμfd. mica.
C_3 — 8-μfd. 250-volt electrolytic.
R — 50,000-ohm variable.
R_1 — 0.25 megohm variable.
R_2 — 25,000 ohms, ½-watt.
R_3 — 2000 ohms, ½-watt.
R_4 — 1000 ohms, 1-watt.
R_5 — 40,000 ohms, 25-watt.
R_6 — 6000 ohms, 2-watt.
R_7 — 10 megohms, ½-watt.

as a properly adjusted vacuum tube, for R_1. In Fig. 2032 a 6SK7 pentode is employed as a constant-current resistor, permitting approximately 75 per cent of the supply voltage to be utilized with a linearity of about 2 per cent.

A complete oscilloscope — The circuit of Fig. 2033 shows a complete cathode-ray oscilloscope, including the basic cathode-ray tube and power supply circuit, vertical and horizontal amplifiers, and a linear sweep generator.

The cathode-ray tube used may be either the 913 (1-inch screen), 902 (2-inch screen), or the 3-inch 3AP1/906. The sweep is of the relaxation-oscillator type, employing an 884 tube, with a range of about 10 to 30,000 cycles. The high-gain amplifier stages, employing 6SJ7 tubes, give reasonably flat response between 10 to 100,000 cycles providing a sensitivity of better than 0.5 volts per inch (for a 3AP1/906) at maximum gain. Each of these circuit units is shown grouped separately in the circuit diagram and should be similarly separated in construction.

The following are panel controls: R_1, the vertical-input level control, is ganged with the vertical-input switch, S_1. The horizontal-input control, R_2, similarly is ganged with the vertical-input switch, S_2. The sweep selector switch, S_3S_4, in position (1) is connected internally to supply a synchronized sweep voltage from the 884, controlled by the input voltage to the vertical plates; (2) locked 60-cycle sweep voltage derived internally, and (3) locked sweep voltage from the 884 synchronized with any external source. R_{16} is the sweep-synchronizing or locking control. R_{14} is the sweep-frequency vernier control, while S_5 is the sweep range selector. R_{18} is the beam-intensity control and R_{20} the beam-focusing control.

The vertical and horizontal amplifiers have shunt compensation in the 6SJ7 plate circuits to extend the high-frequency range. The amplifier input and output leads should be direct and placed well clear of other components. R.f. input should be applied directly to the deflection plates.

In the diagram separate power supplies are shown for the cathode-ray tube and for the associated circuits. A special combination transformer for oscilloscope work is suitable, if available. In the arrangement shown, replacement-type transformers are used, T_1 being required to deliver between 400 to 600 volts d.c. (depending on the type of tube used) and T_2 approximately 300 volts d.c. at the filter output. The nominal d.c. current rating is unimportant, since the actual drain is small.

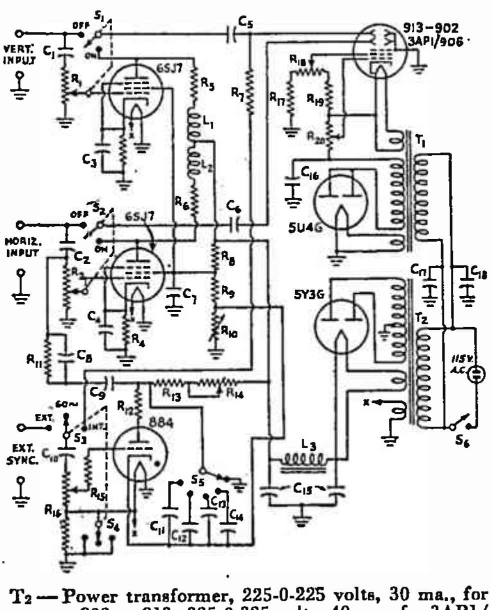


Fig. 2033 — A complete oscilloscope with linear sweep.

C_1, C_2, — 0.1-μfd. 400-volt paper.
C_2, C_9, C_{13}, C_{17}, C_{18} — 0.05-μfd. 400-volt paper.
C_5, C_6, C_{14} — 0.25-μfd. 400-volt paper.
C_8 — 3–30-μμfd. mica trimmer.
C_{11} — 500-μμfd. mica.
C_{12} — 0.005-μfd. mica.
C_{15} — Dual 8-μfd. 450-volt electrolytic.
C_{16} — 2-μfd. 800-volt paper (or dual 450-volt electrolytic in series).
R_1, R_2 — 1-megohm variable.
R_3, R_4 — 750 ohms, ½-watt.
R_5, R_6 — 0.1 megohm, ½-watt.
R_7, R_{11} — 2 megohms, ½-watt.
R_8 — 25,000 ohms, 2 watts
R_9, R_{15} — 10,000 ohms, ½-watt.
R_{10} — 1500-ohm 10-watt semi-variable.
R_{12} — 500 ohms, ½-watt.
R_{13} — 0.2 megohm, ½-watt.
R_{14} — 2.-megohm variable.
R_{16} — 0.1-megohm variable and 250 ohms, 1-watt.
R_{17} — 0.25 megohm, 1 watt.
R_{18} — 0.25-megohm variable.
R_{19} — 0.1 megohm, 1 watt.
R_{20} — 50,000-ohm variable.
L_1, L_2 — 50 mh. r.f. chokes.
S_1 — S.p.s.t. snap switch (mounted on R_1).
S_2 — S.p.s.t. snap switch (mounted on R_2).
S_3, S_4 — Two-pole 3-position rotary switch.
S_5 — Single-pole 5-position rotary switch.
S_6 — S.p.s.t. toggle switch.
T_1 — Power transformer, 275-0-275 volts, 40 ma.; 5 volts, 2 amperes; 6.3 volts, 0.6 amperes; 6.3 volts, 1.5 amperes.
T_2 — Power transformer, 225-0-225 volts, 30 ma., for 902 or 913; 325-0-325 volts, 40 ma., for 3AP1/906; 5 volts, 3 amperes; 6.3 volts, 1.5 amperes, or 2.5 volts, 2 amperes.

When used for measuring voltage, the signal is applied to the vertical plates and its amplitude measured in terms of the height of the resulting trace. Approximate measurements can be made by calibrating the sensitivity of the cathode-ray tube in volts per inch. The sensitivity varies with the anode voltage and type of tube; typical figures for small tubes are 25 to 75 volts per inch, peak-to-peak. The initial calibration can be made with a variable d.c. voltage source and a comparison voltmeter.

Impedance can be measured at any frequency by connecting the circuit or component under measurement in series with a non-reactive resistor across a source of signal voltage of the required frequency. The relative deflection as the oscilloscope is connected across first the resistor and then the impedance will give the value of impedance with respect to the known value of resistance.

Electronic Switch — An electronic switch is used to extend the utility of the cathode-ray oscilloscope by making possible the simultaneous viewing of the waveshapes of two or more voltages on the screen.

The principle underlying the electronic switch is very simple. The voltages to be studied are applied to the grids of two amplifier tubes having a common plate load. These tubes are rendered alternately conducting and non-conducting at a suitable rate by a vacuum-tube switching circuit. The output voltage then corresponds to a series of replicas of the two input waveshapes. When this voltage is applied to the vertical plates of the cathode-ray tube, the two input voltages will appear separately on the screen. The switching is done so rapidly that the two phenomena seem to be present at the same time.

Fig. 2034 shows a simplified form of electronic switch in which two multi-grid tubes perform both switching and amplifying operations. The 6J8G with its triode-heptode construction fulfills the necessary requirements. Switching is effected by voltage applied to the No. 3 grid. This grid is located between two screening grids which are maintained at a fixed positive potential, ensuring against interaction between the signal and the switching voltages. The transition time is reduced to a minimum by the fact that this grid also has a comparatively sharp cut-off with good linearity as well as sensitivity.

Generation of the required switching voltage is the function of the triode sections which, acting in a multivibrator-type switching circuit, provide a square-topped waveform with sharp cross-overs. Since the triode grid of the 6J8G is connected internally to the No. 3 grid of the heptode, no external coupling is required, considerably simplifying the circuit.

The most desirable switching frequency depends on two considerations. Since the output voltage is a square wave, the rate must be sufficiently low to be handled without distortion by the oscilloscope amplifier. If the frequency

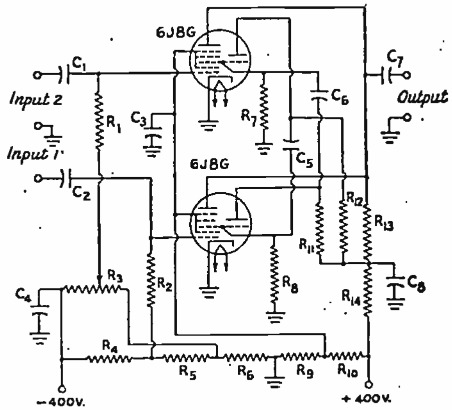


Fig. 2034 — A simple electronic switch circuit.

C_1, C_2, C_7 — 0.5-μfd. 200-volt paper.
C_3 — 40-μfd. 150-volt electrolytic.
C_4 — 50-μfd. 25-volt electrolytic.
C_5, C_6 — 500-μμfd. mica.
C_8 — 16-μfd. 450-volt electrolytic.
R_1, R_2, R_7, R_8 — 1 megohm, ½-watt.
R_3 — 0.25 megohm, ½-watt.
R_4, R_5, R_6 — 50 ohms, ½-watt.
R_9 — 0.25 megohm, 2-watt.
R_{10} — 50,000 ohms, 2-watt.
R_{11}, R_{12} — 50,000 ohms, ½-watt.
R_{13} — 15,000 ohms, 1-watt.
R_{14} — 3500 ohms, ½-watt.

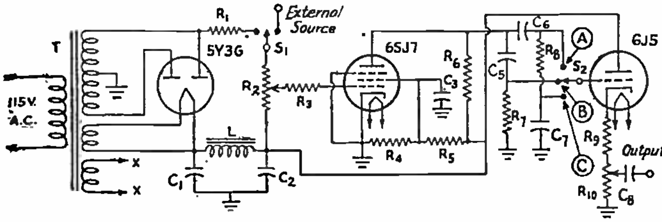
of the square wave is 50 cycles (about the minimum for good images) the amplifier must have a response flat between 0.5 and 500 cycles. At a 5000-cycle switching rate, to preserve the flatness of the tops of the waves and to handle the transient cross-overs satisfactorily, the amplifier would have to be perfectly flat from 5 to 50,000 cycles.

The switching rate of the multivibrator with the values indicated is about 500 cycles, which is the frequency commonly used. It can be varied by changing the values of C_1C_2 and R_1R_2.

To avoid hazy figures the change-over from one tube to the other must be instantaneous. Otherwise both amplifier tubes will draw current during the transition stage, resulting in a sudden drop in output voltage and causing background haze.

Provided the transition period can be kept to less than 0.1 millisecond, or 25 per cent of one cycle at 500 c.p.s. the vertical change-over trace will scarcely be apparent — a condition adequately fulfilled in the circuit shown, which in itself is capable of functioning satisfactorily at any frequency between 1 and 2000 c.p.s.

The input sensitivity for satisfactory output waveshape is about 0.1 volt, based on a maximum gain in the heptode amplifiers of about 100. The actual positions of the traces depends on the steady plate currents of the respective tubes. If these are the same, the two images will appear to have a common base line: varying the relative d.c. level by means of R_3, on the other hand, provides a convenient way of shifting one trace with respect to the other for identification of the individual traces.



Tracings of output waveforms:

(A) Square-wave output from the clipper-tube plate.

(B) Sharp pulse obtained from differentiating circuit.

(C) Triangular waveshape from integrating circuit.

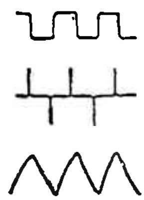


Fig. 2035 **— General-purpose 60-cycle time-base waveshaping circuit.**

C1, C2 — 8-μfd. 450-volt electrolytic.
C3 — 10-μfd. 250-volt electrolytic.
C5 — 0.005-μfd. mica.
C6, C8 — 0.5-μfd. 200-volt paper.
C7 — 0.1-μfd. 600-volt paper.
R1 — 5 megohms, 1-watt.
R2 — 1 megohm variable.
R3 — 1 megohm, ½-watt.
R4, R5 — 0.1 megohm, 1-watt.
R6 — 0.1 megohm, ½-watt.
R7 — 1000 ohms, ½-watt.
R8 — 3 megohms, ½-watt.
R9 — 500 ohms, ½-watt.
R10 — 2500-ohm variable.

60-cycle time-base circuit — The general-purpose time-base circuit shown in Fig. 2035 utilizes 60-cycle sinusoidal a.c. to generate three types of waveforms.

The 6SJ7 blocking-type clipper tube operates with zero bias. A large negative pulse across R_2 drives the grid far beyond cut-off, blocking the tube, while a positive pulse, in turn, drives the tube into the grid-current region where the bias developed across R_3 limits the maximum plate current flow. Thus both positive and negative halves of the input waveform are sharply limited, only a fraction of the input voltage being transferred.

The resultant square-wave voltage across R_6 appears at position (A) on the 3-position selector switch, S_2. When the square wave is applied to the differentiating circuit, C_5R_7, the output of this circuit at position (B) is in the form of very short, sharp pulses. Integrating the square wave gives the triangular wave at position (C).

A cathode-coupled 6J5 output stage isolates the generator circuit from the output load. Since the output voltage from positions (B) and (C) is much less than at (A), an output level control, R_{10}, is required.

Pulse-generator circuit — The simple pulse generator in Fig. 2036 produces rectangular pulses which may be varied over a wide range of duration, voltage and frequency.

Sawtooth waves generated by the 884 relaxation oscillator are differentiated or narrowed and squared by the 6J7 limiting tube. Output is taken from the pair of polarity-reversing 6F6 clipper-amplifiers.

With the values shown for the frequency-controlling elements of the circuit, R_5 and C_1–C_9, a continuous frequency range of from 1 to 40,000 pulses per second is available. The pulse width control, S_2, provides a range of from 0.04 to 3 milliseconds with the time constants established by the capacity values listed. The output amplitude is controlled by R_{21}. S_3 is a reversing switch which changes the polarity of the output pulse without changing its waveform or requiring the shifting of connections.

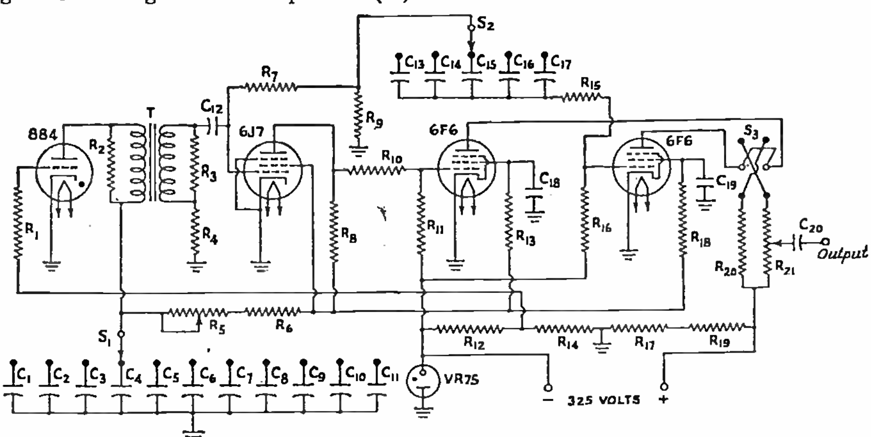


Fig. 2036 **— Wide-range pulse generator circuit.**

C1 — 2-μfd. 200-volt paper.
C2 — 1-μfd. 200-volt paper.
C3, C18, C19 — 0.5-μfd. 200-volt paper.
C4 — 0.25 μfd. 200-volt paper.
C5 — 0.1 μfd. 400-volt paper.
C6, C20 — 0.05-μfd. 400-volt paper.
C7 — 0.025-μfd. 400-volt paper.
C8, C13 — 0.01-μfd. mica.
C9 — 0.005 μfd. mica.
C10 — 700-μμfd. mica.
C11, C15 — 500-μμfd. mica.
C12 — 0.001-μfd. mica.
C14 — 0.0015-μfd. mica.
C16 — 250-μμfd. mica.
C17 — 100-μμfd. mica.
R1 — 25,000 ohms, ½-watt.
R2 — 1000 ohms, ½-watt.
R3 — 50,000 ohms, ½-watt, R4
R5 — 1-megohm variable.
R6 — 0.5 megohm, ½-watt.
R7, R8, R9 — 0.1 megohm, ½-watt.
R10, R15 — 1-megohm, ½-watt.
R11, R16 — 0.5 megohm, ½-watt.
R12 — 3000 ohms, 5-watt.
R13, R18 — 15,000 ohms, ½-watt.
R14 — 200 ohms, 1-watt.
R17 — 7500 ohms, 10-watt.
R19 — 300 ohms, 2-watt.
R20 — 50,000 ohms, ½-watt.
R21 — 5000-ohm variable.
S1 — 11-point rotary switch.
S2 — 5-point rotary switch.
S3 — D.p.d.t. toggle switch.
T — Interstage audio transformer.

Signal Generators

Test oscillators — An extremely simple test oscillator for receiver checking (§ 7-17) and similar uses which requires no external power supply is shown in Fig. 2037. The tetrode section of the 117L7GT is connected as a triode in a simple Hartley circuit, while the rectifier portion supplies plate voltage through a resistance-capacity filter. R.f. output is taken from C_5. A variable resistance r.f. attenuator may be added as in Fig. 2038, if desired. With additional plug-in coils, any required frequency range can be covered.

The more elaborate test oscillator circuit in Fig. 2038 consists of an e.c.o. with provision for suppressor-grid a.f. modulation. The output attenuator is a potentiometer so connected as to present a constant input resistance to the receiver.

For suppressor-grid modulation, apply approximately 10 volts of audio voltage (for 50 per cent modulation), as shown in the diagram. The suppressor-grid is biased 10 volts negative for modulated use; if an unmodulated signal is desired, the upper terminal may be grounded as indicated. This will increase the output

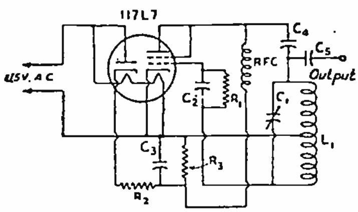


Fig. 2037 — Simple test oscillator for receiver alignment.

C_1 — 350-$\mu\mu$fd. variable.
C_2, C_5 — 250-$\mu\mu$fd. mica.
C_3 — 40-μfd. 150-volt electrolytic.
C_4 — 0.001 to 0.005 μfd. mica.
R_1 — 0.1 megohm, ½-watt.
R_2, R_3 — 50,000 ohms, ½-watt.
L — 440–510 kc.: 140 turns No. 30 e. close-wound on 1½-inch diameter plug-in form. Cathode tap at 35 turns from ground end.
1400–1550 kc.: 42 turns No. 20 d.s.c. on 1½-inch diameter form, tapped at 10 turns.
4500–5500 kc.: 11 turns No. 18 e. on 1½-inch diameter form, spaced wire diameter, tapped at 3 turns.
RFC — 10 mh. or larger r.f. choke.

from the oscillator. Conversely, if the output potentiometer does not attenuate the signal sufficiently, additional d.c. negative bias may be applied between the modulation terminals.

The oscillator should be shielded so that direct pick-up is minimized. Make all ground returns to a heavy copper strap connected to the cabinet at the output ground terminal. The plug-in coil should be separately shielded.

The i.f. ranges of the test oscillator can be calibrated by beating against signals of known frequency in the b.c. band. Frequencies between 465 kc. and 275 kc. can be spotted by using the second harmonic of the oscillator, the remainder of the range to 175 kc. being checked by using the third harmonic.

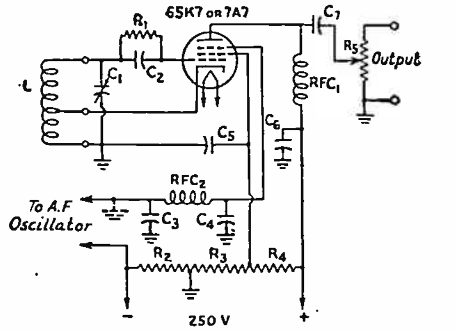


Fig. 2038 — Electron-coupled test oscillator circuit diagram.

C_1 — 100-$\mu\mu$fd. variable with 200-$\mu\mu$fd. fixed silver-mica zero-drift in parallel.
C_2 — 100-$\mu\mu$fd. midget mica.
C_3, C_4 — 250-$\mu\mu$fd. midget mica.
C_5 — 0.005-μfd. mica.
C_6 — 0.1-μfd. 400-volt paper.
C_7 — 500-$\mu\mu$fd. midget mica.
R_1 — 50,000 ohms, ½-watt.
R_2 — 2000 ohms, ½-watt.
R_3 — 20,000 ohms, 1-watt.
R_4 — 20,000 ohms, 2-watt.
R_5 — 500-ohm carbon potentiometer.
L_1 — See Fig. 2037.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 25-mh. r.f. choke.

The a.f. modulating source for the test oscillator can be any audio oscillator capable of delivering 10 to 20 volts at the standard receiver-checking frequency of 400 cycles.

A useful audio oscillator circuit is shown in Fig. 2039. It employs a two-terminal or "transitron" circuit (§ 3-7) using a pentagrid tube. A frequency of approximately 400 cycles is generated with the tuned-circuit values shown. A variable-frequency oscillator can be made by inserting a resistance, *R*, in the tuned circuit, between *L* and ground. The highest frequency available is determined by *L* and *C* alone, with *R* at zero. Increasing *R* will decrease the frequency. If *R* is made 5000 ohms, a frequency ratio of about 5 to 1 can be obtained. A good-quality wire-wound variable resistor should be used. If difficulty is had making the tube oscillate over the entire range, different values of R_1 and C_2 should be tried.

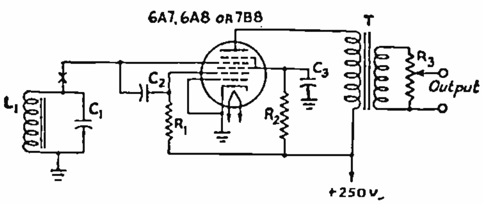


Fig. 2039 — Simple negative-resistance audio oscillator.

C_1 — 0.15-μfd. 400-volt paper.
C_2 — 0.1-μfd. 400-volt paper.
C_3 — 0.25-μfd. 200-volt paper.
R_1, R_2 — 50,000 ohms, 1-watt.
R_3 — 50,000-ohm volume control.
L_1 — 1.2-henry choke (Thordarson T-14C61 with iron core removed).
T — Output transformer (interstage audio, 1:3 ratio).

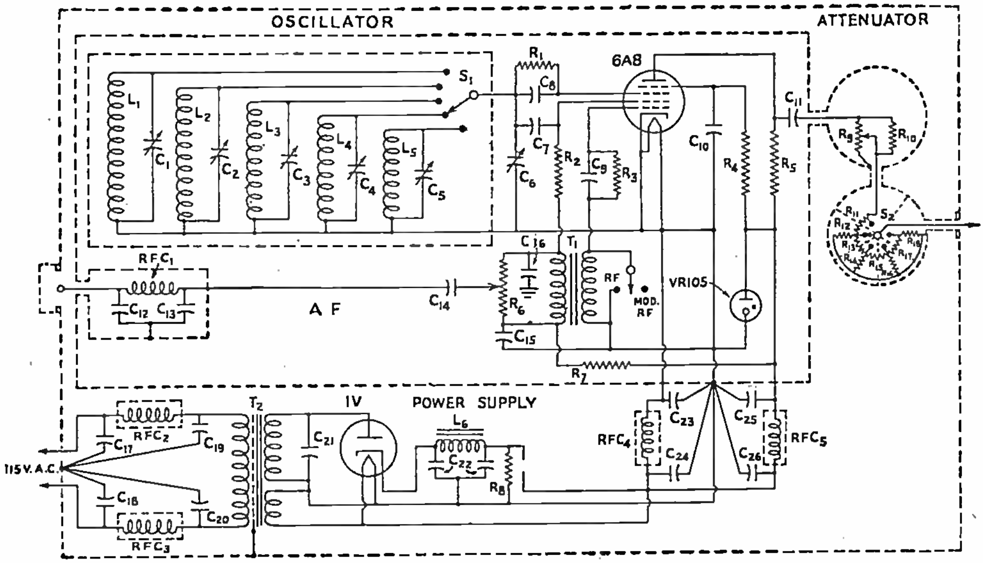


Fig. 2040 — Circuit diagram of a complete amplitude-modulated signal generator with resistance-step attenuator.

C_1, C_2, C_3, C_4, C_5 — 25-$\mu\mu$fd. air trimmers.
C_6 — 410-$\mu\mu$fd. variable (b.c. type with ceramic insulation).
C_7, C_9, C_{21} — 0.001-μfd. mica.
C_8 — 100-$\mu\mu$fd. mica.
C_{10}, C_{14} — 0.1-μfd. 400-volt paper.
C_{11} — 10$\mu\mu$fd. mica.
C_{12}, C_{13} — 250-$\mu\mu$fd. mica.
C_{15} through C_{26} — 0.01-μfd. 400-volt paper.
C_{16} — See text.
R_1, R_4 — 50,000 ohms, ½-watt.
R_2, R_5 — 25,000 ohms, ½-watt.
R_3 — 0.5-megohm variable.
R_6 — 50,000-ohm variable.
R_7, R_8 — 0.1 megohm, ½-watt.
R_9 — 500-ohm carbon variable.
R_{10}, R_{18} — 500 ohms, ½-watt.
R_{11}, R_{13}, R_{15}, R_{17} — 2500 ohms, ½-watt.
R_{12}, R_{14}, R_{16}, — 300 ohms, ½-watt.
L_1 — 2 mh. (455-kc. i.f. winding).
L_2 — 300μh. (200 turns No. 34 e).
L_3 — 30 μh. (62 turns No. 26 e).
L_4 — 5 μh. (25 turns No. 20 e).
L_5 — 0.6 μh. (7 turns No. 12 e).
All wound on ¾-inch diameter forms; winding length 1½-inches.
L_6 — 30-henry 40-ma. 500-ohm filter choke.
RFC_1 — 250-mh. r.f. choke.
RFC_2, RFC_3, RFC_4, RFC_5 — 125-mh. r.f. choke.
S_1 — 5-position ceramic rotary switch (shorting type).
S_2 — 5-position ceramic rotary switch (non-shorting type).
T_1 — Midget interstage audio transformer (1:3 ratio).
T_2 — Isolating power transformer, 115-volt and 6.3-volt a.c. secondaries.

Complete a.m. signal generator — A signal generator is a device for producing simulated radio signals of known characteristics, consisting of a thoroughly shielded oscillator which can be modulated, together with a calibrated output attenuator.

In the signal generator shown in Fig. 2040, a 6A8 pentagrid converter serves as both r.f. and a.f. oscillator. The basic r.f. oscillator circuit is of the two-terminal transitron type. Five tuned circuits enable continuous coverage of the frequency range from 170 kc. to 30 Mc.

A.f. modulation is by means of tuned-plate feed-back-type audio oscillations utilizing the oscillator grids of the 6A8 for plate and grid. The modulation frequency (usually 400 cycles) is determined by the value of C_{16}, which, depending on the primary inductance of T_1, may be between 500 $\mu\mu$fd. and 0.002 μfd. For unmodulated r.f., the modulation is removed by a switch short-circuiting the secondary of T_1. As a separate audio source, output is taken from the a.f. output terminal via C_{14}. External modulation may be applied through this terminal.

In a satisfactory signal generator the external field at minimum setting should not exceed 1 microvolt. This requires the following:

1) Spaced concentric shields, connected electrically at only one point, with a copper inner shield enclosing the r.f. coils and bandswitch, a larger copper or aluminum box around the oscillator unit, and a metal cabinet, which may be of steel, enclosing the whole assembly.

2) Shielding material having adequate wall thickness to prevent direct leakage, with well-made joints of good electrical conductivity so that eddy currents in the shield will not be disturbed. Where possible, joints and seams should be soldered; a literally water-tight joint is required for complete elimination of leakage. Removable covers must have a tight fit. The overlap between the lid and the box must be wide enough to ensure that leakage gaps are long compared with the aperture.

3) All ground connections returned to a common ground point, with by-pass condensers and d.c. returns arranged as shown in Fig. 2040.

4) Arrangement of circuits so the shields carry a minimum of current. Currents in the shield cause different parts to be at different potentials, greatly increasing the possibility of leakage. Insulated shafts must be used for tuning condensers and other controls in the inner compartments. Leakage will be further reduced if the insulated shaft is shielded by a metal tube connected to the external shielding. For maximum isolation the length of this cylinder should be several times its diameter.

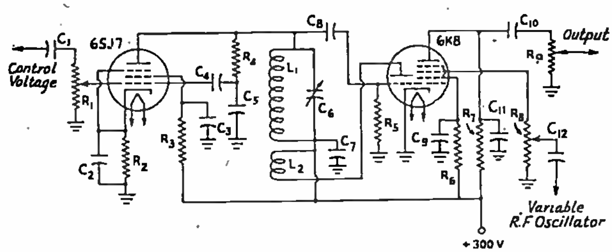


Fig. 2041 — Frequency-modulated beat-frequency signal generator or "wobbulator."

C_1 — 0.25-μfd. 200-volt paper.
C_2 — 0.01-μfd. 400-volt paper.
C_3 — 10-μfd. 200-volt electrolytic.
C_4 — 0.002-μfd. mica.
C_5 — 3-30-μμfd. mica trimmer.
C_6 — 50-μμfd. variable.
C_7, C_{10}, C_{12} — 0.005-μfd. mica.
C_8 — 250-μμfd. mica.
C_9 — 0.1-μfd. 200-volt paper.
C_{11} — 25-μμfd. mica.
R_1 — 0.5-megohm variable.
R_2 — 5000 ohms, ½-watt.
R_3, R_4 — 50,000 ohms, ½-watt.
R_5 — 0.1 megohm, ½-watt.
R_7 — 30,000 ohms, ½-watt.
R_8, R_9 — 25,000-ohm variable.
L_1 — 150 turns No. 28 e., 1½-inch diameter.
L_2 — 20 turns No. 28 e, 1½-inch diameter.

5) Isolating filters in outgoing leads, arranged so that the condensers are returned by short separate leads to the common ground point on each shield. Shielded condensers and coils help prevent currents being induced in the leads on the output side of the filter.

The external-modulation terminal is isolated by a low-pass filter which passes audio frequencies while attenuating the r.f.

Attenuator — The r.f. output attenuator must be capable of accurately controlling the output voltage from the signal generator without affecting the frequency of the oscillator.

The construction of an attenuator with an accuracy of 20 per cent above 10 Mc. requires great care. The resistance units must be non-reactive; above 50 ohms, metallized-filament carbon resistors are suitable. Stray capacity between sections and the inductance of connecting wires must be minimized. All leads should be reduced to minimum length, and all returns made directly to a common bus wire, grounded at the output ground terminal.

In a simple test oscillator such as that in Fig. 2039 an ordinary variable resistor (carbon, not wire-wound) may be used. For maximum attenuation it should be completely shielded and grounded at the output terminal.

Laboratory signal generators commonly use a ladder-type resistance-step attenuator, as shown in Fig. 2040. The multiplier switch, S_2, attenuates the output voltage by decimal (10 ×) steps, giving a theoretical attenuation range of 100,000 to 1. The voltage across the input of the multiplier is continuously varied by R_9, with R_{10} assisting to maintain a constant-impedance lead for the oscillator. Loose coupling between the attenuator and the oscillator, as well as compensation for falling off of the oscillator output at the higher frequencies, is obtained by using a very small coupling condenser (C_{14}). The input section, R_9R_{10}, is mounted in a copper shield can, connected through a short length of copper tubing to a second shield can which houses the multiplier switch, S_2. The fixed shunt resistors (R_{12}, R_{14}, etc.) are assembled vertically on the terminals of S_2 and isolated from each other by copper-strip shields soldered to the walls of the can. The series resistors (R_{11}, R_{13}, etc.) pass through clearance holes in these shields. The attenuator assembly is supported by additional lengths of copper tubing, soldered to the oscillator shield on the input side and to the coaxial output-line ground terminal on the panel. The resistor and the switch shafts should be insulated.

F.m. signal generators — Signal generators in which the oscillator is frequency- rather than amplitude-modulated are used in visual curve tracing with the aid of an oscilloscope.

In an f.m. signal generator the frequency of the oscillator is varied by means of a reactance tube, to the grid of which a 60-cycle sweep voltage is applied. The frequency deviation

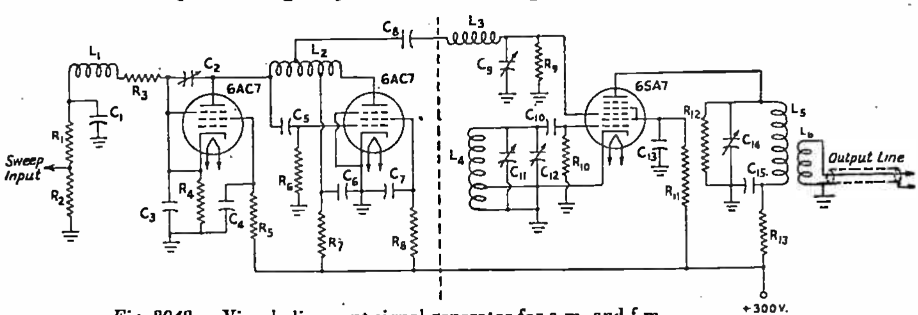


Fig. 2042 — Visual alignment signal generator for a.m. and f.m.

C_1, C_4, C_7 — 500-μμfd. mica.
C_2, C_9 — 3-30-μμfd. mica trimmer.
C_3, C_5, C_8, C_{10} — 100-μμfd. mica.
C_6 — 250-μμfd. mica.
C_{11}, C_{14} — 100-μμfd. variable.
C_{12} — 25-μμfd. variable.
C_{13} — 0.1-μfd. 200-volt paper.
C_{15} — 0.001-μfd. mica.
R_1, R_7 — 1000 ohms, ½-watt.
R_2 — 2 megohms, ½-watt.
R_3 — 100 ohms, ½-watt.
R_4 — 500 ohms, ½-watt.
R_5 — 3500 ohms, ½-watt.
R_6, R_{12}, R_{13} — 10,000 ohms, ½-watt.
R_8 — 25,000 ohms, ½-watt.
R_9, R_{10} — 0.1 megohm, ½-watt.
R_{11}, — 50,000 ohms, ½-watt.
L_1 — 3 turns No. 32 e., ¾-inch diameter.
L_2 — 31 turns No. 20, ¾-inch diam., 2 inches long; tapped 2¼ and 4¼ turns from grid end.
L_3 — 35 turns No. 32e., ½-inch diameter, ½-inch long.
L_4 — 6 turns No. 14, ¾-inch diameter, ¾-inch long.
L_5, L_6 — Plug-in coils to tune to desired i.f. or r.f. output frequency.

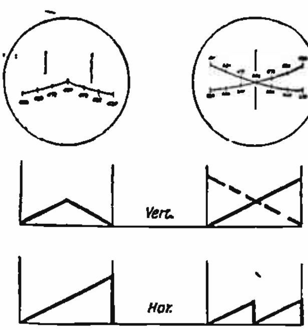


Fig. 2043 — Left — "Single-trace" and, *right*, "double-trace" methods of alignment. In the double-trace system the sweep frequency is twice the modulating frequency, resulting in a pattern showing two superimposed selectivity curves.

must remain constant, being determined only by the amplitude of the control voltage applied to the reactance tube. To accomplish this the desired output frequency is obtained by heterodyning the f.m. signal with a variable-frequency unmodulated oscillator and obtaining the difference frequency from a mixer tube. Apart from maintaining the frequency deviation constant regardless of the output frequency, a general improvement in stability results from using the beat-frequency principle.

Fig. 2041 shows a circuit suitable for general receiver alignment, employing a 6SJ7 reactance tube and a 6K8 mixer tube, the triode section of which serves as the f.m. oscillator. The low-frequency sweep source and the variable r.f. oscillator source are obtained externally. The f.m. frequency oscillator operates on 1000 kc., so that the external oscillator must have a range such as to supply any desired difference frequency. A conventional a.m. signal generator or test oscillator will serve the purpose. The sweep voltage should have a triangular waveshape, obtainable from a generator such as that of Fig. 2039.

In the high-frequency f.m. signal generator shown in Fig. 2042, a self-contained variable-frequency oscillator and mixer (shown to the right of the dotted line) are included. The fixed-frequency f.m. oscillator functions on approximately 20 Mc. Only the capacity of the tubes and wiring is used across the tank-circuit coil, L_2. The resulting high L/C ratio makes possible a large frequency variation. The frequency-modulated signal is heterodyned with the variable signal in the 6SA7 mixer. C_{12} and C_{14} may be ganged and the tuning calibrated directly in terms of the output difference frequency, if desired.

Receiver Characteristics

Measurements in connection with receiving equipment come under two heads: (1) over-all performance, and (2) servicing and alignment. While the measurement of receiver performance requires precision equipment, sufficient apparatus for servicing receivers (§ 7-17) should be available in every amateur station. The first requirement is a multi-range volt-ohm-milliammeter. For aligning tuned circuits a test oscillator is required, preferably one which can be modulated at 400 cycles. A rectifier-type voltmeter can be used as an output meter.

Fig. 2044 shows in block diagram form five basic arrangements for checking the characteristics of receivers, based on the principles outlined in Chapter Seven (§ 7-2, 7-17, 7-18).

The basic arrangement for i.f and r.f. alignment is shown at A. The test signal is applied through a dummy antenna — the 400-ohm series resistor, R, which should be of the metallized carbon-filament type.

In Fig. 2044-B an oscilloscope is used as a voltmeter to indicate the i.f. output level at the second detector. The oscilloscope connection should be made directly to the vertical plates to minimize frequency discrimination. An unmodulated carrier will appear as a thin straight line. Keeping the height of the trace constant by varying the attenuator setting of the signal generator as required, the gain of one or more stages, or the over-all gain of the receiver, can be established more accurately than by the output-meter method.

Figs. 2044-C, -D and -E show various ways of producing visual response curves (§ 7-17), based on the following combinations:

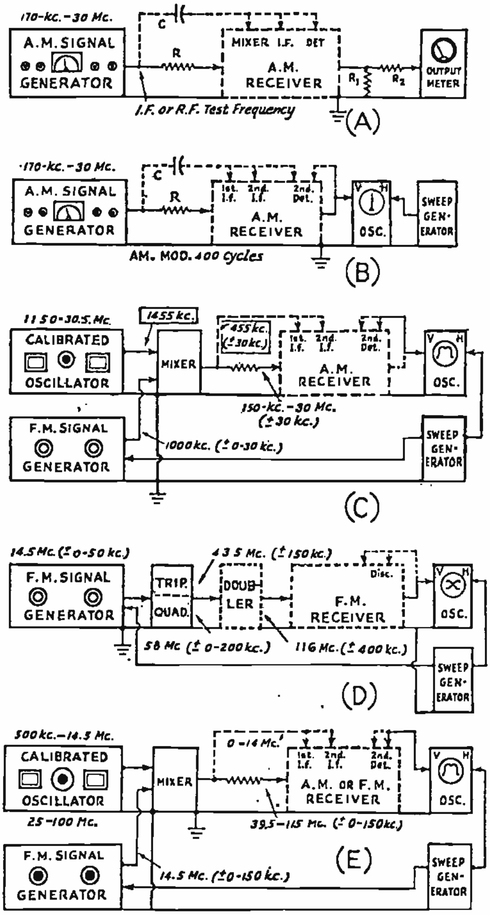


Fig. 2044 — Block diagrams of receiver testing methods.

(C) 1000-kc. f.m. oscillator (Fig. 2041); variable-frequency heterodyne oscillator (Fig. 1239 or 1240); mixer (in Fig. 2041); oscilloscope (Fig. 2033); and sweep-frequency source (Fig. 2036), for triangular wave sweep in the f.m. oscillator and the pulse output connected to the "ext. sync." terminals on the oscilloscope.

(D) 14.5-Mc. f.m. oscillator (the f.m. oscillator portion of Fig. 2042, with L_2 tuned to 14.5 Mc. by a small air trimmer and the frequency deviation adjusted to allow for multiplication on the various frequency ranges.) The 14.5-Mc. fundamental will provide harmonics in all present amateur bands where f.m. is allowed (14.5 × 2 = 29 Mc. 14.5 × 4 = 58 Mc., etc.) as well as the f.m. broadcast band.

(E) Beat-frequency type f.m. signal generator (Fig. 2042) with integral variable oscillator for general a.m. and f.m. coverage; the remainder of the combination as in (C) above.

Amplifier/Modulator Characteristics

Gain and power output — Block diagrams of arrangements for measuring the amplification and frequency response of audio-frequency amplifiers, following the principles given in § 3-3, 5-9, 7-5, and 7-17, are shown in Figs. 2046-A, -B, and -C.

At A, a calibrated audio-frequency source, usually a beat-frequency or resistance-capacity tuned audio oscillator (Fig. 2021) delivers a known voltage to a calibrated attenuator, on which the actual input to the stage under measurement is read. The input level is set to produce the specified output level reading on the output meter, which is arranged to match the load impedance of the amplifier, at a number of frequencies throughout the range. The input-attenuator setting required to produce the specified output at each frequency is noted, the ratio of input to output voltages giving the gain of the amplifier at each frequency.

A more satisfactory method for plotting frequency-response curves is shown at B. The sensitivity of the output meter is adjusted to read the input voltage required to produce the specified output at a given reference frequency (usually 400 cycles). The variable attenuator connected across the output load is so adjusted that, when the output meter is switched, input and output readings coincide at the reference frequency. Varying the input frequency and plotting the resulting output variation above or below the reference level will give an accurate response curve without further manipulation of the attenuator.

The alternative method at C is useful with high-gain amplifiers where the input signal level is too low to be read accurately on the output meter. In this case a large voltage comparable to the output level is delivered by the signal source to a voltage divider so proportioned that R_1/R_2 equals the ratio of input to output voltage. The input frequency is manually varied output variations above or below the reference level recorded as above.

In the diagram a tapped transformer and adjustable load resistor are shown. These represent the impedance-matching system, external to the oscilloscope, which normally would be incorporated in the output meter.

Phase shift — When an oscilloscope is used as a comparison voltmeter, as shown in Fig. 1246-C, equality between the voltages applied to the vertical and horizontal plates (i.e., the source and output voltages) is indicated by equal deflection of the cathode-ray tube beam; ideally, the resulting trace is a straight line at an angle of exactly 45°.

Actually, however, this will be the case only if the phase angle (§ 2-7) between input and output is precisely 0° or 180°. In practice, some phase shift (§ 3-3) is to be expected, altering the pattern from a straight line to an elongated ellipse. (Fig. 2018). By shunting a fixed condenser across the circuit and observing the direction of change in the pattern, the kind of reactance can be determined. A leading phase angle indicates capacitive reactance; lagging phase means inductive reactance.

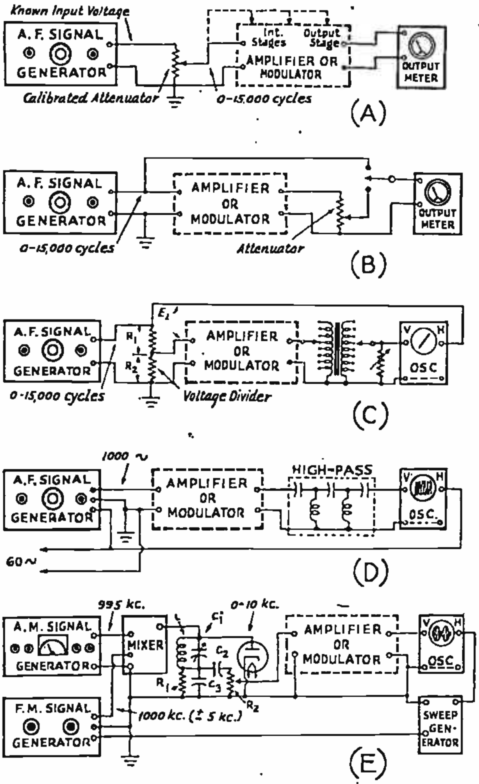


Fig. 2046 — Block diagrams illustrating various audio-frequency amplifier testing and measuring arrangements.

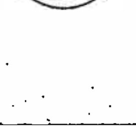
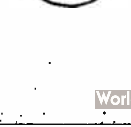


Fig. 2045 — *Left* — Amplitude-modulated r.f. carrier at detector. *Right* — F.m. carrier and sidebands, frequency deviation of two.



Intermodulation — Apart from frequency, amplitude and phase distortion, another less obvious but even more troublesome type, called *intermodulation,* may accur as a result of nonlinearity or rectification in the amplifier.

Fig. 2046-D shows a method of checking for this form of distortion, employing two-frequency input. The low-frequency source is the 60-cycle a.c. supply, the input amplitude being adjusted just below the overload point of the amplifier. The amplitude of higher frequency (1000 cycles) from the audio generator is kept well under this level. A high-pass filter with a cut-off frequency of 400 cycles removes the 60-cycle component from the amplifier output, so that the high-frequency signal appears alone on the screen of the oscilloscope.

Fig. 2047 — Audio amplifier frequency-response characteristics viewed on a visual curve tracer. Trace is reversed by out-of-phase input.

If the amplifier is entirely linear at low frequencies, the 1000-cycle signal will appear on the screen as a rectangular trace of uniform amplitude. If nonlinearity exists, however, the pattern will have a sloping contour resulting from the sinusoidal 60-cycle modulation. An element having a symmetrical nonlinear characteristic, such as an overloaded pentode, will flatten both positive and negative peaks, producing a waveshape that slopes off at either end. An asymmetrical rectifier, such as an over-biased triode, will cause curvature only at one end. If the distortion occurs in an intermediate stage of any point differing in phase from the input signal, the wave envelope divides and assumes an elliptical shape.

Instantaneous curve tracing — The procedures described above for obtaining frequency-response curves require laborious point-by-point plotting. Fig. 2046-E shows the application of an f.m. signal generator, beating against a fixed-frequency oscillator, as an instantaneous visual audio curve tracer.

The two oscillators, one fixed at 995 Mc. and the other swept (at a 60-cycle rate) between 995 and 1005 kc., produce a difference frequency varying between 0 and 10 kc. The resulting a.c. voltage is rectified by the linear diode detector, becoming pulsating d.c. which controls the vertical deflection of the cathode-ray beam.

The image departs in appearance from the familiar plotted curves in that the amplitude scale is linear rather than logarithmic. Thus the vertical amplitude is directly proportional to the output voltage, accentuating the shape over the usual db. scale. On the horizontal scale, the center frequency (5000 cycles) is one-half the maximum deviation.

Using this method, a double trace results, as shown in Fig. 2047. The two curves may begin with either 0 or 10 kc. in the center, depending on the input vs. output phase relationship.

Transmitter Characteristics

The transmitter characteristics ordinarily requiring measurement are d.c., a.c. and r.f. voltages and currents, keying and modulation quality, and modulation percentage. Instruments for the measurement of voltages and currents have been discussed. Keying and modulation checks may be made by several methods; the two commonly used by amateurs are aural checks with monitors and visual checks with the oscilloscope (§ 5-10).

Monitors — A monitor is a miniature receiver, usually having only a single tube, enclosed with its batteries in some sort of metal box which serves as a shield. The requirements for a satisfactory monitor for checking c.w. signals are not difficult to satisfy. It should oscillate steadily over the bands on which the station is to be active; the tuning should not be excessively critical, although the degree of bandspread ordinarily considered desirable for receivers is not essential; the shielding should be complete enough to permit the monitor to be placed near the transmitter and yet give a good beat note when tuned to the fundamental frequency of the transmitter (often impossible with a receiver because the pick-up is too great); and it should be constructed solidly, so that it can be moved around without necessity for retuning.

Any simple regenerative detector covering the required amateur bands may be used for a monitor. Any 1.5- or 2-volt filament-type triode may be used, while the power supply may be dry batteries of a size that will fit into the container selected. A plate-tickler shorting switch will make the monitor nonoscillating when checking 'phone signals. If desired, a regeneration control may be incorporated.

Any simple detector with a means for picking up r.f. from the transmitter can be used as a 'phone monitor. A satisfactory monitor may be made using a simple diode rectifier and an untuned pick-up coil. Headphones are used for listening checks. The monitor may also be employed as an overmodulation indicator by use of a 0–1 milliammeter. The pick-up coil should be loosely coupled to the tank circuit of the final r.f. amplifier until the milliammeter reads approximately 0.9 ma. If the speech amplifier is supplied with a 400-cycle sine-wave tone from an audio oscillator, such as that shown in Fig. 2039, and its gain control turned up until the monitor meter starts to rise, when overmodulation is indicated.

If a copper-oxide rectifier-type a.c. milliammeter is used and the transmitter modulated by a steady tone or speech signal, the modulation percentage will be 140 times the reading of the meter; e.g., for 100 per cent modulation the meter should read approximately 0.7 ma. When measuring percentage of speech modulation, inertia will cause the meter to undershoot on peaks; the swing should, therefore, be limited to somewhat less than 0.7 ma.

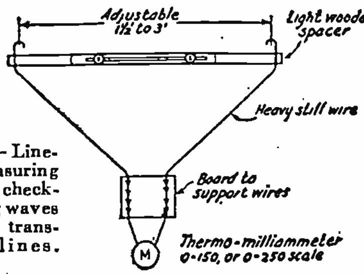


Fig. 2048 — Line-current measuring device for checking standing waves on untuned transmission lines.

Antenna Measurements

Antenna measurements are made for the purpose (a) of securing maximum transfer of power to the antenna from the transmitter, and (b) of adjusting directional antennas to conform with design conditions. Measurements are therefore made of the current (power) in the antenna, voltage and current relationships, resistance, and radiated field intensity. Related to measurements of the antenna system proper is the measurement of the various transmission-line characteristics, chiefly involving impedance and resistance.

The instruments described for r.f. measurement (thermocouple ammeter, vacuum-tube voltmeter, *L*, *C* and *Q* meter) all are applicable in making antenna measurements.

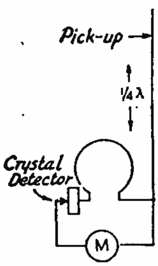


Fig. 2048 — A simple indicator for r.f. at v.h.f.

A number of instruments have been devised for the purpose of measuring stationary waves, the most common being an instrument to run along one wire of the feeder line, and consisting simply of a hot-wire ammeter bridging a portion of the line.

Checking the transmission line for standing waves can be done by measuring the current in the wires, using a device of the type pictured in Fig. 2048. The hooks (which should be sharp enough to cut through the insulation, if any, on the wires) are placed on one of the wires, the spacing between them being adjusted to give a suitable reading on the meter. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a quarter wavelength will indicate whether or not standing waves are present.

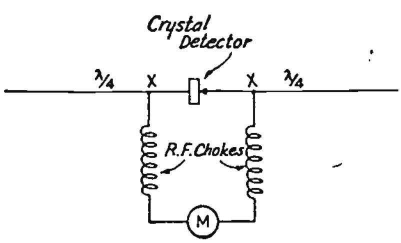


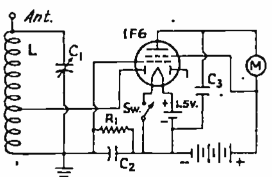
Fig. 2049 — Simple field-strength meter which may also be used as a sensitive indicator when making frequency measurements by connecting Lecher wires at X-X.

Field-intensity meters — In adjusting antenna systems for maximum radiation and in determining radiation patterns, use is made of field-intensity meters. Fundamentally the field-intensity meter is a vacuum-tube voltmeter provided with a tuned input circuit. It is used to indicate the relative intensity of the radiation field under actual *radiating* conditions. It is particularly useful on the very-high frequencies and in adjusting directional antennas. Field-intensity checks should be made at points several wavelengths distant from the antenna and at heights corresponding with the desired angle of radiation.

The absorption frequency meter shown in Figs. 2002–2003 may be used as a field strength meter if it is provided with a pick-up antenna. This can be short length of brass rod connected to the stator of the tuning condenser through a small trimmer. Alternatively, a simple loop can be used as shown in Fig. 2048. The crystal detector does not have a linear characteristic, and a given increase in rectified current does not, therefore, indicate a directly proportional increase in field strength.

As shown in Fig. 2049, a crystal detector in the center of a half-wave pick-up antenna, coupled through r.f. chokes to a 0-1 ma. meter, will serve as a field-strength meter of ample sensitivity. When a signal is tuned in rectification occurs, and the rectified current is read on the microammeter. With this type of indicator, good readings may be obtained at distances of four or five wavelengths from a low-power transmitter with only a few watts input.

Fig. 2050 — Sensitive diode-triode field-intensity meter.



C_1 — 50-μμfd. midget variable.
C_2 — 250-μμfd. midget mica.
C_3 — 0.002-μfd. midget mica.
R_1 — 1 megohm, ½-watt.
M — 0–500 μa. d.c. microammeter.

L — 1.5– Mc.: 58 turns No. 28 d.s.c., close-wound.
3–6 Mc.: 29 turns No. 20 e., close-wound.
6–12 Mc.: 15 turns No. 20 e., spaced.
11–22 Mc.: 8 turns No. 20 e., spaced.
20–40 Mc.: 4 turns No. 20 e., spaced.
All wound on 1½-inch coil forms, winding length 1½ inches; diode tap in center of coil.

A more sensitive field-intensity meter of use in examining the field-strength patterns of lower-frequency antenna systems, employing a diode rectifier and d.c. amplifier in the same envelope, is shown in Fig. 2050. The initial plate current reading is about 1.4 ma.; with signal input, the current dips downward. The scale reading is linear with signal voltage. Radiated power variations will, of course, be as the square of the field-voltage indication.

Power gain in antenna systems usually is expressed in terms of decibels. A field-intensity meter which reads directly in db. is shown in Fig. 2051. It consists of a self-biased linear

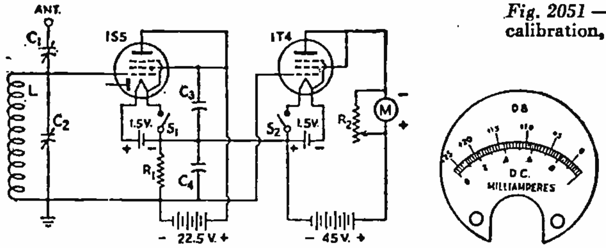


Fig. 2051 — A logarithmic field-intensity meter with db. calibration, employing a variable-μ d.c. amplifier tube.

C_1 — 3–30-μμfd. mica trimmer.
C_2 — 50-μμfd. midget variable.
C_3, C_4 — 500-μμfd. midget mica.
R_1 — 10 megohms, ½-watt.
R_2 — 1000-ohm wire-wound.
S_1, S_2 — Two s.p.s.t. toggle switches or a single d.p.s.t.
L — See the coil data given under Fig. 2050.
M — 0–1 ma. d.c. milliammeter.

triode voltmeter followed by a variable-μ d.c. amplifier tube. Because of the nearly logarithmic grid-voltage/plate-current characteristic of this tube, a 0-1 ma. milliammeter in its plate circuit can be calibrated arbitrarily with a linear db. scale, as shown. For extreme accuracy an individual calibration should be made, applying known values of a.c. voltage to the 155 grid. The arbitrary scale shown will be found sufficiently accurate to be useful, however.

The scale covers approximately 25 db. and is linear over a range of about 20 db. At very small signals it departs from linearity, and therefore 0 db. is placed at 90 per cent of the scale. A variable meter shunt compensates for variations in tubes and battery voltages. In use, the balancing resistor is adjusted to give a full-scale reading of 1 ma. The signal pick-up is then made such as to cause the meter to indicate 0 db. Alternatively, the initial reading may be set arbitrarily at 10 db.; adjustments will then be indicated as losses or gains in relation to that figure.

The range of the instrument may be extended to +45 db. by inserting a 2-point tap switch in the connecting lead to the 1T4 amplifier from the self-biasing resistor R_1 and tapping that resistor by adding a1-megohm unit to provide a 10-to-1 multiplier. Add 20 db. to all readings when the multiplier is used.

Vacuum Tube Characteristics

The best check on a receiving or transmitting tube is by a direct comparison in its own socket with a new tube of known quality under actual operating conditions. Any other method of testing a tube falls short of an actual performance test.

For convenience, however, an auxiliary tube checker is desirable. A number of commercial tube checkers of the type used by servicemen are on the market. In purchasing one, the following qualifications should be sought: (1) complete facilities for checking shorts between any pair of electrodes; (2) a transconductance rather than an "emission" test (the emission of a tube may vary widely with no effect on its performance, while genuinely faulty tubes may show rated emission); (3) provision for checking plate and screen currents under typical conditions (at rated voltages); (4) gas and noise tests.

The construction of a comprehensive tube-checker is an elaborate project. However, for an occasional need the amateur can assemble a circuit using an existing power source, as in Fig. 2052, which will serve to make a reasonably accurate standard transconductance test.

A pentode tube is shown; for other types, omit or add electrode connections as required. The electrode voltages applied should correspond with those listed under "Typical Operating Conditions" in the vacuum-tables in the Appendix. The voltages should be accurate to within 5 per cent (especially grid voltage, plate voltage for triodes and screen voltage for pentodes): With the switch in the No. 2 position, the plate- and screen-current readings should be near the rated values; wide variations from normal indicate a defective tube.

To make the transconductance test, note the plate current with the grid switch alternately on positions 3 and 1, which changes the bias from exactly 0.5 volt less than rated bias to exactly 0.5 volt more. The resulting plate current change multiplied by 1000 ($\Delta I \times 10^3$) equals the transconductance in micromhos. This value should be checked against the vacuum-tube tables. Tubes usually will operate satisfactorily until the transconductance falls to about 70 per cent of rating.

Pentagrid and heptode frequency converters may be checked by this same method if the rated d.c. electrode voltages are applied. The oscillator section of a mixer tube may be checked separately by noting the oscillator-anode current change.

Diodes can be checked by applying 50 volts of 60-cycle a.c. between plate and cathode, in series with a 0.25-megohm load shunted by a 2-μfd. condenser. The resulting rectified current may be read on a 0–1 ma. d.c. meter. A reading of 0.2 to 0.25 ma. indicates a satisfactory diode unit.

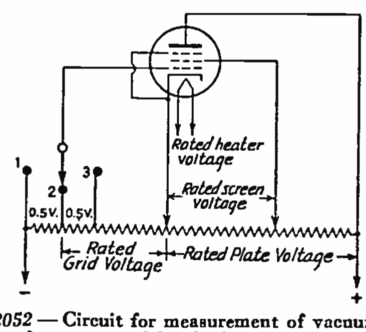


Fig. 2052 — Circuit for measurement of vacuum-tube transconductance, used for checking condition of tubes.

Chapter Twenty-One

Amateur Radio

COUNTLESS thousands of persons all over the world have enjoyed the thrills and pleasures of amateur radio. The story of how this personalized communication came into being and how it grew into one of the most universally beneficial activities on earth is a fascinating one. This chapter recounts that story, tells what amateur radio is, and how one may become an amateur.

Amateur radio is as old as the art itself.

There were amateurs before the present century. Shortly after the late Guglielmo Marconi astounded the world with his first experiments proving that telegraph messages actually could be sent between distant points without wires, they were attempting to duplicate his results. Marconi himself probably was the first amateur — indeed, the distinguished inventor so liked to style himself. But amateur radio, as it has come to be known, was born when private citizens first saw in the new marvel a means for personal communication with others, and set about learning enough of the new art to build homemade stations.

Amateur radio's subsequent development may be divided into two periods: the period up to our entrance into World War I, in 1917, and the period between that war and our entrance into the present conflict — 1919–1941.

Amateur radio prior to 1917 bore little resemblance to radio as we know it today, except in principle. The equipment, both transmitting and receiving, was of a type now long obsolete. No United States amateur had ever heard the signals of a foreign amateur, nor had any foreigner ever reported hearing an American. The oceans were a wall of silence, impenetrable, isolating us from every signal abroad. Even cross-country communication could be accomplished only by relays. "Short waves" meant 200 meters; the entire wavelength spectrum below 200 meters was a vast silence — no signal ever disturbed it. Years were to pass before its phenomenal possibilities were to be suspected.

Yet the period was notable for a number of accomplishments. It saw the number of amateurs in the United States increase to approximately 4,000 by 1917. It witnessed the first appearance of radio laws, licensing, wavelength specifications for the various services. ("Amateurs? — oh, yes — well, stick 'em on 200 meters; they'll never get out of their backyards with it.") It saw an increase in the range of amateur stations to such unheard-of distances as 500 and, in some cases, even 1,000 miles. U. S. amateurs were beginning to wonder, just before the war, if there were other amateurs in other countries across the seas and if — daring thought! — it might some day be possible to span the Atlantic with 200-meter equipment. Because all long-distance messages had to be relayed, this period saw relaying developed to a fine art — an ability that turned out to be a priceless accomplishment later when the government suddenly needed hundreds of skilled operators for war service in 1917. Most important of all, the period witnessed the birth of the American Radio Relay League, the amateur organization whose fame was to travel to all parts of the world and whose name was to be virtually synonymous with subsequent amateur progress and short-wave development. Conceived and formed by the famous inventor and scientist, the late Hiram Percy Maxim, the League was formally launched in early 1914. It was just beginning to exert its full force in amateur activities when the United States declared war in 1917, by that act sounding the knell for amateur radio for the next two and one-half years. By presidential direction, every amateur station was dismantled. Within a few months three-fourths of the amateurs of the country were serving with the armed forces of the United States as operators and instructors — a movement that was to be duplicated in striking manner a quarter of a century later.

Few amateurs today realize that World War I not only marked the close of the first phase of amateur development but came very near marking its end for all time. The fate of amateur radio was in the balance in the days immediately following the signing of the Armistice, in 1918. The government, having had a taste of supreme authority over all communications in wartime, was more than half inclined to keep it; indeed, the war had not been ended a month before Congress was considering legislation that would have made it impossible for the amateur radio of old ever to be resumed. President Maxim rushed to Washington, pleaded, argued; the bill was defeated. But there was still no amateur radio; the war ban continued in effect. Repeated representations to Washington met only with silence; it was to be nearly a year before licenses again were issued.

In the meantime, however, there was much to be done. The League's offices had been closed for a year and a half, its records stored away. Three-fourths of the former amateurs had gone to France; many of them would never come back. Would those who had returned be interested, now, in such things as amateur radio? Mr. Maxim determined to find out, and called a meeting of such members of the old board of directors of the League as he could locate. Eleven men, several still in uniform, met in New York and took stock of the situation. It wasn't very encouraging: amateur radio still banned by law, former members of the League scattered no one knew where, no organization, no membership, no funds. But those eleven men financed the publication of a notice to all the former amateurs that could be located, hired Kenneth B. Warner as the League's first paid secretary, floated a bond issue among old League members to obtain money for immediate running expenses, bought the magazine *QST* to be the League's official organ, and dunned officialdom until the wartime ban was lifted and amateur radio resumed again. Even before the ban was lifted, in October, 1919, old-timers all over the country were flocking back to the League, renewing friendships, planning for the future. When licensing resumed, there was a headlong rush to get back on the air.

From the start, postwar amateur radio took on new aspects. Wartime needs had stimulated technical development in radio. There were new types of equipment. The vacuum tube was being used both for receiving and transmitting. Amateurs immediately adapted the new apparatus to 200-meter work. Ranges promptly increased; it became possible to bridge the continent with but one intermediate relay. Soon stations on one coast were hearing those on the other, direct!

These developments had an inevitable result. Watching DX come to represent 1,000 miles, then 1,500 and then 2,000, amateurs began to dream of trans-Atlantic work. Could they get across? In December, 1921, the ARRL sent abroad one of its most prominent amateurs, Paul F. Godley, with the best amateur receiving equipment available. Tests were run, and thirty American amateur stations were heard in Europe. The news electrified the amateur world. In 1922 another trans-Atlantic test was carried out; this time 315 American calls were logged by European amateurs and, what was more, one French and two British stations were heard on this side.

Everything now was centered on one objective: two-way communication across the Atlantic by amateur radio! It *must* be possible — but somehow they couldn't quite make it. Further increases in power were out of the question; many amateurs already were using the legal maximum of one kilowatt. Better receivers? They already had the superheterodyne; it didn't seem possible to make any very great advance in that direction.

Then how about trying another wavelength, they asked? What about those wavelengths *below* 200 meters? The engineering world thought they were worthless — but then, that had been said about 200 meters too. There have been many wrong guesses in history. And so, in 1922, the assistant technical editor of *QST* (Boyd Phelps, now a commander in the United States Naval Reserve) carried on tests between Hartford and Boston on 130 meters. The results were encouraging. Early in 1923 the ARRL sponsored a series of organized tests on wavelengths down to 90 meters, and it was noted that as the wavelength dropped the reported results were better. A growing excitement began to filter into the amateur ranks.

Finally, in November, 1923, after some months of careful preparation, two-way amateur communication across the Atlantic became a reality, when Schnell, 1MO, and Reinartz, 1XAM (now W9UZ and W3IBZ respectively, and both commanders in the Naval Reserve), worked for several hours with Deloy, 8AB, in France, all three stations using a wavelength of 110 meters! Additional stations dropped down to 100 meters and found that they, too, could easily work two-way across the Atlantic. The exodus from the 200-meter region started.

By 1924 the entire radio world was agog, and dozens of commercial companies were rushing stations into the 100-meter region. Chaos threatened, until the first of a series of national and international radio conferences partitioned off various bands of frequencies for the different services clamoring for assignments. Although thought still centered on 100 meters, League officials at the first of these conferences, in 1924, came to the conclusion that the surface had only been scratched, and wisely obtained amateur bands not only at 80 meters but at 40 and 20 and 10 and even 5 meters.

Many amateurs promptly jumped down to the 40-meter band. A pretty low wavelength, to be sure, but you never could tell about these short waves. "Forty" was given a try, and responded by enabling two-way communication with Australia, New Zealand and South Africa.

How about 20 meters? This new band immediately showed entirely unexpected possibilities by enabling an East Coast amateur to communicate with another on the West Coast, direct, at high noon. The dream of amateur radio — daylight DX! — had come true.

From that time to the advent of World War II — when amateur radio again was closed down "for the duration" — represents a period of unparalleled accomplishment. The short waves proved a veritable gold mine. Country after country came on the air, until the confusion became so great that it was necessary to devise a system of international intermediates in order to distinguish the nationality of calls. The League began issuing what are known as WAC certificates to stations proving that they had worked all the continents. Over five thousand such certificates have been issued. Representatives of the ARRL went to Paris and deliberated with the amateur representatives of twenty-two other nations. On April 17, 1925, this conference formed the International Amateur Radio Union — a federation of national amateur societies.

Amateur radio is one of the finest of hobbies, but this fact alone would hardly merit such whole-hearted support as was given it by the United States government at past international conferences. There must be other reasons to justify such backing. One of these is a thorough appreciation by the Army and Navy of the value of the amateur as a source of skilled radio personnel in time of war. The other is best described as "public service."

We have already seen 3,500 amateurs contributing their skill and ability to the American cause in the Great War. After the war it was only natural that cordial relations should prevail between the Army and Navy and the amateur. Several things occurred in the next few years to strengthen these relations. In 1924, when the U. S. dirigible *Shenandoah* made a tour of the country, amateurs provided continuous contact between the big ship and the ground. In 1925, when the United States battle fleet made a cruise to Australia and the Navy wished to test out short-wave apparatus for future communication purposes, it was the League's Traffic Manager who was in complete charge of an experimental amateur-type set on the U.S.S. *Seattle* and proved for all time the superiority of the high frequencies.

Definite friendly relations between the amateur and the armed forces of the Government were cemented in 1925. In this year both the Army and the Navy came to the League with proposals for amateur coöperation. The radio Naval Reserve and the Army-Amateur Net are the outgrowth of these proposals. Thousands of amateurs in the Naval Reserve now are on active duty with the Navy, from the rank of captain on down, while other thousands are serving in the Army, Air Forces, Coast Guard and Marine Corps. Altogether, more than 25,000 radio amateurs are in the armed forces of the United States, while many more thousands are engaged in vital electronic research, development and manufacturing.

The public service record of the amateur is a brilliant one. These services can be roughly divided into two classes: emergencies and expeditions. It is regrettable that space limitations preclude detailed mention of amateur work in both these classes, for the stories constitute highlights of amateur accomplishment.

Amateur coöperation with expeditions goes back to 1923, when a League member, Don Mix of Bristol, Conn., (now acting technical editor of *QST*) accompanied MacMillan to the Arctic on the schooner *Bowdoin* with an amateur station. Amateurs in Canada and the United States provided the home contact. The success of this venture was such that other explorers made inquiry of the League regarding similar arrangements for their journeys. In 1924 another expedition secured amateur coöperation; in 1925 there were three, and by 1928 the figure had risen to nine for that year alone; altogether, during subsequent years, a total of perhaps two hundred voyages and expeditions were thus assisted.

Since 1913, amateur radio has been the principal, and in many cases the only, means of outside communication in more than one hundred storm, flood and earthquake emergencies in this country. Among the most noteworthy were the Florida hurricanes of 1926, 1928 and 1935, the Mississippi and New England floods of 1927, and the California dam break of 1928. During 1931 there were the New Zealand and Nicaraguan earthquakes, and in 1932 floods in California and Texas. Outstanding in 1933 was the earthquake in southern California. In 1934 further floods in California and Oklahoma resulted in notable amateur accomplishment. The 1936 eastern states flood, the 1937 Ohio River valley flood, and the 1938 southern California flood and Long Island-New England hurricane disaster saw the greatest emergency effort ever performed by amateurs. In these disasters and many others — tornadoes, sleet storms, forest fires, blizzards — amateurs played a major rôle in the rescue work and earned wide commendation for their resourcefulness in effecting communication where all other means had failed.

During 1938 the ARRL inaugurated a new emergency-preparedness program, providing for the appointment of regional and local Emergency Coördinators to organize amateur facilities and establish liaison with other agencies. This was in addition to the registration of personnel and equipment in the Emergency Corps. A comprehensive program of coöperation with the Red Cross, Western Union and others was put into effect. During the three years before the wartime close-down of amateur activity, this emergency organization proved the effectiveness of its planning and the proficiency of its personnel in more than a dozen important emergencies.

Although normal participation in such activity now is impossible, because of restrictions on amateur operation, the peculiar ability of the amateur to perform in such work has been notably recognized by the government in providing for amateur participation in the War Emergency Radio Service, established by the Federal Communications Commission to furnish emergency communication to local communities in connection with the Office of Civilian Defense. The background and functions of WERS are described in detail in Chapter Seventeen. Here it need only be noted that, by official statement, without the reservoir of amateur operators in this country to serve as a nucleus, the War Emergency Radio Service would have been an impossibility.

In mid-1943 the scope of WERS, limited before to war-created emergencies, was broadened to include the supplying of emergency communications in connection with natural disasters or other situations involving civilian defense or national security. Under this extension of its activities, amateurs in WERS again are in position to render emergency communications service in their traditional fashion and have done so whenever the need arose.

Nor was experimental development lost sight of in the enthusiasm incident to international amateur communication. The experimentally minded amateur was constantly at work conducting tests in new frequency bands, devising improved apparatus for amateur receiving and transmitting, learning how to operate two and three and even four stations where previously there was room enough for only one.

In particular, the amateur experimenter pressed on to the development of the higher frequencies represented by the wavelengths below 10 meters, territory only a few years ago regarded even by most amateurs as comparatively unprofitable operating ground.

The amateur's experience with five meters is especially representative of his initiative and resourcefulness and of his ability to make the most of what is at hand. In 1924, first amateur experiments in the vicinity of 56 Mc. indicated that band to be practically worthless for distance work; signals at such frequencies appeared capable of being heard only to "horizon range." But the amateur turns even such apparent disadvantages to use. If not suitable for long-distance work, at least the band was ideal for "short-haul" communication. Beginning in 1931, then, there was tremendous activity in 56-Mc. work by hundreds of amateurs all over the country, and a complete new line of transmitters and receivers was developed to meet the special conditions incident to communicating at these very-high frequencies (then known as the "ultrahighs"). In 1934 additional impetus was given to this band when experiments by the ARRL with directive antennas resulted in remarkably consistent two-way communication over distances of more than 100 miles, without the aid of "hilltop" locations. While atmospheric conditions still are found to affect 5-meter DX, thousands of amateurs, as of the time of the close-down in December, 1941, were spending much of their time on the 56- and 112-Mc. bands, many of them having worked hundreds of stations at distances up to several thousand miles; even transcontinental distances were being spanned when conditions were right. Today's concept of v.h.f. propagation was developed almost entirely through amateur research.

The amateur is constantly in the forefront of technical progress. Many developments by amateurs have come to represent valuable contributions to the art, and the articles about them are as widely read in professional circles as by amateurs. At a time when only a few broadcast engineers in the country knew what was meant by "100 per cent modulation," the technical staff of the ARRL published articles in *QST* urging amateur 'phones to embrace it and showing them how to do it. This is only one example; the complete record of such accomplishments would more than fill this chapter alone. From the League's laboratory in 1932 came the "single-signal" superheterodyne — the world's most advanced high-frequency radiotelegraph receiver. In 1936 the "noise-silencer" circuit for superheterodynes was developed, permitting for the first time satisfactory high-frequency reception through the more common forms of man-made electrical interference.

Currently, hundreds of skilled amateurs are contributing their knowledge to the development of secret wartime radio devices, both in government and in private laboratories. Equally as important, the prewar technical progress by amateurs has provided the keystone for the development of modern military communications equipment. The sets now in use by the armed forces closely resemble the best prewar amateur equipment. This is only logical, because the problems of military communications in modern warfare — extreme congestion, special operating requirements, the need for compactness and efficiency — closely resemble the problems peculiar to the amateur service before the war, and for which the amateurs were forced to devise their own specialized solutions. The fact that amateurs on the staffs of the Signal Corps and Naval Research laboratories have been closely allied with the design of this new equipment has been a vital factor.

On the fighting fronts their operating skill is equally valuable. Sharpened to the highest degree by years of communicating experience under the severest conditions of congestion, and with low-powered equipment, the amateur has the ability to hear signals so faint that they are inaudible to the average ear; and to read signals so confused with interference that for ordinary operators they are completely garbled. These abilities make the amateur a key figure in military communications.

Emergency relief, expeditionary contact, experimental work and countless instances of other forms of public service — rendered, as they always have been and always will be, without hope or expectation of material reward — made amateur radio an integral part of our peacetime national life. Today, the importance of amateur participation in the armed forces and other aspects of national defense emphasize more strongly than ever that amateur radio is vital to our national existence.

¶ The American Radio Relay League

The American Radio Relay League is to-day not only the spokesman for amateur radio in this country but it is the largest amateur organization in the world. It is strictly of, by and for amateurs, is non-commercial and has no stockholders. The members of the League are the owners of the ARRL and *QST*.

The League is organized to represent the amateur in legislative matters. It is pledged to promote interest in two-way amateur communication and experimentation. It is interested in the relaying of messages by amateur radio. It is concerned with the advancement of the radio art. It stands for the maintenance of fraternalism and a high standard of conduct. One of its principal purposes is to keep amateur activities so well conducted that the amateur will continue to justify his existence.

With normal amateur activity suspended for the duration of the war, the ARRL Headquarters establishment is largely devoted to activities designed to advance the war effort — in training, and in personnel and apparatus procurement.

The operating territory of the League is divided into fourteen United States and six Canadian divisions. The affairs of the League are managed by a Board of Directors. One director is elected every two years by the membership of each United States division, and a Canadian General Manager is elected every two years by the Canadian membership. These directors then choose the president and vice-president, who are also members of the Board. No one commercially engaged in selling or manufacturing radio apparatus or literature can become a member of the Board or an officer of the League.

The president, vice-president, secretary, treasurer and communications manager of the League are elected or appointed by the Board of Directors. These officers constitute an Executive Committee which, under certain restrictions, decides how to apply Board policies to matters arising between Board meetings.

The League owns and publishes the amateur's magazine, *QST*. *QST* goes to all members of the League each month. It acts as a monthly bulletin of the League's organized activities. It serves as a medium for the exchange of ideas. It fosters amateur spirit. Its technical articles are renowned. *QST* has grown to be the "amateur's bible," as well as one of the foremost radio magazines in the world. The profits *QST* makes are used in supporting League activities. Membership dues to the League include a subscription to *QST* for the same period.

Members of the League are entitled to write to Headquarters for information of any kind, whether it concerns membership, legislation, or general questions on the construction or operation of amateur apparatus. If you don't find the information you want in *QST* or the *Handbook*, write to ARRL Headquarters, West Hartford 7, Connecticut, telling us your problem. All replies are made directly by letter; no charge is made for the service.

If you come to Hartford, drop out to Headquarters at 38 LaSalle Road, West Hartford. Visitors are always welcome.

From 1927 to 1936 the League operated its headquarters station, W1MK, at Brainard Field, Hartford's municipal airport on the Connecticut River. During the disastrous flood of 1936 this station was devastated. From the spring of 1936 until early summer of 1938 a temporary station was operated at the headquarters offices, at first under the old auxiliary call, W1INF, and later as W1AW. The call W1AW, held until his death by Hiram Percy Maxim, was issued to the League by special order of the FCC for use as the official headquarters station call.

From September, 1938, until the wartime closing of all amateur stations, the Hiram Percy Maxim Memorial Station at Newington, Conn., was in operation as the headquarters station. Operating on all amateur bands, with separate transmitters rated at the legal maximum input of one kilowatt and elaborate antenna systems, this station was regularly heard with good strength in every part of the world. The building in which it is housed was designed by order of the Board of Directors as a permanent memorial to the League's founder-president, Hiram Percy Maxim.

ARRL Operating Organization

The American Radio Relay League maintains, at its headquarters in West Hartford, Connecticut, a Communications Department normally concerned with the practical operating activities of League members. A large field organization, headed by Section Communications Managers in each of the seventy-one sections into which the country is divided, consists of amateur stations especially selected for skill in certain phases of amateur communications work. There are appointments as Official Relay Station or Official 'Phone Station for traffic-handling; as Official Observer for monitoring of frequency and quality of transmissions; as Route Manager and 'Phone Activities Manager for the establishment of trunk lines and networks; as Emergency Coordinator for the promotion of amateur preparedness in the event of loss of commercial communications facilities through natural disaster. Mimeographed bulletins keep appointees informed of the latest developments. Special activities promote operating skill and thereby add to the ability of amateur radio to function "in the public interest, convenience and necessity." A special section is reserved each month in *QST*, the League's official organ, for amateur news from every section of the country.

With the suspension of amateur activities as a result of the war, all such appointments have been "frozen" for the duration excepting those of Emergency Coördinators, who are engaged in promoting WERS.

Complete information on all peacetime appointments and League awards for operating skills is included in the booklet, *Operating an Amateur Radio Station*. Members of the League may obtain a copy of this booklet from League Headquarters free upon request; to others, the cost is 10 cents.

ℂ The Amateur Bands

Discussion of the frequency bands assigned to and formerly used by amateurs may seem academic at a time when all amateur operation is prohibited; yet, although a special order of the FCC has temporarily suspended amateur operation, the bands normally open for amateur operation still appear in the regulations of the Commission and a knowledge of them is necessary in order to pass the amateur operator examination, which is still being given to interested persons who may wish to qualify for their amateur operator licenses.

As will be observed in Fig. 2101, the amateur bands constitute narrow segments in that part of the radio spectrum lying between 1750 kc. and 30,000 kc. (or 300 Mc.). The experimental region above this point was open to amateurs and another exclusive amateur band lies between 400 and 401 Mc. (Not shown in diagram.) In peacetime amateurs use these frequency bands according to their operating objectives and the special operating characteristics of the bands themselves. Briefly, these were as follows:

The 1750-kc. band, which carried all amateur activity before the higher-frequency bands were opened, always served well for general contact all over the country, although during the height of the higher-frequency development there was some dwindling of activity. Popular for radiotelephone work, it also was used for short-haul c.w. nets and code-practice transmissions for beginners. It was useful primarily for distances up to 400–500 miles, at night, but much longer distances were covered under good conditions.

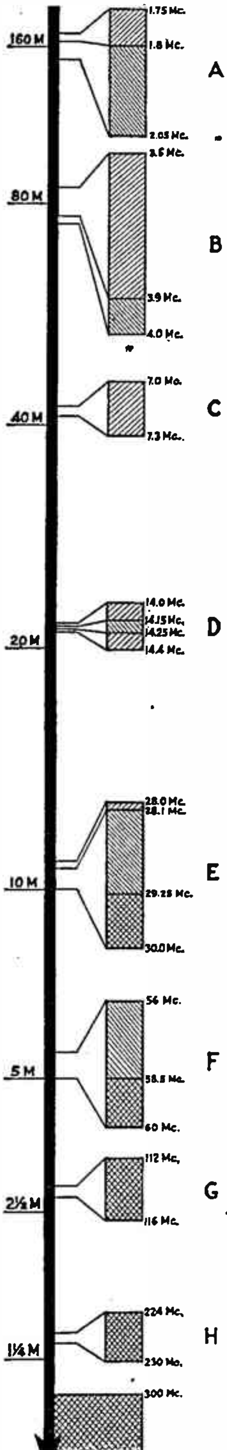


Fig. 2101 — The amateur bands. Areas shaded with diagonal lines sloping to left were open to c.w. telegraphy only. Areas with diagonal lines sloping to right were also open to amplitude-modulated telephony (and c.w.). Cross-hatched areas were open to frequency-modulated 'phone (as well as to regular a.m. 'phone and c.w.).

The 3500-kc. band was regarded, in recent years, as best for consistent domestic communication and as good for coast-to-coast work at night except for a few summer months. Much of the friendly human contact between amateurs and most of their domestic message-handling took place in this band.

The 7000-kc. band was the most popular band for general amateur work for years, both domestic and international, and was useful mainly at nights for contacts over considerable distances as well as being satisfactory for distances of several hundred miles in daylight.

The 14,000-kc. or 14-Mc. band was the one used mostly for covering great distances in daylight, and in fact was the only band generally useful for daylight contacts over coast-to-coast and greater distances. It was, however, subject to sudden changes in transmitting conditions.

The 28,000-kc. (28-Mc.) band combined both the long-distance characteristics of the 14-Mc. band and some of the local advantages of the 56-Mc. band, but was popular chiefly because of its remarkable long-distance characteristics. Its disadvantage was lack of reliability because of seasonal effects and more sudden changes in transmitting conditions even than on 14 Mc.

The 56,000-kc. (56-Mc.) band was used largely for local and short-distance work over distances of ten to fifty miles. Because of compactness and ease of construction of the necessary apparatus, hundreds of stations operated "locally" there. Experiments by the ARRL technical staff beginning in 1934 disclosed that consistent two-way work could be done over distances of a hundred miles or more with suitable conditions and equipment, and such contacts became common by 1940–41. Occasional periodic "sky-wave" work over several thousand miles also was accomplished.

The 112,000-kc. (112-Mc.) band was the newest addition to the amateur spectrum, and before the close-down was attaining widespread popularity for the local work previously carried on in the 56-Mc. band. This band now figures prominently as the chief field of operations for the War Emergency Radio Service (WERS), in which hundreds of amateurs are employing their apparatus and skill on behalf of their communities for civilian-defense work.

The 224-Mc. band and the experimental region above 300 Mc. were not in widespread use for general communication, but were becoming increasingly of interest to the pioneering experimenter. The 224-Mc. band may be called on to carry part of the WERS load.

Amateur Licensing in the United States

The Communications Act lodges in the Federal Communications Commission authority to classify and license radio stations and to prescribe regulations for their operation. Pursuant to the law, FCC has issued detailed regulations for the amateur service.

An amateur is a duly authorized person interested in radio technique solely with a personal aim and without pecuniary interest. Amateur operator licenses are given to United States citizens who pass an examination on radio operation and apparatus and on the provisions of law and regulations affecting amateurs, and who demonstrate ability to send and receive the International Morse Code at 13 words per minute. Amateur station licenses are granted only to licensed amateur operators and permit radio communication between such amateur stations for amateur purposes, i.e., for personal and noncommercial aims flowing from an interest in radio technique. An amateur station may not be used for material compensation of any sort nor for broadcasting. Narrow bands of frequencies beginning at 1,750, 3,500, 7,000, 14,000, 28,000, 56,000, 112,000, 224,000 and 400,000 kc., as shown in Fig. 2101, are allocated exclusively for use by amateur stations, and amateurs also may operate on any frequency above 300,000 kc. Amateur transmissions may be on any frequency within the assigned bands. All the frequencies may be used for c.w. telegraphy and some are available for radiotelephony by any amateur, while others are reserved for radiotelephone use by persons having at least a year's experience and who pass the examination for a Class A license. The input to the final stage of amateur stations is limited to 1,000 watts and on frequencies below 60,000 kc. must be adequately filtered direct current. Emissions must be free from spurious radiations. The licensee must provide for the measurement of the transmitter frequency and establish a procedure for checking it regularly. A complete log of station operation must be maintained, with specified data. The amateur station license also authorizes the holder to operate portable and portable-mobile amateur stations on certain frequencies, subject to certain further regulations. An amateur station may be operated only by an amateur operator licensee, but any licensed amateur operator may operate any amateur station. All radio licensees are subject to heavy penalties for violation of regulations.

Amateur licenses are issued entirely free of charge. They can be issued only to citizens of the United States, but the requirement of citizenship is the only limitation, and licenses are issued without regard to age or physical condition to anyone who successfully completes the examination.

For the duration of the war, amateur operation in this country is prohibited. Amateur frequencies have been temporarily withdrawn and, although amateur station licenses have not been cancelled, no new ones have been issued and no applications for renewal or modification are entertained. However, every effort will be made to reassign station calls to previous holders, after the war is over.

Amateur operator licenses remain in force, being valued by the military services as an attestation of radio proficiency, and new operator licenses are still being issued at the numerous FCC examining points.

It may occur to many readers that there is little point in obtaining an amateur operator license when amateur radio is not permitted. Far from it! An amateur operator license is a valuable possession, as many people engaged in the war effort have learned. In the Army, it may serve as a passport to a preferred position in the Signal Corps or Air Forces; in the Navy and Marine Corps, the holder of an amateur license (provided he also has had a high-school education and can pass the physical requirements) may be eligible for a rating as a petty officer. Among officer candidates, in some branches, possession of an amateur operator license is accepted as indicating proficiency in respect to special radio qualifications. This also applies to positions in various branches of the radio industry engaged in war work. Among women, possession of an amateur operator license is one of the requirements for certain government positions open to feminine applicants. Both industry and Civil Service give preferred attention to amateur licensees.

When you are able to copy 13 words per minute, have studied basic transmitter theory and have familiarized yourself with the radio laws and amateur regulations, you are ready to give serious thought to securing the government amateur operator license which is issued you, after examination at a local district office, through the Federal Communications Commission at Washington, D. C.

A complete discussion of license requirements, and a study guide for those preparing for the examination, are to be found in a League publication, *The Radio Amateur's License Manual*, available from the American Radio Relay League, West Hartford 7, Conn., for 25¢, postpaid. This publication, which is frequently revised, contains also the text of the U. S. regulations governing amateur stations and operators, and pertinent extracts from the basic Communications Act of 1934. It should be studied carefully by anyone intending to enter amateur radio or planning to apply for an amateur operator's license. One of the most valuable features of this book is its representative examination questions with their correct answers.

The frequency changes in amateur regulations, and the new requirements which may come up during the war under special orders of FCC, are regularly reported in the League's magazine, *QST*.

Chapter Twenty-Two

Radio Operating

THE efficient and successful operation of a radio station is, in itself, an exacting accomplishment. It is a task that requires skill and specialized training. For this reason, by federal regulation as well as by international treaty it is rigidly stipulated that no radio station, amateur or otherwise, may transmit except under the supervision of a licensed operator having qualifications appropriate to the class of station. To acquire such qualifications requires training in the various specialized practices employed.

The basic object of most radio communication is the transmission of intelligence from one point to another, accurately and in as short a time as possible. For efficiency in communication, each class of radio service has set up operating methods and procedure which provide the most expeditious handling of radio traffic. Skilled operators need not only to be expert in transmitting and receiving code or voice signals, but also must be thoroughly familiar with the uniform practices observed in the particular class of service concerned. The material following, although generally that of the amateur service, is typical of the basic operating procedure employed in nearly all services with necessary modifications.

Memorizing the Code

Apart from the technical and regulatory phases of the examination, the most important requirement for obtaining an amateur operator's license is ability to send and receive the Continental (International Morse) code at the rate of 13 words per minute. Aside from this consideration, a knowledge of the code is especially desirable during wartime; it is not putting it too strongly to say that everyone should know the code and be able to use it.

The serious student of code — sending, receiving, operating practices, copying on the typewriter, etc. — would be best advised to purchase a copy of the ARRL booklet, *Learning the Radiotelegraph Code* (price, 25 cents, postpaid), and, in fact, anyone desirous of learning the code is advised to do so via the modern and effective method outlined in this booklet. However, the following suggestions will suffice to enable one to acquire the rudiments of code ability.

The first step is to *memorize* the code. This is no task at all if you simply make up your mind to apply yourself to the job and get it over as quickly as possible. The complete Continental alphabet, including the most-used punctuation marks and numerals, is shown in the table of Fig. 2101. All of the characters shown should be learned, starting with the basic alphabet and then going on to the numerals and punctuation marks. Take a few at a time. As progress is made it is helpful to review at intervals all the letters learned up to that time.

One suggestion: Learn to think of the letters in terms of *sound* rather than their appearance as printed dot-and-dash combinations. This is an important point; in fact, successful mastery of the code can be acquired only if one thinks always in terms of the sound of a letter, right from the start. Think of A as the sound "di*dah*" — not as a printed "dot-dash." The sound "dit" is pronounced as "it" with a "d" before it. The sound "dah" is pronounced with "ah" as in "father." The sound "dah" is always stressed or accented — not in a different tone of voice, but slightly drawn out and the least bit louder. The sound "dit" is pronounced as rapidly and sharply as possible; for purposes of easy combination, as a prefix, it is often shortened to "di." When combinations of the sounds appear as one letter, say them smoothly but rapidly, remembering to make the sound "di" staccato, and allowing equal stress to fall on every dah. There should never be a space or hesitation between dits and dahs of the same letter.

If someone can be found to "send" to you, either by whistling or by means of a buzzer or code oscillator, the best way is to enlist his cooperation and learn the code by listening to it. It is best to have someone do this who is familiar with the code and who can be depended on to send the characters correctly.

Learning the code is like learning a new language, and the sooner you learn to understand the language without the necessity for mental "translation" the easier it will be for you to attain speed and proficiency. You don't think of the spoken letter U, for example, as being composed of two separate and distinct sounds — yet actually it is made up of the pure sounds "ee" and "$\overline{oo}$," spoken in rapid succession. You learned the letter U as a sound unit itself. Similarly, you should learn code letters as individual sounds themselves, and not as combinations of other sounds.

Don't think about speed at first; the first requirement is to learn every character to the point where you can recognize each of them without hesitation. Concentrate on any difficult letters until they become as familiar as the rest.

A	• —	didah
B	— • • •	dahdididit
C	— • — •	dahdidahdit
D	— • •	dahdidit
E	•	dit
F	• • — •	dididahdit
G	— — •	dahdahdit
H	• • • •	dididit
I	• •	didit
J	• — — —	didahdahdah
K	— • —	dahdidah
L	• — • •	didahdidit
M	— —	dahdah
N	— •	dahdit
O	— — —	dahdahdah
P	• — — •	didahdahdit
Q	— — • —	dahdahdidah
R	• — •	didahdit
S	• • •	dididit
T	—	dah
U	• • —	dididah
V	• • • —	didididah
W	• — —	didahdah
X	— • • —	dahdididah
Y	— • — —	dahdidahdah
Z	— — • •	dahdahdidit

1	• — — — —	didahdahdahdah
2	• • — — —	dididahdahdah
3	• • • — —	didididahdah
4	• • • • —	dididididah
5	• • • • •	dididididit
6	— • • • •	dahdididit
7	— — • • •	dahdahdididit
8	— — — • •	dahdahdahdidit
9	— — — — •	dahdahdahdahdit
0	— — — — —	dahdahdahdahdah

Period	• — • — • —
Comma	— — • • — —
Question mark	• • — — • •
Error	• • • • • • • •
Double dash	— • • • —
Wait	• — • • •
End of message	• — • — •
Invitation to transmit	— • —
End of work	• • • — • —

Fig. 2101—The Continental (International Morse) Code.

Acquiring Speed by Buzzer Practice

When the code is thoroughly memorized, you can start to develop speed in receiving code transmission. Perhaps the best way to do this is to have two people learn the code together and send to each other by means of a buzzer-and-key outfit. An advantage of this system is that it develops sending ability, too, for the person doing the receiving will be quick to criticize uneven or indistinct sending. If possible it is a good idea to obtain the assistance of an experienced operator for the first few sessions, so that you will learn how well-sent characters should sound.

Either the buzzer set shown in Figs. 2102 and 2103 or one of the audio oscillators described will give satisfactory results as a practice set. The oscillator more closely simulates actual radio signals.

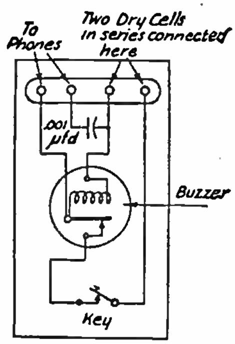


Fig. 2102—Wiring diagram of a buzzer code-practice set. The headphones are connected across the coils of the buzzer, with a condenser in series. The size of this condenser determines the strength of the signal in the 'phones. If the value shown gives an excessively loud signal, it may be reduced to 500 μμfd. or even 250 μμfd.

The battery-operated audio oscillator in Figs. 2104 and 2105 is easy to construct. However, it employs batteries, which are difficult to acquire in wartime. If nothing is heard in the headphones when the key is depressed, reverse the leads going to *either* transformer winding (not both).

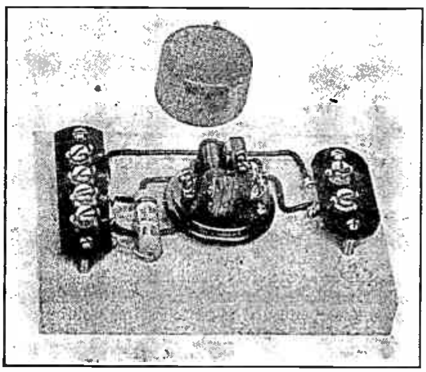


Fig. 2103—The cover of the buzzer unit has been removed in this view of the buzzer code-practice set.

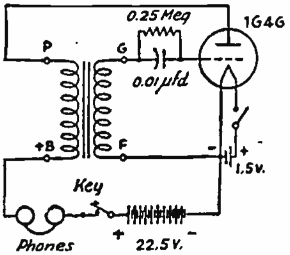


Fig. 2104 — Wiring diagram of a simple vacuum-tube audio-frequency oscillator for use as a code-practice set.

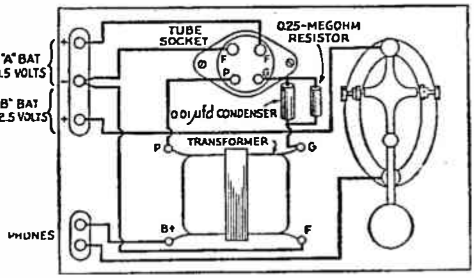


Fig. 2105 — Layout of the audio-oscillator code-practice set. All parts may be mounted on a wooden baseboard, approximately 5 × 7 inches in size.

The a.c.-d.c. oscillator in Fig. 2106 utilizes a combination diode-pentode tube, the pentode section being used as a vacuum-tube oscillator. This set operates directly from the 115-volt a.c. or d.c. power line and must be installed and handled so that electrical shock will be impossible. The set should be enclosed in a protective cabinet or box. If a metal chassis is used, the components *must* be insulated from the chassis.

The headphone circuit is insulated from the power line, as shown in Fig. 2106, but care is required to prevent contact with ground or power line while touching the key.

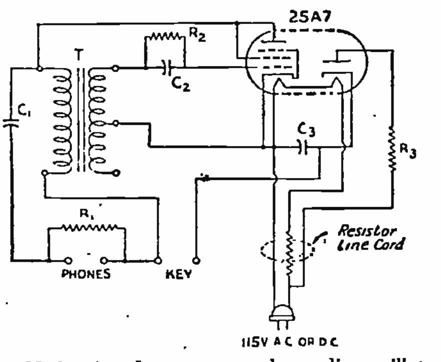


Fig. 2106 — A.c.-d.c. vacuum-tube audio oscillator.
C_1 — 100-μμfd. midget mica.
C_2 — 250-μμfd. midget mica.
C_3 — 8-μfd. 200-volt electrolytic.
R_1 — 0.5 megohm, ½-watt. (A lower value, or a variable resistor, may be used to reduce volume.)
R_2 — 1 megohm, ½-watt.
R_3 — 50 ohms, 1-watt.
T — 3:1-ratio midget push-pull audio transformer.
Line cord resistor — 310 ohms. (A 300-ohm, 50-watt wire-wound fixed resistor may be used instead.)

After the practice set has been built, and another operator's help secured, practice sending turn and turn about to each other. Send single letters at first, the listener learning to recognize each character quickly, without hesitation. Following this, start slow sending of complete words and sentences, always trying to have the material sent at just a little faster rate than you can copy easily; this speeds up your mind. Write down each letter you recognize. Do *not* try to write down the dots and dashes; write down the letters. Don't stop to compare the sounds of different letters, or think too long about a letter or word that has been missed. Go right on to the next one, or each "miss" will cause you to lose several characters you might otherwise have gotten. If you exercise a little patience you will soon be getting every character, and in a surprisingly short time will be receiving at a good rate of speed. When you can receive 13 words a minute (65 letters a minute), have the sender transmit code groups rather than English text. This will prevent you from recognizing a word "on the way" and filling it in before you've really listened to the letters themselves.

After you have acquired a reasonable degree of proficiency, concentrate on the less common characters, as well as the numerals and punctuation marks. These prove the downfall of many applicants taking the code examination under the handicap of nervous stress.

Using a Key

The correct way to grasp the key is important. The knob of the key should be about eighteen inches from the edge of the operating table and about on a line with the operator's right shoulder, allowing room for the elbow to rest on the table. A table about thirty inches in height is best. The spring tension of the key varies with different operators. A fairly heavy spring at the start is desirable. The back adjustment of the key should be changed until there is a vertical movement of about one-sixteenth inch at the knob. After an operator has mastered the use of the hand key the tension should be changed and can be reduced to the minimum spring tension that will cause the key to open immediately when the pressure is released. More spring tension than necessary causes the expenditure of unnecessary energy. The contacts should be spaced by the rear screw on the key only and not by allowing play in the side screws, which are provided merely for aligning the contact points. These side screws should be screwed up to a setting which prevents appreciable side play but not adjusted so tightly that binding is caused. The gap between the contacts should always be at least a thirty-second of an inch, since a too-finely spaced contact will cultivate a nervous style of sending which is highly undesirable. On the other hand too-wide spacing (much over one-sixteenth inch) may result in unduly heavy or "muddy" sending.

Do not hold the key tightly. Let the hand rest lightly on the key. The thumb should be against the left side of the key. The first and second fingers should be bent a little. They should hold the middle and right sides of the knob, respectively. The fingers are partly on top and partly over the side of the knob. The other two fingers should be free of the key. Fig. 2107 shows the correct way to hold a key.

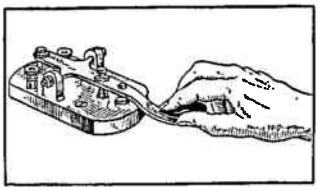


Fig. 2107 — This sketch illustrates the correct position of the hand and fingers for good sending with a telegraph key.

A wrist motion should be used in sending. The whole arm should not be used. One should not send "nervously" but with a steady flexing of the wrist. The grasp on the key should be firm, but not tight, or jerky sending will result. None of the muscles should be tense but they should all be under control. The arm should rest lightly on the operating table with the wrist held above the table. An up-and-down motion without any sideways action is best. The fingers should never leave the key knob.

Good sending may *seem* easier than receiving, but don't be deceived. A beginner should not attempt to send fast. Keep your transmitting speed down to the receiving speed, and bend your efforts to sending *well.* Do not try to speed things up too soon. A slow, even rate of sending is the mark of a good operator. Speed will come with time alone. Leave special types of keys alone until you have mastered the knack of handling the standard key. Because radio transmissions are seldom free from interference, a "heavier" style of sending is best to develop for radio work. A rugged, heavy key will help in developing this characteristic.

Radiotelegraph Operation

The radiotelegraph code is used for *record* communication. Aside from his ability to copy at high speeds, a good operator is noted for his neatness and accuracy of copy. It is evident that a radio operator should copy exactly what is sent, and if there is any doubt about a letter or word he should query the transmitting operator about it.

An operator with a clean-cut, slow, steady method of sending has a big advantage over a poor operator. Good sending is a matter of practice, but patience and judgment are just as important qualities in an operator as is a good fist. Very often, transmission at moderate speeds moves traffic more quickly than faster but erratic sending. In hand operating any unusual words should be sent twice, the word being repeated following transmission of "?". A transmitting operator who is notified of interference on his frequency, either by static or man-made, should adjust his speed of sending to require the least number of "fills." Every operator should have facilities for monitoring to check the accuracy of his sending. Accuracy of transmission comes *first.*

To this end, an operator copying in longhand should use extreme care in writing, so there will be no chance of confusing an "l" with an "e", and the like. On a typewriter, best practice is always to double-space between lines, write ten words to a line with an extra space or two after the fifth word in each line, triple-space between lines every fifth line. This is for the purpose of rapidly determining the number of words in a message as it is sent. As the operator gains mill-copying skill he will be able to type subconsciously in this pattern.

General procedure — (1) Calls should be made by transmitting not more than three times the call signal of the station called, and DE, followed by one's own call signal sent not more than three times, thus: VE2BE VE2BE VE2BE DE W1AW W1AW W1AW. In amateur practice this form is repeated completely once or twice. The call signal of the calling station must be inserted at frequent intervals for identification purposes. Repeating the call signal of the called station five times and signing not more than twice has proved excellent practice in connection with break-in operation (the receiver being kept tuned to the frequency of the called station). The use of a break-in system is highly recommended to save time and reduce unnecessary interference.

2) Answering a call: Call three times (or less); send DE; sign three times (or less); and after contact is established decrease the use of the call signals of both stations to once or twice. Example:

W1GNF DE W1AW GE OM GA K (meaning, "Good evening, old man, go ahead").

3) Ending signals and sign off: The proper use of $\overline{\text{AR}}$, K and $\overline{\text{VA}}$ ending signals is as follows: $\overline{\text{AR}}$ (end of transmission) shall be used at the end of messages during communication; and also at the end of a call, indicating when so used that communication is not yet established. In the case of CQ calls, the international regulations recommend that K shall follow. K (invitation to transmit) shall also be used at the end of each transmission when answering or working another station, carrying the significance of "go ahead." $\overline{\text{VA}}$ (or $\overline{\text{SK}}$) shall be used by each station only when signing off, this followed by the call of the station being worked and your own call sent once for identification purposes. Examples:

($\overline{\text{AR}}$) — W1KQY DE W1CTI $\overline{\text{AR}}$ (showing that W1CTI has not yet gotten in touch with W1KQY but has called and is now listening for his reply). Used after the signature between messages, it indicates the end of one message. There may be a slight pause before starting the second of the series of messages. The courteous and thoughtful operator allows time for the receiving operator to enter the time on the message and put another blank in readiness for the traffic to come. If K is added, it means that the operator wishes his first message acknowledged before going on with the second message. If no K is heard, preparations should be made to continue copying.

(K) — W1JEQ DE W6AJM R K. (This arrangement is very often used for the acknowledgment of a transmission. When anyone overhears this he knows at once that the two stations are in touch, communicating with each other, that W1JEQ's transmission was all understood by W6AJM, and that W6AJM is telling W1JEQ to go ahead with more of what he has to say.) W9KJY DE W7NH NR 23 R K. (Evidently W9KJY is sending messages to W7NH. The contact is good. The message was all received correctly. W7NH tells W9KJY to "go ahead" with more.)

($\overline{VA}$) — R NM NW CUL VY 73 $\overline{AR}$ $\overline{VA}$ W6TI DE W7WY. (W7WY says, "I understand OK, no more now, see you later, very best regards. I am through with you for now and will listen for whomever wishes to call. W7WY 'signing off' with W6TI.")

4) If a station sends test signals, to adjust the transmitter or at the request of another station to permit the latter to adjust its receiving apparatus, the signals must be composed of a series of Vs with the call signal of the transmitting station inserted at frequent intervals.

5) When a station receives a call without being certain that the call is intended for it, it should not reply until the call has been repeated and is understood. If it receives the call but is uncertain of the call signal of the sending station, it should answer using the signal ··— —·· (?) instead of the call signal of this latter station. QRZ? (see page 454) is the appropriate signal to use, followed by your call, to ask who is calling and get this station to call again.

6) Receipting for conversation or traffic: Never send a single acknowledgment until the transmission has been entirely received. "R" means, "All right, OK, I understand completely." When a poor operator, commonly called a "lid," has only received part of a message, he will answer, "R R R R R R R R R R, sorry, missed address and text, pse repeat" and every good operator who hears will rave inwardly. Use "R" only when all is received correctly. Example:

When all the message has been received correctly, a short call with "NR 155 R K" or simply "155 K" is sufficient.

Abbreviations — To speed up radiotelegraph communication, a number of standard and special abbreviations have been devised. As time is a factor, uniform practices in operating are necessary to insure a ready understanding by both operators. Therefore proficiency in the commonly used abbreviations is to be desired. Some of those prescribed by the regulations attached to the International Telecommunications Convention and used by all radio services follow:

C	Yes.
N	No.
W	Word(s).
AA	All after (used after a question mark to request a repetition).
AB	All before (similarly).
AL	All that has just been sent (similarly).
BN	All between (similarly).
BQ	Announcement of reply to a request for rectification.
CL	I am closing my station.
GA	Go ahead (or resume sending).
JM	If I may send, make a series of dashes. To stop my transmission, make a series of dots.
MN	Minute(s) (to indicate duration of a wait).
NW	I resume transmission.
OK	We are in agreement.
RQ	Announcement of a request for rectification.
UA	Do you agree?
WA	Word after (to be used after a question mark to request a repetition).
WB	Word before (similarly).
ADR	Address (similarly).
PBL	Preamble (similarly).
SIG	Signature (similarly).
TXT	Text (similarly).
XS	Atmospherics.
YS	See your service advice.
ABV	Use abbreviations.
CFM	Confirm, or I confirm.
ITP	The punctuation counts.
MSG	Prefix to radiotelegram.
REF	Refer to, or referring to.
RPT	Repeat, or I repeat (to be used to ask or to give repetition of such traffic as is indicated after the abbreviation).
SVC	Prefix to service message.
TFC	Traffic.
P	Indicator or private telegram in the mobile service (to be used as a prefix).
NIL	I have nothing for you.

XXX XXX XXX DE . . . , urgent signal indicating message to follow regarding safety of mobile station or persons in sight therefrom (PAN is similarly used by aircraft); TTT TTT TTT DE . . . , safety signal sent before meteorological warning messages and those concerning safety of navigation; SOS SOS SOS DE . . . , distress signal sent only by mobile stations in grave danger when requesting assistance (MAYDAY is the radiotelephone distress call similarly used).

In the text of a message, no words should be abbreviated by the operator unless they are so written by the sender. If the text includes punctuation, it should be spelled out in English.

Message Handling — Each service — commercial, military, amateur — prescribes its own message form, but all are generally similar to the example here given. A message is broadly divided into four parts: (1) the preamble; (2) the address; (3) the text; (4) the signature. The preamble contains the following:

a) Number (of this message).
b) Station of origin.
c) Check (number of words in text).
d) Place of origin.
e) Time filed.
f) Date.

Therefore, it might look like this:

NR 34 WLTK JH 13
CHICAGO ILL 450 PM MAY 12 1942
CAPT WM MONTGOMERY
MUNITIONS BLDG
WASHINGTON DC $\overline{BT}$
SIXTH CORPS AREA HAS 68
MEN AVAILABLE FOR ACTIVE DUTY
FIXED SERVICE REGARDS $\overline{BT}$
HUNTER WLTK

This is obviously the 34th message (of that day or that month, as the policy of the station prescribes) from station WLTK. The "JH 13" is the "sine" of the operator plus the number of words in the message text. All operators designate themselves with a personal sine to be used on message traffic and on the air; in most cases it consists of the operator's initials. The signal $\overline{\text{BT}}$ (double-dash) is used to separate the text from address and signature.

Several radiograms may be transmitted in series (QSG. . . . .) with the consent of the station which is to receive them. As a general rule long radiograms should be transmitted in sections of approximately fifty words, each ending with ·· — — ·· (?), meaning, "Have you received the message correctly thus far?"

If the first part of a message is received but substantially all of the latter portions lost, the request for the missing parts is simply RPT TXT AND SIG, meaning, "Repeat text and signature." PBL and ADR may be used similarly for the preamble and address of a message. RPT ALL or RPT MSG should not be sent unless nearly all of the message is lost. When a few word-groups in conversation or message handling have been missed, a selection of one or more of the following abbreviations are used to ask for a repeat on the parts in doubt.

Abbreviation	*Meaning*
?AA	Repeat all after
?AB	Repeat all before
?AL	Repeat all that has been sent
?BN...AND	Repeat all between..and...
?WA	Repeat the word after
?WB	Repeat the word before

The good operator will ask only for what fills are needed, separating different requests for repetition by using the break sign or double dash (—·-·-—) between these parts. There is seldom any excuse for repeating a whole message just to get a few lost words.

Another interrogation method is sometimes used, the question signal (·· — — ··) being sent between the last word received correctly and the first word (or first few words) received after the interruption.

As an example of what procedure would be followed in the transmission of a commercial message, let us assume that a passenger aboard the S.S. *Coastwise* wishes to notify a friend of his arrival. Station WKCZ aboard the ship would call a shore station (WSC) and the following would ensue:

WSC WSC WSC DE WKCZ WKCZ WKCZ P $\overline{\text{AR}}$ K

WKCZ WKCZ WKCZ DE WSC ANS 700 K

WSC WSC WSC DE WKCZ P 1 CK12 SS COASTWISE 0827 MAY 10 $\overline{\text{BT}}$ MISS JANET SHANNON 18 LAMBERT STREET BOSTON $\overline{\text{BT}}$ ARRIVE PIER 18 TONIGHT LOVE $\overline{\text{BT}}$ JOHN $\overline{\text{AR}}$ K

WKCZ DE WSC R 1 K

WSC DE WKCZ QRU $\overline{\text{SK}}$

WKCZ DE WSC R $\overline{\text{SK}}$

If the receiving operator missed the number of the pier of arrival, he would send:

PIER ?? TONIGHT or ?WA PIER.

whereupon the transmitting operator would say:

PIER 18 TONIGHT

and then would stand by for an acknowledgment of receipt (R).

The service message — When one station has a message to transmit to another concerning the handling of a previous message, the message is titled a "service" and is indicated by "SVC" in the preamble when sent. Such a message may refer to non-delivery, delayed transmission, errors, or to any phase of message handling activity. Words may be abbreviated in the text of the service message to conserve time. Do not abbreviate to the point where misunderstanding may arise.

Provisions in the Communications Act of 1934 make it a misdemeanor to give out information of any sort to any person except the addressee of a message or his authorized agent. When for some reason a message cannot be delivered, a service message should be sent to the station of origin containing information to that effect.

Land-line check — The land-line or "text" count, consisting of count only of the words in the body or text of the message, is probably now most widely used. (The "cable" count covers all words in the address and signature, as well, probably accounting for its unpopularity.) When in the case of a few exceptions to the basic rule in land-line checking, certain words in the address, signature or preamble are counted, they are known as extra words and all such are so designated in the check right after the total number of words.

The check includes:

1) All words, figures and letters in the body, and
2) the following extra words:
   (a) Signature except the first, when there are more than one (a title with signature does not count extra, but an address following a signature does).
   (b) Words "report delivery," or "rush" in the check.
   (c) Alternate names and/or street addresses, and such extras as "personal" or "attention."

Dictionary words in most languages count as one word irrespective of length of the word. In counting figures, a group of five digits or less counts as one word. Bars of division and decimal points may constitute one or more of the digits in such a group. It is recommended that, where feasible, words be substituted for figures to reduce the possibility of error in transmission. Detailed examples of word counting are about as difficult in one system of count as another.

Radiotelephone Operation

Procedure to be used in radiotelephone operation follows the foregoing general principles closely. The operator makes little use of the special abbreviations available for code work, of course, since he may directly speak out their full meaning. Radiotelephony is used principally for command and control purposes, such as communication between ground stations and aircraft, where recorded message traffic is at a minimum. Transmissions consist mostly of short bursts with little variety in form or content, and each operator must become familiar with procedure methods adopted by the particular service.

Unusual words should be avoided in the interest of accuracy if possible when drafting messages. When they unavoidably turn up difficult words may be repeated, or *repeated and spelled.* The operator says "I will repeat" when thus retransmitting a difficult word or expression. It is recommended that use of *Q* code and special abbreviations be minimized in voice work insofar as possible, and the full expression (with conciseness) be substituted.

The speed of radiotelephone transmission (with perfect accuracy) depends almost entirely upon the skill of the two operators involved. One must learn to speak at a rate allowing perfect understanding as well as permitting the receiving operator to copy down the message text, if that is necessary. Because of the similarity of many English speech sounds, the use of alphabetical word lists has been found necessary. One such list — that used by the Army and Navy — is given in Chapter Seventeen. All voice-operated stations should use a *standard* list as needed to identify call signals or unfamiliar expressions.

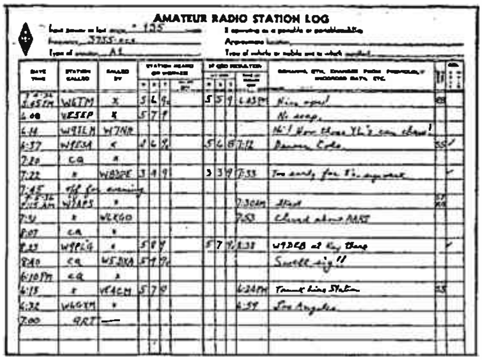
Net Operation

In field work many military communications units operate in "net" fashion, wherein one station (at the headquarters of the unit) is designated as net-control station (NCS) to direct the business of the net. The operation of all stations in the same net is on one single frequency, so that any one operator may hear any other station(s) without retuning his receiver. "Break-in" is advantageously employed here — the receiver is kept running during transmissions, so that nearly simultaneous two-way communication is possible.

Briefly, the procedure in net operation is as follows: The NCS calls the net together at a pre-announced time and using a predetermined call. Immediately, station members of the net reply in alphabetical (or some other predetermined) order, reporting on the NCS's signal strength and stating what traffic is on hand and for whom. The NCS acknowledges, meanwhile keeping an account of all traffic on hand, by stations. He then directs the transfer of messages from one station to another, giving preference to any urgent traffic so indicated at roll call. When all traffic has been distributed and it is apparent there is no further business, the NCS will close the net, in most cases maintaining watch on the net frequency for any special traffic which might appear.

Keeping a Log

FCC regulations require nearly every radiocommunication station to keep a complete operating record or "log," including such data as times and dates of transmissions, stations contacted, message traffic handled, input power to the transmitter, frequency used, and signature or "sine" of the operator in charge.



Log-keeping procedure differs with each class of communications service. A typical page from an amateur radio station log, prepared on the standard ARRL form, is shown above and is illustrative of the form and data generally required.

Time Systems

While most continental-commercial telegraph and radio circuits use local standard (or war) time in log-keeping and message-handling, international radiocommunication stations and the military services now use a 24-hour system of time-keeping. One is Greenwich Civil Time, a 24-hour clock system used in international radiocommunication work. All figures are based on the time in Greenwich, England, the city of 0° meridian fame. 0000 represents midnight in Greenwich; 0600 represents 6 A.M. there; 1200 is noon; 1800 is 6 P.M.; 2400 is again midnight and the same as 0000 of the following day. The figures must be corrected to each individual time zone. The Central War Time zone is five hours behind Greenwich, so that 0630 GCT (6:30 A.M. in Greenwich) would represent 1:30 A.M. CWT, for example. As an example of reverse translation, 9:30 A.M. CWT would be designated in the log as 1430 GCT. EWT is four hours behind GCT; MWT, six hours; PWT, seven.

At present the military services use simply a 24-hour clock, based on local time, without correcting to Greenwich or any other longitude. Then 6 A.M. CWT becomes 0600; 6 A.M. EWT is 0600, and so on. The principal advantage of this system is an elimination of the necessity for the use of P.M. or A.M. abbreviations.

FCC Frequency Allocations

The following is a condensed table of frequency allocations established by the Federal Communications Commission in the United States, as they existed prior to the war. Certain departures from it under wartime conditions are, of course, to be expected.

Frequencies (Kc.)	*Allocation*
10–103	Fixed, government.
103–141	Coastal telegraph, government.
143–193	Maritime calling, ship telegraph, fixed and coastal telegraph (190 kc. to state police and government).
194–391	Government, fixed, airport, aircraft (375 kc. to direction finding).
392–548	Coastal telegraph, government, ship telegraph, aircraft, intership 'phone (500 kc. to maritime calling and government).
550–1,600	Broadcasting (1,592 to Alaska services).
1,600–1,746	Geophysical, relay, police, government, experimental, marine fire, aviation, motion picture.
1,750–2,050	*Amateur.*
2,052–2,500	Experimental visual and relay broadcast, police, government, ship harbor, fixed, miscellaneous.
2,504–3,497.5	Coastal harbor, government, aviation, fixed, miscellaneous.
3,500–4,000	*Amateur.*
4,005–6,000	Government, aviation, fixed.
6,020–6,190	International broadcast, government.
6,200–6,990	Coastal telegraph and 'phone, government, fixed, miscellaneous.
7,000–7,300	*Amateur.*
7,305–9,490	Government, fixed, aviation, ship telegraph, coastal telegraph, miscellaneous.
9,510–9,690	International broadcast.
9,710–11,000	Government, fixed aviation.
11,010–11,685	Ship telegraph, maritime calling, government, coastal telegraph, fixed, aviation, miscellaneous.
11,710–11,890	International broadcast, government.
11,910–13,990	Aviation, fixed, government, ship telegraph, coastal telegraph, miscellaneous.
14,000–14,400	*Amateur.*
14,410–15,085	Fixed.
15,110–15,330	International broadcast, government.
15,355–17,740	Fixed, government, aviation, ship and coastal telegraph, miscellaneous.
17,760–17,840	International broadcast.
17,860–21,440	Fixed, government, aviation.
21,460–21,650	International broadcast, government.
21,650–23,175	Coastal telegraph, government, ship telegraph, miscellaneous.
23,200–25,000	Aviation, government, miscellaneous.
25,025–26,975	Broadcast, government.
27,000–27,975	Government, general communication.
28,000–30,000	*Amateur.*
30,000–42,000	Police, government, relay broadcast, coastal and ship harbor, miscellaneous.
42,000–50,000	Broadcast and educational (FM).
50,000–56,000	Television, fixed.
56,000–60,000	*Amateur.*
60,000–112,000	Government, television.
112,000–116,000	*Amateur.*
116,110–139,960	Broadcast, government, aviation, police, miscellaneous.
140,100–143,880	Aviation.
144,000–224,000	Government, television, fixed.
224,000–230,000	*Amateur.*
230,000–400,000	Government, television, fixed.
401,000 and above	Amateur and experimental.

"Q CODE"

In the regulations accompanying the existing International Radiotelegraph Convention, there is a very useful internationally agreed code designed to meet the major needs in international radio communication. This code is given in the following table. The abbreviations themselves have the meanings shown in the "answer" column. When an abbreviation is followed by an interrogation mark (?), it assumes the meaning shown in the "question" column.

Abbreviation	*Question*	*Answer*
QRA	What is the name of your station?	The name of my station is ....
QRB	How far approximately are you from my station?	The approximate distance between our stations is .... nautical miles (or .... kilometers).
QRC	What company (or Government Administration) settles the accounts for your station?	The accounts for my station are settled by the .... company (or by the Government Administration of ....).
QRD	Where are you bound and where are you from?	I am bound for ...., from ....
QRG	Will you tell me my exact frequency (wave-length) in kc/s (or m)?	Your exact frequency (wave-length) is .... kc/s (or .... m).
QRH	Does my frequency (wave-length) vary?	Your frequency (wave-length) varies.
QRI	Is my note good?	Your note varies.
QRJ	Do you receive me badly? Are my signals weak?	I cannot receive you. Your signals are too weak.
QRK	What is the legibility of my signals (1 to 5)?	The legibility of your signals is .... (1 to 5).
QRL	Are you busy?	I am busy (or I am busy with ....). Please do not interfere.
QRM	Are you being interfered with?	I am being interfered with.
QRN	Are you troubled by atmospherics?	I am troubled by atmospherics.
QRO	Shall I increase power?	Increase power.
QRP	Shall I decrease power?	Decrease power.
QRQ	Shall I send faster?	Send faster (........ words per minute).
QRS	Shall I send more slowly?	Send more slowly (............ words per minute).
QRT	Shall I stop sending?	Stop sending.
QRU	Have you anything for me?	I have nothing for you.
QRV	Are you ready?	I am ready.
QRW	Shall I tell .... that you are calling him on .... kc/s (or .... m)?	Please tell .... that I am calling him on ...... kc/s (or .... m).
QRX	Shall I wait? When will you call me again?	Wait (or wait until I have finished communicating with ....) I will call you at .... o'clock (or immediately).
QRY	What is my turn?	Your turn is No. .... (or according to any other method of arranging it).
QRZ	Who is calling me?	You are being called by ....
QSA	What is the strength of my signals (1 to 5)?	The strength of your signals is .... (1 to 5).
QSB	Does the strength of my signals vary?	The strength of your signals varies.
QSD	Is my keying correct; are my signals distinct?	Your keying is incorrect; your signals are bad.
QSG	Shall I send .... telegrams (or one telegram) at a time?	Send .... telegrams (or one telegram) at a time.
QSJ	What is the charge per word for .... including your internal telegraph charge?	The charge per word for .... is .... francs, including my internal telegraph charge.
QSK	Shall I continue with the transmission of all my traffic. I can hear you through my signals?	Continue with the transmission of all your traffic, I will interrupt you if necessary.
QSL	Can you give me acknowledgment of receipt?	I give you acknowledgment of receipt.
QSM	Shall I repeat the last telegram I sent you?	Repeat the last telegram you have sent me.
QSO	Can you communicate with .... direct (or through the medium of ....)?	I can communicate with .... direct (or through the medium of ....).
QSP	Will you retransmit to .... free of charge?	I will retransmit to .... free of charge.
QSR	Has the distress call received from .... been cleared?	The distress call received from .... has been cleared by ....
QSU	Shall I send (or reply) on .... kc/s (or m) and/or on waves of Type A1, A2, A3, or B?	Send (or reply) on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3, or B.
QSV	Shall I send a series of VVV ....?	Send a series of VVV ....
QSW	Will you send on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3, or B?	I am going to send (or I will send) on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3, or B.
QSX	Will you listen for .... (call sign) on .... kc/s (or .... m)?	I am listening for .... (call sign) on .... kc/s (or .... m).
QSY	Shall I change to transmission on .... kc/s (or .... m) without changing the type of wave? or Shall I change to transmission on another wave?	Change to transmission on .... kc/s (or .... m) without changing the type of wave or Change to transmission on another wave.
QSZ	Shall I send each word or group twice?	Send each word or group twice.
QTA	Shall I cancel telegram No. .... as if it had not been sent?	Cancel telegram No. .... as if it had not been sent.
QTB	Do you agree with my number of words?	I do not agree with your number of words; I will repeat the first letter of each word and the first figure of each number.
QTC	How many telegrams have you to send?	I have .... telegrams for you (or for ....).

Abbreviation	Question	Answer
QTE	What is my true bearing in relation to you? or What is my true bearing in relation .... (call sign)? What is the true bearing of .... (call sign) in relation to .... (call sign)?	Your true bearing in relation to me is .... degrees or Your true bearing in relation to .... (call sign) is .... degrees at .... (time) or The true bearing of .... (call sign) in relation to .... (call sign) is .... degrees at ...... (time).
QTF	Will you give me the position of my station according to the bearings taken by the direction-finding stations which you control?	The position of your station according to the bearings taken by the direction-finding stations which I control is .... latitude .... longitude.
QTG	Will you send your call sign for fifty seconds followed by a dash of ten seconds on .... kc/s (or .... m) in order that I may take your bearing?	I will send my call sign for fifty seconds followed by a dash of ten seconds on .... kc/s (or .... m) in order that you may take my bearing.
QTH	What is your position in latitude and longitude (or by any other way of showing it)?	My position is .... latitude .... longitude (or by any other way of showing it).
QTI	What is your true course?	My true course is .... degrees.
QTJ	What is your speed?	My speed is .... knots (or .... kilometers) per hour.
QTM	Send radioelectric signals and submarine sound signals to enable me to fix my bearing and my distance.	I will send radioelectric signals and submarine sound signals to enable you to fix your bearing and your distance.
QTO	Have you left dock (or port)?	I have just left dock (or port).
QTP	Are you going to enter dock (or port)?	I am going to enter dock (or port).
QTQ	Can you communicate with my station by means of the International Code of Signals?	I am going to communicate with your station by means of the International Code of Signals.
QTR	What is the exact time?	The exact time is ....
QTU	What are the hours during which your station is open?	My station is open from .... to ....
QUA	Have you news of .... (call sign of the mobile station)?	Here is news of .... (call sign of the mobile station).
QUB	Can you give me in this order, information concerning: visibility, height of clouds, ground wind for .... (place of observation)?	Here is the information requested ....
QUC	What is the last message received by you from .... (call sign of the mobile station)?	The last message received by me from .... (call sign of the mobile station) is ....
QUD	Have you received the urgency signal sent by .... (call sign of the mobile station)?	I have received the urgency signal sent by .... (call sign of the mobile station) at ....(time).
QUF	Have you received the distress signal sent by .... (call sign of the mobile station)?	I have received the distress signal sent by .... (call sign of the mobile station) at .... (time).
QUG	Are you being forced to alight in the sea (or to land)?	I am forced to alight (or land) at .... (place).
QUH	Will you indicate the present barometric pressure at sea level?	The present barometric pressure at sea level is .... (units).
QUJ	Will you indicate the true course for me to follow, with no wind, to make for you?	The true course for you to follow, with no wind, to make for me is .... degrees at .... (time).
QUK	Can you tell me the condition of the sea observed at .... (place or coördinates)?	The sea at .... (place or coördinates) is ....
QUL	Can you tell me the swell observed at .... (place or coördinates)?	The swell at .... (place or coördinates) is ....
QUM	Is the distress traffic ended?	The distress traffic is ended.

Special abbreviations adopted by the ARRL:

QST General call preceding a message addressed to all amateurs and ARRL Members. This is in effect "CQ ARRL."

QRR Official ARRL "land SOS." A distress call for use by stations in emergency zones only.

Scales Used in Expressing Signal Strength and Readability

(See QRK and QSA in the Q Code)

Strength		*Readability*	
QSA1	Barely perceptible.	QRK1	Unreadable.
QSA2	Weak.	QRK2	Readable now and then.
QSA3	Fairly good.	QRK3	Readable with difficulty.
QSA4	Good.	QRK4	Readable.
QSA5	Very good.	QRK5	Perfectly readable.

Vacuum Tube Characteristics and Miscellaneous Data

This chapter contains a compilation of miscellaneous data useful to the practising radio amateur. The first part contains reference information intended to illustrate and supplement the basic material in the remainder of this *Handbook*. The larger part of the chapter is devoted to data on different types of receiving and transmitting tubes.

Inductance and Capacity

Inductance (L) — The formula for computing the inductance of air-core coils is:

$$L = \frac{0.2\ a^2n^2}{3a + 9b + 10c}\ \mu\text{h}.$$

where a is the mean diameter of the coil in inches, b is the length of the winding in inches, c is the radial depth of the winding in inches, and n is the number of turns. The quantity c may be neglected if the coil is a single-layer solenoid.

For example, assume a coil having 35 turns of No. 30 d.s.c. wire on a form 1.5 inches in diameter. Consulting the wire table (page 461), 35 turns of No. 30 d.s.c. will occupy 0.5 inch. Therefore, $a = 1.5$, $b = 0.5$, $n = 35$, and

$$L = \frac{0.2 \times (1.5)^2 \times (35)^2}{(3 \times 1.5) + (9 \times 0.5)} = 61.25\ \mu\text{h}.$$

To calculate the number of turns of a single-layer coil for a required value of inductance:

$$N = \sqrt{\frac{3a + 9b}{0.2a^2} \times L}$$

More rapid and convenient calculations in designing coils can be made with the ARRL *Lightning Radio Calculator* (Type A).

The range of this *Calculator* may be extended to either higher or lower frequencies or dimensions than appears on the scales by employing the following relationships:

1) For any given dimensions of diameter and length, the inductance is proportional to n^2.

2) For any given value of n^2, the inductance is proportional to an extension of its dimensions.

Straight round wires:

To calculate the high frequency inductance of a straight round wire:

$$L = 0.00508\ l\ (2.303 \log_{10} \frac{4l}{d} - 1)$$

l = length in inches
d = diameter in inches
L = inductance in microhenrys

Condenser capacity (C) — The formula for determining the capacity of a condenser is:

$$C = 0.2235 \frac{KA}{d} (n - 1)\ \mu\mu\text{fd}.$$

where A is the area of one side of one plate in square inches, n is the total number of plates, d is the separation between plates in inches, and K is the dielectric constant (= 1 for air; see the table on page 458 for values for other materials).

The dielectric constant is the ratio of the capacity of a condenser with a given dielectric to its capacity with air dielectric.

Linear Circuits

At very-high and ultrahigh frequencies transmission lines are used as linear resonant circuits. The following formulas cover the design of such circuits. All dimensions are in inches.

Characteristic impedance — The characteristic or surge impedance, Z_o, of various types of transmission lines may be computed as follows:

Single conductor to perfect ground:

$$Z_o = 138 \log \frac{2D}{a}$$

where D is the height of the conductor above ground and a is the radius of the conductor.

Concentric or coaxial conductor:

$$Z_o = \frac{138}{\sqrt{K}} \log \frac{b}{a}$$

where b is the *inside* radius of the outer conductor, a is the *outside* radius of the inner conductor and K is the dielectric constant (=1 for air; see table for values for other materials).

When the dielectric consists of spaced insulating washers or beads with air between, a corrected value for K (K') must be used:

$$K' = \frac{K - 1}{s} w$$

where K is the dielectric constant of the spacer material, s is the distance between adjacent spacers, and w is the width of one spacer.

For coaxial lines with square or trough-type outer conductors, the value for b is taken as the height of one side × 1.079.

Parallel conductors:

$$Z_o = 276 \log \frac{D}{a}$$

where D is the center-to-center spacing between conductors and a is the radius of the conductors.

Parallel conductors in a concentric shield:

$$Z_o = \frac{276}{\sqrt{K}} \log \left[\frac{D}{a} \times \frac{1 - \left(\frac{D}{2b}\right)^2}{1 + \left(\frac{D}{2b}\right)^2} \right]$$

where K, D, b and a are the same as above.

Inductance, capacity and resistance— The capacity, inductance and resistance of transmission lines are linear functions, and are computed in terms of unit length. Except as noted, the symbols are the same as before.

Concentric and coaxial conductors:

$$L = 0.14 \log \frac{b}{a} \ \mu\text{h./ft.}$$

$$C = \frac{7.35}{\log \frac{b}{a}} \ \mu\mu\text{fd./ft.}$$

$$R = \frac{\sqrt{f}}{2a} \ \text{microhms/ft.}$$

where f is the operating frequency in cycles.

Parallel conductors:

$$L = 0.279 \log \frac{D}{a} \ \mu\text{h./ft.}$$

$$C = \frac{3.66}{\log \frac{D}{a}} \ \mu\mu\text{fd./ft.}$$

$$R = \frac{\sqrt{f}}{a} \ \text{microhms/ft.}$$

Reactance, resonant impedance and Q— For lines less than one-quarter of a wavelength long at the operating frequency,

$$X_L = Z_o \tan \frac{360°\, l}{\lambda}$$

where l is the length of the line and λ is one wavelength at the operating frequency (both measured in the same units).

To tune such a line to resonance, capacitive reactance equal to the inductive reactance must be connected across the line. The capacity required to tune a given line to resonance is

$$C = \frac{1}{Z_o \tan \frac{360°\, l}{\lambda}}$$

The parallel-resonant or input (sending-end) impedance of the line is

$$Z_s = \frac{X_L^2}{R} = \frac{L}{RC} = X_L Q$$

For high parallel-resonant impedance (maximum gain), the b/a or D/a ratio should be made between 6 and 12.

The Q of a resonant line is

$$Q = \frac{2\pi f L}{R} = \frac{1}{R}\sqrt{\frac{L}{C}}$$

For high Q (maximum selectivity) the b/a or D/a ratio should be made between 2.5 and 5. Where the conductor spacing is within these limits,

$$Q = 0.22\, b\sqrt{f}$$

as an approximation accurate to within ± 10 per cent.

Ohm's Law Formulas

Ohm's Law for D. C.

W	EI	I^2R	$\frac{E^2}{R}$			
I			$\frac{E}{R}$	$\frac{W}{R}$	$\frac{W}{E}$	
R	$\frac{E}{I}$				$\frac{E^2}{W}$	$\frac{W}{I^2}$
E		IR		WR		$\frac{W}{I}$

Ohm's Law for A. C. (Single Phase)

W	$\frac{E^2 \cos\theta}{Z}$	$EI \cos\theta$	$I^2Z \cos\theta$	I^2R		
I	$\frac{W}{E \cos\theta}$		$\frac{E}{Z}$	$\frac{W}{R}$	$\frac{W}{Z \cos\theta}$	
R	$\frac{E^2 \cos^2\theta}{W}$	$\frac{E \cos\theta}{I}$	$Z \cos\theta$		$\frac{W}{I^2}$	$Z^2 - X^2$
E		$\frac{W}{I \cos\theta}$	IZ	$\frac{WR}{\cos\theta}$	$\frac{WZ}{\cos\theta}$	
X	$(X_L - X_C)$		$(2\pi fL - \frac{1}{2\pi fC})$			$Z^2 - R^2$
Z	$\frac{E}{I}$	$\frac{W}{I^2 \cos\theta}$		$\frac{R}{\cos\theta}$	$\frac{E^2 \cos\theta}{W}$	$R^2 + X^2$
$\cos\theta$	$\frac{IR}{E}$	$\frac{W}{I^2Z}$	$\frac{WZ}{E^2}$	$\frac{R}{Z}$	$\frac{W}{EI}$	$\frac{R}{R^2+X^2}$

Z = Impedance in ohms
X_L = Inductive reactance in ohms
X_C = Capacitive reactance in ohms
$\cos\theta$ = Power Factor
L = Inductance in henries
C = Capacity in farads
θ = Angle of lead or lag
f = Frequency in cycles per second

Table of Dielectric Characteristics

Dielectric material[1]	*Dielectric constant (K)*	*Power factor*[2] *60 cycles*	*1 kc.*	*1 Mc.*	*10 Mc.*	*100 Mc.*	*Dielectric strength (puncture voltage)*[2]	*Volume resistivity*[3] *(ρ) ohms per cm.*
Air (normal pressure)	1.0						19.8–22.8	
Aniline formaldehyde	3–5	1–6					400	
Asphalts	2.7–3.1		2.3				25–30	
Bakelite — *See* Phenol								
Beeswax	2.9–3.2							
Casein plastics[4]	6.1–6.4			5.2–6			165	
Castor oil	4.3–4.7			7			380	
Celluloid	4–16			5–10				
Cellulose acetate[5]	6–8	3–6	4–6	4–6	5.5		300–1000	4.5×10^{10}
Cellulose nitrate[6]	4–7			2.8–5			300–780	$2\text{–}30 \times 10^{10}$
Ceresin wax	2.5–2.6			0.12–0.21				
Cresol formaldehyde	6	10					400	
Ethyl cellulose	2–2.7	0.7	1.2	1.5			1500	10^{15}
Fiber	5–7.5			4.5–5			150–180	5×10^{9}
Glass:								
Cobalt	7.3			0.7				
Common window	7.6–8			1.4			200–250	
Crown	6.2–7		1	1[3]			500	
Electrical	4–5			0.5			2000	8×10^{14}
Flint	7–10		0.45	0.4				
Nonex	4.2			0.25		0.28		
Photographic	7.5			0.8–1				
Plate	6.8–7.6			0.6–0.8				
Pyrex	4.2–4.9		0.5	0.7		0.54	335	10^{14}
Gutta percha	2.5–4.9						200–500	$5 \times 10^{14}\text{–}10^{15}$
Lucite[7]	2.5–3	7	5	1.5–3	1.9		480–500	
Melamine formaldehyde	8	16					300	
Mica	2.5–8	0.2	0.3	0.2–6	0.02			2×10^{17}
Mica (clear India)	6.4–7.5	2	2	2	2		600–1500	
Mycalex	7.4			0.18			250	10^{13}
Mycalex (British)	6			0.3			350	
Nylon	3.6			2.2				
Paper	2.0–2.6						1250	
Paraffin wax (solid)	1.9–2.6			0.1–0.3			300	$10^{15}\text{–}10^{19}$
Phenol:[8]								
Pure	5			1			400–475	1.5×10^{12}
Asbestos base	7.5			15			90–150	
Black molded	5–5.5			3.5			400–500	
Fabric base	5–6.5			3.5–11			150–500	
Mica-filled	5–6			0.8–1			475–600	
Paper base	3.8–5.5			2.5–4			650–750	$10^{10}\text{–}10^{13}$
Yellow	5.3–5.4			0.36–0.7			500	
Polyethylene	2.2	0.06					600	10^{17}
Polyindene	3	0.04						
Polyisobutylene	2.4–2.5	0.04–5	0.05				500	10^{16}
Polystyrene[9]	2.4–2.9(2.6)	0.02	0.018	0.02	0.02	0.02	500–2500	10^{20}
Porcelain (dry process)	6.2–7.5			0.7–15			40–100	5×10^{8}
Porcelain (wet process)	6.5–7			0.6			150	
Pressboard (untreated)	2.9–4.5						125–300	
Pressboard (oiled)	5						750	
Quartz (fused)	3.5–(3.8)	0.01	0.01	0.015–0.03	0.01	0.05	200	$10^{14}\text{–}10^{18}$
Rubber (hard)[10]	2–3.5(3)			0.5–1			450	$10^{12}\text{–}10^{15}$
Shellac	2.5–4			0.09			900	10^{16}
Steatite:[11]								
"Commercial" grade	4.9–6.5	0.02	0.2	0.2	0.4	0.5		
"Low-loss" grade	4.4	0.02	0.2	0.2	0.18	0.13	150–315	$10^{14}\text{–}10^{15}$
Titanium dioxide[12]	90–170		0.1	0.1				
Urea formaldehyde[13]	5–7	3–5	2–3	2–4	4		300–550	$10^{12}\text{–}10^{13}$
Varnished cloth[14]	2–2.5			2–3			440–550	
Vinyl resins	4			1.4–1.7			400–500	10^{14}
Vitrolex	6.4			0.3				
Wire coverings:								
Enamel, ordinary							500–750	
Enamel, plastic							4500	
Cotton							140	
Silk							450	
Wood (dry oak)	2.5–6.8(3)		3.8	4.2				
Wood (paraffined maple)	4.1						115	

[1] Most data taken at 25° C.

[2] Puncture voltage, in volts per mil. Most data applies to relatively thin sections and cannot be multiplied directly to give breakdown for thicker sections without added safety factor.

[3] In ohms per cm.

[4] Includes such products as Aladdinite, Ameroid, Galalith, Erinoid, Lactoid, etc.

[5] Includes Fibestas, Lumerith, Nixonite, Plastacele, Tenite, etc.

[6] Includes Amerith, Nitron, Nixonoid, Pyralin, etc.

[7] Methylmethacrylate resin.

[8] Phenolaldehyde products include Acrolite, Bakelite, Catalin, Celeron, Dielecto, Durez, Durite, Formica, Gemstone, Heresite, Indur, Makalot, Marblette, Micarta, Opalon, Prystal, Resinox, Synthane, Textolite, etc. Yellow bakelite is so-called "low-loss" bakelite.

[9] Includes Amphenol 912A, Distrene, Intelin IN 45, Loalin, Lustron, Quartz Q, Rezoglas, Rhodolene M, Ronilla L, Styraflex, Styron, Trolitul, Victron, etc.

[10] Also known as Ebonite.

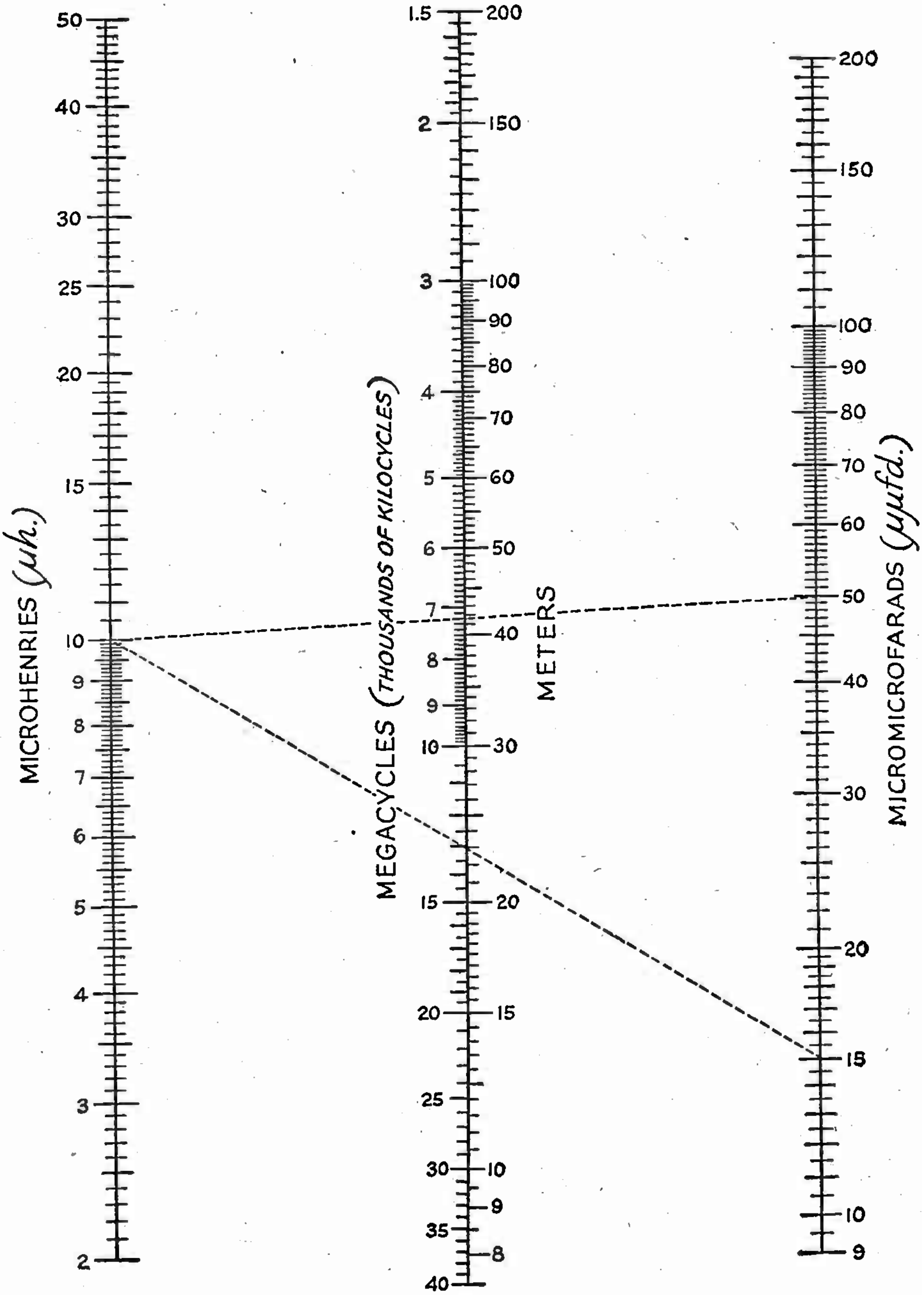
[11] Soapstone — Alberene, Alsimag, Isolantite, Lava, etc.

[12] Rutile. Used in low temperature-coefficient fixed condensers.

[13] Includes Aldur, Beetle, Plaskon, Pollopas, Prystal, etc.

[14] Includes Empire cloth.

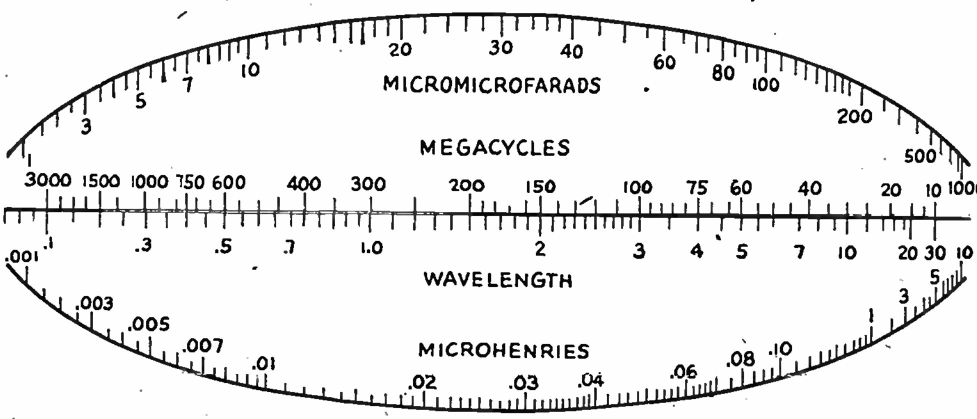
INDUCTANCE, CAPACITY AND FREQUENCY — CHART I, 1.5–30 MC.



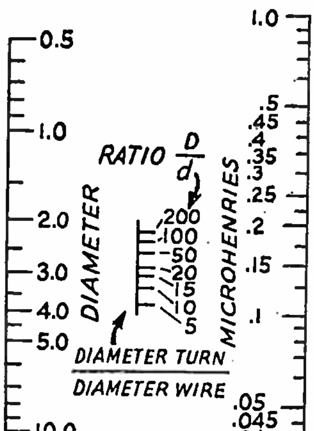
This chart may be used to find the values of inductance and capacity required to resonate at any given frequency in the medium- or high-frequency ranges; or, conversely, to find the frequency to which any given coil-condenser combination will tune. In the example shown by the dashed lines, a condenser has a minimum capacity of 15 μμfd and a maximum capacity of 50 μμfd. If it is to be used with a coil of 10-μh. inductance, what frequency range will be covered? The straight-edge is connected between 10 on the left-hand scale and 15 on the right, giving 13 Mc. as the high-frequency limit. Keeping the straight-edge at 10 on the left-hand scale, the other end is swung to 50 on the right-hand scale, giving a low-frequency limit of 7.1 Mc. The tuning range would, therefore, be from 7.1 Mc. to 13 Mc., or 7100 kc. to 13,000 kc. The center scale also serves to convert frequency to wavelength.

The range of the chart can be extended by multiplying each of the scales by 0.1 or 10. In the example above, if the capacities are 150 and 500 μμfd. and the inductance 100 μh., the range becomes approximately 231 to 422 meters or 0.7 to 1.3 Mc. Alternatively, 1.5 to 5 μμfd. and 1 μh. will give a range of approximately 71 to 130 Mc.

INDUCTANCE, CAPACITY AND FREQUENCY — CHART II, 30–300 MC.



The chart above is an extension of the chart on the preceding page, by means of which the same values of L, C and F can be found for the very-high- and ultrahigh-frequency ranges. It is used in the same manner as Chart I.

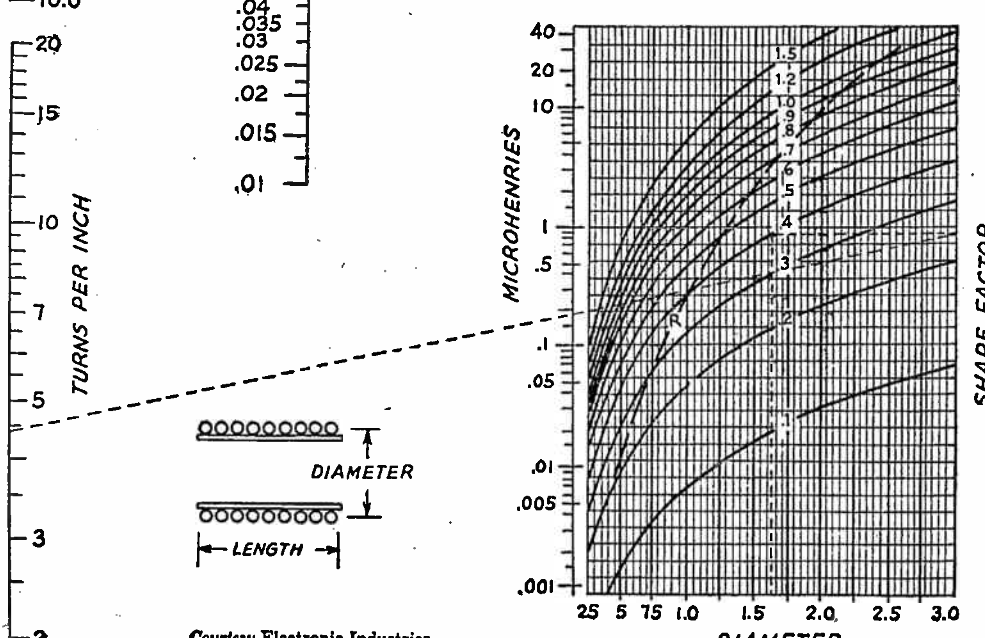


The charts on this page, prepared by Ralph R. Batcher, make possible the design of small coils for very-high frequency use with reasonable accuracy. The inductance of coils with two or more turns, or the dimensions required for a given value of inductance, may be determined from the chart above. The inductance of single-turn coils or loops can be obtained from the chart at the left.

To find the inductance of a given coil from the chart above, first locate the diameter in inches on the abscissa at the bottom. Draw a vertical line upward to intersect the curved scale corresponding to the length in inches. From this intersection project a line at right angles to locate the "shape factor" on the right-hand ordinate. Draw a line from this point to the "turns per inch" scale at the left; the point where this line crosses the "microhenries" scale is the inductance of the coil. In the example shown by the dashed lines, a coil of 1⅝-inch diameter wound with 1¾ turns, 0.4 inches long, or 4.4 turns per inch, has an inductance of approximately 0.18 μh.

To find the inductance of a single-turn coil or loop, draw a line from the diameter scale on the chart at the left through the point on the center scale which corresponds to the ratio of turn diameter to conductor diameter. The point where this line crosses the "microhenries" scale will give the inductance. Thus, a 2-inch loop of No. 12 wire has an inductance of approximately 0.11 μh.

In computing dimensions, the diameter of the coil is taken between centers of turns (diameter of form plus diameter of wire), while the length is the overall length of the coil (number of turns times distance between centers).



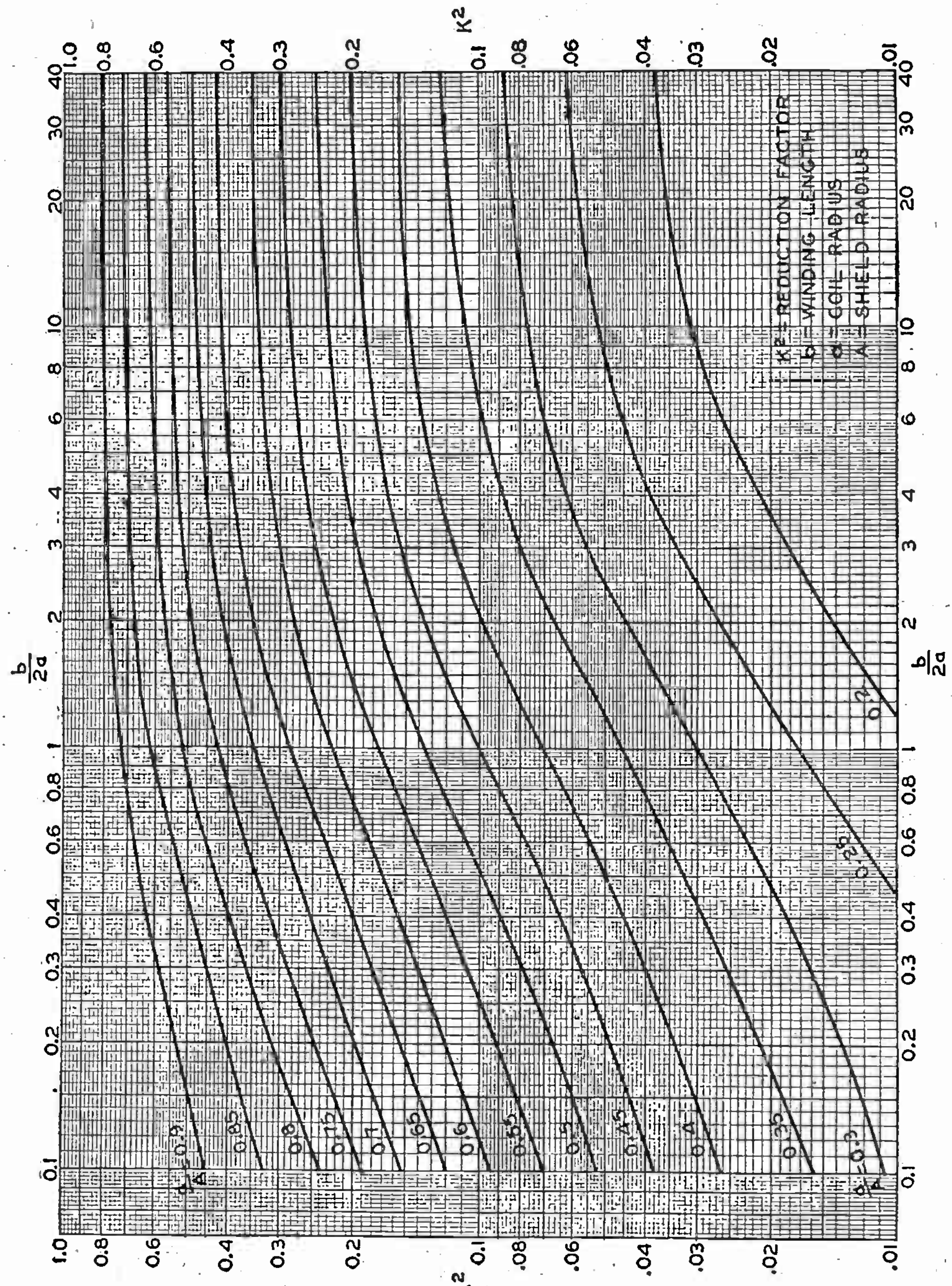
Courtesy Electronic Industries

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.	Current Carrying Capacity at 1500 C.M. per Amp.[3]	Diam. in mm.	Nearest British S.W.G. No.
			Enamel	S.C.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel S.C.C.	D.C.C.	Bare	D.C.C.				
1	289.3	83690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348	1
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544	3
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827	4
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189	5
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	22.0	4.621	7
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115	8
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665	9
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264	10
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906	11
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	12
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305	13
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053	14
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828	15
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628	16
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450	17
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291	18
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
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
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116	20
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118	21
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230	22
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438	23
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733	24
24	20.10	404.0	46.3	35.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106	25
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547	26
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049	27
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606	29
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211	30
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859	31
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546	33
31	8.928	79.70	101	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268	34
32	7.950	63.21	113	101	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019	36
33	7.080	50.13	127	110	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798	37
34	6.305	39.75	143	120	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601	38
35	5.615	31.52	158	132	104	73.5	10900	8700	—	10480	6737	335.0	.021	.1426	38–39
36	5.000	25.00	175	143	111	77.0	12200	10700	—	13210	7877	423.0	.017	.1270	39–40
37	4.453	19.83	198	154	118	80.3	—	—	—	16660	9309	533.4	.013	.1131	41
38	3.965	15.72	224	166	126	83.6	—	—	—	21010	10666	672.6	.010	.1007	42
39	3.531	12.47	248	181	133	86.6	—	—	—	26500	11907	848.1	.008	.0897	43
40	3.145	9.88	282	194	140	89.7	—	—	—	33410	14222	1069	.006	.0799	44

[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.

EFFECT OF COIL SHIELDS ON INDUCTANCE



Enclosing a coil in a shield decreases the inductance of the coil (§ 2-11). Considering the shield as a single turn having low resistance compared to its reactance, the following formula gives the actual inductance of the coil within the shield: $L = L_o(1-K^2)$, where L is the desired inductance, L_o is the inductance of the coil outside the shield, and K^2 is a factor from the chart above. b — length of winding of coil; a — radius of coil; A — radius of shield. The curves are sufficiently accurate for all practical purposes throughout the range shown *when the length of the shield is greater than that of the coil by at least the radius of the coil.* If the shield can is square instead of circular, A may be taken as 0.6 times the width of one side. The reduction factor, K^2, is plotted against $b/2a$ (ratio of length to diameter of coil), for a series of values of a/A, the ratio of coil radius to shield radius (or coil diameter to shield diameter). The reduction in inductance does not become serious with coils of $b/2a$ ratios of 2 or less, until the shield diameter becomes less than twice the coil diameter. With an a/A ratio of 0.5, the reduction will be of the order of 15 per cent.

RMA Radio Color Codes

Standard color codes have been adopted by the Radio Manufacturers Association for the identification of the values and connections of standard components.

RESISTOR-CONDENSER COLOR CODE

Color	*Significant Figure*	*Decimal Multiplier*	*Tolerance (%)*	*Voltage Rating**
Black	0	1	–	—
Brown	1	10	1*	100
Red	2	100	2*	200
Orange	3	1000	3*	300
Yellow	4	10,000	4*	400
Green	5	100,000	5*	500
Blue	6	1,000,000	6*	600
Violet	7	10,000,000	7*	700
Gray	8	100,000,000	8*	800
White	9	1,000,000,000	9*	900
Gold	–	0.1	5	1000
Silver	–	0.01	10	2000
No color	–	—	20	500

* Applies to condensers only.

Condensers:

If one row of three colored markers appears on the condenser, the voltage rating is 500 volts and the capacity is expressed to two significant figures in micromicrofarads as follows (for the usual tolerance of 20 per cent): First dot on left, first significant figure. Second dot, second significant figure. Third dot, decimal multiplier.

For example — A condenser having one row of colored markers as follows: *brown, black* and *brown* would have a capacity of 100 $\mu\mu$fd.

When two rows of three colored markers appear on the condenser, the top row represents the significant figures, read from left to right; the bottom row indicates the decimal multiplier, tolerance, and voltage rating, read from right to left. Capacity is in micromicrofarads.

For example — A condenser has two rows of colored markers as follows: Top row — left, *brown;* center, *black;* right, *no color;* Bottom row — right, *brown;* center, *green;* left, *blue.* Its ratings are 100 $\mu\mu$fd. ± 5%, 600 volts.

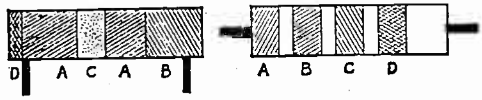
Two groups of colored bands are used on tubular condensers. Viewed with the wide bands on the right, the wide bands indicate the significant figures read from left to right; the narrow bands indicate the decimal multiplier, tolerance and voltage rating, from right to left, respectively.

Resistors:

Two different types of resistors are commonly used, one having radial leads, and the other axial leads.

The resistor and condenser color code shown in the preceding paragraph is used in determining the values of resistance and tolerance indicated by the colored dots, bands or stripes on the resistor.

The following illustration and chart shows the two types of resistors and the method of interpretation.



Radial Leads	*Axial Leads*	*Color*
Body A	*Band A*	*Indicates first significant figure of resistance value in ohms.*
End B	*Band B*	*Indicates second significant figure.*
Band C (or Dot)	*Band C*	*Indicates decimal multiplier.*
Band D	*Band D*	*If shown indicates tolerance in per cent. If no color is shown tolerance is ± 20 per cent.*

I.f. transformers:

Blue — plate lead.
Red — "B" + lead.
Green — grid (or diode) lead.
Black — grid (or diode) return.

NOTE: If the secondary of the i.f.t. is center-tapped, the second diode plate lead is green-and-black striped, and black is used for the center-tap lead.

A.f. transformers:

Blue — plate (finish) lead of primary.
Red — "B" + lead (this applies whether the primary is plain or center-tapped).
Brown — plate (start) lead on center-tapped primaries. (Blue may be used for this lead if polarity is not important.)
Green — grid (finish) lead to secondary.
Black — grid return (this applies whether the secondary is plain or center-tapped).
Yellow — grid (start) lead on center-tapped secondaries. (Green may be used for this lead if polarity is not important.)

NOTE: These markings apply also to line-to-grid, and tube-to-line transformers.

Loudspeaker voice coils:

Green — finish.
Black — start.

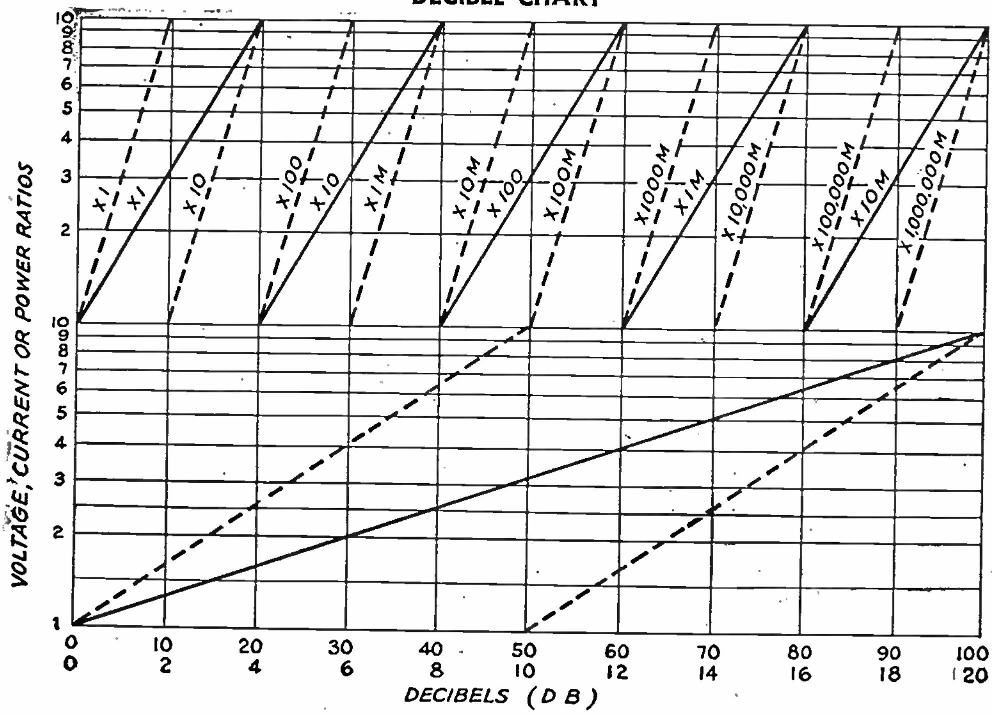
Field coils:

Black and red — start.
Yellow and red — finish.
Slate and Red — tap (if any).

Power transformers:

1) Primary Leads .... *Black*
  If tapped:
    Common .... *Black*
    Tap .... *Black and Yellow Striped*
    Finish .... *Black and Red Striped*
2) High-Voltage Plate Winding .... *Red*
    Center-Tap ... *Red and Yellow Striped*
3) Rectifier Filament Winding .... *Yellow*
    Center-Tap .. *Yellow and Blue Striped*
4) Filament Winding No. 1 .... *Green*
    Center-Tap .. *Green and Yellow Striped*
5) Filament Winding No. 2 .... *Brown*
    Center-Tap . *Brown and Yellow Striped*
6) Filament Winding No. 3 .... *Slate*
    Center-Tap ... *Slate and Yellow Striped*

DECIBEL CHART



– – – – POWER ———— VOLTAGE OR CURRENT

The chart above is direct-reading in terms of decibels for all power, voltage or current ratios. The top scale goes from 0 to 100 db. and is useful for very large ratios; the lower scale permits closer reading between 0 and 20 db., or one cycle of the extended scale. Solid lines show voltage or current ratios; dotted lines, power ratios. To find db. gain, divide output power by corresponding input power and read db. value for this ratio, using the appropriate curve (i.e., " × 1" for ratios from 1 to 10, " × 10" for ratios from 10 to 100, " × 100" for ratios from 100 to 1000, and so on). To find db. loss, as where output is less than input, divide input value by output value. Current and voltage ratios in db. can be found similarly, provided the input and output impedances are the same. Power, voltage and current values must be in the same units (watts, millivolts, microamperes, etc.).

DECIBEL TABLE

Power Ratio	*Voltage and Current Ratio*	*Decibels*	*Power Ratio*	*Voltage and Current Ratio*	*Decibels*
1.0000	1.0000	0			
1.0233	1.0116	0.1	19.953	4.4668	13.0
1.0471	1.0233	0.2	25.119	5.0119	14.0
1.0715	1.0315	0.3	31.623	5.6234	15.0
1.0965	1.0471	0.4	39.811	6.3096	16.0
1.1220	1.0593	0.5	50.119	7.0795	17.0
1.1482	1.0715	0.6	63.096	7.9433	18.0
1.1749	1.0839	0.7	79.433	8.9125	19.0
1.2023	1.0965	0.8	100.00	10.0000	20.0
1.2303	1.1092	0.9	158.49	12.589	22.0
1.2589	1.1220	1.0	251.19	15.849	24.0
1.3183	1.1482	1.2	398.11	19.953	26.0
1.3804	1.1749	1.4	630.96	25.119	28.0

1.4454	1.2023	1.6	1000.0	31.623	30.0
1.5136	1.2303	1.8	1584.9	39.811	32.0
1.5849	1.2589	2.0	2511.9	50.119	34.0
1.6595	1.2882	2.2	3981.1	63.096	36.0
1.7328	1.3183	2.4	6309.6	79.433	38.0
1.8198	1.3490	2.6	10^4	100.000	40.0
1.9055	1.3804	2.8	$10^4 \times 1.585$	125.89	42.0
1.9953	1.4125	3.0	$10^4 \times 2.512$	158.49	44.0
2.2387	1.4962	3.5	$10^4 \times 3.981$	199.53	46.0
2.5119	1.5849	4.0	$10^4 \times 6.31$	251.19	48.0
2.8184	1.6788	4.5	10^5	316.23	50.0
3.1623	1.7783	5.0	$10^5 \times 1.585$	398.11	52.0
3.5480	1.8836	5.5	$10^5 \times 2.512$	501.19	54.0
3.9811	1.9953	6.0	$10^5 \times 3.981$	630.96	56.0
5.0119	2.2387	7.0	$10^5 \times 6.31$	794.33	58.0
6.3096	2.5119	8.0	10^6	1,000.00	60.0
7.9433	2.8184	9.0	10^7	3,162.3	70.0
10.0000	3.1623	10.0	10^8	10,000.0	80.0
12.589	3.5480	11.0	10^9	31,623.0	90.0
15.849	3.9811	12.0	10^{10}	100,000.0	100.0

ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.	Medium frequency	m.f.
Ampere (amperes)	a.	Megacycles (per second)	Mc.
Amplitude modulation	a.m.	Megohm	MΩ
Antenna	ant.	Meter	m.
Audio frequency	a.f.	Microfarad	μfd.
Centimeter	cm.	Microhenry	μh.
Continuous waves	c.w.	Micromicrofarad	μμfd.
Cycles per second	c.p.s.	Microvolt	μv.
Decibel	db.	Microvolt per meter	μv/m.
Direct current	d.c.	Microwatt	μw.
Electromotive force	e.m.f.	Milliampere	ma.
Frequency	f.	Millivolt	mv.
Frequency modulation	f.m.	Milliwatt	mw.
Ground	gnd.	Modulated continuous waves	m.c.w.
Henry	h.	Ohm	Ω
High frequency	h.f.	Power	P.
Intermediate frequency	i.f.	Power factor	p.f.
Interrupted continuous waves	i.c.w.	Radio frequency	r.f.
Kilocycles (per second)	kc.	Ultrahigh frequency	u.h.f.
Kilovolt	kv.	Very-high frequency	v.h.f.
Kilowatt	kw.	Volt (volts)	v.
Magnetomotive force	m.m.f.	Watt (watts)	w.

Symbols for Electrical Quantities

Admittance	Y, y
Angular velocity (2πf)	ω
Capacitance	C
Conductance	G, g
Conductivity	γ
Current	I, i
Difference of potential	E, e
Dielectric constant	K
Dielectric flux	Ψ
Energy	W
Frequency	f
Impedance	Z, z
Inductance	L
Magnetic intensity	H
Magnetic flux	Φ
Magnetic flux density	B
Magnetomotive force	F
Mutual inductance	M
Number of conductors or turns	N
Period	T
Permeability	μ
Phase displacement	θ
Power	P, p
Quantity of electricity	Q, q
Reactance	$X, \bar{x}$
Reactance, Capacitive	X_c
Reactance, Inductive	X_L
Reluctivity	v
Resistance	R, r
Resistivity	ρ
Susceptance	b
Speed of rotation	n
Voltage	E, e
Work	W

Greek Alphabet

Since Greek letters are used to stand for many electrical and radio quantities, the names and symbols of the Greek alphabet with the equivalent English characters are given.

Greek Letter	*Greek Name*	*English Equivalent*
Α α	Alpha	a
Β β	Beta	b
Γ γ	Gamma	g
Δ δ	Delta	d
Ε ε	Epsilon	e
Ζ ζ	Zeta	z
Η η	Eta	é
Θ θ	Theta	th
Ι ι	Iota	i
Κ κ	Kappa	k
Λ λ	Lambda	l
Μ μ	Mu	m
Ν ν	Nu	n
Ξ ξ	Xi	x
Ο ο	Omicron	ŏ
Π π	Pi	p
Ρ ρ	Rho	r
Σ σ	Sigma	s
Τ τ	Tau	t
Υ υ	Upsilon	u
Φ φ	Phi	ph
Χ χ	Chi	ch
Ψ ψ	Psi	ps
Ω ω	Omega	ō

Standard Metal Gauges

Gauge No.	*American or B. & S.*[1]	*U. S. Standard*[2]	*Birmingham or Stubs*[3]
1	.2893	.28125	.300
2	.2576	.265625	.284
3	.2294	.25	.259
4	.2043	.234375	.238
5	.1819	.21875	.220
6	.1620	.203125	.203
7	.1443	.1875	.180
8	.1285	.171875	.165
9	.1144	.15625	.148
10	.1019	.140625	.134
11	.09074	.125	.120
12	.08081	.109375	.109
13	.07196	.09375	.095
14	.06408	.078125	.083
15	.05707	.0703125	.072
16	.05082	.0625	.065
17	.04526	.05625	.058
18	.04030	.05	.049
19	.03589	.04375	.042
20	.03196	.0375	.035
21	.02846	.034375	.032
22	.02535	.03125	.028
23	.02257	.028125	.025
24	.02010	.025	.022
25	.01790	.021875	.020
26	.01594	.01875	.018
27	.01420	.0171875	.016
28	.01264	.015625	.014
29	.01126	.0140625	.013
30	.01003	.0125	.012
31	.008928	.0109375	.010
32	.007950	.01015625	.009
33	.007080	.009375	.008
34	.006350	.00859375	.007
35	.005615	.0078125	.005
36	.005000	.00703125	.004
37	.004453	.006640626	....
38	.003965	.00625	....
39	.003531	......	....
40	.003145	......	....

[1] Used for aluminum, copper, brass and non-ferrous alloy sheets, wire and rods.
[2] Used for iron, steel, nickel and ferrous alloy sheets, wire and rods.
[3] Used for seamless tubes; also by some manufacturers for copper and brass.

Metric Prefixes

μ	$\frac{1}{1,000,000}$	One-millionth	micro-
m	$\frac{1}{1,000}$	One-thousandth	milli-
c	$\frac{1}{100}$	One-hundredth	centi-
d	$\frac{1}{10}$	One-tenth	deci-
	1	One	uni-
dk	10	Ten	deka-
h	100	One hundred	hekto-
k	1,000	One thousand	kilo-
	10,000	Ten thousand	myria-
M	1,000,000	One million	mega-

Multiples and Sub-Multiples

Ampere	= 1,000,000 microamperes
Ampere	= 1,000 milliamperes
Cycle	= 0.000,001 megacycle
Cycle	= 0.001 kilocycle
Farad	= 1,000,000,000,000 micro-microfarads
Farad	= 1,000,000 microfarads
Farad	= 1,000 millifarads
Henry	= 1,000,000 microhenrys
Henry	= 1,000 millihenrys
Kilocycle	= 1,000 cycles
Kilovolt	= 1,000 volts
Kilowatt	= 1,000 watts
Megacycle	= 1,000,000 cycles
Megohm	= 1,000,000 ohms
Mho	= 1,000,000 micromhos
Mho	= 1,000 millimhos
Microampere	= 0.000,001 ampere
Microfarad	= 0.000,001 farad
Microhenry	= 0.000,001 henry
Micromho	= 0.000,001 mho
Micro-ohm	= 0.000,001 ohm
Microvolt	= 0.000,001 volt
Microwatt	= 0.000,001 watt
Micromicrofarad	= 0.000,000,000,001 farad
Micromicro-ohm	= 0.000,000,000,001 ohm
Milliampere	= 0.001 ampere
Millihenry	= 0.001 henry
Millimho	= 0.001 mho
Milliohm	= 0.001 ohm
Millivolt	= 0.001 volt
Milliwatt	= 0.001 watt
Volt	= 1,000,000 microvolts
Volt	= 1,000 millivolts
Watt	= 1,000,000 microwatts
Watt	= 1,000 milliwatts
Watt	= 0.001 kilowatt

Decimal Equivalents of Fractions

1/32...........	.03125	17/32...........	.53125
1/16........	.0625	9/16........	.5625
3/32...........	.09375	19/32...........	.59375
1/8.........	.125	5/8.........	.625
5/32...........	.15625	21/32...........	.65625
3/16........	.1875	11/16........	.6875
7/32...........	.21875	23/32...........	.71875
1/4.........	.25	3/4.........	.75
9/32...........	.28125	25/32...........	.78125
5/16........	.3125	13/16........	.8125
11/32...........	.34375	27/32...........	.84375
3/8.........	.375	7/8.........	.875
13/32...........	.40625	29/32...........	.90625
7/16........	.4375	15/16........	.9375
15/32...........	.46875	31/32...........	.96875
1/2.........	.5	1............	1.0

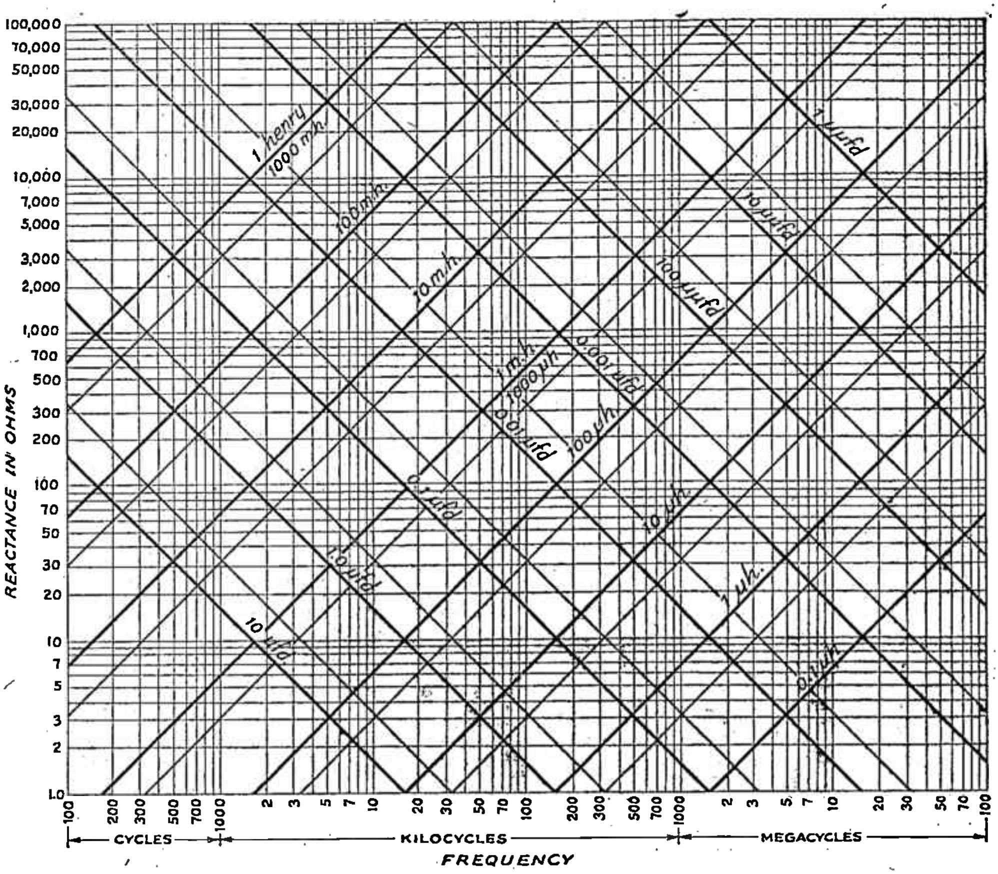
Units of Length

English		*Metric*	
1 mil	= 0.001 inch	1 millimeter	= 39.37 mils
	= 0.0254 millimeter		
1 inch	= 2.54 centimeters	1 centimeter	= 0.3937 inch
			= 0.0328 foot
1 foot	= 30.48 centimeters	1 meter	= 3.28 feet
1 yard	= 0.9144 meter		= 1.094 yards
1 mile	= 1.6093 kilometers	1 kilometer	= 0.6214 mile

1 micron = 10^{-6} meter
= 0.0001 centimeter
= 10,000 Angstrom units (A°)

1 Angstrom = 10^{-10} meter
= 10^{-8} centimeter
= 0.0001 micron

INDUCTIVE AND CAPACITIVE REACTANCE VS. FREQUENCY CHART



By use of the chart above, the approximate reactance of any capacity from 1.0 μμfd. to 10 μfd. at any frequency from 100 cycles to 100 megacycles, or the reactance of any inductance from 0.1 μh. to 1.0 henry, can be read directly. Intermediate values can be estimated by interpolation. In making interpolations, remember that the rate of change between lines is logarithmic. Use the frequency or reactance scales as a guide in estimating intermediate values on the capacity or inductance scales.

This chart also can be used to find the approximate resonance frequencies of *LC* combinations, or the frequency to which a given coil and condenser will tune. First locate the respective slanting lines for the capacity and inductance. The point where they intersect, i.e., where the reactances are equal, is the resonant frequency (projected downward and read on the frequency scale).

Electrical Conductivity of Metals

	Relative Conductivity [1]	*Temperature Coefficient* [2] *of Resistance*
Aluminum (2S; pure)	59	0.0049
Aluminum (alloys):		
Soft-annealed	45–50	
Heat-treated	30–45	
Brass	28	0.002–0.007
Cadmium	19	
Chromium	55	
Climax	1.83	
Cobalt	16.3	
Constantin	3.24	0.00002
Copper (hard drawn)	89.5	0.004
Copper (annealed)	100	
Everdur	6	
German Silver (18%)	5.3	0.00019
Gold	65	
Iron (pure)	17.7	0.006
Iron (cast)	2–12	
Iron (wrought)	11.4	
Lead	7	0.0041
Manganin	3.7	0.00002

	Relative Conductivity [1]	*Temperature Coefficient* [2] *of Resistance*
Mercury	1.66	0.00089
Molybdenum	33.2	0.0033
Monel	4	0.0019
Nichrome	1.45	0.00017
Nickel	12–16	0.005
Phosphor Bronze	36	0.004
Platinum	15	
Silver	106	0.004
Steel	3–15	
Tin	13	0.0042
Tungsten	28.9	0.0045
Zinc	28.2	0.0035

Approximate relations:

An increase of 1 in A. W. G. or B. & S. wire size increases resistance 25%.
An increase of 2 increases resistance 60%.
An increase of 3 increases resistance 100%.
An increase of 10 increases resistance 10 times.

[1] At 20° C., based on copper as 100.
[2] Per °C. at 20° C.

TABLE OF REACTANCES AT COMMONLY USED AUDIO AND RADIO FREQUENCIES

Capacitive Reactance in Ohms at:	1	2	2.5	4	5	6	8	10	12	16 μfd.
60 cycles	2630	1310	1060	655	530	442	332	265	221	166
100 "	1590	796	637	398	318	265	199	159	131	99.5
400 "	389	199	159	99.5	79.6	66.3	49.7	38.9	33.2	24.9
1000 "	159	79.6	63.7	39.8	31.8	26.5	19.9	15.9	13.1	9.95
5000 "	31.8	15.9	12.7	7.96	6.37	5.31	3.98	3.18	2.65	1.99

Capacitive Reactance in Ohms at:	10	20	30	40	50	60	70	80	90	100 $\mu\mu$fd.
100 kc.	159000	79600	53100	39800	31800	26500	22700	19900	17700	15900
455 "	35000	17500	11700	8750	7000	5830	5000	4370	3890	3500
1600 "	9960	4980	3320	2490	1990	1660	1420	1240	1110	996
1750 "	9100	4550	3030	2270	1820	1510	1300	1140	1010	910
2000 "	7960	3980	2650	1990	1590	1330	1140	995	885	796
5000 "	3180	1590	1090	795	635	530	454	398	353	318

Inductive Reactance in Ohms at:	1	2	3	4	5	6	7	8	9	10 henries
60 cycles	376	752	1130	1500	1880	2260	2630	3000	3380	3760
100 "	628	1260	1880	2520	3140	3770	4400	5020	5650	6280
400 "	2510	5030	7540	10500	12600	15100	17600	21000	22600	25100
1000 "	6280	12600	18800	25200	31400	37700	44000	50200	56500	62800
5000 "	31400	62800	94200	126000	157000	188000	220000	252000	283000	314000

Inductive Reactance in Ohms at:	10	20	30	40	50	60	70	80	90	100 μh.
100 kc.	6.28	12.6	18.8	25.2	31.4	37.7	44.0	50.2	56.5	62.8
455 "	28.6	57.2	85.8	114	143	171	200	228	257	286
1600 "	100	200	300	400	500	600	700	800	900	1000
1750 "	110	220	330	440	550	660	770	880	990	1100
2000 "	126	252	378	504	630	756	882	1008	1134	1260
5000 "	314	628	942	1260	1570	1880	2200	2520	2830	3140

To find reactance for other values of *L* or *C*, move the decimal point in the reactance figure to correspond with the difference in the *L* or *C* figure. Move the decimal point to the left for increasing capacity and to the right for increasing inductance, and vice versa.

To find reactance for higher-frequency multiples multiply the reactance figure by the multiple of the frequency, according to the sign of the reactance (i.e., multiply for inductive reactance and divide for capacitive reactance).

Example 1: The reactance of a 1000-$\mu\mu$fd. condenser at 1750 kc. is the value for 100 $\mu\mu$fd. with the decimal one place to the left, or 99.1 ohms. At 7 Mc. the reactance of this condenser is its value at 1750 kc. divided by 4, or 24.3 ohms.

Example 2: The reactance of a 60-henry choke at 60 cycles is the value for 6 henries with the decimal moved one place to the right, or 22600 ohms. At 120 cycles the reactance of this choke is its value at 60 cycles multiplied by 2, or 45200 ohms.

Numerical Values

$\pi = 3.1416$

$2\pi = 6.28$

$4\pi = 12.6$

$(2\pi)^2 = 39.4784$

$\log_{10} \pi = 0.4971$

$\log_{10} (\pi/2) = 0.1961$

$\log_{10} \pi^2 = 0.9943$

$\log_{10} \sqrt{\pi} = 0.2486$

$\log_{10} \epsilon = 0.4343$

$\epsilon = 2.7183$

$1/\epsilon = 0.3679$

$\epsilon^2 = 7.3890$

$\sqrt{\epsilon} = 1.6487$

$\sqrt{2} = 1.4142$

$\sqrt{3} = 1.7321$

$1/\sqrt{2} = 0.7071$

$1/\sqrt{3} = 0.5773$

$1/\pi = 0.3183$

$\pi^2 = 9.8696$

$1/\pi^2 = 0.1013$

$\pi^3 = 31.0063$

$1/\pi^3 = 0.0323$

$\sqrt{\pi} = 1.7725$

$1/\sqrt{\pi} = 0.5642$

$\sqrt{\pi/2} = 1.2533$

$2\pi = 6.2832$

$\dfrac{1}{\log_{10} \epsilon} = 2.3026$

$1/2\pi = 0.1592$

$(1/2\pi)^2 = 0.0253$

1 radian = 57° 17′ 44″ .8

= 57° 17′ .7468

= 57° .29578

1° = 0.001745 radian

360 degrees = 2π radians

Sine 1′ = .0002929

Area of circle = $(\pi/4) \times D^2 = 0.7854\ D^2$

Volume of sphere = $(\pi/6) \times D^3 = 0.5246\ D^3$

Area of triangle = Base × ½ height

Area of ellipse = Major axis × minor axis × .7854

Area of parabola = Base × ⅔ perpendicular height

Surface area of sphere = $4\pi r^2$

Side of square = .707 diagonal of square

Volume of pyramid = Area of base × ⅓ of height or cone

Letter Symbols for Vacuum-Tube Notation

Grid potential	E_g, e_g
Grid current	I_g, i_g
Grid conductance	g_g
Grid resistance	r_g
Grid bias voltage	E_c
Plate potential	E_p, e_p
Plate current	I_b, I_p, i_p
Plate conductance	g_p
Plate resistance	r_p
Plate supply voltage	E_b
Cathode current	I_c
Emission current	I_s
Mutual conductance	g_m
Amplification factor	μ
Filament terminal voltage	E_f
Filament current	I_f
Grid-plate capacity	C_{gp}
Grid-cathode capacity	C_{gk}
Plate-cathode capacity	C_{pk}
Grid capacity (input)	C_g
Plate capacity (output)	C_p

NOTE. — Small letters refer to instantaneous values.

Pilot Lamp Data

Lamp No.	Bead Color	Base (Miniature)	Bulb Type	RATING	
				Volts	Amps.
40	Brown	Screw	T-3¼	6–8	0.15
40A[1]	Brown	Bayonet	T-3¼	6–8	0.15
41	White	Screw	T-3¼	2.5	0.5
42	Green	Screw	T-3¼	3.2	**
43	White	Bayonet	T-3¼	2.5	0.5
44	Blue	Bayonet	T-3¼	6–8	0.25
45	*	Bayonet	T-3¼	3.2	**
46[2]	Blue	Screw	T-3¼	6–8	0.25
47[1]	Brown	Bayonet	T-3¼	6–9	0.15
48	Pink	Screw	T-3¼	2.0	0.06
49[3]	Pink	Bayonet	T-3¼	2.0	0.06
[4]	White	Screw	T-3¼	2.1	0.12
49A[2]	White	Bayonet	T-3¼	2.1	0.12
50	White	Screw	G-3½	6–8	0.2
51[2]	White	Bayonet	G-3½	6–8	0.2
—	White	Screw	G-4½	6–8	0.4
55	White	Bayonet	G-4½	6–8	0.4
292[5]	White	Screw	T-3¼	2.9	0.17
292A[5]	White	Bayonet	T-3¼	2.9	0.17
1455	Brown	Screw	G-5	18.0	0.25
1455A	Brown	Bayonet	G-5	18.0	0.25

* White in G.E. and Sylvania; Green in National Union Raytheon and Tung-Sol.

** 0.35 in G.E. and Sylvania; 0.5 in National Union Raytheon and Tung-Sol.

[1] 40A and 47 are interchangeable.

[2] Have frosted bulb.

[3] 49 and 49A are interchangeable.

[4] Replace with No. 48.

[5] Use in 2.5 volt sets where regular bulb burns out too frequently.

Receiving-Tube Ratings

The data in the classified tube tables on pages 481–512 are of two kinds — maximum ratings, and typical operating conditions.

As explained in §3-1, vacuum tubes are designed to be operated within definite maximum (and minimum) ratings. These ratings are the maximum safe operating voltages and currents for the electrodes, based on inherent limiting factors such as permissible cathode temperature, emission, and power dissipation in electrodes. In addition to the maximum ratings for each type performance data is given in the form of typical operating conditions showing applications and circuit-design considerations.

In the transmitting-tube tables, maximum ratings for electrode voltage, current and dissipation are given separately from the typical operating conditions for the recommended classes of operation. In the receiving-tube tables, because of space limitations, ratings and operating data are combined. Where only one set of operating conditions appears, the positive electrode voltages shown (plate, screen, etc.) are, in general, also the maximum rated voltages for those electrodes.

The maximum ratings are intended to apply under conditions of normal operation. In practice, power-supply or primary-source voltage fluctuations result in appreciable variation around the normal or designed operating conditions. The maximum limits of these variations must be taken into account in the design of equipment. Recommended practice is to assume a "design-center voltage," which is the normal voltage supplied by the primary power source. The over-all design is made such that the maximum upward variation (design-maximum voltage) will not cause the maximum ratings of components to be exceeded, while operation at the design-minimum voltage will not result in impaired performance. The usual ranges of variation for the three common power sources are given in the following table.

Power-supply design voltages:

Source	Design-Center Voltage	Design-Minimum Voltage	Design-Maximum Voltage
"115-volt" a.c. power line[1] . .	117[1]	105	130
"6-volt" storage battery	6	5.25	6.3[2]
	6.3[3]	. . . .	7–8[3]
"1.5-volt" dry cell[4]	1.4	1.1[5]	1.6

[1] Also applies to a.c.-d.c. equipment which may be operated from "110-volt" d.c. lines.

[2] Maximum terminal potential without load for fully charged battery with specific gravity of 1.300.

[3] For storage battery connected to charger, as in automotive service. Maximum voltage will vary with system; equipment should be rated to withstand 7 volts continuously, 8 volts intermittently.

[4] Values given are for filament operation. For "B" batteries, use standard rating as design-center voltage (45, 90, etc.); minimum and maximum voltages will be as shown multiplied by the number of cells.

[5] In some types of equipment, design-minimum voltages can be carried as low as 0.9 volts per cell.

RECEIVING TUBE CLASSIFICATION CHART

Cathode Volts	1.4	2.0	2.5 to 5.0	6.3	12.6 to 117
DIODE DETECTORS & RECTIFIERS					
Detectors — single	1A3				
Detectors — twin				(6H6, 6H6-GT/G), 7A6	12H6
Rectifiers — half-wave				1-v	12Z3, 35Z3, 35Z4-GT, 35Z5-GT/G, 45Z3, 45Z5-GT
Rectifiers — half-wave, with beam power amplifier					32L7-GT, 70L7-GT, 117L/M7-GT, 117N7-GT, 117P7-GT
Rectifiers — half-wave, with power pentode					12A7, 25A7-GT/G
Rectifiers — full-wave — vacuum			(5T4, 5U4-G, 5X4-G, 5Z3), (5W4, 5W4-GT/G, 5Y3-GT/G, 5Z4, 5Y4-G, 80), (5V4-G, 83-v)	(6X5, 6X5-GT/G, 84), 6Y5, 6Z5, 6ZY5-G, 7Y4	
Rectifiers — full-wave — mercury			82, 83		
Rectifiers — full-wave — gas	Cold-Cathode Types: 0Z4, 0Z4-G.				
Rectifier-Doublers					(25Z6, 25Z6-GT/G, 25Z5, 25Y5), 50Y6-GT/G, 50Z7-G, 117Z6-GT/G
DIODE DETECTORS with AMPLIFIERS					
One Diode — with high-mu triode	(1H5-G, 1H5-GT), 1LH4				
One Diode — with high-mu triode, r-f pentode	3A8-GT*				
One Diode — with medium-mu triode, power pentode	1D8-GT				
One Diode — with pentode	1S5				
One Diode — with power pentode	1N6-G				
Two Diodes — with medium mu-triode		(1B5, 1H6-G)	55	(6SR7, 6R7, 6R7-G, 6R7-GT, 6ST7, 6V7-G, 85), 6C7, 7E6	12SR7
Two Diodes — with high-mu triode			2A6	(6SQ7, 6SQ7-GT/G, 6Q7, 6Q7-G, 6Q7-GT, 6B6-G, 6T7-G, 75), 7B6, 7C6	(12SQ7, 12SQ7-GT/G, 12Q7-GT)
Two Diodes — with pentode		(1F7-G, 1F6)	2B7	(6B8, 6B8-G, 6B7, 6B7-S), 6SF7, 7E7	12C8, 12SF7
CONVERTERS & MIXERS					
Pentagrid Converters	(1A7-G, 1A7-GT), 1R5, 1B7-GT, 1LA6	(1C7-G, 1C6), (1D7-G, 1A6)	2A7	(6SA7, 6SA7-GT/G, 6A8, 6A8-G, 6A8-GT, 6D8-G, 6A7, 6A7S), 7B8, 7Q7	(12SA7, 12SA7-GT/G, 12A8-GT)
Triode-Hexode Converters				(6K8, 6K8-G, 6K8-GT),	12K8
Triode-Heptode Converters				6J8-G, 7J7	
Octode Converters				7A8	
Pentagrid Mixers				(6L7, 6L7-G)	

Courtesy of R.C.A.

VOLTAGE AMPLIFIERS, DETECTORS, OSCILLATORS					
Triodes — medium-mu — single unit	1G4-GT/G	(1H4-G, 30)	27, 56, 485	(6C5, 6C5-GT/G), (6J5, 6J5-GT/G, 7A4), (6P5-GT/G, 76), 6L5-G, 6AE5-GT/G, 37	12J5-GT
Triodes — medium-mu — twin unit	3A5*			6C8-G, 6F8-G 6J6, 6SN7-GT	12AH7-GT 12SN7-GT
Triodes — medium-mu — twin plate				6AE6-G	
Triodes — medium-mu — twin input				6AE7-GT	
Triodes — medium-mu — with power pentode				6AD7-G	
Triodes — medium-mu — with diode, power pentode	1D8-GT				
Triodes — high-mu — single unit				6SF5, 6SF5-GT, 6F5, 6F5-G, 6F5-GT), 6K5-G, 7B4	(12SF5, 12SF5-GT, 12F5-GT)
Triodes — high-mu — twin unit				(6SC7, 7F7), 6SL7-GT	12SC7, 12SL7-GT
Triodes — high-mu — with diode, r-f pentode	3A8-GT*				
Tetrodes — remote cut-off		1D5-GT	35		
Tetrodes — sharp cut-off		32	24-A	36	
Pentodes — remote cut-off	1T4, 1P5-GT	(1D5-GP, 1A4-P), 34	58	6SS7, (6SK7, 6SK7-GT/G, 6K7, 6K7-G, 6K7-GT, 7B), (6S7, 6S7-G), (6U7-G, 6D6, 6E7), 6W7-G, 39/44, 7A7, 6AB7, 6AC7, 7H7, 7B7	(12SK7, 12SK7-GT, 12K7-GT), 14A7/12B7
Pentodes — remote cut-off, with triode				6F7, 6P7-G	12B8-GT, 25B8-GT
Pentodes — semi-remote cut-off				6SG7	12SG7
Pentodes — sharp cut-off	(1N5-G, 1N5-GT), 1L4, 1LN5	(1E5-GP, 1B4-P), 15	57	6AG5, 6SH7, (6SJ7, 6SJ7-GT, 6J7, 6J7-G, 6J7-GT, 6D7), 77, 6C6, 7C7, 7G7/1232	12SH7, (12SJ7, 12SJ7-GT, 12J7-GT)
Pentodes — sharp cut-off, with diode, high-mu triode	3A8-GT*				
POWER AMPLIFIERS					
Triodes — low-mu — single unit		31	2A3, 45, 1B3/483	6A3, 6B4-G	
Triodes — low-mu — twin unit				6E6	
Triodes — high-mu — single unit		49	46	6AC5-GT/G, 6C4	25AC5-GT/G
Triodes — high-mu — twin unit	1G6-GT/G	(1J6-G, 19)	53	(6N7, 6N7-GT/G, 6A6), (6Y7-G, 79), 6Z7-G	
Beam Tubes — without rectifier	(1Q5-GT/G, 3Q5-GT/G*), 1T5-GT			(6L6, 6L6-G), (6V6, 6V6-GT/G), 6Y6-G, 7A5, 7C5	(25L6, 25L6-GT/G), 25C6-G, 35A5, 35L6-GT/G, 50L6-GT
Beam Tubes — with rectifier					32L7-GT, 70L7-GT, 117L/M7-GT, 117N7-GT, 117P7-GT
Pentodes — single unit	1A5-GT/G, (1S4, 3S4*), 1C5-GT/G, 1LA4, 1LB4, (3A4*, 3Q4*)	(1F5-G, 1F4), (1G5-G, 1J5-G), 33	2A5, 47, 59	(6F6, 6F6-G, 42), (6K6-GT/G, 41), 6G6-G, 38, 6A4, 89, 7B5	12A5, (25A6, 25A6-GT/G, 43), 25B6-G
Pentodes — twin unit		1E7-G★			
Pentodes — with diode & triode	1D8-GT				
Pentodes — with medium-mu triode				6AD7-G	
Pentodes — with rectifier					12A7, 25A7-GT/G
Pentodes — video				6AG7	
Direct-Coupled Amplifiers				6B5, 6N6-G	(25B5, 25N6-G)
ELECTRON-RAY TUBES					
Single — with remote cut-off triode				6AB5/6N5, 6U5/6G5	
Single — with sharp cut-off triode			2E5	6E5	
Twin, without triode				6AD6-G, 6AF6-G	
GAS-TRIODES			2A4-G		

★ Two 1F5-G's in one bulb.

* Filament arranged for either 1.4 volt or 2.8-volt operation.

ARMY-NAVY PREFERRED LIST OF RADIO TUBES

RECEIVING

Filament Voltage	Diodes	Diode Triodes	Triodes	Twin Triodes	Pentodes		Converters	Power Output	Indicators	Rectifiers	Miscellaneous	
					Remote	Sharp					Cathode Ray	Crystals
1.4	1A3	1S5**	1LE3	3A5	1T4	1L4 1LN5 1S5**	1LC6 1R5	1LB4 3A4 3S4		1006	2AP1 3BP1 3DP1 3FP7 5CP1 5CP7 5FP7 5JP1 7BP7 12DP7 12GP7 913	1N21B 1N23B 1N25 1N26 1N27 1N28
5.0										5U4G 5Y3GT		
6.3	6AL5 6H6* 559	6AQ6 6SQ7* 6SR7*	2C22 2C40 6C4 6F4 6J4 6J5* 7E5/1201 9002	6J6 6SL7GT 6SN7GT 7F8	6AB7 6AG7 6SK7* 9003	6AC7 6AG7 6AJ5† 6AK5 6AS6 6SJ7* 7W7 9001	6SA7	6B4G 6G6G 6L6GA 6N7GT/G 6V6GT/G 6Y6G	6AF6 6E5	6X5GT/G 1005		*Phototubes* 927 929 930 931A *Voltage Regulators*
12.6	12H6*	12SQ7* 12SR7* 14E6†	12J5GT	12SL7GT 12SN7GT 14N7†	12SG7* 12SK7* 14R7†	12SJ7* 14H7† 14W7	12SA7* 14J7†	12A6*	1629			OA3/VR75 OC3/VR105 OD3/VR150
25 and above								25L6GT/G 28D7†	991	25Z6GT/G		

TRANSMITTING

Triodes	Tetrodes	Twin Tetrodes	Pentodes	Pulse Modulators	Rectifiers			Clipper Tubes	Gas Switching
					Vacuum	Gas	Grid Control		
2C26A 2C39 2C43 3C24 CV92(Br) 100TH 250TH 304TH 527 811 826 833A 862A 880 889R 893A 1626 8025	807 813 814 827R 1625	815 829B 832A	2E22 4E27 803 837	3D21 3E29 6C21 715C	2X2A 3B24 5R4GY 371B 705A 836 1616 8016 8020	4B25 83 857B 866A 869B 872A	2D21 3C23 3C31/C1B C5B 6D4 394A 884 2050	3B26 719A	1B32/532A

* Where direct interchangeability is assured, "GT" and "L" counterparts of the preferred metal tube may be used.
** Diode-pentode.
† These tubes are the only types with characteristics specified for 28-volt plate supply, and may be used in this type of application. Any of the types listed under "Receiving Diodes" may be used for 28-volt plate supply applications.

VACUUM-TUBE CLASSIFIED DATA TABLES AND INDEX

In the tables on pages 480–512 will be found essential characteristics and typical operating data for all U. S.-made standard receiving, transmitting and special-purpose vacuum tubes on which data is available, classified by use. Base diagrams are shown on pages 476–480. For convenience in locating types whose essential characteristics are not known, the index below lists all types in numerial/alphabetical order with the page on which data is to be found.

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2002	495	HK54	501	RK16	486	TZ40	500
2005	495	HK154	501	RK17	486	UH35	502
2050	494	HK158	501	RK18	500	UH50	501
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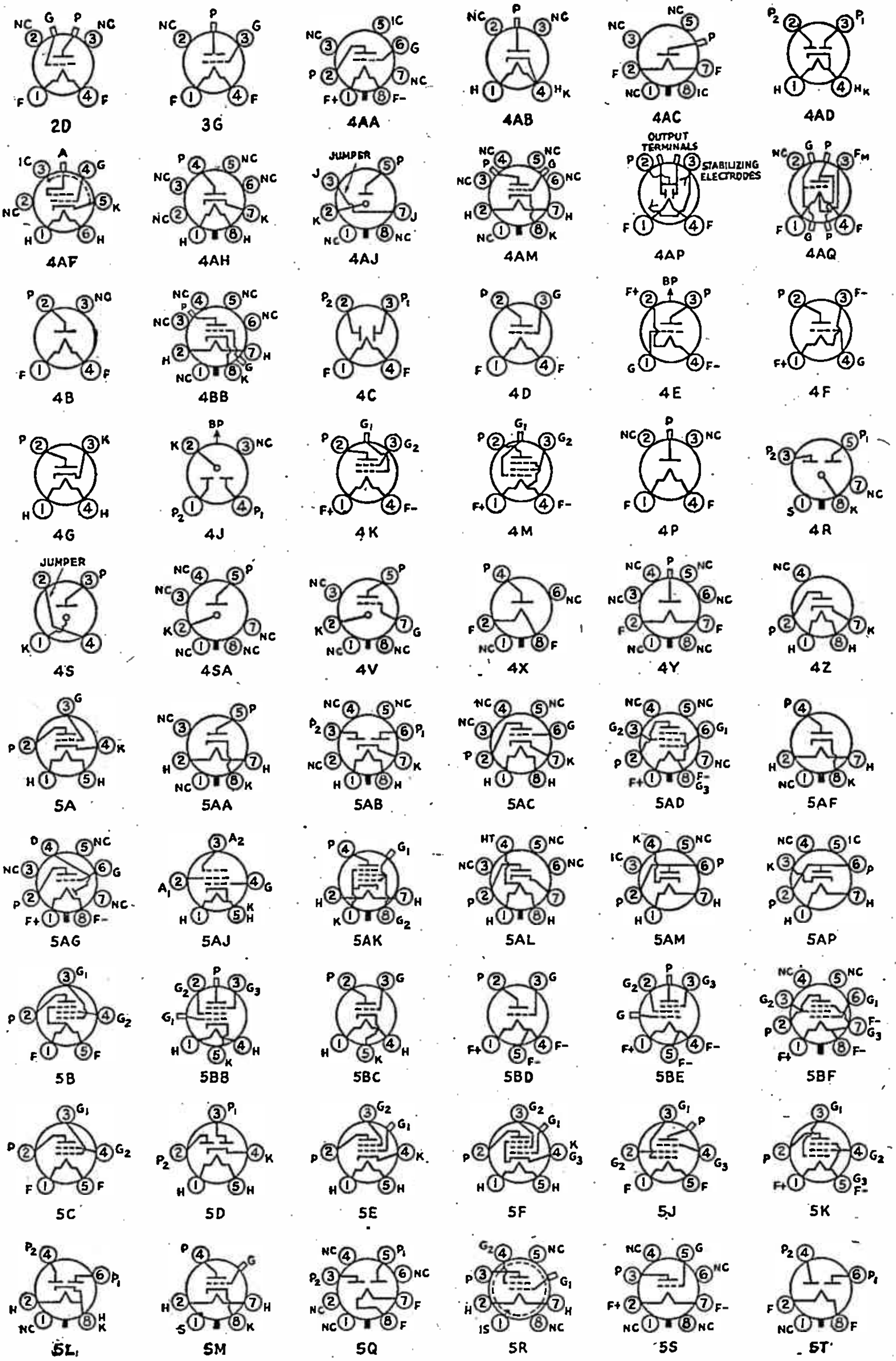
VACUUM-TUBE BASE DIAGRAMS

The diagrams on the following five pages show standard socket connections corresponding to the base designations given in the column headed "Socket Connections" in the classified tube data tables on pages 473–475. Footnotes under each table indicate in which group a given base diagram is to be found. In pages (476–479) are contained all receiving tubes having base designations corresponding to the standard RMA registry system. Transmitting tube diagrams are given on page 480. All diagrams are in uniform style. Bottom views are shown throughout. Terminal designations are as follows:

- A = Anode
- BP = Bayonet Pin
- D = Deflecting Plate
- F = Filament
- G = Grid
- H = Heater
- IC = Internal Connection
- IS = Internal Shield
- K = Cathode
- NC = No Connection
- P = Plate (Anode)
- P_1 = Starter-Anode
- P_{BF} = Beam-Forming Plates
- RC = Ray-Control Electrode
- S = Shell
- TA = Target
- ● = Gas-Type Tube
- U = Unit
- SH = Internal Shield

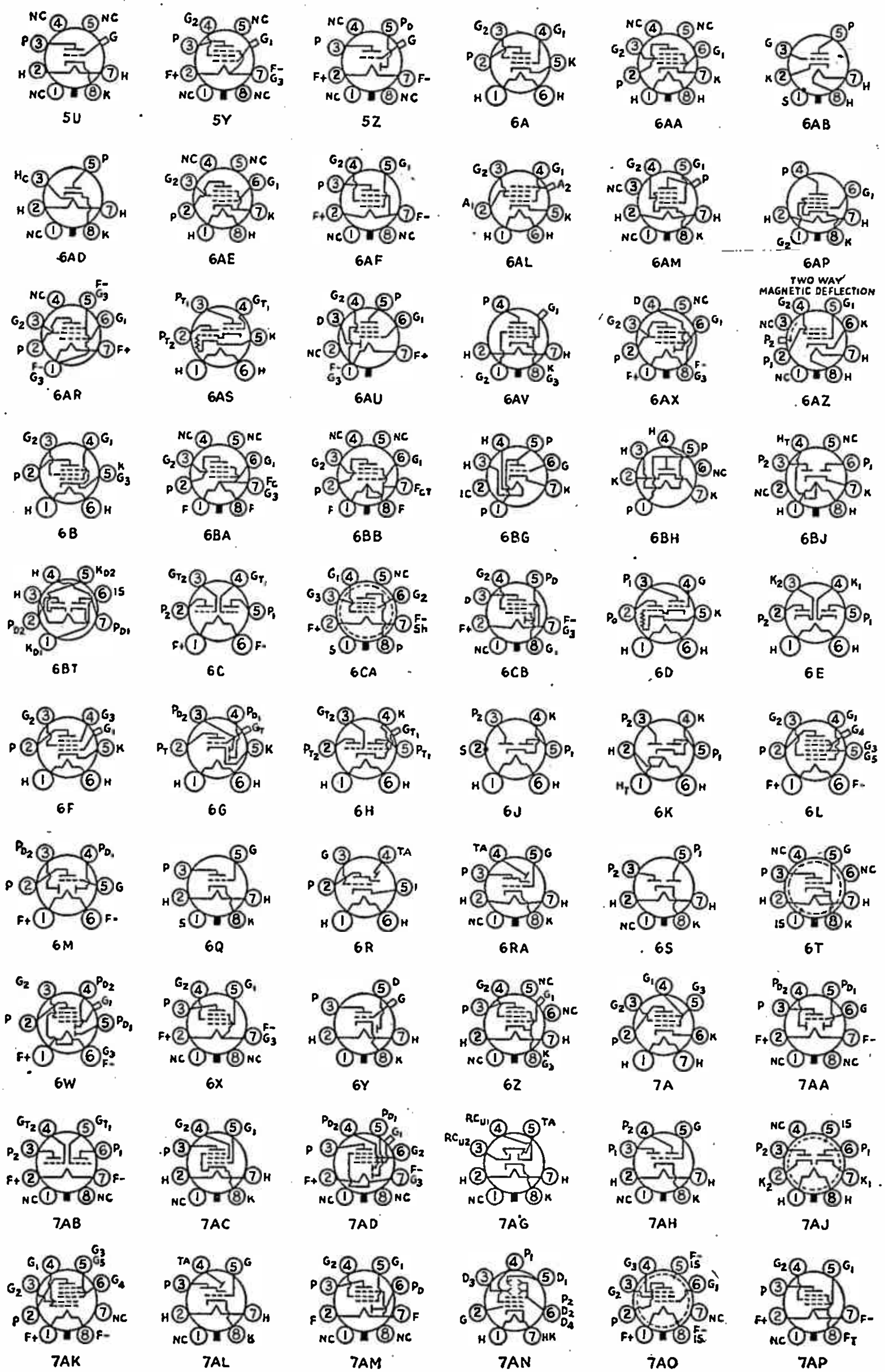
Alphabetical subscripts D, P, T and HX indicate, respectively, diode unit, pentode unit, triode unit or hexode unit in multi-unit types. Subscript M, T or CT indicates filament or heater tap.

Wherever the No. 1 pin of a metal-type tube in Table I is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.



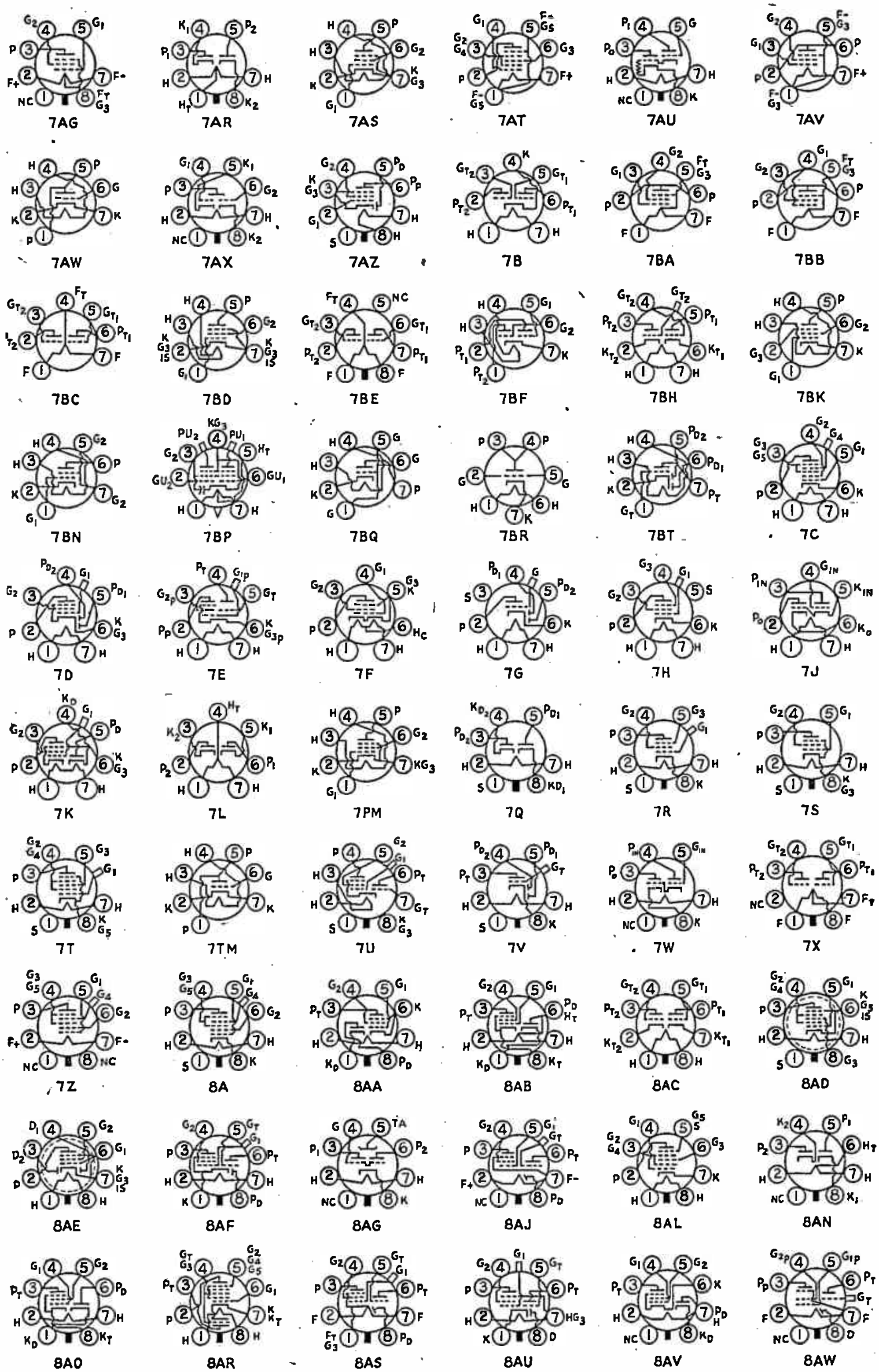
RECEIVING TUBE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 474.



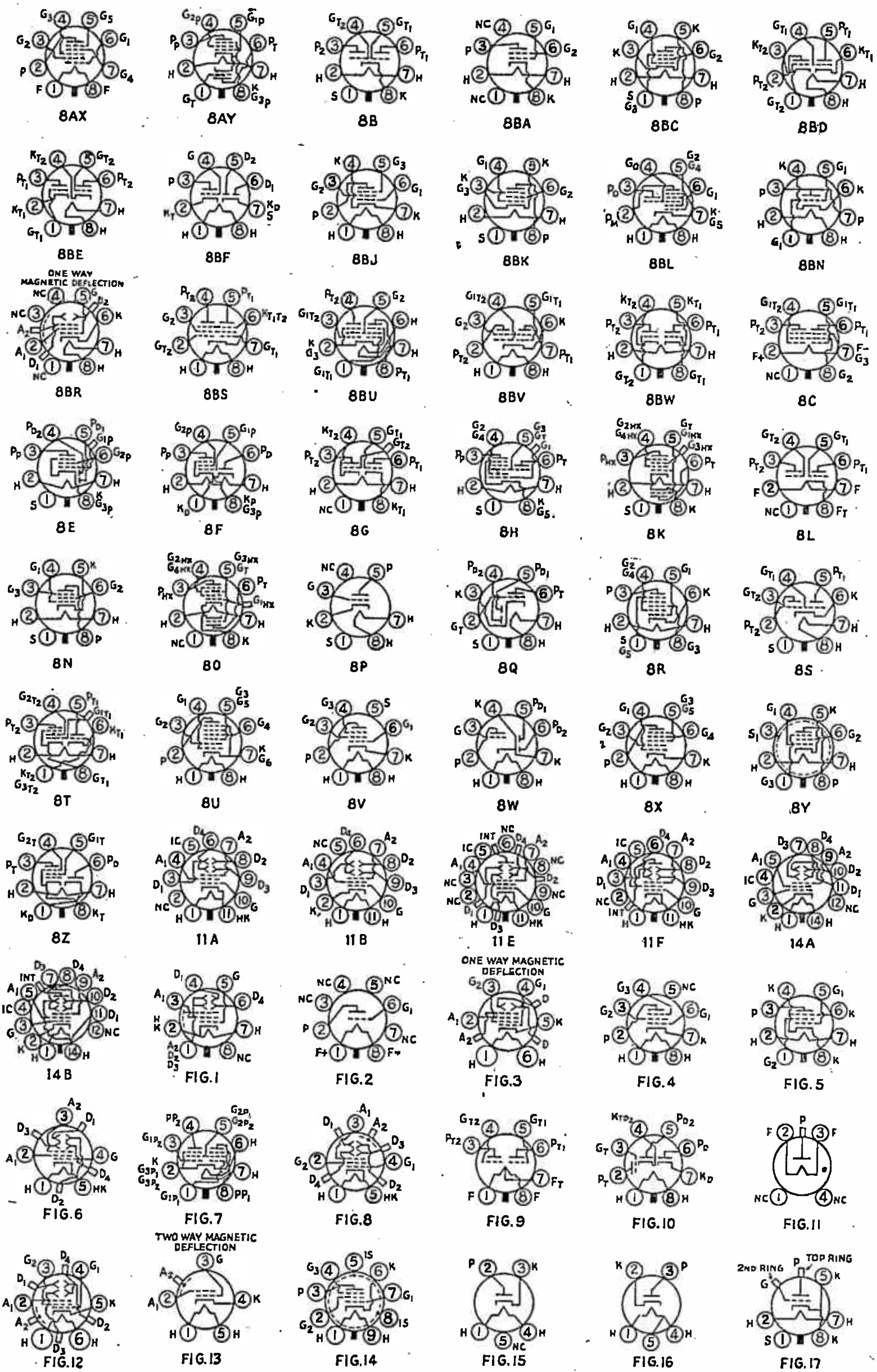
RECEIVING TUBE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 474.



RECEIVING TUBE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 474.



SUPPLEMENTARY BASE DIAGRAMS

(*Continued on page 479*)

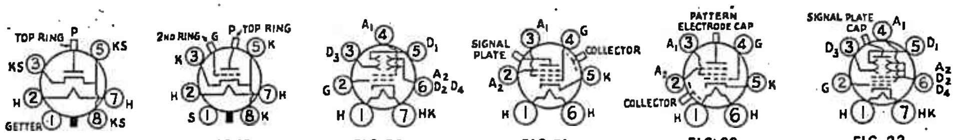
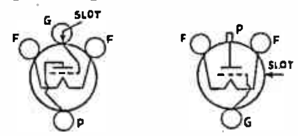


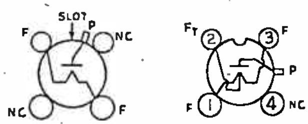
FIG. 18 FIG. 19 FIG. 20 FIG. 21 FIG. 22 FIG. 23

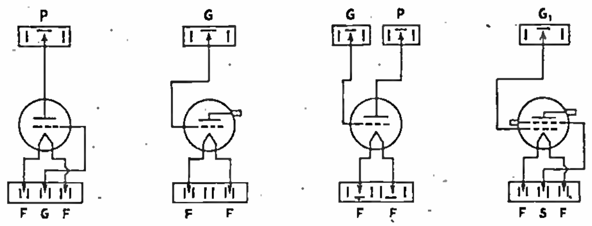
TRANSMITTING TUBE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 475.

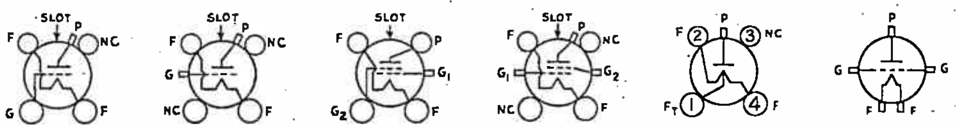


T-2A T-2AA

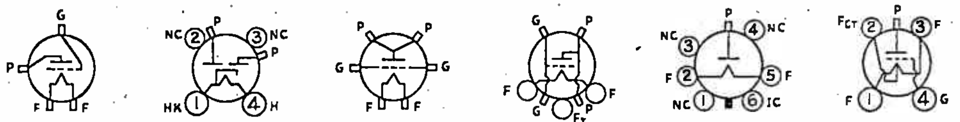




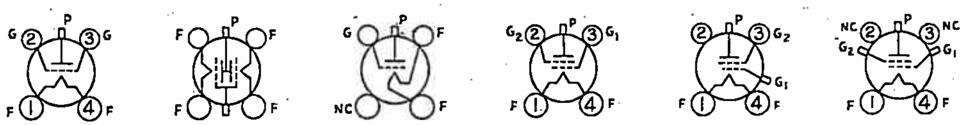
T-1A T-1AA T-1AB T-1B T-3A T-3AA



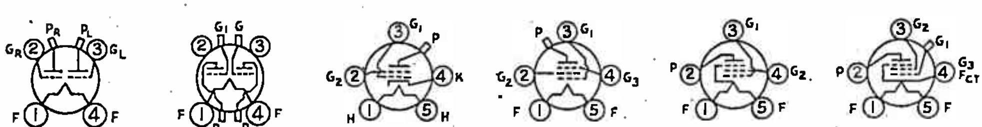
T-3AB T-3AC T-3B T-3BC T-4A T-4D



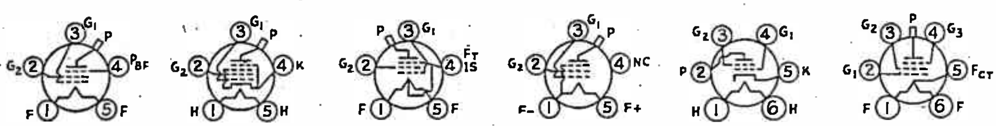
T-4AF T-4AG T-4B T-4BB T-4BC T-4BD



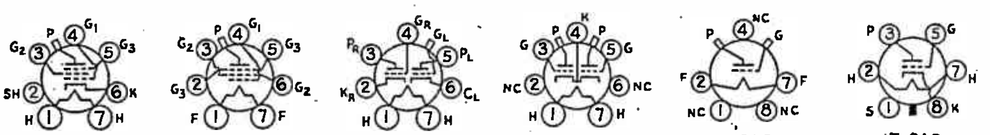
T-4BE T-4BF T-4BG T-4C T-4CB T-4CE



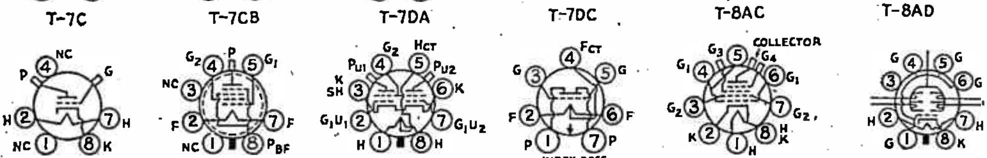
T-4D T-4DB T-5BB T-5C T-5CA T-5CB



T-5D T-5DA T-5DB T-5DC T-6B T-6C



T-7C T-7CB T-7DA T-7DC T-8AC T-8AD



T-8AG T-8DB T-8FA T-9A T-9C T-9D

TABLE I.—METAL RECEIVING TUBES

Characteristics given in this table apply to all tubes having type numbers shown, including metal tubes, glass tubes with "G" suffix, and bantam tubes with "GT" suffix. For "G" and "GT" tubes not listed (not having metal counterparts), see Tables II, VII, VIII and IX.

Type	Name	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6A8	Pentagrid Converter	8A	Htr.	6.3	0.3	Osc.-Mixer	250	−3.0	100	3.2	3.3	Anode-grid (No. 2) 250 volts max. thru 20,000-ohms					6A8
6AB7 1853	Television Amp. Pentode	8N	Htr.	6.3	0.45	Class-A Amplifier	300	−3.0	200 [2]	3.2	12.5	700000	5000	3500	—	—	6AB7 1853
6AC7 1852	Television Amp. Pentode	8N	Htr.	6.3	0.45	Class-A Amplifier	300	−2.0 [4]	150 [2]	2.5	10	750000	9000	6750	—	—	6AC7 1852
6AG7	Video Beam Power Amp.	8Y	Htr.	6.3	0.65	Class-A_1 Amplifier [5]	300	−3.0	150	7/9	30/30.5	130000	11000	—	10000	3.0	6AG7
6B8	Duplex-Diode Pentode	8E	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	125	2.3	9.0	650000	1125	730	—	—	6B8
6C5 [15]	Triode Detector, Amplifier	6Q	Htr.	6.3	0.3	Class-A Amplifier	250	−8.0	—	—	8.0	10000	2000	20	—	—	6C5
						Bias Detector	250	−17.0	—	—	Plate current adjusted to 0.2 ma. with no signal						
6F5	High-μ Triode	5M	Htr.	6.3	0.3	Class-A Amplifier	250	−1.3	—	—	0.2	66000	1500	100	—	—	6F5
6F6	Pentode Power Amplifier	7S	Htr.	6.3	0.7	Class-A Pentode	250 315	−16.5 −22.0	250 315	6.5 8.0	34 42	80000 75000	2500 2650	200 200	7000 7000	3.0 5.0	6F6
						Class-A Triode [3]	250	−20.0	—	—	31	2600	2700	7.0	4000	0.85	
						Push-Pull Class-AB Pentode Triode Connection [3]	375 350	−26.0 −38.0	250 —	2.5 —	17 22.5	Power output for 2 tubes at stated load, plate-to-plate			10000 6000	19.0 18.0	
6H6 [15]	Twin Diode	7Q	Htr.	6.3	0.3	Rectifier	Max. a.c. voltage per plate=100 r.m.s. Max. output current 4.0 ma. d.c.										6H6
6J5	Detector Amplifier Triode	6Q	Htr.	6.3	0.3	Class-A Amplifier	250	−8.0	—	—	9	7700	2600	20	—	—	6J5
6J7 [15]	Triple-Grid Detector, Amplifier	7R	Htr.	6.3	0.3	R.F. Amplifier	250	−3.0	100	0.5	2.0	1.5 meg.	1225	1500	—	—	6J7
						Bias Detector	250	−4.3	100	Cathode current 0.43 ma.			—	—	0.5 meg.	—	
6K7	Triple-Grid Variable-μ Amplifier	7R	Htr.	6.3	0.3	R.F. Amplifier	250	−3.0	125	2.6	10.5	600000	1650	990	—	—	6K7
						Mixer	250	−10.0	100	—	—	—	Oscillator peak volts=7.0				
6K8	Triode Hexode Converter	8K	Htr.	6.3	0.3	Osc.-Mixer	250	−3.0	100	6	2.5	Triode Plate (No. 2) 100 volts, 3.8 ma.					6K8
6L6	Beam Power Amplifier	7AC	Htr.	6.3	0.9	Single-Tube A_1 [6] Cathode Bias	250 300	— [7] — [8]	250 200	5.4–7.2 3.0–4.6	75–78 51–54.5	— —	— —	— —	2500 4500	6.5 6.5	6L6
						Single-Tube A_1 [6] Fixed Bias	250 350	−14.0 −18.0	250 250	5.0–7.3 2.5–7.0	72–79 54–66	22500 33000	6000 5200	— —	2500 4200	6.5 10.8	
						Push-Pull A_1 [5] Cathode Bias	270	— [9]	270	11–17	134–145	—	—	—	5000	18.5	
						Push-Pull A_1 [5] Fixed Bias	250 270	−16.0 −17.5	250 270	10–16 11–17	120–140 134–155	24500 23500	5500 5700	— —	5000 5000	14.5 17.5	
						Push-Pull AB_1 [6] Cathode Bias	360	— [10]	270	5–17	88–100	Power output for 2 tubes. Load plate-to-plate			9000	24.5	
						Push-Pull AB_1 [5] Fixed Bias	360	−22.5	270	5–15	88–132				6600 [11]	26.5	
						Push-Pull AB_2 [5] Fixed Bias	360 360	−18.0 −22.5	225 270	3.5–11 5–16	78–142 88–205				6000 3800	31.0 47.0	
6L7	Pentagrid Mixer Amplifier	7T	Htr.	6.3	0.3	R.F. Amplifier	250	−3.0	100	5.5	5.3	800000	1100	—	—	—	6L7
						Mixer	250	−6.0	150	8.3	3.3	Over 1 meg.	Oscillator-grid (No. 3) voltage=−15.0				
6N7	Twin Triode	8B	Htr.	6.3	0.8	Class-B Amplifier	300	0	—	—	35–70	—	—	—	8000	10.0	6N7
6Q7	Duplex-Diode Triode	7V	Htr.	6.3	0.3	Triode Amplifier	250	−3.0	—	—	1.1	58000	1200	70	—	—	6Q7
6R7	Duplex-Diode Triode	7V	Htr.	6.3	0.3	Triode Amplifier	250	−9.0	—	—	9.5	8500	1900	16	10000	0.28	6R7
6S7 [15]	Triple-Grid Variable-μ	7R	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	2.0	8.5	1000000	1750	1750	—	—	6S7
6SA7	Pentagrid Converter	8R [12]	Htr.	6.3	0.3	Osc.-Mixer	250	0 [13]	100	8.0	3.4	800000	Grid No. 1 Resistor 20000 ohms				6SA7
6SC7	Twin Triode Amplifier	8S	Htr.	6.3	0.3	Class-A Amplifier	250	−2.0	—	—	2.0	53000	1325	70	—	—	6SC7
6SF5	High-μ Triode	6AB	Htr.	6.3	0.3	Class-A Amplifier	250	−2.0	—	—	0.9	66000	1500	100	—	—	6SF5
6SF7	Diode Variable-μ Pentode	7AZ	Htr.	6.3	0.3	Class-A Amplifier	250	−1.0	100	3.3	12.4	700000	2050	—	—	—	6SF7
6SG7	Triple-Grid Semi-Variable-μ	8BK	Htr.	6.3	0.3	H. F. Amplifier	250	−2.5	150	3.4	9.2	Over 1 meg.	4000	—	—	—	6SG7
6SH7	Triple-Grid Amplifier	8BK	Htr.	6.3	0.3	H. F. Amplifier	250	−1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7

TABLE I—METAL RECEIVING TUBES—*Continued*

Type	Name	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6SJ7[16]	Triple-Grid Amplifier	8N	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	100	0.8	3	1500000	1650	2500	—	—	6SJ7
6SK7	Triple-Grid Variable-μ	8N	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	100	2.4	9.2	800000	2000	1600	—	—	6SK7
6SQ7	Duplex-Diode Triode	8Q	Htr.	6.3	0.3	Class-A Amplifier	250	−2.0	—	—	0.8	91000	1100	100	—	—	6SQ7
6SR7	Duplex-Diode Triode	8Q	Htr.	6.3	0.3	Class-A Amplifier	250	−9.0	—	—	9.5	8500	1900	16	—	—	6SR7
6SS7	Triple-Grid Variable-μ	8N	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	2.0	9.0	1000000	1850	—	—	—	6SS7
6ST7	Duplex-Diode Triode	8Q	Htr.	6.3	0.15	Class-A Amplifier	250	−9.0	—	—	9.5	8500	1900	16	—	—	6ST7
6T7	Duplex-Diode Triode	7V	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	—	—	1.2	62000	1050	65	—	—	6T7
6V6	Beam Power Amplifier	7AC	Htr.	6.3	0.45	Class-A Amplifier	250	−12.5	250	4.5/7.0	45–47	52000	4100	218	5000	4.5	6V6
						Class-AB Amplifier 2 Tubes	250	−15.0	250	5/13	70–79	60000	3750	—	10000	10.0	
							285	−19.0	285	4/13.5	70–92	65000	3600	—	8000	14.0	
1611	Pentode Power Amplifier	7S	Htr.	6.3	0.7	Relay Tube	Characteristics same as 6F6										1611
1612	Pentagrid Amplifier	7T	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	100	6.5	5.3	600000	1100	880	—	—	1612
1620	Triple-Grid Det.-Amp.	7R	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as 6J7										1620
1621	Power Amplifier Pentode	7S	Htr.	6.3	0.7	Class-A, Pentode P. P.	300	−30.0	300	6.5/13	38/69	—	—	—	4000	5.0	1621
						Class-A Triode[3] P. P.	327.5	−27.5[14]	—	—	55/59	—	—	—	5000	2.0	
1622	Beam Power Amplifier	7AC	Htr.	6.3	0.9	Class-A Amplifier	300	−20.0	250	4/10.5	86/125	—	—	—	4000	10.0	1622
1851	Television Amp. Pentode	7R	Htr.	6.3	0.45	Class-A Amplifier	300	−2.0[4]	150[2]	2.5	10	750000	9000	6750	—	—	1851

[1] See Receiving Tube Diagrams.
[2] From fixed screen supply. If series resistor from plate supply is used, value for 6AB7/1853 is 30,000 ohms, for 6AC7/1852 and 1851 60,000 ohms. Series resistor gives variable-μ characteristic, fixed screen supply gives sharp cut-off.
[3] Screen tied to plate.
[4] Cathode bias resistor should be adjusted for plate current of 10 ma.; minimum value 160 ohms.
[5] Typical operation for 4-Mc.-bandwidth video voltage amplifier; 70 volts output with 4 volts input.
[6] Subscript 1 indicates no grid-current flow. Subscript 2 indicates grid-current flow over part of input cycle.
[7] Cathode resistor 170 ohms. [8] Cathode resistor 220 ohms.
[9] Cathode resistor 125 ohms. [10] Cathode resistor 250 ohms.
[11] Output 18 watts with 3800-ohm load.
[12] For 6SA7GT, use Base Diagram 8AD.
[13] Grid bias −2 volts if separate oscillator excitation is used.
[14] Cathode resistor 500 ohms.
[15] Types G or GT have internal shield connected to number one pin. [16] Also type "6SJ7Y".

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES

(For "G" and "GT" -Type Tubes Not Listed Here, See Equivalent Type in Table I; Characteristics and Connections Will Be Identical)

Type	Name	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2C22	Triode Amplifier	4AM	Htr.	6.3	0.3	Class-A Amplifier	300	−10.5	—	—	11	6600	3000	20	—	—	2C22
6A5G	Triode Power Amplifier	6T	Htr.	6.3	1.0	Class-A Amplifier	250	−45.0	—	—	60	800	5250	4.2	2500	3.75	6A5G
						Push-Pull Class AB	325	−68.0	—	—	80[2]	—		—	3000	15.0	
						Push-Pull Class AB	325	850-ohm cathode resistor			80[2]	—		—	5000	10.0	
6AB6G	Direct-Coupled Amplifier	7W	Htr.	6.3	0.5	Class-A Amplifier	250	0	Input		5.0	40000	1800	72	8000	3.5	6AB6G
							250	0	Output		34						
6AC5G	High-μ Power Amplifier Triode	6Q	Htr.	6.3	0.4	Push-Pull Class-B	250	0	—	—	5.0[2]	36700	3400	125	10000	8.0	6AC5G
						Dynamic-Coupled Amp.	250	—	—	—	32				7000	3.7	
6AC6G	Direct-Coupled Amplifier	7W	Htr.	6.3	1.1	Class-A Amplifier	180	0	Input		7.0	—	3000	54	4000	3.8	6AC6G
							180	0	Output		45						
6AD5G	High-μ Triode	6Q	Htr.	6.3	0.3	Class-A Amplifier	250	−2.0	—	—	0.9	—	1500	100	—	—	6AD5G
6AD6G	Electron-Ray Tube	7AG	Htr.	6.3	0.15	Indicator Tube	100	0 for 90°; −23 for 135°, 45 for 0°. Target current 1.5 ma.									6AD6G
6AD7G	Triode-Pentode	8AY	Htr.	6.3	0.85	Triode Amplifier	250	−25.0	—	—	4.0	19000	325	6.0	—	—	6AD7G
						Pentode Amplifier	250	−16.5	250	6.5	34	80000	2500	—	7000	3.2	
6AE5G	Triode Amplifier	6Q	Htr.	6.3	0.3	Class-A Amplifier	95	−15.0	—	—	7.0	3500	1200	4.2	—	—	6AE5G
6AE6G	Twin-Plate Triode	7AH	Htr.	6.3	0.15	Indicator Control	250	−1.5	—	—	6.5[4]	25000	1000	25	—	—	6AE6G
							250	−1.5			4.5[5]	35000	950	33			

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6AE7GT	Twin-Input Triode[7]	7AX	Htr.	6.3	0.5	Driver Amplifier	250	−13.5	—	—	5.0	9300	1500	14	—	—	6AE7GT
6AF5G	Triode Amplifier	6Q	Htr.	6.3	0.3	Class-A Amplifier	180	−18.0	—	—	7.0	—	1500	7.4	—	—	6AF5G
6AF7G	Twin Electron Ray	8AG	Htr.	6.3	0.3	Indicator Tube											6AF7G
6AG6G	Power Amplifier Pentode	7S	Htr.	6.3	1.25	Class-A Amplifier	250	−6.0	250	6.0	32	—	10000	—	8500	3.75	6AG6G
6AH5G	Beam Power Amplifier	6AP	Htr.	6.3	0.9	Class-A Amplifier	350	−18	250	—	—	33000	5200	—	4200	10.8	6AH5G
6AH7GT	Twin Triode	8BE	Htr.	6.3	0.3	Converter and Amp.	250	−9.0	—	—	12[3]	6600	2400	16	—	—	6AH7GT
6AL6G	Beam Power Amplifier	6AM	Htr.	6.3	0.9	Class-A Amplifier	250	−14.0	250	5.0	72	22500	6000	—	2500	6.5	6AL6G
6B4G	Triode Power Amplifier	5S	Fil.	6.3	1.0	Power Amplifier	Characteristics same as Type 6A3—Table IV							—	—	—	6B4G
6B6G	Duplex-Diode High-μ Triode	7V	Htr.	6.3	0.3	Detector-Amplifier	Characteristics same as Type 75—Table IV							—	—	—	6B6G
6C8G	Twin Triode	8G	Htr.	6.3	0.3	Amp. 1 Section	250	−4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6D8G	Pentagrid Converter	8A	Htr.	6.3	0.15	Converter	250	−3.0	100	Cathode current 13.0 Ma.			Anode grid (No. 2) Volts = 250[9]				6D8G
6E8G	Triode-Hexode Converter	8O	Htr.	6.3	0.3	Osc.-Mixer	250	−2.0	Triode Plate 150 volts								6E8G
6F8G	Twin Triode	8G	Htr.	6.3	0.6	Amplifier	250	−8.0	—	—	9[3]	7700	2600	20	—	—	6F8G
6G6G	Pentode Power Amplifier	7S	Htr.	6.3	0.15	Class-A Amplifier	180	−9.0	180	2.5	15	175000	2300	400	10000	1.1	6G6G
						Class-A Amplifier[11]	180	−12.0	—	—	—	4750	2000	9.5	12000	0.25	
6H4GT	Diode Rectifier	5AF	Htr.	6.3	0.15	Detector	100	—	—	—	4.0	—	—	—	—	—	6H4GT
6H8G	Duo-Diode High-μ Pentode	8E	Htr.	6.3	0.3	Class-A Amplifier	250	−2.0	100	—	8.5	650000	2400	—	—	—	6H8G
6J8G	Triode Heptode	8H	Htr.	6.3	0.3	Converter	250	−3.0	100	2.8	1.2	Anode-grid (No. 2) 250 volts max.[9] 5 ma.					6J8G
6K5G	High-μ Triode	5U	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	—	—	1.1	50000	1400	70	—	—	6K5G
6K6G	Pentode Power Amplifier	7S	Htr.	6.3	0.4	Class-A Amplifier	Characteristics same as Type 41—Table IV										6K6G
6L5G	Triode Amplifier	6Q	Htr.	6.3	0.15	Class-A Amplifier	250	−9.0	—	—	8.0	—	1900	17	—	—	6L5G
6M6G	Power Amplifier Pentode	7S	Htr.	6.3	1.2	Class-A Amplifier	250	−6.0	250	4.0	36	—	9500	—	7000	4.4	6M6G
6M7G	Triple-Grid Amplifier	7R	Htr.	6.3	0.3	R. F. Amplifier	250	−2.5	125	2.8	10.5	900000	3400	—	—	—	6M7G
6M8GT	Diode Triode Pentode	8AU	Htr.	6.3	0.6	Triode Amplifier	100	—	—	—	0.5	91000	1100	—	—	—	6M8GT
						Pentode Amplifier	100	−3.0	100	—	8.5	200000	1900	—	—	—	
6N6G[10]	Direct-Coupled Amplifier	7AU	Htr.	6.3	0.8	Power Amplifier	Characteristics same as Type 6B5—Table IV							—	—	—	6N6G
6P5G	Triode Amplifier	6Q	Htr.	6.3	0.3	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	6P5G
6P7G	Triode-Pentode	7U	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as 6F7—Table IV										6P7G
6P8G	Triode-Hexode Converter	8K	Htr.	6.3	0.8	Osc.-Mixer	250	−2.0	75	1.4	1.5	Triode Plate 100 v. 2.2 ma.					6P8G
6Q6G	Diode-Triode	6Y	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	—	—	1.2	—	1050	65	—	—	6Q6G
6R6G	Pentode Amplifier	6AA	Htr.	6.3	0.3	Class-A Amplifier	250	−3.0	100	1.7	7.0	—	1450	1160	—	—	6R6G
6S6GT	Triple-Grid Variable-μ	5AK	Htr.	6.3	0.45	R.F. Amplifier	250	−2.0	100	3.0	13	350000	4000	—	—	—	6S6GT
6SD7GT	Triple-Grid Semi-Variable-μ	8N	Htr.	6.3	0.3	R.F. Amplifier	250	−2.0	100	1.9	6.0	1000000	3600	—	—	—	6SD7GT
6SE7GT	Triple-Grid Amplifier	8N	Htr.	6.3	0.3	R.F. Amplifier	250	−1.5	100	1.5	4.5	1100000	3400	3750	—	—	6SE7GT
6SL7GT	Twin Triode	8BD	Htr.	6.3	0.3	Amplifier	250	−2.0	—	—	2.3[3]	44000	1600	70	—	—	6SL7GT
6SN7GT	Twin Triode	8BD	Htr.	6.3	0.3	Amplifier	250	−8.0	—	—	9.0[3]	7700	2600	20	—	—	6SN7GT
6T6GM[8]	Triple-Grid Amplifier	6Z	Htr.	6.3	0.45	R.F. Amplifier	250	−1.0	100	2.0	10	1000000	5500	—	—	—	6T6GM
6U6GT	Beam Power Amplifier	7AC	Htr.	6.3	0.75	Class-A Amplifier	200	−14.0	135	3.0	56	20000	6200	—	3000	5.5	6U6GT
6U7G	Triple Grid Variable-μ	7R	Htr.	6.3	0.3	R.F. Amplifier	Characteristics same as Type 6D6—Table III										6U7G
6V7G	Duplex Diode-Triode	7V	Htr.	6.3	0.3	Detector-Amplifier	Characteristics same as Type 85—Table III										6V7G
6W6GT	Beam Power Amplifier	7AC	Htr.	6.3	1.25	Class-A Amplifier	135	−9.5	135	12.0	61.0	—	9000	215	2000	3.3	6W6GT
6W7G	Triple-Grid Det. Amp.	7R	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	2.0	0.5	1500000	1225	1850	—	—	6W7G
6X6G	Electron-Ray Tube	7AL	Htr.	6.3	0.3	Indicator Tube	250	0 v. for 300°, 2 ma. −8 v. for 0°, 0 ma. Vane grid 135 v.									6X6G
6Y6G	Beam Power Amplifier	7AC	Htr.	6.3	1.25	Class-A Amplifier	135	−13.5	135	3.0	60.0	9300	7000	—	2000	3.6	6Y6G
6Y7G	Twin Triode Amplifier	8B	Htr.	6.3	0.3	Class-B Amplifier	Characteristics same as Type 79—Table IV										6Y7G

TABLE II — 6.3-VOLT GLASS TUBES WITH OCTAL BASES — *Continued*

Type	Name	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6Z7G	Twin Triode Amplifier	8B	Htr.	6.3	0.3	Class-B Amplifier	180	0	—	—	8.4 [2]	—	—	—	12000	4.2	6Z7G
							135	0	—	—	6.0 [2]	—	—	—	9000	2.5	
717A	Pentode Amplifier	8BK [12]	Htr.	6.3	0.175	Class-A Amplifier	120	– 2.0	120	2.5	7.5	390000	4000	—	—	—	717A
1223	Pentode Amplifier	7R	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as 6C6 — Table IV										1223
1231	Pentode Amplifier	8V	Htr.	6.3	0.45	Class-A Amplifier	300	– 2.5 [6]	150	2.5	10	700000	5500	3850	—	—	1231
1635	Twin Triode Amplifier	8B	Htr.	6.3	0.6	Class-B Amplifier	400	0	—	—	10 [2]/63	—	—	—	14000	17	1635
2C21/1642	Twin-Triode Amplifier	7BH	Htr.	6.3	0.6	Class-A Amplifier	250	–16.5	—	—	8.3	7600	1375	10.4	—	—	2C21/1642
7000	Low-Noise Amplifier	7R	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as Type 6J7 — Table I										7000

[1] Refer to Receiving Tube Diagrams. No connection to Pin No. 1.
[2] No-signal value for 2 tubes.
[3] Per plate.
[4] Plate No. 1, remote cut-off.
[5] Plate No. 2, sharp cut-off.
[6] Through 200-ohm cathode resistor.
[7] Common plate.
[8] Metal-sprayed glass envelope.
[9] Through 20,000-ohm dropping resistor.
[10] Also type MG.
[11] Screen tied to plate.
[12] Low-loss phenolic base.

TABLE III — 7-VOLT LOKTAL-BASE TUBES

For other loktal-base types see Tables VIII, IX, X and XIII.

Type	Name	Socket Connections	Cathode	Heater* Volts	Heater* Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen [1] Current Ma.	Plate [1] Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7A4	Triode Amplifier	5 AC	Htr.	7.0	0.32	Class-A Amplifier	250	– 8.0	—	—	9.0	7700	2600	20	—	—	7A4
7A5	Beam Power Amplifier	6 AT	Htr.	7.0	0.75	Class-A₁ Amplifier	125	– 9.0	125	3.2/8	37.5/40	17000	6100	—	2700	1.9	7A5
7A6	Twin Diode	7 AJ	Htr.	7.0	0.16	Rectifier	Max. A.C. volts per plate — 150. Max. Output current — 10 ma.										7A6
7A7	Remote Cut-off Pentode	8 V	Htr.	7.0	0.32	R.F. Amplifier	250	– 3.0	100	2.0	8.6	800000	2000	1600	—	—	7A7
7A8	Multigrid Converter	8 U	Htr.	7.0	0.16	Osc.-Mixer	250	– 3.0	100	3.1	3.0	50000	Anode-grid 250 volts max.[2]				7A8
7B4	High-μ Triode	5 AC	Htr.	7.0	0.32	Class-A Amplifier	250	– 2.0	—	—	0.9	66000	1500	100	—	—	7B4
7B5	Pentode Power Amplifier	6 AE	Htr.	7.0	0.43	Class-A₁ Amplifier	250	–18.0	250	5.5/10	32/33	68000	2300	—	7600	3.4	7B5
7B6	Duo-Diode Triode	8 W	Htr.	7.0	0.32	Class-A Amplifier	250	– 2.0	—	—	1.0	91000	1100	100	—	—	7B6
7B7	Remote Cut-off Pentode	8 V	Htr.	7.0	0.16	R.F. Amplifier	250	– 3.0	100	2.0	8.5	700000	1700	1200	—	—	7B7
7B8	Pentagrid Converter	8 X	Htr.	7.0	0.32	Osc.-Mixer	250	– 3.0	100	2.7	3.5	360000	Anode-grid 250 volts max.[2]				7B8
7C5	Tetrode Power Amplifier	6 AA	Htr.	7.0	0.48	Class-A₁ Amplifier	250	–12.5	250	4.5/7	45/47	52000	4100	—	5000	4.5	7C5
7C6	Duo-Diode Triode	8 W	Htr.	7.0	0.16	Class-A Amplifier	250	– 1.0	—	—	1.3	100000	1000	100	—	—	7C6
7C7	Pentode Amplifier	8 V	Htr.	7.0	0.16	R.F. Amplifier	250	– 3.0	100	0.5	2.0	2 meg.	1300	—	—	—	7C7
7D7	Triode-Hexode Converter	8 AR	Htr.	7.0	0.48	Osc.-Mixer	250	– 3.0	Triode Plate (No. 3) 150 v. 3.5 ma.								7D7
7E6	Duo-Diode Triode	8 W	Htr.	7.0	0.32	Class-A Amplifier	250	– 9.0	—	—	9.5	8500	1900	16	—	—	7E6
7E7	Duo-Diode Pentode	8W	Htr.	7.0	0.32	Class-A Amplifier	250	– 3.0	100	1.6	7.5	700000	1300	—	—	—	7E7
7F7	Twin Triode	8AC	Htr.	7.0	0.32	Class-A Amplifier [3]	250	– 2.0	—	—	2.3	44000	1600	70	—	—	7F7
7F8	Twin Triode	8BW	Htr.	6.3	0.30	R.F. Amplifier	250	– 2.5	—	—	10.0	10400	5000	—	—	—	7F8
							180	– 1.0			12.0	8500	7000				
7G7/1232	Triple-Grid Amplifier	8 V	Htr.	7.0	0.48	Class-A Amplifier	250	– 2.0	100	2.0	6.0	800000	4500	—	—	—	7G7/1232
7G8/1206	Dual Pentode	8BV	Htr.	6.3	0.30	R.F. Amplifier [3]	250	– 2.5	100	0.8	4.5	225000	2100	—	—	—	7G8/1206
7H7	Triple-Grid Semi-Variable-μ	8 V	Htr.	7.0	0.32	R.F. Amplifier	250	– 2.5	150	2.5	9.0	1000000	3500	—	—	—	7H7
7J7	Triode-Hexode Converter	8 AR	Htr.	7.0	0.32	Osc.-Mixer	250	– 3.0	100	2.9	1.3	Triode Plate 250 v. Max.[2]					7J7
7K7	Duo-Diode High-μ Triode	8BF	Htr.	7.0	0.32	Class-A Amplifier	250	– 2.0	—	—	2.3	44000	1600	70	—	—	7K7
7L7	Triple-Grid Amplifier	8 V	Htr.	7.0	0.32	Class-A Amplifier	250	– 1.5	100	1.5	4.5	100000	3100	Cathode Resistor 250 ohms			7L7

TABLE III — 7-VOLT LOKTAL-BASE TUBES — *Continued*

Type	Name	Socket Connections	Cathode	Heater Volts	Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen[1] Current Ma.	Plate[1] Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7N7	Twin Triode	8AC	Htr.	7.0	0.6	Class-A Amplifier[3]	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7N7
7Q7	Pentagrid Converter	8AL	Htr.	7.0	0.32	Osc.-Mixer	250	0	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				7Q7
7R7	Duo-Diode Pentode	8AE	Htr.	7.0	0.32	Class-A Amplifier	250	− 1.0	100	1.7	5.7	1000000	3200	—	—	—	7R7
7S7	Triode Hexode Converter	8AR	Htr.	7.0	0.32	Osc.-Mixer	250	− 2.0	100	2.2	1.7	2000000	Triode Plate 250 v. Max.[2]				7S7
7T7	Triple-Grid Amplifier	8V	Htr.	7.0	0.32	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	7T7
7V7	Triple-Grid Amplifier	8V	Htr.	7.0	0.48	Amplifier	300	—[4]	150[5]	3.9	9.6	300000	5800	—	—	—	7V7
7W7	Triple-Grid Variable-μ	8BJ	Htr.	7.0	0.48	Class-A Amplifier	300	− 2.2	150	3.9	10	300000	5800	—	—	—	7W7
XXL	Triode Oscillator	5AC	Htr.	7.0 *	0.32	Oscillator	250	− 8.0	—	—	8.0	—	2300	20	—	—	XXL

* Maximum rating, corresponding to 130-volt line condition; normal rating is 6.3 v. for 117-v. line.
[1] Values to left of diagonal lines are for "no-signal" condition; values to right are "with signal." [2] Applied through 20000-ohm dropping resistor. [3] Each triode. [4] Cathode bias resistor, 160 ohms.
[5] From fixed screen supply. If series resistor from plate supply is used, value should be 40,000 ohms. Series resistor gives extended cut-off (variable-μ) characteristic, fixed screen supply gives sharp cut-off.

TABLE IV — 6.3-VOLT GLASS RECEIVING TUBES

Type	Name	Base[4]	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6A3	Triode Power Amplifier	4-pin M.	4D	Fil.	6.3	1.0	Class-A Amplifier Push-Pull Amplifier	250 300 300	− 45 − 62 —	— Fixed Bias Self Bias [E]	—	60 40 40	800 Power output for 2 tubes load plate-to-plate	5250	4.2	2500 3000 5000	3.5 15 10	6A3
6A4[3]	Pentode Power Amplifier	5-pin M.	5B	Fil.	6.3	0.3	Class-A Amplifier	180	−12.0	180	3.9	22	45500	2200	100	8000	1.4	6A4
6A6	Twin Triode Amplifier	7-pin M.	7B	Htr.	6.3	0.8	Class-B Amplifier	250 300	0 0	—	—	Power output is for one tube at stated load, plate-to-plate				8000 10000	8.0 10.0	6A6
6A7[11]	Pentagrid Converter	7-pin S.	7C	Htr.	6.3	0.3	Converter	250	− 3.0	100	2.2	3.5	360000	Anode grid (No. 2) 200 volts max.				6A7
6AB5/6N5	Electron-Ray Tube	6-pin S.	6R	Htr.	6.3	0.15	Indicator Tube	180	Cut-off Grid Bias = −12 v.			0.5	Target Current 2 ma.				—	6AB5/6N5
6AF6G	Electron-Ray Tube Twin Indicator Type	7-pin S.	7AG	Htr.	6.3	0.15	Indicator Tube	135 100	Ray Control Voltage = 81 for 0° Shadow Angle. Target current 1.5 ma. " " " = 60 " " " " " 0.9 "									6AF6G
6B5	Direct-Coupled Power Amplifier	6-pin M.	6AS	Htr.	6.3	0.8	Class-A Amplifier Push-Pull Amplifier	300 400	0 −13.0	— —	6[5] 4.5[5]	45 40	241000 —	2400 —	58 —	7000 10000	4.0 20	6B5
6B7[11]	Duplex-Diode Pentode	7-pin S.	7D	Htr.	6.3	0.3	Pentode R.F. Amplifier	250	− 3.0	125	2.3	9.0	650000	1125	730	—	—	6B7
6C6	Triple-Grid Amplifier	6-pin S.	6F	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	6C6
6C7	Duplex Diode Triode	7-pin S.	7G	Htr.	6.3	0.3	Class-A Amplifier	250	− 9.0	—	—	4.5	—	20	1250	—	—	6C7
6D6	Triple-Grid Variable-μ	6-pin S.	6F	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	6D6
6D7	Triple-Grid Amplifier	7-pin S.	7H	Htr.	6.3	0.3	Class-A Amplifier	250	− 3.0	100	0.5	2.0	—	1600	1280	—	—	6D7
6E5	Electron-Ray Tube	6-pin S.	6R	Htr.	6.3	0.3	Indicator Tube	250	0	—	—	0.25	Target Current 4 ma.				—	6E5
6E6	Twin Triode Amplifier	7-pin M.	7B	Htr.	6.3	0.6	Class-A Amplifier	250	−27.5	Per plate — 18.0			3500	1700	6.0	14000	1.6	6E6
6E7	Triple-Grid Variable-μ	7-pin S.	7H	Htr.	6.3	0.3	R.F. Amplifier	Characteristics same as 6U7G — Table II										6E7
6F7[11]	Triode Pentode	7-pin S.	7E	Htr.	6.3	0.3	Triode Unit Amplifier Pentode Unit Amplifier	100 250	− 3.0 − 3.0	— 100	— 1.5	3.5 6.5	16000 850000	500 1100	8 900	— —	— —	6F7
6U5/ 6G5	Electron-Ray Tube	6-pin S.	6R	Htr.	6.3	0.3	Indicator Tube	250 100	Cut-off Grid Bias = −22 v. " " " = − 8 v.			0.24 0.19	Target Current 4 ma. " " 1 "				—	6U5/ 6G5
6H5	Electron-Ray Tube	6-pin S.	6R	Htr.	6.3	0.3	Indicator Tube	Same characteristics as Type 6G5 — Circular Pattern										6H5
6T5	Electron-Ray Tube	6-pin S.	6R	Htr.	6.3	0.3	Indicator Tube	250	Cut-off Grid Bias = −12 v.			0.24	Target Current 4 ma.				—	6T5
36	Tetrode R.F. Amplifier	5-pin S.	5E	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	90	1.7	3.2	550000	1080	595	—	—	36
37	Triode Detector Amplifier	5-pin S.	5A	Htr.	6.3	0.3	Class-A Amplifier	250	−18.0	—	—	7.5	8400	1100	9.2	—	—	37
38	Pentode Power Amplifier	5-pin S.	5F	Htr.	6.3	0.3	Class-A Amplifier	250	−25.0	250	3.8	22.0	100000	1200	120	10000	2.5	38

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base[4]	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
39/44	Variable-μ R.F. Amplifier	5-pin S.	5F	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	90	1.4	5.8	1000000	1050	1050	—	—	39/44
41	Pentode Power Amplifier	6-pin S.	6B	Htr.	6.3	0.4	Class-A Amplifier	250	−18.0	250	5.5	32.0	68000	2200	150	7600	3.4	41
42	Pentode Power Amplifier	6-pin M.	6B	Htr.	6.3	0.7	Class-A Amplifier	250	−16.5	250	6.5	34.0	100000	2200	220	7000	3.0	42
56AS[11]	Triode Amplifier	5-pin S.	5A	Htr.	6.3	0.4	Class-A Amplifier	Characteristics same as 56										56AS
57AS[11]	Pentode	6-pin S.	6F	Htr.	6.3	0.4	R.F. Amplifier	Characteristics same as 57										57AS
58AS[11]	Triple-Grid Variable-μ	6-pin S.	6F	Htr.	6.3	0.4	R.F. Amplifier	Characteristics same as 58										58AS
75	Duplex-Diode Triode	6-pin S.	6G	Htr.	6.3	0.3	Triode Amplifier	250	− 1.35	—	—	0.4	91000	1100	100	—	—	75
76	Triode Detector Amplifier	5-pin S.	5A	Htr.	6.3	0.3	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	76
77	Triple-Grid Detector	6-pin S.	6F	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	100	0.5	2.3	1500000	1250	1500	—	—	77
78	Triple-Grid Variable-μ	6-pin S.	6F	Htr.	6.3	0.3	R.F. Amplifier	250	− 3.0	100	1.7	7.0	800000	1450	1160	—	—	78
79	Twin Triode Amplifier	6-pin S.	6H	Htr.	6.3	0.6	Class-B Amplifier	250	0	—	—	Power output is for one tube				14000	8.0	79
85	Duplex-Diode Triode	6-pin S.	6G	Htr.	6.3	0.3	Class-A Amplifier	250	−20.0	—	—	8.0	7500	1100	8.3	20000	0.35	85
85AS[11]	Duplex-Diode Triode	6-pin S.	6G	Htr.	6.3	0.3	Class-A Amplifier	250	− 9.0	—	—	5.5	—	1250	20	—	—	85 AS
89	Triple-Grid Power Amplifier	6-pin S.	6F	Htr.	6.3	0.4	Triode Amplifier[6]	250	−31.0	—	—	32.0	2600	1800	4.7	5500	0.9	89
							Pentode Amplifier[7]	250	−25.0	250	5.5	32.0	70000	1800	125	6750	3.4	
1603[9]	Triple-Grid Amplifier	6-pin M.	6F	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as 6C6										1603
7700[9]	Triple-Grid Amplifier	6-pin S.	6F	Htr.	6.3	0.3	Class-A Amplifier	Characteristics same as 6C6										7700
RK100	Mercury-vapor Triode	6-pin M.	6A	Htr.	6.3	0.6	Amplifier	100	−2.5	Cathode (G_1) current 250 ma.				20000	50	—	—	RK100

[1] Refer to Receiving Tube Diagrams.
[2] Suppressor grid, connected to cathode inside tube, not shown on base diagram.
[3] Also known as Type LA.
[4] S. — small; M. — medium.
[5] Current to input plate (P_1).
[6] Grids Nos. 2 and 3 connected to plate.
[7] Grid No. 2, screen; grid No. 3, suppressor.
[8] Cathode resistor, 780 ohms.
[9] Low noise, non-microphonic, tubes.
[10] Cathode bias resistor-ohms. Fixed bias not recommended.
[11] Types with final letter "S" have external shield connected to cathode pin.

TABLE V—2.5-VOLT RECEIVING TUBES

Type	Name	Base[3]	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2S/4S	Duodiode	5-pin M.	5D	Htr.	2.5	1.35	Detector	At 50 D.C. Volts per plate, cathode ma. = 80										2S/4S
2A3	Triode Power Amplifier	4-pin M.	4D	Fil.	2.5	2.5	Class-A Amplifier	Characteristics same as Type 6A3, Table IV										2A3
2A5	Pentode Power Amplifier	6-pin M.	6B	Htr.	2.5	1.75	Class-A Amplifier	Characteristics same as Type 42, Table IV										2A5
2A6	Duplex-Diode Triode	6-pin S.	6G	Htr.	2.5	0.8	Class-A Amplifier	Characteristics same as Type 75, Table IV										2A6
2A7[8]	Pentagrid Converter	7-pin S.	7C	Htr.	2.5	0.8	Osc.-Mixer	Characteristics same as Type 6A7, Table IV										2A7
2B6	Direct-Coupled Amplifier	7-pin M.	7J	Htr.	2.5	2.25	Amplifier	250	−24.0	—	—	40.0	5150	3500	18.0	5000	4.0	2B6
2B7[8]	Duplex-Diode Pentode	7-pin S.	7D	Htr.	2.5	0.8	Pentode Amplifier	Characteristics same as Type 6B7—Table IV										2B7
2E5	Electron-Ray Tube	6-pin S.	6R	Htr.	2.5	0.8	Indicator Tube	Characteristics same as Type 6E5—Table IV										2E5
2G5	Electron-Ray Tube	6-pin S.	6R	Htr.	2.5	0.8	Indicator Tube	Characteristics same as 6U5/6G5—Table IV										2G5
24-A	Tetrode R.F. Amplifier	5-pin M.	5E	Htr.	2.5	1.75	Screen-Grid R.F. Amp.	250	− 3.0	90	1.7	4.0	600000	1050	630	—	—	24-A
							Bias Detector	250	− 5.0	20/45	Plate current adjusted to 0.1 ma. with no signal							
27[8]	Triode Detector-Amplifier	5-pin M.	5A	Htr.	2.5	1.75	Class-A Amplifier	250	−21.0	—	—	5.2	9250	975	9.0	—	—	27
							Bias Detector	250	−30.0	—	Plate current adjusted to 0.2 ma. with no signal							
35/51[8]	Variable-μ Amplifier	5-pin M.	5E	Htr.	2.5	1.75	Screen-Grid R.F. Amp.	250	− 3.0	90	2.5	6.5	400000	1050	420	—	—	35/51
45	Triode Power Amplifier	4-pin M.	4D	Fil.	2.5	1.5	Class-A Amplifier	275	−56.0	—	—	36.0	1700	2050	3.5	4600	2.00	45
46	Dual-Grid Power Amplifier	5-pin M.	5C	Fil.	2.5	1.75	Class-A Amplifier[4]	250	−33.0	—	—	22.0	2380	2350	5.6	6400	1.25	46
							Class-B Amplifier[5]	400	0	—	—	Power output for 2 tubes				5800	20.0	

TABLE V—2.5-VOLT RECEIVING TUBES—*Continued*

Type	Name	Base [3]	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
47	Pentode Power Amplifier	5-pin M.	5B	Fil.	2.5	1.75	Class-A Amplifier	250	−16.5	250	6.0	31.0	60000	2500	150	7000	2.7	47
53	Twin Triode Amplifier	7-pin M.	7B	Htr.	2.5	2.0	Class-B Amplifier	Characteristics same as Type 6A6, Table IV										53
55 [8]	Duplex-Diode Triode	6-pin S.	6G	Htr.	2.5	1.0	Class-A Amplifier	Characteristics same as Type 85, Table IV										55
56 [8]	Triode Amplifier, Detector	5-pin S.	5A	Htr.	2.5	1.0	Class-A Amplifier	Characteristics same as Type 76, Table IV										56
57 [8]	Triple-Grid Amplifier	6-pin S.	6F	Htr.	2.5	1.0	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	——	——	57
58 [8]	Triple-Grid Variable-μ	6-pin S.	6F	Htr.	2.5	1.0	Screen-Grid R.F. Amp.	250	− 3.0	100	2.0	8.2	800000	1600	1280	——	——	58
59	Triple-Grid Power Amplifier	7-pin M.	7A	Htr.	2.5	2.0	Class-A Triode [6]	250	−28.0	——	——	26.0	2300	2600	6.0	5000	1.25	59
							Class-A Pentode [7]	250	−18.0	250	9.0	35.0	40000	2500	100	6000	3.0	
RK15	Triode Power Amplifier	4-pin M.	4D [2]	Fil.	2.5	1.75	Characteristics same as Type 46 with Class-B connections											RK15
RK16	Triode Power Amplifier	5-pin M.	5A	Htr.	2.5	2.0	Characteristics same as Type 59 with Class-A triode connections											RK16
RK17	Pentode Power Amplifier	5-pin M.	5F	Htr.	2.5	2.0	Characteristics same as Type 2A5											RK17

[1] Refer to Receiving Tube Diagrams.
[2] Grid connection to cap; no connection to No. 3 pin.
[3] S. — small; M. — medium.
[4] Grid No. 2 tied to plate.
[5] Grids Nos. 1 and 2 tied together.
[6] Grids Nos. 2 and 3 connected to plate.
[7] Grid No. 2, screen; grid No. 3, suppressor.
[8] Types with final letter "S" indicate external shield connected to cathode pin.

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES

Type	Name	Base [2]	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1A4P	Variable-μ Pentode	4-pin S.	4M	Fil.	2.0	0.06	R.F. Amplifier	180	− 3.0	67.5	0.8	2.3	1000000	750	750	——	——	1A4P
1A4T	Variable-μ Tetrode	4-pin S.	4K	Fil.	2.0	0.06	R.F. Amplifier	180	− 3.0	67.5	0.7	2.3	960000	750	720	——	——	1A4T
1A6	Pentagrid Converter	6-pin S.	6L	Fil.	2.0	0.06	Converter	180	− 3.0	67.5	2.4	1.3	500000	Anode grid (No. 2) 180 max. volts				1A6
1B4P/951	Pentode R.F. Amplifier	4-pin S.	4M	Fil.	2.0	0.06	R.F. Amplifier	180 90	− 3.0 − 3.0	67.5 67.5	0.6 0.7	1.7 1.6	1500000 1000000	650 600	1000 550	——	——	1B4P/951
1B5/25S	Duplex-Diode Triode	6-pin S.	6M	Fil.	2.0	0.06	Triode Class-A Amplifier	135	− 3.0	——	——	0.8	35000	575	20	——	——	1B5/25S
1C6	Pentagrid Converter	6-pin S.	6L	Fil.	2.0	0.12	Converter	180	− 3.0	67.5	2.0	1.5	750000	Anode grid (No. 2) 135 max. volts				1C6
1F4	Pentode Power Amplifier	5-pin M.	5K	Fil.	2.0	0.12	Class-A Amplifier	135	− 4.5	135	2.6	8.0	200000	1700	340	16000	0.34	1F4
1F6	Duplex-Diode Pentode	6-pin S.	6W	Fil.	2.0	0.6	R.F. Amplifier	180	− 1.5	67.5	0.6	2.0	1000000	650	650	——	——	1F6
							A.F. Amplifier	135	− 1.0	135	Plate, 0.25 megohm; screen, 1.0 megohm				Amp. =48			
15	R.F. Pentode	5-pin S.	5F	Htr.	2.0	0.22	R.F. Amplifier	135	− 1.5	67.5	0.3	1.85	800000	750	600	——	——	15
19	Twin-Triode Amplifier	6-pin S.	6C	Fil.	2.0	0.26	Class-B Amplifier	135	0	——	——	——	Load plate-to-plate			10000	2.1	19
30	Triode Detector Amplifier	4-pin S.	4D	Fil.	2.0	0.06	Class-A Amplifier	180	−13.5	——	——	3.1	10300	900	9.3	——	——	30
31	Triode Power Amplifier	4-pin S.	4D	Fil.	2.0	0.13	Class-A Amplifier	180	−30.0	——	——	12.3	3600	1050	3.8	5700	0.375	31
32	Tetrode R.F. Amplifier	4-pin M.	4K	Fil.	2.0	0.06	R.F. Amplifier	180	− 3.0	67.5	0.4	1.7	1200000	650	780	——	——	32
33	Pentode Power Amplifier	5-pin M.	5K	Fil.	2.0	0.26	Class-A Amplifier	180	−18.0	180	5.0	22.0	55000	1700	90	6000	1.4	33
34	Variable-μ Pentode	4-pin M.	4M	Fil.	2.0	0.06	R.F. Amplifier	180	− 3.0	67.5	1.0	2.8	1000000	620	620	——	——	34
49	Dual-Grid Power Amplifier	5-pin M.	5C	Fil.	2.0	0.12	Class-A Amplifier [3]	135	−20.0	——	——	6.0	4175	1125	4.7	11000	0.17	49
							Class-B Amplifier [4]	180	0	——	——	Power output for 2 tubes				12000	3.5	
840	R.F. Pentode	5-pin S.	5J	Fil.	2.0	0.13	Class-A Amplifier	180	− 3.0	67.5	0.7	1.0	1000000	400	400	——	——	840
950	Pentode Power Amplifier	5-pin M.	5B	Fil.	2.0	0.12	Class-A Amplifier	135	−16.5	135	2.0	7.0	100000	1000	100	13500	0.45	950
RK24	Triode Amplifier	4-pin M.	4D	Fil.	2.0	0.12	Class-A Amplifier	180	−13.5	——	——	8.0	5000	1600	8.0	12000	0.25	RK24

[1] See Receiving Tube Diagrams.
[2] S.—small; M.—medium.
[3] Grid No. 2 tied to plate.
[4] Grids Nos. 1 and 2 tied together.

TABLE VII — 2.0-VOLT BATTERY TUBES WITH OCTAL BASES

Type	Name	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1C7G	Pentagrid Converter	7Z	Fil.	2.0	0.06	Converter			Characteristics same as Type 1C6 — Table VI								1C7G
1D5GP	Variable-μ R.F. Pentode	5Y	Fil.	2.0	0.06	R.F. Amplifier			Characteristics same as Type 1A4P — Table VI								1D5GP
1D5GT	Variable-μ R.F. Tetrode	5R	Fil.	2.0	0.06	R.F. Amplifier	180	− 3.0	67.5	0.7	2.2	600000	650				1D5GT
1D7G	Pentagrid Converter	7Z	Fil.	2.0	0.06	Converter			Characteristics same as Type 1A6 — Table VI								1D7G
1E5GP	R.F. Amplifier Pentode	5Y	Fil.	2.0	0.06	R.F. Amplifier			Characteristics same as Type 1B4 — Table VI								1E5GP
1E7G	Double Pentode Power Amp.	8C	Fil.	2.0	0.24	Class-A Amplifier	135	− 7.5	135	2.0 [2]	6.5 [2]	220000	1600	350	24000	0.65	1E7G
1F5G	Pentode Power Amplifier	6X	Fil.	2.0	0.12	Class-A Amplifier			Characteristics same as Type 1F4 — Table VI								1F5G
1F7GV [3]	Duplex-Diode Pentode	7AD	Fil.	2.0	0.06	Detector-Amplifier			Characteristics same as Type 1F6 — Table VI								1F7GV
1G5G	Pentode Power Amplifier	6X	Fil.	2.0	0.12	Class-A Amplifier	135	− 13.5	135	2.5	8.7	1600000	1550	250	9000	0.55	1G5G
1H4G	Triode Amplifier	5S	Fil.	2.0	0.06	Detector-Amplifier			Characteristics same as Type 30 — Table VI								1H4G
1H6G	Duplex-Diode Triode	7AA	Fil.	2.0	0.06	Detector-Amplifier			Characteristics same as Type 1B5 — Table VI								1H6G
1J5G	Pentode Power Amplifier	6X	Fil.	2.0	0.12	Class-A Amplifier	135	−16.5	135	2.0	7.0	——	950	100	13500	0.45	1J5G
1J6G	Twin Triode	7AB	Fil.	2.0	0.24	Class-B Amplifier			Characteristics same as Type 19 — Table VI								1J6G

[1] Refer to Receiving Tube Diagrams. [2] Total current for both sections; no signal. [3] Also type G or GH.

TABLE VIII — 1.5-VOLT FILAMENT DRY-CELL TUBES

See also Table X for Special 1.4-volt Tubes

Type	Name	Base	Socket Connections [1]	Filament Volts	Filament Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Milliwatts	Type
1A3	H. F. Diode	7-pin B.[10]	5AP	1.4	0.15	Detector F.M. Discriminator	Max. a.c. voltage per plate — 117. Max. output current — 0.5 ma.										1A3
1A5G	Pentode Power Amplifier	7-pin O.	6X	1.4	0.05	Class-A_1 Amplifier	90 [3]	−4.5 [3]	90	0.8	4.0	300000	850	240	25000	115	1A5G
1A7G	Pentagrid Converter	8-pin O.	7Z	1.4	0.05	Osc.-Mixer	90	0	45 [4]	0.6	0.55	600000	Anode-grid volts 90				1A7G
1AB5	Pentode R.F. Amplifier	8-pin O.	5BF	1.2	0.05	R.F. Amplifier	90 150	0 −1.5	90 150	0.8 2.0	3.5 6.8	275000 125000	1100 1350	——	——	——	1AB5
1B7G	Pentagrid Converter	6-pin O.	7Z	1.4	0.1	Osc.-Mixer	90	0	45	1.3	1.5	350000	Grid No. 1 resistor 200,000 ohms				1B7G
1B8GT	Diode Triode Tetrode	8-pin O.	8AW	1.4	0.1	Triode Amplifier Tetrode Amplifier	90 90	0 −6.0	—— 90	—— 1.4	0.15 6.3	240000 ——	275 1150	—— ——	—— 14000	—— 210	1B8GT
1C5G	Pentode Power Amplifier	7-pin O.	6X	1.4	0.1	Class-A_1 Amplifier	90 [3]	−7.5 [3]	90	1.6	7.5	115000	1550	165	8000	240	1C5G
1D8GT	Diode Triode Pentode	8-pin O.	8AJ	1.4	0.1	Triode Amplifier Pentode Amplifier	90 90	0 −9.0	—— 90	—— 1.0	1.1 5.0	43500 200000	575 925	25 ——	—— ——	—— ——	1D8GT
1E4G	Triode Amplifier	8-pin O.	5S [12]	1.4	0.05	Class-A Amplifier	90 90	0 −3.0	——	——	4.5 1.5	11000 17000	1325 825	14.5 14	——	——	1E4G
1G4G	Triode Amplifier	7-pin O.	5S	1.4	0.05	Class-A Amplifier	90	−6.0	——	——	2.3	10700	825	8.8	——	——	1G4G
1G6G	Twin Triode	6-pin O.	7AB	1.4	0.1	Class-A Amplifier Class-B Amplifier	90 90	0 0	—— ——	—— ——	1.0 1/7 [5]	45000 34 volts input per grid	675	30	—— 12000	—— 675	1G6G
1H5G	Diode High-μ Triode	7-pin O.	5Z	1.4	0.05	Class-A Amplifier	90	0	——	——	0.14	240000	275	65	——	——	1H5G
1L4 [16]	R. F. Pentode Amplifier	7-pin B.[10]	6AR	1.4	0.05	Class-A Amplifier	90	0	90	2.0	4.5	350000	1025	——	——	——	1L4
1LA4	Pentode Power Amplifier	8-pin L.	5AD	1.4	0.05	Class-A Amplifier	Characteristics same as 1A5G										1LA4
1LA6	Pentagrid Converter	8-pin L.	7AK	1.4	0.05	Osc.-Mixer	90	0	45	0.6	0.55	Anode Grid Volts 90					1LA6
1LB4	Pentode Power Amplifier	8-pin L.	5AD	1.4	0.05	Class-A Amplifier	90	−9	90	1.0	5.0	200000	925	——	12000	200	1LB4
1LB6GL	Heptode Converter	8-pin L.	8AX	1.4	0.05	Osc.-Mixer	90	0	67.5	2.2	0.4	Grid No. 4 — 67.5 v., No. 5 — 0 v.					1LB6GL
1LC5	Triple-Grid Variable-μ	8-pin L.	7AO	1.4	0.05	R.F. Amplifier	90	0	45	0.2	1.15	1500000	775	——	——	——	1LC5
1LC6	Pentagrid Converter	8-pin L.	7AK	1.4	0.05	Osc.-Mixer	90	0	35 [9]	0.7	0.75	Anode Grid Volts 45					1LC6

TABLE VIII — 1.5-VOLT FILAMENT DRY-CELL TUBES — *Continued*

Type	Name	Base	Socket Connections[1]	Filament Volts	Filament Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Milliwatts	Type
1LD5	Diode Pentode	7-pin L.	6AX	1.4	0.05	Class-A Amplifier	90	0	45	0.1	0.6	950000	600	—	—	—	1LD5
1LE3	Triode Amplifier	8-pin L.	4AA	1.4	0.05	Class-A Amplifier	90 90	0 −3	—	—	4.5 1.3	11200 19000	1300 760	14.5	—	—	1LE3
1LH4	Diode High-μ Triode	8-pin L.	5AG	1.4	0.05	Class-A Amplifier	90	0	—	—	0.15	240000	275	65	—	—	1LH4
1LN5	Triple-Grid Amplifier	8-pin L.	7AO	1.4	0.05	Class-A Amplifier	90	0	90	0.3	1.2	1500000	750	—	—	—	1LN5
1N5G	Pentode R.F. Amplifier	7-pin O.	5Y	1.4	0.05	Class-A Amplifier	90	0	90	0.3	1.2	1500000	750	1160	—	—	1N5G
1N6G	Diode-Power-Pentode	6-pin O.	7AM	1.4	0.05	Class-A Amplifier	90	−4.5	90	0.6	3.1	300000	800	—	25000	100	1N6G
1P5G	Triple-Grid Pentode	5-pin O.	5Y	1.4	0.05	R.F. Amplifier	90	0	90	0.7	2.3	800000	800	640	—	—	1P5G
1Q5G	Tetrode Power Amplifier	5-pin O.	6AF	1.4	0.1	Class-A Amplifier	85 90	−5.0 −4.5	85 90	1.2 1.6	7.2 9.5	70000 75000	1950 2100	—	9000 8000	250 270	1Q5G
1R4/ 1294	U.h.f. Diode	8-pin L.	4AH	1.4	0.15	Rectifier	Max. r.m.s. voltage per plate — 30					Max. d.c. output current — 340 μa.					1R4/ 1294
1R5	Pentagrid Converter	7-pin B.[10]	7AT	1.4	0.05	Osc.-Mixer	90	0	67.5	3.0	1.7	500000	300	Grid No. 1 100000 ohms			1R5
1S4	Pentagrid Power Amplifier	7-pin B.[10]	7AV	1.4	0.1	Class-A Amplifier	90	−7.0	67.5	1.4	7.4	100000	1575	—	8000	270	1S4
1S5	Diode Pentode	7-pin B.[10]	6AU	1.4	0.05	Class-A Amplifier	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	1S5
						Resistor-Coupled Amp.	90	0	90	Screen resistor 3 meg., grid 10 meg.					1 meg.	50[13]	
1SA6GT	R.F. Pentode	8-pin O.	6CA	1.4	0.05	R.F. Amplifier	90	0	67.5	0.68	2.45	800000	970	—	—	—	1SA6GT
1SB6GT	Diode Pentode	7-pin O.	6CB	1.4	0.05	Class-A Amplifier	90	0	67.5	0.38	1.45	700000	665	—	—	—	1SB6GT
						Resistance-Coupled Amp.	90	0	90	Screen resistor 5 meg., grid 10 meg.					1 meg.	110[13]	
1T4[15]	Triple-Grid Variable-μ	7-pin B.[10]	6AR	1.4	0.05	Class-A Amplifier	90	0	45	0.65	2.0	800000	750	—	—	—	1T4
1T5GT	Beam Power Amplifier	7-pin O.	6AF	1.4	0.05	Class-A Amplifier	90	−6.0	90	1.4	6.5	—	1150	—	14000	170	1T5GT
3B7/ 1291	U.h.f. Twin Triode	8-pin L.	7BE	1.4	0.22	Class-A Amplifier	90	0	—	—	5.2	11350	1850	21	—	—	3B7/ 1291
1293	U.h.f. Triode	8-pin L.	Fig. 2[14]	1.4	0.11	Class-A Amplifier	90	0	—	—	4.7	10750	1300	14	—	—	1293
3D6/ 1299	U.h.f. Tetrode	8-pin L.	6BB	1.4	0.22	Class-A Amplifier	135	−6	90	0.7	5.7	—	2200	—	13000	0.5	3D6/ 1299
CK501	Pentode Voltage Amplifier	None[6]	—[16]	1.25	0.033	Class-A Amplifier	30 45	0 −1.25	30 45	0.06 0.055	0.3 0.28	1000000 1500000	325 300	—	—	—	CK501
CK502	Pentode Output Amplifier	None[6]	—[16]	1.25	0.033	Class-A Amplifier	30	0	30	0.13	0.55	500000	400	—	60000	3	CK502
CK503	Pentode Output Amplifier	None[6]	—[16]	1.25	0.033	Class-A Amplifier	30	0	30	0.33	1.5	150000	600	—	20000	6[7]	CK503
CK504	Pentode Output Amplifier	None[6]	—[16]	1.25	0.033	Class-A Amplifier	30	−1.25	30	0.09	0.4	500000	350	—	60000	3[7]	CK504
CK505	Pentode Voltage Amplifier	None[6]	—[16]	0.625[11]	0.03	Class-A Amplifier	30 45	0 −1.25	30 45	0.07 0.08	0.17 0.2	1100000 2000000	140 150	—	—	—	CK505
CK506	Pentode Output Amplifier	None[6]	—[16]	1.25	0.05	Class-A₁ Amplifier	45	−4.5	45	0.4	1.25	120000	500	—	30000	25	CK506
CK507	Pentode Output Amplifier	None[6]	—[16]	1.25	0.05	Class-A₁ Amplifier	45	−2.5	45	0.21	0.6	360000	500	—	50000	12	CK507
CK509	Triode Voltage Amplifier	None[6]	—[16]	0.625[11]	0.03	Class-A Amplifier	45	0	—	—	0.15	150000	160	16	1000000	—	CK509
HY113 HY123	Triode Amplifier	5-pin P.[8]	5K[8]	1.4	0.07	Class-A Amplifier	45	−4.5	—	—	0.4	25000	250	6.3	40000	6.5	HY113 HY123
HY115 HY145	Pentode Voltage Amplifier	5-pin P.[8]	5K	1.4	0.07	Class-A Amplifier	45 90	−1.5 −1.5	22.5 45	0.008 0.1	0.03 0.48	5200000 1300000	58 270	300 370	—	—	HY115 HY145
HY125 HY155	Pentode Power Amplifier	5-pin P.[8]	5K	1.4	0.07	Class-A Amplifier	45 90	−3.0 −7.5	45 90	0.2 0.5	0.9 2.6	825000 420000	310 450	255 190	50000 28000	11.5 90	HY125 HY155
RK42	Triode Amplifier	4-pin S.	4D	1.5	0.6	Class-A Amplifier	Characteristics same as Type 30—Table VI										RK42
RK43	Twin Triode Amplifier	6-pin S.	6C	1.5	0.12	Twin Triode Amplifier	135	−3	—	—	4.5	14500	900	13	—	—	RK43

[1] Refer to Receiving Tube Diagrams.
[2] M. — medium; S. — small; O. — octal; L. — loktal.
[3] Series bias is recommended.
[4] Obtained from 90-volt supply through 70,000-ohm dropping resistor.
[5] Per tube. Values to left of diagonal line for no-signal condition; values to right are with signal.
[6] Special miniature peanut base.
[7] With 5-megohm grid resistor and 0.02-μfd. grid coupling condenser.
[8] No screen connection.
[9] Through series resistor. Screen voltage must be at least 10 volts lower than oscillator anode.
[10] Special 7-pin "button" base, miniature type.
[11] Two tubes connected in series for 1.4-volt operation.
[12] Internal shield connected to pin 1.
[13] Voltage gain.
[14] See Supplementary Base Diagrams.
[15] No external shield needed.
[16] Tinned wire leads extend from bottom of tube. Connections are labeled on tube.

TABLE IX — HIGH-VOLTAGE HEATER TUBES

Type	Name	Base [3]	Socket Connections [1]	Heater Volts	Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
12A5	Pentode Power Amplifier	7-pin M.	7F	12.6 6.3	0.3 0.6	Class-A_1 Amplifier	100 180	−15 −25	100 180	3/6.5 8/14	17/19 45/48	50000 35000	1700 2400	— —	4500 3300	0.8 3.4	12A5
12A6	Beam Power Amplifier	7-pin O.	7AC	12.6	0.15	Class-A Amplifier	250	−12.5	250	3.5	30	70000	3000	—	7500	3.4	12A6
12A7	Rectifier-Amplifier [5]	7-pin M.	7K	12.6	0.3	Class-A Amplifier	135	−13.5	135	2.5	9.0	102000	975	100	13500	0.55	12A7
12A8GT	Pentagrid Converter	8-pin O.	8A	12.6	0.15	Osc.-Mixer	Characteristics same as 6A8—Table I										12A8GT
12AH7GT	Twin Triode	8-pin O.	8BE	12.6	0.15	Converter and Amplifier	Characteristics same as 6AH7GT — Table II										12AH7GT
12B6M [3]	Diode Triode	6-pin O.	6Y	12.6	0.15	Class-A Amplifier	250	− 2.0	—	—	0.9	91000	1100	100	—	—	12B6M
12B7ML	Pentode Amplifier	8-pin O.	8V	12.6	0.15	Class-A Amplifier	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	12B7ML
12B8GT	Triode-Pentode	8-pin O.	8T	12.6	0.3	Class-A Triode	100	− 1	—	—	0.6	73000	1500	110	—	—	12B8GT
						Class-A Pentode	100	− 3	100	2	8	170000	2100	360	—	—	
12C8	Duplex-Diode Pentode	8-pin O.	8E	12.6	0.15	Class-A Amplifier	Characteristics same as 6B8—Table I										12C8
12E5GT	Triode Amplifier	6-pin O.	6Q	12.6	0.15	Class-A Amplifier	250	−13.5	—	—	50	—	1450	13.8	—	—	12E5GT
12F5GT	Triode Amplifier	5-pin O.	5M	12.6	0.15	Class-A Amplifier	Characteristics same as 6F5—Table I										12F5GT
12G7G	Duplex-Diode Triode	7-pin O.	7V	12.6	0.15	Class-A Amplifier	250	− 3.0	—	—	—	58000	1200	70	—	—	12G7G
12H6	Twin Diode	7-pin O.	6Q	12.6	0.15	Rectifier	Characteristics same as 6H6—Table I										12H6
12J5GT	Triode Amplifier	6-pin O.	6Q	12.6	0.15	Class-A Amplifier	Characteristics same as 6J5—Table I										12J5GT
12J7GT	Pentode Voltage Amplifier	7-pin O.	7R	12.6	0.15	Class-A Amplifier	Characteristics same as 6J7—Table I										12J7GT
12K7GT	Remote Cut-off Pentode	7-pin O.	7R	12.6	0.15	R.F. Amplifier	Characteristics same as 6K7—Table I										12K7GT
12K8	Triode Hexode Converter	8-pin O.	8K	12.6	0.15	Osc.-Mixer	Characteristics same as 6K8—Table I										12K8
12L8GT	Twin Pentode	8-pin O.	8BU	12.6	0.15	Class-A_1 Amplifier [13]	180	− 9.0	180	2.8	13.0	160000	2150	—	10000	1.0	12L8GT
12Q7GT	Duplex-Diode Triode	7-pin O.	7V	12.6	0.15	Class-C Amplifier	Characteristics same as 6Q7—Table I										12Q7GT
12SA7	Pentagrid Converter	8-pin O.	8R	12.6	0.15	Osc.-Mixer	Characteristics same as 6SA7—Table I										12SA7
12SC7	Twin Triode	8-pin O.	8S	12.6	0.15	Class-A Amplifier	Characteristics same as 6SC7 — Table I										12SC7
12SF5	High-μ Triode	6-pin O.	8P	12.6	0.15	Class-A Amplifier	Characteristics same as 6SF5 — Table I										12SF5
12SF7	Diode Variable-μ Pentode	8-pin O.	7AZ	12.6	0.15	Class-A Amplifier	Characteristics same as 6SF7 — Table I										12SF7
12SG7	Triple-Grid Variable-μ	8-pin O.	8BC	12.6	0.15	Class-A Amplifier	Characteristics same as 6SG7 — Table I										12SG7
12SH7	H-F Amplifier Pentode	8-pin O.	8BK	12.6	0.15	H-F Amplifier	Characteristics same as 6SH7 — Table I										12SH7
12SJ7	Pentode Voltage Amplifier	8-pin O.	8N	12.6	0.15	Class-A Amplifier	Characteristics same as 6SJ7—Table I										12SJ7
12SK7	Remote Cut-off Pentode	8-pin O.	8N	12.6	0.15	R.F. Amplifier	Characteristics same as 6SK7—Table I										12SK7
12SL7GT	Twin Triode	8-pin O.	8BD	12.6	0.15	Class-A Amplifier	Characteristics same as 6SL7GT—Table II										12SL7GT
12SN7GT	Twin Triode	8-pin O.	8BD	12.6	0.3	Class-A Amplifier	Characteristics same as 6SN7GT—Table II										12SN7GT
12SQ7	Duplex-Diode Triode	8-pin O.	8Q	12.6	0.15	Class-A Amplifier	Characteristics same as 6SQ7—Table I										12SQ7
12SR7	Duplex-Diode Triode	8-pin O.	8Q	12.6	0.15	Class-A Amplifier	Characteristics same as 6R7—Table I										12SR7
14A4	Triode Amplifier	8-pin L.	5AC	14 [4]	0.16	Class-A Amplifier	Characteristics same as 7A4—Table III										14A4
14A5	Beam Power Amplifier	8-pin L.	6AA	14 [4]	0.16	Class-A_1 Amplifier	250	−12.5	250	3.5/5.5	30/32	70000	3000	—	7500	2.8	14A5
14A7/ 12B7	Triple-Grid Variable-μ	8-pin L.	8V	14 [4]	0.16	Class-A Amplifier	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	14A7/ 12B7
14AF7	Twin Triode	8-pin L.	8AC	14	0.16	Class-A Amplifier	250	−10	—	—	9	7600	2100	16	—	—	14AF7
14B6	Duplex-Diode Triode	8-pin L.	8W	14 [4]	0.16	Class-A Amplifier	Characteristics same as 7B6—Table III										14B6
14B8	Pentagrid Converter	8-pin L.	8X	14 [4]	0.16	Osc.-Mixer	Characteristics same as 7B8—Table III										14B8
14C5	Beam Power Amplifier	8-pin L.	6AA	14 [4]	0.24	Class-A Amplifier	Characteristics same as 6V6—Table I										14C5
14C7	Triple-Grid Amplifier	8-pin L.	8V	14 [4]	0.16	Class-A Amplifier	250	− 3.0	100	0.7	2.2	1000000	1575	—	—	—	14C7
14E6	Duplex-Diode Triode	8-pin L.	8W	14 [4]	0.16	Class-A Amplifier	Characteristics same as 7E6—Table III										14E6
14E7	Duplex-Diode Pentode	8-pin L.	8AE	14 [4]	0.16	Class-A Amplifier	Characteristics same as 7E7—Table III										14E7

TABLE IX — HIGH-VOLTAGE HEATER TUBES — *Continued*

Type	Name	Base [3]	Socket Connections [1]	Heater Volts	Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
14F7	Twin Triode	8-pin L.	8AC	14[4]	0.16	Class-A Amplifier	Characteristics same as 7F7—Table III										14F7
14H7	Triple-Grid Semi-Variable-μ	8-pin L.	8V	14[4]	0.16	Class-A Amplifier	250	− 2.5	150	3.5	9.5	800000	3800	—	—	—	14H7
14J7	Triode-Hexode Converter	8-pin L.	8AR	14[4]	0.16	Osc.-Mixer	Characteristics same as 7J7—Table III										14J7
14N7	Twin Triode	8-pin L.	8AC	14[4]	0.32	Class-A Amplifier	Characteristics same as 7N7—Table III										14N7
14Q7	Heptode Pentagrid Converter	8-pin L.	8AL	14[4]	0.16	Osc.-Mixer	Characteristics same as 7Q7—Table III										14Q7
14R7	Duplex-Diode Pentode	8-pin L.	8AE	14[4]	0.16	Class-A Amplifier	Characteristics same as 7R7—Table III										14R7
14S7	Triode Heptode	8-pin L.	8BL	14[4]	0.16	Osc.-Mixer	250	− 2.0	100	3	1.8	1250000	525	—	—	—	14S7
14W7	Pentode	8-pin L.	8BJ	14[4]	0.24	Class-A Amplifier	300	− 2.2	150	3.9	10	300000	5800	—	—	—	14W7
18	Pentode	6-pin M.	6B	14[4]	0.30	Class-A Amplifier	Characteristics same as 6F6G										18
20J8GM[5]	Triode Heptode Converter	8-pin O.	8H	20	0.15	Osc.-Mixer	250	− 3.0	100	3.4	1.5	Triode Plate (No. 6) 100 v. 1.5 ma.					20J8GM
21A7	Triode Hexode Converter	8-pin L.	8AR	21	0.16	Osc.-Mixer	250 150	− 3.0 − 3.0	100 Triode	2.8	1.3 3.5	— —	275 1900	— 32	— —	— —	21A7
25A6	Pentode Power Amplifier	7-pin O.	7S	25	0.3	Class-A Amplifier	135	−20.0	135	8	37	35000	2450	85	4000	2.0	25A6
25A7G	Rectifier-Amplifier[6]	8-pin O.	8F	25	0.3	Class-A Amplifier	100	−15.0	100	4	20.5	50000	1800	90	4500	0.77	25A7G
25AC5G	Triode Power Amplifier	6-pin O.	6Q	25	0.3	Class-A Amplifier	110 165	+15.0 Used in dynamic-coupled circuit with 6AF5G driver	—	—	45	—	3800	58	2000 3500	2.0 3.3	25AC5G
25B5	Direct-Coupled Triodes	6-pin S.	6D	25	0.3	Class-A Amplifier	110	0	110	7	45	11400	2200	25	2000	2.0	25B5
25B6G	Pentode Power Amplifier	7-pin O.	7S	25	0.3	Class-A Amplifier	95	−15.0	95	4	45	—	4000	—	2000	1.75	25B6G
25B8GT	Triode Pentode	8-pin O.	8T	25	0.15	Class-A Amplifier	Characteristics same as 12B8GT										25B8GT
25C6G	Beam Power Amplifier	7-pin O.	7AC	25	0.3	Class-A₁ Amplifier	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	25C6G
25D8GT	Diode Triode Pentode	8-pin O.	8AF	25	0.15	Triode Amplifier Pentode Amplifier	100 100	− 1.0 − 3.0	— 100	— 2.7	0.5 8.5	91000 200000	1100 1900	100 —	— —	— —	25D8GT
25L6	Beam Power Amplifier	7-pin O.	7AC	25	0.3	Class-A₁ Amplifier	110	− 8.0	110	3.5/10.5	45/48	10000	8000	80	2000	2.2	25L6
25N6G	Direct-Coupled Triodes	7-pin O.	7W	25	0.3	Class-A Amplifier	110	0	110	7	45	11400	2200	25	2000	2.0	25N6G
32L7GT	Diode-Beam Tetrode[6]	8-pin O.	8F	32.5	0.3	Class-A Amplifier	110	− 7.5	110	3	40	15000	6000	—	2500	1.5	32L7GT
35A5	Beam Power Amplifier	8-pin L.	6AA	35	0.15	Class-A₁ Amplifier	110	− 7.5	110	3/7	40/41	14000	5800	—	2500	1.5	35A5
35L6G	Beam Power Amplifier	7-pin O.	7AC	35	0.15	Class-A₁ Amplifier	110	− 7.5	110	3/7	40/41	13800	5800	—	2500	1.5	35L6G
43	Pentode Power Amplifier	6-pin M.	6B	25	0.3	Class-A Amplifier	95	−15.0	95	4.0	20.0	45000	2000	90	4500	0.90	43
48	Tetrode Power Amplifier	6-pin M.	6A	30	0.4	Class-A Amplifier	96	−19.0	96	9.0	52.0	—	3800	—	1500	2.0	48
50A5	Beam Power Amplifier	8-pin L.	6AA	50	0.15	Class-A₁ Amplifier	110	− 7.5	110	4/11	49/50	10000	8200	—	2000	2.2	50A5
50C6G	Beam Power Amplifier	7-pin O.	7AC	50	0.15	Class-A₁ Amplifier	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	50C6GT
50L6GT	Beam Power Amplifier	7-pin O.	7AC	50	0.15	Class-A₁ Amplifier	110	− 7.5	110	4/11	49/50	—	8200	82	2000	2.2	50L6GT
70A7GT	Diode-Beam Tetrode[6]	8-pin O.	8AB[7]	70	0.15	Class-A Amplifier	110	− 7.5	110	3.0	40	—	5800	80	2500	1.5	70A7GT
70L7GT	Diode-Beam Tetrode[6]	8-pin O.	8AA	70	0.15	Class-A₁ Amplifier	110	− 7.5	110	3/6	40/43	15000	7500	—	2000	1.8	70L7GT
117L7GT/ 117M7GT	Rectifier-Amplifier[6]	8-pin O.	8AO	117	0.09	Class-A Amplifier	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117L7GT/ 117M7GT
117N7GT	Rectifier-Amplifier[6]	8-pin O.	8AV	117	0.09	Class-A Amplifier	100	− 6.0	100	5.0	51	16000	7000	—	3000	1.2	117N7GT
117P7GT	Rectifier-Amplifier[6]	8-pin O.	8AV	117	0.09	Class-A Amplifier	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117P7GT
1284	U.h.f. Pentode	8-pin O.	Fig. 4[8]	12.6	0.15	Class-A Amplifier	250	− 3.0	100	2.5	9.0	800000	2000	—	—	—	1284
1629	Electron-Ray Tube	7-pin O.	6RA	12.6	0.15	Indicator Tube	Characteristics same as 6E5—Table IV										1629
1631	Beam Power Amplifier	7-pin O.	7AC	12.6	0.45	Class-A Amplifier	Characteristics same as 6L6—Table I										1631
1632	Beam Power Amplifier	7-pin-O.	7AC	12.6	0.6	Class-A Amplifier	Characteristics same as 25L6										1632
1633	Twin Triode	8-pin-O.	8BD	25	0.15	Class-A Amplifier	Characteristics same as 6SN7GT—Table II										1633
1634	Twin Triode	8-pin-O.	8S	12.6	0.15	Class-A Amplifier	Characteristics same as 6SC7—Table I										1634
1644	Twin Pentode	8-pin O.	Fig. 7[8]	12.6	0.15	Class-A Amplifier	180	− 9.0	180	2.8/4.6	13	160000	2150	—	10000	1.0	1644

TABLE IX — HIGH-VOLTAGE HEATER TUBES — *Continued*

Type	Name	Base[3]	Socket Connections[1]	Heater Volts	Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
XXD	Twin Triode	8-pin L.	8AC	12.6	0.15	Class-A Amplifier	250	— 10	—	—	9.0	—	2100	16	—	—	XXD
28D7	Double Beam Power Amplifier	8-pin L.	8BS	28.0	0.4	Class-A$_2$ Amplifier	28	390[10] 180[10]	28[8] 28[9]	0.7[8] 1.2[9]	9.0[8] 18.5[9]	— —	— —	— —	4000[9] 6000[11]	0.08[9] 0.175[9]	28D7

[1] Refer to Receiving Tube Diagrams.
[2] M.—medium; S.—small; O.—octal; L.—loktal.
[3] Metal-sprayed glass envelope.
[4] Maximum rating, corresponding to 130-volt line condition; normal rating is 12.6 v. for 117-v. line.
[5] For rectifier data, see Table XIII.
[6] See Supplementary Base Diagrams.
[7] 6.3-volt pilot lamp must be connected between pins 6 and 7.
[8] Per section (except heater)—resistance coupled.
[9] P. P. operation—values for both sections, resistance coupled.
[10] Cathode resistor—ohms.
[11] Plate to plate.
[12] Type Y has micanol base.
[13] Each unit.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base[2]	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
00-A	Triode Detector	4-pin M.	4D	Fil.	5.0	0.25	Grid Leak Detector	45	—	—	—	1.5	30000	666	20	—	—	00-A
01-A	Triode Detector Amplifier	4-pin M.	4D	Fil.	5.0	0.25	Class-A Amplifier	135	— 9.0	—	—	3.0	10000	800	8.0	—	—	01-A
3A4	Power Amplifier Pentode	7-pin B.	7BB	Fil.[6]	1.4 2.8	0.2 0.1	Class-A Amplifier	135 150	— 7.5 — 8.4	90 90	2.6 2.2	14.8 13.3	90000 100000	1900	—	8000	0.6 0.7	3A4
3A5	H.F. Twin Triode	7-pin B.	7BC	Fil.[6]	1.4 2.8	0.22 0.11	Class-A Amplifier	90	— 2.5	—	—	3.7	8300	1800	15	—	—	3A5
3A8GT	Diode Triode Pentode	8-pin O.	8AS	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Triode Class-A Pentode	90 90	0 0	— 90	— 0.3	0.15 1.2	240000 600000	275 750	65 —	— —	— —	3A8GT
3B5GT	Beam Power Amplifiers	7-pin O.	7AP	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	67.5	— 7.0	67.5	0.6 0.5	8.0 6.7	100000	1650 1500	—	5000	0.2 0.18	3B5GT
3C5GT	Power Output Pentode	7-pin O.	7AQ	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	90	— 9.0	90	1.4	6.0	—	1550 1450	—	8000 10000	0.24 0.26	3C5GT
3LE4	Power Amplifier Pentode	8-pin L.	6BA	Fil.	2.8	0.05	Class-A Amplifier	90	— 9.0	90	1.8	9.0	110000	1600	—	6000	0.30	3LE4
3LF4	Power Amplifier Tetrode	8-pin L.	6BB	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	90	— 4.5	90	1.3 1.0	9.5 8.0	75000 80000	2200 2000	—	8000 7000	0.27 0.23	3LF4
3Q4	Power Amplifier Pentode	7-pin B.[8]	7BA	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	90	— 4.5	90	2.1 1.7	9.5 7.7	100000 120000	2150 2000	—	10000	0.27 0.24	3Q4
3Q5GT	Beam Power Amplifier	7-pin O.	7AP	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	90	— 4.5	90	1.6 1.0	9.5 7.5	—	2100 1800	—	8000	0.27 0.25	3Q5GT
3S4	Power Amplifier Pentode	7-pin B.[8]	7BA	Fil.[6]	1.4 2.8	0.1 0.05	Class-A Amplifier	90	— 7.0	67.5	1.4 1.1	7.4 6.1	100000	1575 1425	—	8000	0.27 0.235	3S4
10[24]	Triode Power Amplifier	4-pin M.	4D	Fil.	7.5	1.25	Class-A Amplifier	425	—39.0	—	—	18.0	5000	1600	8.0	10200	1.6	10
11/12	Triode Detector Amplifier	4-pin M.	4D	Fil.	1.1	0.25	Class-A Amplifier	135	—10.5	—	—	3.0	15000	440	6.6	—	—	11/12
20	Triode Power Amplifier	4-pin S.	4D	Fil.	3.3	0.132	Class-A Amplifier	135	—22.5	—	—	6.5	6300	525	3.3	6500	0.11	20
22	Tetrode R.F. Amplifier	4-pin M.	4K	Fil.	3.3	0.132	R.F. Amplifier	135	— 1.5	67.5	1.3	3.7	325000	500	160	—	—	22
26	Triode Amplifier	4-pin M.	4D	Fil.	1.5	1.05	Class-A Amplifier	180	—14.5	—	—	6.2	7300	1150	8.3	—	—	26
40	Triode Voltage Amplifier	4-pin M.	4D	Fil.	5.0	0.25	Class-A Amplifier	180	— 3.0	—	—	0.2	150000	200	30	—	—	40
4A6G	Twin Triode Amplifier	8-pin O.	8L	Fil.[5]	4[3] 2[3]	0.06 0.12	Class-A Amplifier[4] Class-B Amplifier	90 90	— 1.5 0	— —	— —	2.2 4.6[5]	13300 —	1500 —	20 —	— 8000	— 1.0	4A6G
50	Triode Power Amplifier	4-pin M.	4D	Fil.	7.5	1.25	Class-A Amplifier	450	—84.0	—	—	55.0	1800	2100	3.8	4350	4.6	50
6AG5[27]	Pentode R.F. Amplifier	7-pin B.[8]	7BD	Htr.	6.3	0.3	Class-A Amplifier	250 100	200[16] 100[16]	150 100	2.0 1.6	7.0 5.5	800000 300000	5000 4750	— —	— —	—	6AG5

TABLE X — SPECIAL RECEIVING TUBES — *Continued*

Type	Name	Base [2]	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6AK5	H.F. Pentode	7-pin B.[8]	7BD	Htr.	6.3	0.175	R.F. Amplifier	180	200[16]	120	2.4	7.7	690000	5100	3500	—	—	6AK5
								150	330[16]	140	2.2	7.0	420000	4300	1800	—	—	
								120	200[16]	120	2.5	7.5	340000	5000	1700	—	—	
6AK6	Power Amplifier Pentode	7-pin B.[8]	7BK	Htr.	6.3	0.15	Class-A Amplifier	180	−9.0	180	2.5	15.0	200000	2300	—	10000	1.1	6AK6
6AL5	U.H.F. Twin Diode [25]	7-pin B.[8]	6BT	Htr.	6.3	0.3	Detector	Max. r.m.s. voltage — 150. Max. d.c. output current — 9 ma.[26]										6AL5
6AQ6	Duodiode Hi-mu Triode	7-pin B.[8]	7BT	Htr.	6.3	0.15	Class-A Triode	250	−3.0	—	—	1.0	58000	1200	70	—	—	6AQ6
								100	−1.0	—	—	0.8	61000	1150	70	—	—	
6C4	Triode Amplifier	7-pin B.[8]	6BG	Htr.	6.3	0.15	Class-A Amplifier	250	−8.5	—	—	10.5	7700	2200	17	—	—	6C4
6F4	Acorn Triode	Acorn	7BR	Htr.	6.3	0.225	Class-A1 Amplifier	80	150[16]	—	—	13.0	2900	5800	17	—	—	6F4
6J4	U.H.F. Grounded-Grid R.F. Amplifier	7-pin B.[8]	7BQ	Htr.	6.3	0.4	Grounded-Grid Class-A Amplifier	150	100[16]	—	—	15.0	4500	12000	55	—	—	6J4
								100	100[16]	—	—	10.0	5000	11000	55	—	—	
6J6 [28]	Twin Triode	7-pin B.[8]	7BF	Htr.	6.3	0.45	Class-A Amplifier Mixer, Oscillator	100	50[16]	—	—	8.5	7100	5300	38	—	—	6J6
71-A	Triode Power Amplifier	4-pin M.	4D	Fil.	5.0	0.25	Class-A Amplifier	180	−43.0	—	—	20.0	1750	1700	3.0	4800	0.79	71-A
99 [23]	Triode Detector Amplifier	4-pin S.	4D	Fil.	3.3	0.063	Class-A Amplifier	90	−4.5	—	—	2.5	15500	425	6.6	—	—	99
112A	Triode Detector Amplifier	4-pin M.	4D	Fil.	5.0	0.25	Class-A Amplifier	180	−13.5	—	—	7.7	4700	1800	8.5	—	—	112A
182B/482B	Triode Amplifier	4-pin M.	4D	Fil.	5.0	1.25	Class-A Amplifier	250	−35.0	—	—	18.0	—	1500	5.0	—	—	182B/482B
183/483	Power Triode	4-pin M.	4D	Fil.	5.0	1.25	Class-A Amplifier	250	−60.0	—	—	25.0	18000	1800	3.2	4500	2.0	183/483
257	Power Pentode	5-pin M.	5B	Fil.	5.0	0.3	Class-A Amplifier	110	−21.5	110	7.0	20.0	41000	1350	55	6000	0.8	257
485	Triode	5-pin S.	5A	Htr.	3.0	1.3	Class-A Amplifier	180	−9.0	—	—	6.0	9300	1350	12.5	—	—	485
864	Triode Amplifier	4-pin S.	4D	Fil.	1.1	0.25	Class-A Amplifier	90	−4.5	—	—	2.9	13500	610	8.2	—	—	864
954 [7]	Pentode Detector, Amplifier	Special	5BB	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	0.7	2.0	1.5 megohms	1400	2000	—	—	954
							Bias Detector	250	−6.0	100	—	Plate current to be adjusted to 0.1 ma. with no signal					—	
955 [7]	Triode Detector, Amplifier, Oscillator	Special	5BC	Htr.	6.3	0.15	Class-A Amplifier	250	−7.0	—	—	6.3	11400	2200	25	—	—	955
								90	−2.5	—	—	2.5	14700	1700	25	—	—	
956 [7]	Triple-Grid Variable-μ R.F. Amplifier	Special	5BB	Htr.	6.3	0.15	R.F. Amplifier	250	−3.0	100	2.7	6.7	700000	1800	1440	—	—	956
							Mixer	250	−10.0	100	—	—	Oscillator peak volts — 7 min.					
957 [7]	Triode Det., Amp., Osc.	Special	5BD	Fil.	1.25	0.05	Class-A Amplifier	135	−5.0	—	—	2.0	20800	650	13.5	—	—	957
958 [7] 958-A	Triode A.F. Amp., Osc.	Special	5BD	Fil.	1.25	0.1	Class-A Amplifier	135	−7.5	—	—	3.0	10000	1200	12	—	—	958 958-A
959 [7]	Pentode Det., Amplifier	Special	5BE	Fil.	1.25	0.05	Class-A Amplifier	135	−3.0	67.5	0.4	1.7	800000	600	480	—	—	959
7E5/1201	U.H.F. Triode	8-pin L.	8BN	Htr.	6.3	0.15	Class-A Amplifier	180	−3	—	—	5.5	12000	—	36	—	—	7E5/1201
7C4/1203	U.H.F. Diode	8-pin L.	4AH	Htr.	6.3	0.15	Rectifier	Max. r.m.s. voltage — 150					Max. d.c. output current — 8 ma.					7C4/1203
1204	U.H.F. Pentode	8-pin L.	Fig. 5[11]	Htr.	6.3	0.15	Class-A Amplifier	250	−2	100	0.6	1.75	800000	1200	—	—	—	1204
1609	Pentode Amplifier	5-pin S.	5B	Fil.	1.1	0.25	Class-A Amplifier	135	−1.5	67.5	0.65	2.5	400000	725	300	—	—	1609
9001	Triple-Grid Detector, Amplifier	7-pin B.[8]	7PM	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	0.7	2.0	Over 1 meg.	1400	—	—	—	9001
							Mixer	250	−5.0	100	Osc. peak voltage 4 volts			550	—	—	—	
9002	Triode Det., Amp., Osc.	7-pin B.[8]	7TM	Htr.	6.3	0.15	Class-A Amplifier	250	−7.0	—	—	6.3	11400	2200	25	—	—	9002
								90	−2.5	—	—	2.5	14700	1700	25	—	—	
9003	Triple-Grid Variable-μ R.F. Amplifier	7-pin B.[8]	7PM	Htr.	6.3	0.15	Class-A Amplifier	250	−3.0	100	2.7	6.7	700000	1800	—	—	—	9003
							Mixer	250	−10.0	100	Osc. peak voltage 9 volts			600	—	—	—	
9004	U.H.F. Diode	Special	Fig. 15[11]	Htr.	6.3	0.15	Detector	Max. a.c. voltage — 117. Max. d.c. output current — 5 ma.										9004
9005	U.H.F. Diode	Special	Fig. 16[11]	Htr.	3.6	0.165	Detector	Max. a.c. voltage — 117. Max. d.c. output current — 1 ma.										9005
9006	U.H.F. Diode	7-pin B.[8]	6BH	Htr.	6.3	0.15	Detector	Max. a.c. voltage — 270. Max. d.c. output current — 5 ma.										9006

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base[2]	Socket Connections[1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance, Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
EF-50	High Frequency Pentode Amplifier	9-Pin L.	Fig. 14[11]	Htr.	6.3	0.3	I.F.-R.F. Amplifier	250	150[16]	250	3.1	10	600000	6300	—	—	—	EF-50
GL-2C44 GL-464A	U.H.F. Triode[29]	6-pin O.	Fig. 17[11]	Htr.	6.3	0.75	Class-A Amplifier and Modulator	250	100[16]	—	—	25.0	—	7000	—	—	—	GL-2C44 GL-464A
GL-446A GL-446B[22]	U.H.F. Triode[29]	6-pin O.	Fig. 19[11]	Htr.	6.3	0.75	Oscillator, Amplifier or Converter	250	200[16]	—	—	15.0	—	4500	45	—	—	GL-446A GL-446B
GL-559	U.H.F. Diode[29]	6-pin O.	Fig. 18[11]	Htr.	6.3	0.75	Detector or transmission line switch	5.0	—	—	—	24.0	—	—	—	—	—	GL-559
M54	Tetrode Power Amplifier	None[9]	—	Fil.	0.625[10]	0.04	Class-A Amplifier	30	0	30	0.06	0.5	130000	200	26	35000	0.005	M54
M64	Tetrode Voltage Amplifier	None[9]	—	Fil.	0.625[10]	0.02	Class-A Amplifier	30	0	—	—	0.03	200000	110	25	—	—	M64
M74	Tetrode Voltage Amplifier	None[9]	—	Fil.	0.625[10]	0.02	Class-A Amplifier	30	0	7.0	0.01	0.02	500000	125	70	—	—	M74
XXB	Twin Triode Frequency Converter	7-pin L.	Fig. 9[11]	Fil.[6], [21]	2.8/1.4	0.05/0.10	Frequency Converter[17]	90[12]	0	—	—	4.5[19] 4.5[20]	11200[19] 11200[20]	1300[19] 1300[20]	14.5[12]	—	—	XXB
					3.2[18]/1.6	—			−3	—	—	1.4[19] 1.4[20]	1900[19] 1900[20]	760[19] 760[20]	14.5[12]	—	—	
XXFM	Twin-Diode Triode	8-pin L.	Fig. 10[11]	Htr.	6.3	0.3	Special Detector Amplifier	250[13]	−1	—	—	1.9	6700	1500	100	—	—	XXFM
								100[13]	0	—	—	1.2	85000	1000	85	—	—	
								100[14]	—	—	—	4[15]	—	—	—	—	—	

1 Refer to Receiving Tube Diagrams.
2 M.—medium, S.—small, O. — octal, L. — loktal.
3 Cathode terminal is mid-point of filament; use series connection with 4 volts, parallel with 2 volts.
4 Triodes connected in parallel.
5 Idling current, both plates.
6 Filament mid-point tap permits series or parallel connection.
7 "Acorn" type, miniature unbased tubes for ultrahigh frequencies.
8 Special 7-pin "button" base, miniature type.
9 No base, tinned wire leads. Dimensions 0.36" x 1.10".
10 Intended for series-parallel operation on 1.4-volt dry cell.
11 See Supplementary Base Diagrams.
12 Both Sections.
13 Amplifier plate.
14 Diode plates (A.C. max. volts per plate).
15 Max. D.C. output.
16 Cathode resistor ohms.
17 Section No. 2 recommended for h.f.o.
18 Dry battery operation.
19 Section No. 1.
20 Section No. 2.
21 Series operation, pin 8 is negative & pin 9 positive. Parallel operation, pins 1 & 8 tied together for positive.
22 Highest frequency oscillator. Use 10,000 to 20,000 ohm grid-leak in this service.
23 Same as X99. Type V99 is same, but socket connections are 4E.
24 Type 210-T has ceramic base.
25 Resonant frequency 700 Mc.
26 Per plate.
27 Useful up to 400 Mc.
28 Useful up to 600 Mc.
29 "Lighthouse" tube. Has special ring contacts.

TABLE XI—CONTROL AND REGULATOR TUBES

Type	Name	Base[1]	Socket Connections[2]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Peak Anode Voltage	Max. Anode Current[3]	Minimum Starting Voltage	Operating Voltage	Operating Current	Grid Resistor	Tube Voltage Drop	Type
0A4G	Gas Triode Starter-Anode Type	6-pin O.	4V	Cold	—	—	Cold-Cathode Starter-Anode Relay Tube	With 105–120-volt a.c. anode supply, peak starter-anode a.c. voltage is 70, peak r.f. voltage 55. Peak D.C. ma = 100. Average D.C. ma = 25							0A4G
1C21	Gas Triode Glow-Discharge Type	6-pin O.	4V	Cold	—	—	Relay Tube	125–145	25	66[15]	—	—	—	73	1C21
							Voltage Regulator		0.1[15]	180[17]				55[15]	
2A4G	Gas Triode Grid Type	7-pin O.	5S	Fil.	2.5[16]	2.5	Control Tube	200	100	—	—	—	—	15	2A4G
2D21	Gas Tetrode	7-pin B.[21]	7BN	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	650	100	0.1–10[15]	8	2D21
							Relay Tube	400	—	—	—	—	1.0[15]	—	
3C23	Gas and Mercury Vapor Grid Type	4-pin M.	3G	Fil.[19]	2.5	7.0	Grid-Controlled Rectifier	1000	6000	—	500	1500	−4.5[20]	15	3C23
										—	100	1500	−2.5[20]	15	
874	Voltage Regulator	4-pin M.	4S	—	—	—	Voltage Regulator[5]	—	—	125	90	10–50	—	—	874
876	Current Regulator	Mogul	—	—	—	—	Current Regulator[5]	—	—	—	40–60	1.7	—	—	876
884	Gas Triode Grid Type	6-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	2	25000[4]	—	884
							Grid-Controlled Rectifier	350	300	—	—	75	25000[4]	—	
885	Gas Triode Grid Type	5-pin S.	5A	Htr.	2.5	1.4	Same as Type 884	Characteristics same as Type 884							885
886	Current Regulator	Mogul	—	—	—	—	Current Regulator[5]	—	—	—	40–60	2.05	—	—	886
967	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	2500	500	−5[11]	—	—	—	10–24	967

TABLE XI—CONTROL AND REGULATOR TUBES—*Continued*

Type	Name	Base [1]	Socket Connections [2]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Peak Anode Voltage	Max. Anode Current [3]	Minimum Starting Voltage	Operating Voltage	Operating Current	Grid Resistor	Tube Voltage Drop	Type
991	Voltage Regulator	Bayonet [14]	—	—	—	—	Voltage Regulator	—	—	87	55–60	2.0	—	—	991
2050	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	—	100	0.1–10 [18]	8	2050
2051	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	350	375	—	—	75	0.1–10 [18]	14	2051
KY21	Gas Triode Grid Type	4-pin M.	—	Fil.	2.5	10.0	Grid-Controlled Rectifier	—	—	—	3000	500	—	—	KY21
RK62	Gas Triode Grid Type	4-pin S.	4D	Fil.	1.4	0.05	Relay Tube [6]	45	1.5	—	30–45	0.1–1.5	—	15	RK62
RM208	Permatron	4-pin M.	—	Fil.	2.5	5.0	Controlled Rectifier [7]	7500 [8]	1000	—	—	—	—	15	RM208
RM209	Permatron	4-pin M.	—	Fil.	5.0	10.0	Controlled Rectifier [7]	7500 [8]	5000	—	—	—	—	15	RM209
OA3/VR75	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	105	75	5–40 [9]	—	—	OA3/VR75
OB3/VR90	Voltage Regulator	7-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	125	90	5–40 [9]	—	—	OB3/VR90
OC3/VR105	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	135	105	5–40 [9]	—	—	OC3/VR105
OD3/VR150	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	185	150	5–40 [9]	—	—	OD3/VR150
KY866	Mercury Vapor Triode	4-pin M.	Fig. 8 [22]	Fil.	2.5	5.0	Grid-Controlled Rectifier	10000	1000	100–150	—	—	—	—	KY866

[1] M.—medium; S.—small; O.—octal.
[2] Refer to Receiving Tube Diagrams.
[3] In ma.
[4] Not less than 1000 ohms per grid volt; 500,000 ohms max.
[5] For use in series with power transformer primary.
[6] For use as self-quenching super-regenerative detector with high-resistance relay (5000–10000 ohms) in anode circuit.
[7] For use as grid-controlled rectifier or with external magnetic control. RM-208 has characteristics of 866, RM-209 of 872.
[8] When under control peak inverse rating is reduced to 2500.
[9] Sufficient resistance must be used in series with tube to limit current to maximum current rating.
[10] Refer to Transmitting Tube Diagrams.
[11] At 1000 anode volts.
[12] At 350 anode volts and 0 Grid No. 2 volts.
[13] At 650 anode volts and 0 Grid No. 2 volts.
[14] Candelabra type, double contact.
[15] Grid.
[16] Filament voltage should be applied 2 seconds before using.
[17] Grid tied to plate.
[18] Megohms.
[19] Heating time 15 seconds.
[20] Grid voltage.
[21] Special 7-pin button-base miniature.
[22] Refer to supplementary base diagrams.

TABLE XII—CATHODE-RAY TUBES AND KINESCOPES

Type	Name	Socket Connections [1]	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage [2]	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [3]	Screen Input Power [4]	Deflection Sensitivity [5] D1 D2	Deflection Sensitivity [5] D3 D4	Screen Persistence [6]	Pattern Color [6]	Type
2AP1	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillograph Television	2"	1000	250	−60	—	—	660	—	0.11	0.13	P1	Green	2AP1
							500	125	−30	—	—			0.22	0.26			
3AP1/ 906-P1 [15]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph Television	3"	1500	475	−50	—	—	600	10	0.22	0.23	P1	Green	3AP1/ 906-P1
							1000	285	−34	—	—			0.33	0.35			
							600	170	−20	—	—			0.55	0.58	P4	White	
3BP1	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph Television	3"	2000	575	−60	—	—	550	—	0.115	0.155	P1	Green	3BP1
							1500	430	−45	—	—			0.153	0.207			
3EP1/ 1806-P1	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph Television	3"	2000	575	−60	—	—	550	—	0.115	0.154	P1	Green	3EP1/ 1806-P1
							1500	430	−45	—	—			0.153	0.205			
3GP1 [15]	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	3"	1500	350	−50	—	—	500	—	0.32	0.36	P1	Green	3GP1
							1000	234	−33	—	—		—					
5AP1/ 1805-P1 5AP4/ 1805-P4	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5"	2000	575	−35	—	—	500	10	0.17	0.21	P1	Green	5AP1/ 1805-P1 5AP4/ 1805-P4
							1500	430	−27					0.23	0.28	P4	White	
5BP1/ 1802-P1 5BP2 5BP4/ 1802-P4 5BP5	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5"	2000	425	−35	—	—	500	10	0.3	0.33	P1 P2	Green Green	5BP1/ 1802-P1 5BP2 5BP4/ 1802-P4 5BP5
							1500	310	−21	—	—			0.4	0.44	P4 P5	White Blue	

TABLE XII—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections[1]	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage[2]	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage[3]	Screen Input Power[4]	Deflection Sensitivity[5] D_1 D_2	Deflection Sensitivity[5] D_3 D_4	Screen Persistence[6]	Pattern Color[6]	Type
5CP1[16]	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph Television	5"	2000[12] 1500[13] 2000[14]	575 430 575	− 60 − 45 − 60	— — —	— — —	550	— — —	0.28 0.37 0.36	0.32 0.43 0.41	P1	Green	5CP1
5HP1 5HP4	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	5"	2000 1500	425 310	− 40 − 30	—	—	500	—	0.3 0.4	0.33 0.44	P1 P4	Green White	5HP1 5HP4
5JP1[16]	Electrostatic Cathode-Ray	11E	6.3	0.6	Oscillograph	5"	2000 1000	520 260	− 75 − 37	— —	— —	500	— —	0.24 0.48	0.28 0.56	P1	Green	5JP1
5LP1[16]	Electrostatic Cathode-Ray	11F	6.3	0.6	Oscillograph Television	5"	2000[12] 1500[13] 1000[14]	500 375 250	− 60 − 45 − 30	— — —	— — —	500	— — —	0.25 0.33 0.49	0.28 0.37 0.56	P1	Green	5LP1
5MP1[15]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	5"	1500 1000	375 250	− 50 − 33	— —	— —	600	— —	0.58	0.64	P1	Green	5MP1
7AP4	Electromagnetic Picture Tube	5AJ	2.5	2.1	Television	7"	3500	1000	−67.5	—	—	—	2.5	—	—	P4	White	7AP4
7CP1/ 1811-P1	Electromagnetic Cathode-Ray	6AZ	6.3	0.6	Oscillograph	7"	7000 4000	1470 840	− 45 − 45	250 250	— —	—	—	— —	— —	P1	Green	7CP1/ 1811-P1
9AP4/ 1804-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	9"	7000 6000	1425 1225	− 40 − 38	250	25	—	10	—	—	P4	White	9AP4/ 1804-P4
9CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	9"	7000	—	−110	—	25	—	10	—	—	P4	White	9CP4
9JP1/ 1809-P1	Electrostatic Cathode-Ray	8BR	2.5	2.1	Oscillograph	9"	5000 2500	1570 785	− 90 − 45	—	—	3000	—	0.136 0.272	— —	P1	Green	9JP1/ 1809-P1
12AP4/ 1803-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	12"	7000 6000	1460 1240	− 75	250	25	—	10	—	—	P4	White	12AP4/ 1803-P4
12CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	12"	7000	—	−110	—	25	—	10	—	—	P4	White	12CP4
902	Electrostatic Cathode-Ray	Fig. 1[11]	6.3	0.6	Oscillograph	2"	600	150	− 60	—	—	350	5	0.19	0.22	P1	Green	902
903[10]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9"	7000	1360	−120	250	—	—	10	—	—	P1	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3[11]	2.5	2.1	Oscillograph	5"	4600	970	− 75	250	—	4000	10	0.09	—	P1	Green	904
905	Electrostatic Cathode-Ray	Fig. 6[11]	2.5	2.1	Oscillograph	5"	2000	450	− 35	—	—	1000	10	0.19	0.23	P1	Green	905
907	Electrostatic Cathode-Ray	Fig. 6[11]	2.5	2.1	Oscillograph	5"	Characteristics same as Type 905							—	—	P5	Blue	907
908	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1							—	—	P5	Blue	908
909[10]	Electrostatic Cathode-Ray	Fig. 6[11]	2.5	2.1	Oscillograph	5"	Characteristics same as Type 905							—	—	P2	Blue	909
910[10]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1							—	—	P2	Blue	910
911[10]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1[7]							—	—	P1	Green	911
912	Electrostatic Cathode-Ray	Fig. 8[11]	2.5	2.1	Oscillograph	5"	10000	2000	− 66	250	—	7000	10	0.041	0.051	P1	Green	912
913	Electrostatic Cathode-Ray	Fig. 1[11]	6.3	0.6	Oscillograph	1"	500	100	− 65	—	—	250	5	0.07	0.10	P1	Green	913
914	Electrostatic Cathode-Ray	Fig. 12[11]	2.5	2.1	Oscillograph	9"	7000	1450	− 50	250	—	3000	10	0.073	0.093	P1	Green	914
1800[10]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9"	6000	1250	− 75	250	25	—	10	—	—	P3	Yellow	1800
1801[10]	Electromagnetic Kinescope	Fig. 13[11]	2.5	2.1	Television	5"	3000	450	− 35	—	20	—	10	—	—	P3	Yellow	1801
2001	Electrostatic Cathode-Ray	Fig. 2[11]	6.3	0.6	Oscillograph	1"	Characteristics essentially same as 913											2001
2002	Electrostatic Cathode-Ray	Fig. 1[11]	6.3	0.6	Oscillograph	2"	600	120	—	—	—	—	—	0.16	0.17	P1	Green	2002
2005	Electrostatic Cathode-Ray	Fig. 1[11,8]	2.5	2.1	Television	5"	2000	1000	− 35	200	—	—	10	0.5	0.56	—	—	2005
24-XH	Electrostatic Cathode-Ray	Fig. 1[11]	6.3	0.6	Oscilloscope	2"	600	120	− 60	—	—	—	10	0.14	0.16	P5	Blue	24-XH

[1] Refer to Receiving Tube Diagrams.
[2] For current cut-off. In terms of average center values; should be adjustable to ± 50 per cent to take care of individual tubes. Control grid should never be allowed to go positive.
[3] Between Anode No. 2 and any deflecting plate.
[4] In mw./sq. cm., max. [5] In mm./volt d.c.
[6] Phosphorescent material used in screen determines persistence.

P1 is phosphor of medium persistence, P2 long, P3 also medium but especially suited for television, P4 same as P3 but white, and P5 short persistence for oscillographic use.
[7] The 911 is identical to 906 except for the gun material, which is designed to be especially free from magnetization effects.
[8] Cathode connected to pin 7.
[10] Obsolete type.

[11] See Supplementary Base Diagrams.
[12] Intensifier (Anode No. 3) voltage = 4000.
[13] Intensifier (Anode No. 3) voltage = 3000.
[14] Intensifier (Anode No. 3) voltage = 2000.
[15] Also available in P4 and P5.
[16] Also available in P2, P4 and P5.

TABLE XIII—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XI—Control and Regulator Tubes

Type No.	Name	Base [2]	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Max. A.C. Voltage Per Plate	Max. D.C. Output Current Ma.	Max. Inverse Peak Voltage	Max. Peak Plate Current Ma.	Type [7]
BA	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	350	Tube drop 80 v.		G
BH	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	125	Tube drop 90 v.		G
BR	Half-Wave Rectifier	4-pin M.	4J	Cold	—	—	300	50	Tube drop 60 v.		G
CE-220	Half-Wave Rectifier	4-pin-M.	4P	Fil.	2.5	3.0	—	20	20000	100	V
OZ4	Full-Wave Rectifier	6-pin O.	4R	Cold	—	—	350	30–75	1250	200	G
1 [5]	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	1000	400	M
1-V [5]	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	—	—	V
2V3G	Half-Wave Rectifier	6-pin O.	6BA	Fil.	2.5	5.0	—	2.0	16500	12	V
2W3	Half-Wave Rectifier	5-pin O.	4X	Fil.	2.5	1.5	350	55	—	—	V
2X2/879	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4500 [11]	7.5	—	—	V
2Y2	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	1.75	4400 [11]	5.0	—	—	V
2Z2/G84	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	1.5	350	50	—	—	V
3B24	Half-Wave Rectifier	4-pin M.	T-4A [4]	Fil. [21]	5.0	3.0	—	60	20000	300	V
					2.5	3.0	—	30	20000	150	
3B25	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	500	4000	2000	G
5R4GY	Full-Wave Rectifier	5-pin M.	5T	Fil.	5.0	2.0	900 [10] 950 [18]	150 [10] 175 [18]	2800	650	V
5T4 [3]	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	3.0	450	250	1250	800	V
5U4G	Full-Wave Rectifier	8-pin O.	5T	Fil.	5.0	3.0	Same as Type 5Z3				V
5V4G	Full-Wave Rectifier	8-pin O.	5L	Htr.	5.0	2.0	Same as Type 83V				V
5W4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	1.5	350	110	1000	—	V
5X3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	1275	30	—	—	V
5X4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	3.0	Same as 5Z3				V
5Y3G	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				V
5Y4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	2.0	Same as Type 80				V
5Z3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	—	V
5Z4 [3]	Full-Wave Rectifier	5-pin O.	5L	Htr.	5.0	2.0	400	125	1100	—	V
6W5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.9	350	100	1250	350	V
6X5 [3]	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.5	350	75	—	—	V
6Y3	Half-Wave Rectifier	6-pin O.	4AC	Fil.	6.3	0.7	5000	7.5	14000	100	V
6Y5	Full-Wave Rectifier	6-pin S.	6J	Htr.	6.3	0.8	350	50	—	—	V
6Z3	Half-Wave Rectifier	4-pin M.	4G	Fil.	6.3	0.3	350	50	—	—	V
6Z5	Full-Wave Rectifier	6-pin S.	6K	Htr.	6.3	0.6	230	60	—	—	V
6ZY5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.3	350	35	1000	150	V
7Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	7.0 [12]	0.53	350	60	—	—	V
7Z4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	7.0 [12]	0.96	450 [6] 325 [10]	100	1250	300	V
12A7	Rectifier-Pentode [14]	7-pin S.	7K	Htr.	12.6	0.3	125	30	—	—	V
12Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	V
12Z5	Voltage-Doubling Rectifier	7-pin M.	7L	Htr.	12.6	0.3	225	60	—	—	V
14Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	14 [12]	0.32	450 [6] 325 [10]	70	1250	210	V
14Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	14 [12]	0.3	250	60	—	—	V
25A7G	Rectifier-Pentode [14]	8-pin O.	8F	Htr.	25	0.3	125	75	—	—	V
25X6GT	Voltage-Doubling Rectifier	7-pin O.	7Q	Htr.	25	0.15	125	60	—	—	V
25Y4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.15	125	75	—	—	V
25Y5	Voltage-Doubling Rectifier	6-pin S.	6E	Htr.	25	0.3	250	85	—	—	V
25Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	25	0.3	250	50	—	—	V
25Z4	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.3	125	125	—	—	V
25Z5	Rectifier-Doubler	6-pin S.	6E	Htr.	25	0.3	125	100	—	500	V
25Z6	Rectifier-Doubler	7-pin O.	7Q	Htr.	25	0.3	125	100	—	500	V
28Z5	Full-Wave Rectifier	8-pin L.	5AB	Htr.	28	0.24	450 [18] 325 [10]	100	—	300	V
32L7GT	Rectifier-Tetrode [14]	8-pin O.	8F	Htr.	32.5	0.3	125	60	—	—	V
35Y4	Half-Wave Rectifier	8-pin O.	5AL	Htr.	35 [8]	0.15	235	60 100 [19]	700	600	V
35Z3 [23]	Half-Wave Rectifier	8-pin L.	4Z	Htr.	35	0.15	250 [13]	100	700	600	V
35Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	35	0.15	250	100	—	—	V
35Z5G	Half-Wave Rectifier	6-pin O.	6AD	Htr.	35 [8]	0.15	125	60 100 [19]	—	—	V
35Z6G	Voltage Doubler	6-pin O.	7Q	Htr.	35	0.3	125	110	—	500	V
40Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	40 [8]	0.15	125	60 100 [19]	—	—	V
45Z3	Half-Wave Rectifier	7-pin B.	5AM	Htr.	45	0.075	117	65	350	390	V
45Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	45 [8]	0.15	125	60 100 [19]	—	—	V
50Y6GT	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.15	125	85	—	—	V

TABLE XIII — RECTIFIERS — RECEIVING AND TRANSMITTING — *Continued*

See also Table XI — Control and Regulator Tubes

Type No.	Name	Base [2]	Socket Connections [1]	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Max. A.C. Voltage Per Plate	Max. D.C. Output Current Ma.	Max. Inverse Peak Voltage	Max. Peak Plate Current Ma.	Type [7]
50Z6G	Voltage-Doubling Rectifier	7-pin O.	7Q	Htr.	50	0.3	125	150	—	—	V
50Z7G	Voltage-Doubling Rectifier	8-pin O.	8AN	Htr.	50	0.15	117	65	—	—	V
70A7GT	Rectifier-Tetrode [14]	8-pin O.	8AB	Htr.	70	0.15	125	60	—	—	V
70L7GT	Rectifier-Tetrode [14]	8-pin O.	8AA	Htr.	70	0.15	117	70	—	350	V
72	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	30	20000	150	V
73	Half-Wave Rectifier Clipper Tube	8-pin O.	4Y	Fil.	2.5	4.5	—	20	13000	3000	V
80	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	350 400 550	125 110 135	—	—	V
81	Half-Wave Rectifier	4-pin M.	4B	Fil.	7.5	1.25	700	85	—	—	V
82	Full-Wave Rectifier	4-pin M.	4C	Fil.	2.5	3.0	500	125	1400	400	M
83	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	800	M
83-V	Full-Wave Rectifier	4-pin M.	4AD	Htr.	5.0	2.0	400	200	1100	—	V
84/6Z4	Full-Wave Rectifier	5-pin S.	5D	Htr.	6.3	0.5	350	60	1000	—	V
117L7GT/ 117M7GT	Rectifier-Tetrode [14]	8-pin O.	8AO	Htr.	117	0.09	117	75	—	—	V
117N7GT	Rectifier-Tetrode [14]	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	V
117P7GT	Rectifier-Tetrode [14]	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	V
117Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	117	0.04	117	90	350	—	V
117Z6GT	Voltage-Doubling Rectifier	7-pin O.	7Q	Htr.	117	0.075	235	60	700	360	V
217-A	Half-Wave Rectifier	4-pin J.	T-3A [4]	Fil.	10	3.25	—	—	3500	600	V
217-C	Half-Wave Rectifier	4-pin J.	T-3A [4]	Fil.	10	3.25	—	—	7500	600	V
Z225 [17]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250 [10]	10000	1000	M
HK253	Half-Wave Rectifier	4-pin J.	T-3A [4]	Fil.	5.0	10	—	350	10000	1500	V
705A RK-705A [20]	Half-Wave Rectifier	4-pin W	T-3AA [4]	Fil. [21]	2.5 [26] 5.0	5.0 5.0	— —	50 100	35000 35000	375 750	V
816	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	2.0	1750	125	5000	500	M
836	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	5.0	—	—	5000	1000	V
866A/866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250 [10]	10000	1000	M
866B	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.0	—	—	8500	1000	M
866Jr.	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	2.5	1250	250 [9]	—	—	M
HY866Jr.	Half-Wave Rectifier	4-pin M.	4P	Fil. [22]	2.5	2.5	1750	250 [9]	5000	—	M
RK866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250 [10]	10000	1000	M
871 [24]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.0	1750	250	5000	500	M
878 [11]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	7100	5	20000	—	V
879 [11]	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	1.75	2650	7.5	7500	100	V
872A/872	Half-Wave Rectifier	4-pin J.	T-3A [4]	Fil.	5.0	7.5	—	1250	10000	5000	M
975A	Half-Wave Rectifier	4-pin J.	T-3A [4]	Fil.	5.0	10.0	—	1500	15000	6000	M
1616	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	130	6000	800	V
8008	Half-Wave Rectifier	4-pin [18]	Fig. 11 [15]	Fil.	5.0	7.5	—	1250	10000	5000	M
8013A [25]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	20	40000	150	V
8016	Half-Wave Rectifier	6-pin O.	4Y	Fil.	1.25	0.2	—	2.0	10000	7.5	V
8020	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0 5.8	5.5 6.5	10000 12500	100 100	40000 40000	750 750	V
RK19	Full-Wave Rectifier	4-pin M.	T-3A [4]	Htr.	7.5	2.5	1250	200 [10]	3500	600	V
RK21	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	4.0	1250	200 [10]	3500	600	V
RK22	Full-Wave Rectifier	4-pin M.	T-4AG [4]	Htr.	2.5	8.0	1250	200 [10]	3500	600	V
RK60	Full-Wave Rectifier	4-pin M.	T-4AG [4]	Fil.	5	3.0	750	250	2120	—	V

[1] Refer to Receiving Tube Diagrams.
[2] M.—medium; S.—small; O.—octal; L.—loktal; J.—jumbo B.—button. W,—wafer. [3] Metal tube series.
[4] Refer to Transmitting Tube Diagrams.
[5] Types 1 and 1-V interchangeable.
[6] With input choke of at least 20 henrys.
[7] M.—Mercury-vapor type; V.—high vacuum type; G.—Gaseous Type. [8] Tapped for pilot lamps.
[9] Per pair with choke input. [10] Condenser input.
[11] For use with cathode-ray tubes.
[12] Maximum rating, corresponding to 130-volt line condition; normal rating is 12.6 v. for 117-v. line.
[13] With 100 ohms min. resistance in series with plate; without series resistor, maximum r.m.s. plate rating is 117 volts.
[14] For other data, see Table IX.
[15] See Supplementary Base Diagrams.
[16] Same as 872A/872 except for heavy-duty push-type base. Fil. connected to pins 2 and 3, plate to top cap.
[17] Same as 872A/872 except for small envelope.
[18] Choke input. [19] Without panel lamp. [20] Ceramic Base.
[21] Center tapped. [22] Formerly heater type. [23] Formerly type LT.
[24] Obsolete. [25] Formerly 8013.
[26] Using only one-half of filament.

TABLE XIV — TRIODE TRANSMITTING TUBES

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [3]	Approx. Carrier Output Power Watts	Type
958-A****	0.6	1.25	0.1	135	7	1.0	12	0.6	2.6	0.8	Special	5BD	Class-C Amp.-Oscillator	135	— 20	7	1.0	0.035	0.6	958-A
RK24**	1.5	2.0	0.12	180	20	6.0	8.0	3.5	5.5	3.0	4-pin S.	4D	Class-C Amp.-Oscillator	180	— 45	16.5	6.0	0.5	2.0	RK24
6J6[7]***	1.5	6.3	0.45	300	30	16	32	2.2	1.6	0.4	7-pin B.	7BF[15]	Class-C Amp. (Telegraphy)	150	— 10	30	16	0.35	3.5	6J6
9002***	1.6	6.3	0.15	250	8	2	25	1.2	1.4	1.1	7-pin B.	7TM	Class-C Amp.-Oscillator	180	— 35	7	1.5	—	0.5	9002
955***	1.6	6.3	0.15	180	8	2	25	1.0	1.4	0.6	Acorn	5BC	Class-C Amp. (Plate Mod. or Telegraphy)-Osc.	180	— 35	7	1.5	—	0.5	955
HY114B***	1.8	1.4	0.155	180	12	3.0	13	1.0	1.3	1.0	5-pin O.	T-8AC	Class-C Amp.-Oscillator	180	— 30	12	2.0	0.2	1.4 [11]	HY114B
													Class-C Amp. Plate-Mod.	180	— 35	12	2.5	0.3	1.4[11]	
3A5 [7]	2.0	1.4 / 2.8	0.22 / 0.11	135	30	5.0	15	0.9	3.2	1.0	7-pin B.	7BC [15]	Class-C Amp.-Oscillator	135	— 20	30	5.0	0.2	2.0[21]	3A5
6F4****	2.0	6.3	0.225	150	20	8	17	2.0	1.9	0.6	Acorn	7BR	Class-C Amp.-Oscillator	150	— 15 / 550[18] / 2000[19]	20	7.5	0.2	1.8[20]	6F4
HY24* [5]	2.0	2.0	0.13	180	20	4.5	9.3	2.7	5.4	2.3	4-pin S.	4D	Class-C Amp. (Telegraphy)	180	— 45	20	4.5	0.2	2.7	HY24
													Class-C Amp. (Telephony)	180	— 45	20	4.5	0.3	2.5	
RK33* [4] [7]	2.5	2.0	0.12	250	20	6.0	10.5	3–2 [7]	3–2 [7]	2.5	7-pin S.	T-7DA	Class-C Amp.-Oscillator	250	— 60	20	6.0	0.54	3.5	RK33
2C22 7193	3.5	6.3	0.3	500	—	—	20	2.2 [18]	3.6	0.7	8-pin O.	4AM [15]	Class-C Amp. (Telegraphy)	—	—	—	—	—	—	2C22/7193
HY615*** HY-E1148	3.5	6.3	0.175	300	20	4.0	20	1.4	1.6	1.2	5-pin O.	T-8AG	Class-C Amp.-Oscillator	300	— 35	20	2.0	0.4	4.0 [11]	HY615
													Class-C Amp. Plate-Mod.	300	— 35	20	3.0	0.8	3.5 [11]	HY-E1148
HY6J5GTX*	3.5	6.3	0.3	250	20	4.0	20	3.8	2.7	3.0	6-pin O.	T-8AD	Class-C Amp. (Telegraphy)	250	— 30	20	2.0	0.2	3	HY6J5GTX
													Class-C Amp. (Telephony)	250	— 30	20	2.5	0.4	3	
GL-446A**** GL-446B****	3.75	6.3	0.75	400[23]	20	—	45	2.2	1.6	0.02	6-pin O.	Fig. 19	Class-C Amp.-Oscillator	250	10000[19] / 20000[19]	25[23]	—	—	—	GL-446 / GL-446B
GL-2C44**** GL-464A****	5.0	6.3	0.75	500[23]	40[23]	—	—	2.7	2.0	0.1	6-pin O.	Fig. 17	Class-C Amp.-Oscillator	250	—	—	—	—	—	GL-2C44 / GL-464A
6C4	5.0	6.3	0.15	300	25	8.0	17	1.8	1.6	1.3	7-pin B.	6BG [15]	Class-C Amp.-Oscillator	300	— 27	25	7.0	0.35	5.5	6C4
1626	5.0	12.6	0.25	250	25	8.0	5.0	3.2	4.4	3.4	8-pin O.	T-8AD	Class-C Amp.-Oscillator	250	— 70	25	5.0	0.5	4.0	1626
2C26***	10	6.3	1.15	350	—	—	16.3	2.6	2.8	1.1	8-pin O.	4BB	Special Oscillator	350	— 15	16	—	—	—	2C26
2C45	10	7.0	1.18	250	40	0	3.6	5.0	7.7	3.0	4-pin M.	4D	Class-A Modulator	250	— 40	29	0	0	1.0	2C45
RK34 *** [7]	10	6.3	0.8	300	80	20	13	4.2	2.7	0.8	7-pin M.	T-7DC	Class-C Amp.-Oscillator	300	— 36	80	20	1.8	16	RK34
205D	14	4.5	1.6	400	50	10	7.2	5.2	4.8	3.3	4-pin M.	4D	Class-C Amp.-Oscillator	400	—112	45	10	1.5	10	205D
													Class-C Amp. (Plate-Mod.)	350	—144	35	10	1.7	7.1	
2C25	15	7.0	1.18	450	60	15	8.0	6.0	8.9	3.0	4-pin M.	4D	Class-C Amp.-Oscillator	450	—100	65	15	3.2	19	2C25
													Class-C Amp. Plate-Mod.	350	—100	50	12	2.2	12	
10Y	15	7.5	1.25	450	65	15	8	4.1	7.0	3.0	4-pin M.	4D	Class-C Amp.-Oscillator	450	—100	65	15	3.2	19	10Y
													Class-C Amp. Plate-Mod.	350	—100	50	12	2.2	12	
843	15	2.5	2.5	450	40	7.5	7.7	4.0	4.5	4.0	5-pin M.	5A	Class-C Amp.-Oscillator	450	—140	30	5.0	1.0	7.5	843
													Class-C Amp. (Plate-Mod.)	350	—150	30	7.0	1.6	5.0	
RK59 [7]	15	6.3	1.0	500	90	25	25	5.0	9.0	1.0	4-pin M.	T-4D	Class-C Amp.-Oscillator	500	— 60	90	14	1.3	32	RK59
HY75 * [5]	15	6.3	2.5	450	80	20	10	1.6	3.8	0.6	5-pin O.	T-8AC	Class-C Amp.-Oscillator	450	— 50	80	12	—	21 [11]	HY75
													Class-C Amp. Plate-Mod.	450	— 60	80	12	—	16[11]	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
1602	15	7.5	1.25	450	60	15	8.0	4.0	7.0	3.0	4-pin M.	4D	Class-C Amp. (Telegraphy)	450	−115	55	15	3.3	13	1602
													Class-C Amp. (Telephony)	350	−135	45	15	3.5	8.0	
841	15	7.5	1.25	450	60	20	30	4.0	7.0	3.0	4-pin M.	4D	Class-C Amp. (Telegraphy)	450	− 34	50	15	1.8	15	841
													Class-C Amp. (Telephony)	350	− 47	50	15	2.0	11	
10 RK10 *[4]	15	7.5	1.25	450	65	15	8.0	3.0	8.0	4.0	4-pin M.	4D	Class-C Amp. (Telegraphy)	450	−100	65	15	3.2	19	10 RK10
													Class-C Amp. (Telephony)	350	−100	50	12	2.2	12	
RK100 [4]	15	6.3	0.9	150	250	100	40	23	19	3.0	6-pin M.	T-6B	Class-C Oscillator [10]	110	—	80	8.0	—	3.5	RK100
													Class-C Amplifier [10]	110	—	185	40	2.1	12	
1608	20	2.5	2.5	425	95	25	20	8.5	9.0	3.0	4-pin M.	4D	Class-C Amp. (Telegraphy)	425	− 90	95	20	3.0	27	1608
													Class-C Amp. (Telephony)	350	− 80	85	20	3.0	18	
310	20	7.5	1.25	600	70	15	8.0	4.0	7.0	2.2	4-pin M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	25	310
													Class-C Amp. (Telephony)	500	−190	55	15	4.5	18	
801-A/801 *	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	4-pin M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	25	801-A/801
													Class-C Amp. (Telephony)	500	−190	55	15	4.5	18	
HY801-A*	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	4-pin M.	4D	Class-C Amp. (Telegraphy)	600	−200	70	15	4.0	30	HY801-A
													Class-C Amp. (Telephony)	500	−200	60	15	4.5	22	
T20 *[6]	20	7.5	1.75	750	85	25	20	4.9	5.1	0.7	4-pin M.	3G	Class-C Amp. (Telegraphy)	750	− 85	85	18	3.6	44	T20
													Class-C Amp. Plate-Mod.	750	−140	70	15	3.6	38	
TZ20 *[6]	20	7.5	1.75	750	85	30	62	5.3	5.0	0.6	4-pin M.	3G	Class-C Amp. (Telegraphy)	750	− 40	85	28	3.75	44	TZ20
													Class-C Amp. Plate-Mod.	750	−100	70	23	4.8	38	
15E	20	5.0	4.0	12500	—	—	25	1.4	1.15	0.3	None	T-4AF	Oscillator at 400 Mc.	Approximately 15 watts output						15E
25T***	25	5.0	3.0	2000	75	25	24	2.7	1.5	0.3	4-pin M.	3G	Class-C Amp.-Oscillator	2000	−130	63	18	4.0	100	25T
														1500	− 95	67	13	2.2	75	
														1000	− 70	72	9	1.3	47	
3C24***	25	6.3	3.0	2000	75	25	23	1.7	1.5	0.3	4-pin S.	2D	Class-C Amp.-Oscillator	2000	−170	63	17	4.5	100	3C24
														1500	−110	67	15	3.1	75	
														1000	− 80	72	15	2.6	47	
RK11 *[4]	25	6.3	3.0	750	105	35	20	7.0	7.0	0.9	4-pin M.	3G	Class-C Amp. (Telegraphy)	750	−120	105	21	3.2	55	RK11
													Class-C Amp. Plate-Mod.	600	−120	85	24	3.7	38	
RK12 *	25	6.3	3.0	750	105	40	100	7.0	7.0	0.9	4-pin M.	3G	Class-C Amp. (Telegraphy)	750	−100	105	35	5.2	55	RK12
													Class-C Amp. Plate-Mod.	600	−100	85	27	3.8	38	
HK24 *	25	6.3	3.0	2000	75	30	25	2.5	1.7	0.4	4-pin S.	3G	Class-C Amp. (Telegraphy)	2000	−140	56	18	4.0	90	HK24
													Class-C Amp. Plate-Mod.	1500	−145	50	25	5.5	60	
HY25 *	25	7.5	2.25	800	75	25	55	4.2	4.6	1.0	4-pin M.	3G	Class-C Amp. (Telegraphy)	750	− 45	75	15	2.0	42	HY25
													Class-C Amp. Plate-Mod.	700	− 45	75	17	5.0	39	
8025	30	6.3	1.92	1000	65	—	18	2.7	2.8	0.35	4-pin M.	4AQ	Class-C Amp. (Grid. Mod.)	1000	−135	50	4[17]	3.5[17]	20	8025
	20				65	20							Class-C Amp. (Plate Mod.)	800	−105	40	10.5[17]	1.4[17]	22	
	30				80	20							Class-C Amp. (Telegraphy)	1000	− 90	50	14[17]	1.6[17]	35	
Twin 30*[5,7]	30	6.0	4.0	1500	85	25	32	1.9	2.0	0.3	4-pin M.	T-4DB	Class-C Amp. (Telegraphy)	1500	−100	150 [8]	40 [8]	15	225	Twin 30
													Class-C Amp. Plate-Mod.	1250	−100	135 [8]	40 [8]	15	125	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
HY30Z *	30	6.3	2.25	850	90	25	87	6.0	4.9	1.0	4-pin M.	T-4BE	Class-C Amp.-Oscillator	850	−75	90	25	2.5	58	HY30Z
													Class-C Amp. Plate-Mod.	700	−75	90	25	3.5	47	
HY31Z *[7] HY1231Z *[5,7]	30	6.3 12.6	3.5 1.7	500	150	30	45	5.0	5.5	1.9	4-pin M.	T-4D	Class-C Amp. (Telegraphy)	500	−45	150	25	2.5	56	HY31Z HY1231Z
													Class-C Amp. (Telephony)	400	−100	150	30	3.5	45	
316A ****	30	2.0	3.65	450	80	12	6.5	1.2	1.6	0.8	None [9]	—	Class-C Amp. (Telegraphy)	450	—	80	12	—	7.5	316A
													Class-C Amp. Plate-Mod.	400	—	80	12	—	6.5	
809 *[6]	30	6.3	2.5	1000	125	—	50	5.7	6.7	0.9	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	1000	−75	100	25	3.8	75	809
													Class-C Amp. Plate-Mod.	750	−60	100	32	4.3	55	
1623 *[6]	30	6.3	2.5	1000	100	25	20	5.7	6.7	0.9	4-pin M.	3G [15]	Class-C Amp.-Oscillator	1000	−90	100	20	3.1	75	1623
													Class-C Amp. Plate-Mod.	750	−125	100	20	4.0	55	
53A	35	5.0	12.5	15000	—	—	35	3.6	1.9	0.4	None	T-4B	Oscillator at 300 Mc.	Approximately 50 watts output						53A
RK30 *[4]	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	4-pin M.	T-4BC	Class-C Amp. (Telegraphy)	1250	−180	90	18	5.2	85	RK30
													Class-C Amp. Plate-Mod.	1000	−200	80	15	4.5	60	
800 *	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	4-pin M.	T-4BC	Class-C Amp. (Telegraphy)	1250	−175	70	15	4.0	65	800
													Class-C Amp. Plate-Mod.	1000	−200	70	15	4.0	50	
1628 ****[4]	40	3.5	3.25	1000	60	15	23	2.0	2.0	0.4	None [9]	T-4BB	Class-C Amp.-Oscillator	1000	−65	50	15	1.7	35	1628
													Class-C Amp. Plate-Mod.	800	−100	40	11	1.6	22	
													Grid-Modulated Amp.	1000	−120	50	3.5	5.0	20	
8012 ****	40 [14]	6.3	2.0	1000	80	20	18	2.7	2.8	0.35	None [9]	T-4BB	Class-C Amp.-Oscillator	1000	−90	50	14	1.6	35	8012
													Class-C Amp. Plate-Mod.	800	−105	40	10.5	1.4	22	
													Grid-Modulated Amp.	1000	−135	50	4.0	3.5	20	
RK18 *[4]	40	7.5	3.0	1250	100	40	18	6.0	4.8	1.8	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	1250	−160	100	12	2.8	95	RK18
													Class-C Amp. Plate-Mod.	1000	−160	80	13	3.1	64	
RK31	40	7.5	3.0	1250	100	35	170	7.0	1.0	2.0	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	1250	−80	100	30	3.0	90	RK31
													Class-C Amp. Plate-Mod.	1000	−80	100	28	3.5	70	
HY40 *	40	7.5	2.25	1000	125	25	25	6.1	5.6	1.0	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	1000	−90	125	20	5.0	94	HY40
													Class-C Amp. Plate-Mod.	850	−90	125	15	3.5	82	
													Grid-Modulated Amp.	1000	—	125	—	—	20 [12]	
HY40Z *	40	7.5	2.6	1000	125	30	80	6.2	6.3	0.8	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	1000	−27	125	25	5.0	94	HY40Z
													Class-C Amp. Plate-Mod.	850	−30	100	30	7.0	82	
													Grid-Modulated Amp.	1000	—	60	—	—	20 [12]	
T40 *[6]	40	7.5	2.5	1500	150	40	25	4.5	4.8	0.8	4-pin M.	3G [15]	Class-C Amp.-Oscillator	1500	−140	150	28	9.0	158	T40
													Class-C Amp. Plate-Mod.	1250	−115	115	20	5.25	104	
TZ40 *[6]	40	7.5	2.5	1500	150	45	62	4.8	5.0	0.8	4-pin M.	3G [15]	Class-C Amp.-Oscillator	1500	−90	150	38	10	165	TZ40
													Class-C Amp. Plate-Mod.	1250	−100	125	30	7.5	116	
HY57 *	40	6.3	2.25	850	110	25	50	4.9	5.1	1.7	4-pin M.	3G [15]	Class-C Amp. (Telegraphy)	850	−48	110	15	2.5	70	HY57
													Class-C Amp. Plate-Mod.	700	−45	90	17	5.0	47	
													Grid-Modulated Amp.	850	—	70	—	—	20 [12]	
756 [4]	40	7.5	2.0	850	110	25	8.0	3.0	7.0	2.7	4-pin M.	4D	Class-C Amplifier	850	—	110	25	—	—	756

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
830 [4]	40	10	2.15	750	110	18	8.0	4.9	9.9	2.2	4-pin M.	4D	Class-C Amplifier	750	−180	110	18	7.0	55	830
													Grid-Modulated Amp.	1000	−200	50	2.0	3.0	15	
8010-R ****	50	6.3	2.4	1350	150	20	30	2.3	1.5	0.07	Special	—	Class-C Amplifier	—	—	—	—	—	—	8010-R
RK32 ** [4]	50	7.5	3.25	1250	100	25	11	2.5	3.4	0.7	4-pin M.	2D	Class-C Amp. (Telegraphy)	1250	−225	100	14	4.8	90	RK32
													Class-C Amp. Plate-Mod.	1000	−310	100	21	8.7	70	
RK35 * [4]	50	7.5	4.0	1500	125	20	9.0	3.5	2.7	0.4	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−250	115	15	5.0	120	RK35
													Class-C Amp. Plate-Mod.	1250	−250	100	14	4.6	93	
													Grid-Modulated Amp.	1500	−180	37	—	2.0	25	
RK37 *	50	7.5	4.0	1500	125	35	28	3.5	3.2	0.2	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−130	115	30	7.0	122	RK37
													Class-C Amp. Plate-Mod.	1250	−150	100	23	5.6	90	
													Grid-Modulated Amp.	1500	− 50	50	—	2.4	26	
UH50 *	50	7.5	3.25	1250	125	25	10.6	2.2	2.6	0.3	4-pin M.	2D	Class-C Amp. (Telegraphy)	1250	−225	125	20	7.5	115	UH50
													Class-C Amp. Plate-Mod.	1250	−325	125	20	10	115	
													Grid-Modulated Amp.	1250	−200	60	2.0	3.0	25	
UH51 *	50	5.0	6.5	2000	175	25	10.6	2.2	2.3	0.3	4-pin M.	2D	Class-C Amp. (Telegraphy)	2000	−500	150	20	15	225	UH51
													Class-C Amp. Plate-Mod.	1500	−400	165	20	15	200	
													Grid-Modulated Amp.	1500	−400	85	2.0	8.0	65	
HK54 *	50	5.0	5.0	3000	150	30	27	1.9	1.9	0.2	4-pin M.	2D	Class-C Amp. (Telegraphy)	3000	−290	100	25	10	250	HK54
													Class-C Amp. Plate-Mod.	2500	−250	100	20	8.0	210	
													Grid-Modulated Amp.	2000	−150	39	1.5	3.0	28	
HK154 [4]	50	5.0	6.5	1500	175	30	6.7	4.3	5.9	1.1	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−590	167	20	15	200	HK154
													Class-C Amp. Plate-Mod.	1250	−460	170	20	12	162	
													Grid-Modulated Amp.	1500	−450	52	—	5.0	28	
HK158 *	50	12.6	2.5	2000	200	40	25	4.7	4.6	1.0	4-pin M.	2D	Class-C Amp.-Oscillator	2000	−150	125	25	6.0	200	HK158
													Class-C Amp. Plate-Mod.	2000	−140	105	25	5.0	170	
WE304A * [4] 304B *	50	7.5	3.25	1250	100	25	11	2.0	2.5	0.7	4-pin M.	2D	Class-C Amp. (Telegraphy)	1250	−200	100	—	—	85	WE304A 304B
													Class-C Amp. Plate-Mod.	1000	−180	100	—	—	65	
356A *	50	5.0	5.0	1500	120	35	50	2.25	2.75	1.0	Special	T-4BD	Class-C Amp. (Telegraphy)	1500	− 60	100	—	—	100	356A
													Class-C Amp. Plate-Mod.	1250	−100	100	35	—	85	
808	50	7.5	4.0	1500	150	35	47	5.3	2.8	0.15	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−200	125	30	9.5	140	808
													Class-C Amp. Plate-Mod.	1250	−225	100	32	10.5	105	
834 *	50	7.5	3.1	1250	100	20	10.5	2.2	2.6	0.6	4-pin M.	2D	Class-C Amp. (Telegraphy)	1250	−225	90	15	4.5	75	834
													Class-C Amp. Plate-Mod.	1000	−310	90	17.5	6.5	58	
841A [4]	50	10	2.0	1250	150	30	14.6	3.5	9.0	2.5	4-pin M.	3G	Class-C Amplifier	—	—	—	—	—	85	841A
841SW	50	10	2.0	1000	150	30	14.6	—	9.0	—	4-pin M.	3G	Class-C Amplifier	—	—	—	—	—	—	841SW
T55 **	55	7.5	3.0	1500	150	40	20	5.0	3.9	1.2	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	18	6.0	170	T55
													Class-C Amp. Plate-Mod.	1500	−195	125	15	5.0	145	
811 * [5]	55	6.3	4.0	1500	150	50	160	5.5	5.5	0.6	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−113	150	35	8.0	170	811
													Class-C Amp. Plate-Mod.	1250	−125	125	50	11	120	
812 * [5]	55	6.3	4.5	1500	150	35	29	5.3	5.3	0.8	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−175	150	25	6.5	170	812
													Class-C Amp. Plate-Mod.	1250	−125	125	25	6.0	120	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
		Volts	Amps.					Grid to Fil.	Grid to Plate	Plate to Fil.										
RK51 *	60	7.5	3.75	1500	150	40	20	6.0	6.0	2.5	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−250	150	31	10	170	RK51
													Class-C Amp. Plate-Mod.	1250	−200	105	17	4.5	96	
													Grid-Modulated Amp.	1500	−130	60	0.4	2.3	128	
RK52*	60	7.5	3.75	1500	130	50	170	6.6	12	2.2	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−120	130	40	7.0	135	RK52
													Class-C Amp. Plate-Mod.	1250	−120	115	47	8.5	102	
HF60	60	10	2.5	1600	150	—	20	—	5.2	—	4-pin M.	2D	Class-C Amp.-Oscillator	1600	−150	—	—	—	100	HF60
826***	60	7.5	4.0	1000	125	40	31	3.7	2.9	1.4	Special	T-9A	Class-C Amp.-Oscillator	1000	− 70	125	35	5.8	86	826
													Class-C Amp. Plate-Mod.	800	− 98	94	35	6.2	53	
													Class-B Amp. (Telephony)	1000	− 50	65	8.5	3.7	22	
													Grid-Modulated Amp.	1000	−125	65	9.5	8.2	25	
830B 930B	60	10	2.0	1000	150	30	25	5.0	11	1.8	4-pin M.	3G	Class-C Amp.-Oscillator	1000	−110	140	30	7.0	90	830B 930B
													Class-C Amp. Plate-Mod.	800	−150	95	20	5.0	50	
													Class-B Amp. (Telephony)	1000	− 35	85	6.0	6.0	26	
HY51A * HY51B *	65	7.5 10	3.5 2.25	1000	175	25	25	6.5	7.0	1.1	4-pin M.	3G	Class-C Amp. (Telegraphy)	1000	− 75	175	20	7.5	131	HY51A HY51B
													Class-C Amp. Plate-Mod.	1000	−67.5	130	15	7.5	104	
													Grid-Modulated Amp.	1000	—	100	—	—	33 [12]	
HY51Z *	65	7.5	3.5	1000	175	35	85	7.9	7.2	0.9	4-pin M.	T-4BE	Class-C Amp. (Telegraphy)	1000	−22.5	175	35	10	131	HY51Z
													Class-C Amp. Plate-Mod.	1000	− 30	150	35	10	104	
													Grid-Modulated Amp.	1000	—	100	—	—	33 [12]	
UH35 * [6]	70	5.0	4.0	1500	150	35	30	1.4	1.6	0.2	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	30	7.0	170	UH35
													Class-C Amp. Plate-Mod.	1500	−120	100	30	5.0	120	
35T * 35TG	70	5.0	4.0	2000	150	35	30	3.8 1.9	1.9	0.2	4-pin M.	3G 2D	Class-C Amp. (Telegraphy)	1500	−120	150	30	7.0	170	35T 35TG
													Class-C Amp. Plate-Mod.	1500	−120	100	30	5.0	120	
													Grid-Modulated Amp.	2000	−400	60	3.0	3.0	50	
V70 V70B	70	10	2.5	1500	140	25	14	5.0	9.0	2.3	4-pin J. 4-pin M.	T-3AB 3G	Class-C Amp. (Telegraphy)	1500	−215	130	6.0	3.0	140	V70 V70B
													Class-C Amp. Plate-Mod.	1250	−250	130	6.0	3.0	120	
V70A V70C	70	10	2.5	1500	140	20	25	5.0	9.5	2.0	4-pin J. 4-pin M.	T-3AB 3G	Class-C Amp. (Telegraphy)	1000	−110	140	30	7.0	90	V70A V70C
													Class-C Amp. Plate-Mod.	800	−150	95	20	5.0	50	
V70D	70	10	3.0	1500	165	40	20	4.5	4.5	1.75	4-pin M.	3G	Class-C Amp. (Telegraphy)	1500	−200	130	20	6.0	140	V70D
													Class-C Amp. Plate-Mod.	1000	−140	165	30	7.0	120	
50T [4]	75	5.0	6.0	3000	100	30	12	2.0	2.0	0.4	4-pin M.	2D	Class-C Amplifier	3000	−600	100	25	—	250	50T
75T *	75	5.0	6.5	3000	175	30	10.6	2.2	2.3	0.3	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−300	175	30	10	200	75T
													Class-C Amp. Plate-Mod.	1500	−300	175	30	10	200	
													Grid-Modulated Amp.	1500	−400	85	2.0	8.0	65	
HF75*	75	10	3.25	2000	120	—	12.5	—	2.0	—	4-pin M.	2D	Class-C Osc.-Amp.	2000	—	120	—	—	150	HF75
TW75 *	75	7.5	4.15	2000	175	60	20	3.35	1.5	0.7	4-pin M.	2D	Class-C Amp.-Oscillator	2000	−175	150	37	12.7	225	TW75
													Class-C Amp. Plate-Mod.	2000	−260	125	32	13.2	198	
HF100	75	10	2.0	1500	150	30	23	3.5	4.5	1.4	4-pin M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	170	HF100
													Class-C Amp. Plate-Mod.	1250	−250	110	21	8.0	105	
													Grid-Modulated Amp.	1500	−280	72	1.5	6.0	42	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances ($\mu\mu$fd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
111H	75	10	2.25	1500	160	—	23	—	4.6	—	4-pin M.	2D	Class-C Osc.-Amp.	1500	—	160	—	—	175	111H
ZB120	75	10	2.0	1250	160	40	90	5.3	5.2	3.2	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−135	160	23	5.5	145	ZB120
													Class-C Amp. Plate-Mod.	1000	−150	120	21	5.0	95	
													Grid-Modulated Amp.	1250	—	95	8.0	1.5	45	
327B	75	10.5	10.6	15000	—	—	30	3.4	2.45	0.3	None	T-4D	—	—	—	—	—	—	—	327B
242A	85	10	3.25	1250	150	50	12.5	6.5	13	4.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	130	242A
													Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
284D	85	10	3.25	1250	150	100	4.8	6.0	8.3	5.6	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	125	284D
													Class-C Amp. Plate-Mod.	1000	−450	150	50	—	100	
8005* [6]	85	10	3.25	1500	200	45	20	6.4	5.0	1.0	4-pin M.	T-4BB	Class-C Amp.-Oscillator	1500	−130	200	32	7.5	220	8005
													Class-C Amp. Plate-Mod.	1250	−195	190	28	9.0	170	
													Class-B Amp. (Telephony)	1500	− 80	83	1.0	5.0	45	
RK36* [4]	100	5.0	8.0	3000	165	35	14	4.5	5.0	1.0	4-pin M.	2D	Class-C Amp. (Telegraphy)	2000	−360	150	30	15	200	RK36
													Class-C Amp. (Telephony)	2000	−360	150	30	15	200	
													Grid-Modulated Amp.	2000	−270	72	1.0	3.5	42	
													Class-B Amp. (Telephony)	2000	−180	75	3.0	10	50	
RK38* [4]	100	5.0	8.0	3000	165	40	—	4.6	4.3	0.9	4-pin M.	2D	Class-C Amp. (Telegraphy)	2000	−200	160	30	10	225	RK38
													Class-C Amp. (Telephony)	2000	−200	160	30	10	225	
													Grid-Modulated Amp.	2000	−150	80	2.0	5.5	60	
													Class-B Amp. (Telephony)	2000	−100	75	2.0	7.0	55	
100TH	100	5.0	6.5	3000	225	50	30	2.2	2.0	0.3	4-pin M.	2D	Class-C Amp. (Telegraphy)	3000	−210	167	40	18	400	100TH
													Class-C Amp. Plate-Mod.	3000	−210	167	45	18	400	
													Class-B Amp. (Telephony)	3000	− 70	50	2.0	5.0	50	
													Grid-Modulated Amp.	3000	−400	70	3.0	7.0	100	
100TL	100	5.0	6.5	3000	225	35	12	2.0	2.3	0.4	4-pin M.	2D	Class-C Amp. (Telegraphy)	3000	−600	167	30	18	400	100TL
													Class-C Amp. Plate-Mod.	3000	−600	167	35	18	400	
													Class-B Amp. (Telephony)	3000	−280	50	1.0	5.0	50	
													Grid-Modulated Amp.	3000	−560	60	2.0	7.0	90	
VT127A	100	5.0	10.5	16000	—	—	15	—	—	—	None	T-4B	Oscillator at 200 Mc.	Approximately 75 watts output						VT127A
227A	100	10.5	10.6	15000	—	—	30	3.25	2.45	0.3	None	T-4B	Oscillator at 200 Mc.	Approximately 75 watts output						227A
327A	100	10.5	10.6	15000	—	—	30	3.4	2.25	0.3	None	T-4D	Oscillator at 200 Mc.	Approximately 75 watts output						327A
HK254	100	5.0	7.5	4000	200	40	25	3.3	3.4	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	4000	−380	120	35	20	475	HK254
													Class-C Amp. Plate-Mod.	3000	−290	135	40	23	320	
													Class-B Amp. (Telephony)	3000	−125	51	2.0	3.0	54	
													Grid-Modulated Amp.	3000	—	51	3.0	4.0	58	
RK58 [6]	100	10	3.25	1250	175	70	—	8.5	6.5	10.5	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	1250	− 90	150	30	6.0	130	RK58
													Class-C Amp. Plate-Mod.	1000	−135	150	50	16	100	
													Class-B Amp. (Telephony)	1250	—	106	15	6.0	42.5	
HF120	100	10	3.25	1250	175	—	12	—	10.5	—	4-pin J.	—	Class-C Amp.-Oscillator	1250	—	175	—	—	150	HF120
HF125	100	10	3.25	1500	175	—	25	—	11.5	—	4-pin J.	—	Class-C Amp.-Oscillator	1500	—	175	—	—	200	HF125

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
HF140	100	10	3.25	1250	175	—	12	—	12.5	—	4-pin J.	—	Class-C Amp.-Oscillator	1250	—	175	—	—	150	HF140
203A 303A	100	10	3.25	1250	175	60	25	6.5	14.5	5.5	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	25	7.0	130	203A 303A
													Class-C Amp. (Telephony)	1000	−135	150	50	14	100	
													Class-B Amp. (Telephony)	1250	− 45	105	3.0	3.0	42.5	
203H	100	10	3.25	1500	175	60	25	6.5	11.5	1.5	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−200	170	12	3.8	200	203H
													Class-C Amp. (Telephony)	1250	−160	167	19	5.0	160	
													Class-B Amp. (Telephony)	1500	− 48	100	3.0	2.0	52	
211 311 835 [4]	100	10	3.25	1250	175	50	12	6.0 6.0	14.5 9.25	5.5 5.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−225	150	18	7.0	130	211 311 835
													Class-C Amp. (Telegraphy)	1000	−260	150	35	14	100	
													Class-B Amp. (Telephony)	1250	−100	106	1.0	7.5	42.5	
242B 342B	100	10	3.25	1250	150	50	12.5	7.0	13.6	6.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	130	242B 342B
													Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
													Class-B Amp. (Telephony)	1250	− 80	120	—	—	50	
242C	100	10	3.25	1250	150	50	12.5	6.1	13.0	4.7	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	130	242C
													Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
													Class-B Amp. (Telephony)	1250	− 90	120	—	—	50	
261A 361A	100	10	3.25	1250	150	50	12	6.5	9.0	4.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	100	261A 361A
													Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
													Class-B Amp. (Telephony)	1250	−100	125	—	—	50	
276A 376A	100	10	3.0	1250	125	50	12	6.0	9.0	4.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	100	276A 376A
													Class-C Amp. Plate-Mod.	1000	−160	125	50	—	85	
													Class-B Amp. (Telephony)	1250	−100	125	—	—	50	
284B	100	10	3.25	1250	150	100	5.0	4.2	7.4	5.3	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	125	284B
													Class-C Amp. Plate-Mod.	1000	−430	150	50	—	100	
													Class-B Amp. (Telephony)	1250	−270	120	—	—	50	
295A	100	10	3.25	1250	175	50	25	6.5	14.5	5.5	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	—	—	125	295A
													Class-C Amp. Plate-Mod.	1000	−125	150	50	—	100	
													Class-B Amp. (Telephony)	1250	− 75	105	—	—	42.5	
838 938	100	10	3.25	1250	175	70	—	6.5	8.0	5.0	4-pin J.	4E	Class-C Amp. (Telegraphy)	1250	− 90	150	30	6.0	130	838 938
													Class-C Amp. (Telephony)	1000	−135	150	60	16	100	
													Class-B Amp. (Telephony)	1250	0	106	15	6.0	42.5	
852	100	10	3.25	3000	150	40	12	1.9	2.6	1.0	4-pin M.	2D	Class-C Amp. (Telegraphy)	3000	−600	85	15	12	165	852
													Class-C Amp. (Telephony)	2000	−500	67	30	23	75	
													Class-B Amp. (Telephony)	3000	−250	43	0	7.0	40	
8003	100	10	3.25	1500	250	50	12	5.8	11.7	3.4	4-pin J.	T-3AB	Class-C Amp.-Oscillator	1350	−180	245	35	11	250	8003
													Class-C Amp. Plate-Mod.	1100	−260	200	40	15	167	
													Class-B Amp. (Telephony)	1350	−110	110	1.5	8	50	
RK57/ /805	125	10	3.25	1500	210	70	—	6.5	8.0	5.0	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	215	RK57/ /805
													Class-C Amp. (Telephony)	1250	−160	160	60	16	140	
													Class-B Amp. (Telephony)	1500	− 10	115	15	7.5	57.5	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
T125 *	125	10	4.5	2500	250	60	25	6.3	6.0	1.3	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	2500	−200	240	31	11	475	T125
													Class-C Amp. Plate-Mod.	2000	−215	200	28	10	320	
HF130	125	10	3.25	1250	210	—	12.5	—	9.0	—	4-pin J.	—	Class-C Amp.-Oscillator	1250	−210	—	—	—	170	HF130
HF150	125	10	3.25	1500	210	—	12.5	—	7.2	—	4-pin J.	—	Class-C Amp.-Oscillator	1500	—	210	—	—	200	HF150
HF175	125	10	4.0	2000	250	—	18	—	6.3	—	4-pin J.	—	Class-C Amp.-Oscillator	2000	—	250	—	—	300	HF175
GL146	125	10	3.25	1500	200	60	78	7.2	9.2	3.9	4-pin GL	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	150	GL146
													Class-C Amp. Plate-Mod.	1000	−200	160	40	—	100	
													Class-B Amp. (Telephony)	1250	0	132	—	—	55	
GL152	125	10	3.25	1500	200	60	25	7.0	8.8	4.0	4-pin GL	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	150	GL152
													Class-C Amp. Plate-Mod.	1000	−200	160	30	—	100	
													Class-B Amp. (Telephony)	1250	− 40	132	—	—	55	
805	125	10	3.25	1500	210	70	40/60	8.5	6.5	10.5	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	215	805
													Class-C Amp. Plate-Mod.	1250	−160	160	60	16	140	
													Class-B Amp. (Telephony)	1500	− 10	115	15	7.5	57.5	
150T	150	5.0	10	3000	200	50	13	3.0	3.5	0.5	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−600	200	35	—	450	150T
TW150	150	10	4.1	3000	200	60	35	3.9	2.0	0.8	4-pin J.	T-3AC	Class-C Amp.-Oscillator	3000	−170	200	45	17	470	TW150
													Class-C Amp. Plate-Mod.	3000	−260	165	40	17	400	
152TL ** [5] HK252-L ** [5]	150	5/10 [13]	13/6.5	3000	500	75	10	7.0	5.0	0.4	Special	T-4BF	Class-C Amp.-Oscillator	3000	−400	250	30	15	610	152TL HK252-L
													Class-C Amp. Plate-Mod.	2500	−350	250	35	16	500	
HF200 HV18	150	10–11	3.4	2500	200	50	18	5.2	5.8	1.2	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	380	HF200 HV18
													Class-C Amp. Plate-Mod.	2000	−350	160	20	9.0	250	
													Class-B Amp. (Telephony)	2500	−140	90	—	4.0	80	
HD203A	150	10	4.0	2000	250	60	25	—	12	—	4-pin J.	T-3AB	Class-C Amplifier	—	—	—	—	—	375	HD203A
HF250	150	10.5	4.0	2500	200	—	18	—	5.8	—	4-pin J.	T-3AC	Class-C Amp.-Oscillator	2500	—	200	—	—	375	HF250
HK354 HK354C	150	5.0	10	4000	300	50	14	4.5	3.8	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	4000	−690	245	50	48	830	HK354 HK354C
													Class-C Amp. Plate-Mod.	3000	−550	210	50	35	525	
													Class-B Amp. (Telephony)	3000	−205	78	2.0	10	82	
													Grid-Modulated Amp.	3000	−400	78	3.0	12	85	
HK354D	150	5.0	10	4000	300	55	22	4.5	3.8	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3500	−490	240	50	38	690	HK354D
													Class-C Amp. Plate-Mod.	3500	−425	210	55	36	525	
HK354E	150	5.0	10	4000	300	60	35	4.5	3.8	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3500	−448	240	60	45	690	HK354E
													Class-C Amp. Plate-Mod.	3000	−437	210	60	45	525	
HK354F	150	5.0	10	4000	300	75	50	4.5	3.8	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3500	−368	250	75	50	720	HK354F
													Class-C Amp. Plate-Mod.	3000	−312	210	75	45	525	
810 [6] 1627	150	10 5.0	4.5 9.0	2250	275	70	36	8.7	4.8	12	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	2250	−160	275	40	12	475	810 1627
													Class-C Amp. Plate-Mod.	1800	−200	250	50	17	335	
													Class-B Amp. (Telephony)	2250	− 70	100	2.0	4.0	75	
													Grid-Modulated Amp.	2250	−140	100	2.0	4.0	75	
8000 [5]	150	10	4.5	2250	275	40	16.5	5.0	6.4	3.3	4-pin J.	T-3AC	Class-C Amp.-Oscillator	2250	−210	275	25	9.0	475	8000
													Class-C Amp. Plate-Mod.	1800	−320	250	20	8.8	335	
													Class-B Amp. (Telephony)	2250	−145	100	0	5.4	75	
													Grid-Modulated Amp.	2250	−265	100	0	2.5	75	

TABLE XIV — TRIODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [5]	Approx. Carrier Output Power Watts	Type
RK63 RK63A	200	5.0 6.3	10 14	3000	250	60	37	2.7	3.3	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−200	233	45	17	525	RK63 RK63A
													Class-C Amp. Plate-Mod.	2500	−200	205	50	19	405	
													Class-B Amp. (Telephony)	3000	−150	100	1.0	12	100	
													Grid-Modulated Amp.	3000	−250	100	7.0	12.5	100	
T200	200	10	5.75	2500	350	80	16	9.5	7.9	1.6	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	2500	−280	350	54	25	685	T200
													Class-C Amp. Plate-Mod.	2000	−260	300	54	23	460	
HF300	200	11–12	4.0	3000	275	60	23	6.0	6.5	1.4	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−400	250	28	16	600	HF300
													Class-C Amp. Plate-Mod.	2000	−300	250	36	17	385	
													Class-B Amp. (Telephony)	2500	−100	120	0.5	6.0	105	
T814 HV12	200	10	4.0	2500	200	60	12	8.5	12.8	1.7	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	2500	−240	300	30	10	575	T814 HV12
													Class-C Amp. Plate-Mod.	2000	−370	300	40	20	485	
T822 HV27	200	10	4.0	2500	300	60	27	8.5	13.5	2.1	4-pin J.	T-3AB	Class-C Amp. (Telegraphy)	2500	−175	300	50	15	585	T822 HV27
													Class-C Amp. Plate-Mod.	2000	−195	250	45	15	400	
													Class-B Amp. (Telephony)	2500	− 95	125	5.0	8.0	110	
806 [5]	225	5.0	10	3300	300	50	12.6	6.1	4.2	1.1	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3300	−600	300	40	34	780	806
													Class-C Amp. Plate-Mod.	3000	−670	195	27	24	460	
													Class-B Amp. (Telephony)	3300	−280	102	—	10.3	115	
250TH*	250	5.0	10.5	3000	350	100	32	3.5	3.3	0.3	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−210	330	75	42	750	250TH
													Class-C Amp. Plate-Mod.	3000	−210	330	75	42	750	
													Class-B Amp. (Telephony)	3000	− 80	125	4.0	15	125	
													Grid-Modulated Amp.	3000	−160	125	4.0	20	125	
250TL*	250	5.0	10.5	3000	350	50	13	3.0	3.5	0.5	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−600	330	45	42	750	250TL
													Class-C Amp. Plate-Mod.	3000	−600	330	45	42	750	
													Class-B Amp. (Telephony)	3000	−225	125	2.0	15	125	
													Grid-Modulated Amp.	3000	−450	125	2.0	15	125	
GL159	250	10	9.6	2000	400	100	20	11	17.6	5.0	4-pin GL	T-4BG	Class-C Amp.-Oscillator	2000	−200	400	17	6.0	620	GL159
													Class-C Amp. Plate-Mod.	1500	−240	400	23	9.0	450	
													Class-B Amp. (Telephony)	2000	− 90	190	—	2.5	130	
GL169	250	10	9.6	2000	400	100	85	11.5	19	4.7	4-pin GL	T-4BG	Class-C Amp.-Oscillator	2000	−100	400	42	10	620	GL169
													Class-C Amp. Plate-Mod.	1500	−100	400	45	10	450	
													Class-B Amp. (Telephony)	2000	− 10	190	—	3.5	130	
204A 304A	250	11	3.85	2500	275	80	23	12.5	15	2.3	Special	T-1A	Class-C Amp. (Telegraphy)	2500	−200	250	30	15	450	204A 304A
													Class-C Amp. Plate-Mod.	2000	−250	250	35	20	350	
													Class-B Amp. (Telephony)	2500	− 70	160	—	15	100	
308B	250	14	4.0	2250	325	75	8.0	13.6	17.4	9.3	4-pin W.E.	T-2A	Class-C Amp. (Telegraphy)	1750	−400	300	—	—	350	308B
													Class-C Amp. Plate-Mod.	1250	−320	300	75	—	250	
													Class-B Amp. (Telephony)	1750	−230	215	—	—	125	
HK454H* [6]	250	5.0	11	5000	375	85	30	4.6	3.4	1.4	4-pin J.	T-3AC	Class-C Amplifier	3500	−275	270	60	28	760	HK454H
HK454-L* [6]	250	5.0	11	5000	375	60	12	4.6	3.4	1.4	4-pin J.	T-3AC	Class-C Amplifier	3500	−450	270	45	30	760	HK454-L
212E 241B 312E	275	14	4.0	3000	350	75	16	14.9	18.8	8.6	4-pin W.E. 3-pin W.E.	T-2A T-2AA	Class-C Amp. (Telegraphy)	2000	−225	300	—	—	400	212E 241B 312E
													Class-C Amp. Plate-Mod.	1500	−200	300	75	—	300	
													Class-B Amp. (Telephony)	2000	−120	300	—	—	200	

TABLE XIV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [2]	Socket Connections [1]	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts [3]	Approx. Carrier Output Power Watts	Type
300T [4]	300	8.0	11.5	3500	350	75	16	4.0	4.0	0.6	4-pin J.	T-3AC	Class-C Amplifier	3500	−600	300	60	—	800	300T
304TL [6] HK304-L*[6]	300	5/10 [13]	26/13	3000	1000	150	10	12	9.0	0.8	Special	T-4BF	Class-C Amplifier	2000	−300	500	—	—	800	304TL HK304-L
527	300	5.5	135.0	20000	—	—	38	19.0	12.0	1.4	Special	T-4B	Oscillator at 200 Mc.	Approximately 250 watts output						527
HK654	300	7.5	15	4000	600	100	22	6.2	5.5	1.5	4-pin J.	T-3AC	Class-C Amp. (Telegraphy)	2000	−380	500	75	57	720	HK654
													Class-C Amp. Plate-Mod.	2000	−365	450	110	70	655	
													Class-B Amp. (Telephony)	3500	−137	150	13	13	210	
													Grid-Modulated Amp.	3500	−210	150	15	15	210	
833A	300	10	10	3000	500	100	35	12.3	6.3	8.5	Special	T-1AB	Class-C Amp. (Telegraphy)	2000	−200	475	65	25	740	833A
													Class-C Amp. (Telephony)	2500	−300	335	75	30	635	
													Class-B Amp. (Telephony)	3000	− 70	150	2.0	10	150	
270A	350	10	4.0	3000	375	75	16	18	21	2.0	Special	T-1A	Class-C Amp. (Telegraphy)	3000	−375	350	—	—	700	270A
													Class-C Amp. Plate-Mod.	2250	−300	300	80	—	450	
													Class-B Amp. (Telephony)	3000	−180	175	—	—	175	
849	400	11	5.0	2500	350	125	19	17	33.5	3.0	Special	T-1A	Class-C Amp. (Telegraphy)	2500	−250	300	20	8.0	560	849
													Class-C Amp. (Telephony)	2000	−300	300	30	14	425	
													Class-B Amp. (Telephony)	2500	−125	216	1.0	12	180	
831 [4]	400	11	10	3500	350	75	14.5	3.8	4.0	1.4	Special	T-1AA	Class-C Amp. (Telegraphy)	3500	−400	275	40	30	590	831
													Class-C Amp. (Telephony)	3000	−500	200	60	50	360	
													Class-B Amp. (Telephony)	3500	−220	146	—	—	160	

[1] S.—small; M.—medium; J.—jumbo; O.—octal.
[2] Refer to Transmitting Tube Diagrams.
[3] See Chapter Five for discussion of grid driving power.
[4] Obsolete type.
[5] Instant-heating filament for mobile use.
[6] Intermittent commercial and amateur service ratings.
[7] Twin triode. Values, except inter-element capacities, are for both sections, in push-pull.
[8] The 805 has a variable high-μ grid.
[9] All wire leads. Ratings at 500 Mc.
[10] Gaseous discharge tube for use on 110-volt d.c.
[11] Output at 112 Mc.
[12] Calculated at 33% efficiency for 100% modulation.
[13] Multiple-unit tube with dual filaments which can be connected in series or parallel.
[14] Forced-air cooling is recommended at ratings above 75 per cent of maximum.
[15] See Receiving Tube Base Diagrams.
[16] Input resonant frequency approximately 335 Mc.
[17] Subject to wide variation.
[18] Cathode resistor in ohms.
[19] Grid-leak resistor in ohms.
[20] Approximately 45 milliwatts output at 1200 Mc. (With 100 volts and 2000 ohm grid resistor).
[21] At 40 Mc.
[22] At 150 Mc.
[23] Absolute maximum.

Frequency limits:
* May be used at full ratings on 56–60 Mc. band and lower.
** May be used at full ratings on 112-Mc. band and lower.
*** May be used at full ratings on 224-Mc. band and lower.
**** May be used at full rating above 300 Mc.

TABLE XV—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen [3] Resistor Ohms	Approx. Grid Driving Power Watts [4]	Approx. Carrier Output Power Watts	Type
3A4	2.0	1.4 2.8	0.2 0.1	150	135	0.9	4.8	0.2	4.2	7-pin B.	7BB [14]	Class-C Amp.-Oscillator	150	135	0	− 26	18.3	6.5	0.13	2300	—	1.2	3A4
HY63*[5,6]	3.0	2.5 1.25	0.1125 0.225	200	100	0.6	8.0	0.1	8.0	7-pin O.	T-8DB	Class-C Amp.-Osc.	200	100	—	− 22.5	20	4.0	2.0	—	0.1	3.0	HY63
												Class-C Amp. Plate-Mod.	180	100	—	− 35	15	3.0	2.0	—	0.2	2.0	

TABLE XV — TETRODE AND PENTODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen [3] Resistor Ohms	Approx. Grid Driving Power Watts [4]	Approx. Carrier Output Power Watts	Type
RK64*[5]	6.0	6.3	0.5	400	100	3.0	10	0.4	9.0	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	400	100	30	−30	35	10	3.0	—	0.18	10	RK64
												Class-C Amp. Plate-Mod.	300	—	30	−30	26	8.0	4.0	30000	0.2	6.0	
1610	6.0	2.5	1.75	400	200	2.0	8.6	1.2	13	5-pin M.	T-5CA	Class-C Amp.-Oscillator	400	150	—	−50	22.5	7.0	1.5	—	0.1	5.0	1610
RK56*	8.0	6.3	0.55	300	300	4.5	10	0.2	9.0	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	400	300	—	−40	62	12	1.6	—	0.1	12.5	RK56
												Class-C Amp. Plate-Mod.	250	200	—	−40	.50	10	1.6	2800	0.28	8.5	
RK23 RK25 RK25B[5]	10	2.5	2.0	500	250	8	10	0.2	10	7-pin M.	T-7C	Class-C Amp. (Telegraphy)	500	200	45	−90	55	38	4.0	—	0.5	22	RK23 RK25 RK25B
		6.3	0.9									Class-C Amp. (Telephony)	400	150	0	−90	43	30	6.0	8300	0.8	13.5	
												Suppressor-Modulated Amp.	500	200	−45	−90	31	39	4.0	—	0.5	6.0	
1613	10	6.3	0.7	350	275	2.5	8.5	0.5	11.5	7-pin O.	7S[14]	Class-C Amp. (Telegraphy)	350	200	—	−35	50	10	3.5	20000	0.22	9	1613
												Class-C Amp. Plate-Mod.	275	200	—	−35	42	10	2.8	10000	0.16	6.0	
6F6 6F6G	11	6.3	0.7	375	285	3.75	6.5	0.2	13	7-pin O.	7AC[14]	Class-C Amp.-Oscillator	350	200	—	−35	50	10	3.5	—	0.22	9.0	6F6 6F6G
							8.0	0.5	6.5			Class-C Amp. Plate-Mod.	275	200	—	−35	42	10	2.8		0.16	6.0	
837 RK44[6]	12	12.6	0.7	500	300	8	16	0.2[11]	10	7-pin M.	T-7C	Class-C Amp. (Telegraphy)	500	200	40	−70	80	15	4.0	20000	0.4	28	837 RK44
												Class-C Amp. (Telephony)	400	140	40	−40	45	20	5.0	13000	0.3	11	
												Suppressor-Modulated Amp.	500	—	−65	−20	30	23	3.5	14000	0.1	5.0	
802[7]	13	6.3	0.9	600	250	6.0	12	0.15[11]	8.5	7-pin M.	T-7C	Class-C Amp. (Telegraphy)	600	250	40	−120	55	16	2.4	22000	0.30	23	802
												Class-C Amp. Plate-Mod.	500	245	40	−40	40	15	1.5	16300	0.10	12	
												Suppressor-Mod. Amp.	600	250	−45	−100	30	24	5.0	14500	0.6	6.3	
HY6V6-GTX*	13	6.3	0.5	350	225	2.5	9.5	0.7	9.5	7-pin O.	7AC[14]	Class-C Amp.-Osc.	300	200	—	−45	60	7.5	2.5	—	0.3	12	HY6V6-GTX
												Class-C Amp. Plate-Mod.	250	200	—	−45	60	6.0	2.0	15000	0.4	10	
HY60*	15	6.3	0.5	425	225	2.5	10	0.2	8.5	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	425	200	—	−62.5	60	8.5	3.0	—	0.3	18	HY60
												Class-C Amp. Plate-Mod.	325	200	—	−45	60	7.0	2.5	—	0.2	14	
HY65*[5]	15	6.3	0.85	450	250	4.0	9.1	0.18	7.2	7-pin O.	T-8DB	Class-C Amp.-Osc.	450	250	—	−45	75	15	3.0	—	0.5	24	HY65
												Class-C Amp. Plate-Mod.	350	200	—	−45	63	12	3.0	—	0.5	16	
306A	15	2.75	2.0	300	300	6.0	13	0.35	13	5-pin M.	T-5CB	Class-C Amp. (Telephony)	300	180	—	−50	36	15	3.0	8000	—	7.0	306A
307A RK-75	15	5.5	1.0	500	250	6.0	15	0.55	12	5-pin M.	T-5C	Class-C Amp. (Telegraphy)	500	250	0	−35	60	13	1.4	20000	—	20	307A RK-75
												Suppressor-Modulated Amp.	500	200	−50	−35	40	20	1.5	14000	—	6.0	
832**[10]	15	6.3	1.6	500	250	5.0	7.5	0.05[11]	3.8	Special	7BP	Class-C Amp. (Telegraphy)	500	200	—	−65	72	14	2.6	21000	0.18	26	832
		12.6	0.8									Class-C Amp. (Telephony)	425	200	—	−60	52	16	2.4	14000	0.15	16	
832A**[10]	15	6.3	1.6	750	250	5.0	7.5	0.05[11]	3.8	Special	7BP	Class-C Amp. (Telegraphy)	750	200	—	−65	48	15	2.8	36500	0.19	26	832A
		12.6	0.8									Class-C Amp. (Telephony)	600	200	—	−65	36	16	2.6	25000	0.16	17	
844[5]	15	2.5	2.5	500	180	3.0	9.5	0.15	7.5	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	500	175	—	−125	25	—	5.0	—	—	9.0	844
												Class-C Amp. (Telephony)	500	150	—	−100	20	—	—	—	—	4.0	
865	15	7.5	2.0	750	175	3.0	8.5	0.1[11]	8.0	4-pin M.	T-4C	Class-C Amp. (Telegraphy)	750	125	—	−80	40	—	5.5	—	1.0	16	865
												Class-C Amp. (Telephony)	500	125	—	−120	40	—	9.0	—	2.5	10	
1619	15	2.5	2.0	400	300	3.5	10.5	0.35	12.5	7-pin O.	7AC[14]	Class-C Amp. (Telegraphy)	400	300	—	−55	75	10.5	5.0	9500	0.36	19.5	1619
												Class-C Amp. Plate-Mod.	325	285	—	−50	62	7.5	2.8	5000	0.18	13	
254A	20	5.0	3.25	750	175	5.0	4.6	0.1	9.4	4-pin M.	T-4C	Class-C Amplifier	750	175	—	−90	60	—	—	—	—	25	254A
6L6 6L6G	21	6.3	0.9	375	300	3.5	10	0.4	12	7-pin O.	7AC[14]	Class-C Amp.-Oscillator	375	200	—	−35	88	9.0	3.5	—	0.18	17	6L6 6L6G
							11.5	0.9	9.5			Class-C Amp. Plate-Mod.[8]	325	—	—	−70	65	—	9.0	—	0.8	11	

TABLE XV—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen [3] Resistor Ohms	Approx. Grid Driving Power Watts [4]	Approx. Carrier Output Power Watts	Type
6L6GX	21	6.3	0.9	500	300	3.5	11	1.5	7.0	7-pin O.	7AC	Class-C Amp. (Telegraphy)	500	250	—	−50	90	9.0	2.0	—	0.25	30	6L6GX
												Class-C Amp. Plate-Mod.	325	225	—	−45	90	9.0	3.0	—	0.25	20	
HY6L6-GTX*	21	6.3	0.9	500	300	3.5	11	0.5	7.0	7-pin O.	7AC	Class-C Amp.-Osc.	500	250	—	−50	90	9.0	2.0	—	0.5	30	HY6L6-GTX
												Class-C Amp. Plate-Mod.	400	225	—	−45	90	9.0	3.0	16000	0.8	20	
T21*	21	6.3	0.9	400	300	3.5	13	0.7	12	6-pin M.	T-6B	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	25	T21
												Class-C Amp. Plate-Mod.	350	200	—	−45	65	17	5.0	—	0.35	14	
RK49	21	6.3	0.9	400	300	3.5	11.5	1.4	10.6	6-pin M.	T-6B	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	25	RK49
												Class-C Amp. (Telephony)	300	200	—	−45	60	15	5.0	6700	0.34	12	
1614*	21	6.3	0.9	375	300	3.5	10	0.4	12.5	7-pin O.	7AC	Class-C Amp. (Telegraphy)	375	250	—	−40	80	10	2.0	12500	0.1	21	1614
												Class-C Amp. Plate-Mod.	325	—	—	−40	70	8.0	2.0	10000	0.1	15	
RK41*[5] RK39*	25	2.5 6.3	2.4 0.9	600	300	3.5	13	0.2	10	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	600	300	—	−90	93	10	3.0	—	0.38	36	RK41 RK39
												Class-C Amp. (Telephony)	475	250	—	−50	85	9.0	2.5	25000	0.2	26	
HY61/ 807*	25	6.3	0.9	600	300	3.5	11	0.2	7.0	5-pin M.	T-5BB	Class-C Amp. (Telegraphy)	600	250	—	−50	100	9.0	3.0	39000	0.22	40	HY61/ 807
												Class-C Amp. (Telephony)	475	250	—	−50	83	9.0	3.5	25000	0.2	27	
815**[7,10]	25	6.3	1.6	500	200	4.0	13.3	0.2[11]	8.5	8-pin O.	T-8FA[12]	Class-C Amp.-Oscillator	500	200	—	−45	150	17	2.5	—	0.13	56	815
												Class-C Amp. Plate-Mod.	400	175	—	−45	150	15	3.0	—	0.16	45	
254B	25	7.5	3.25	750	150	5.0	11.2	0.085	5.4	4-pin M.	T-4C	Class-C Amplifier	750	150	—	−135	75	—	—	—	—	30	254B
1624*	25	2.5	2.0	600	300	3.5	11	0.25	7.5	5-pin M.	T-5DC	Class-C Amp. (Telegraphy)	600	300	—	−60	90	10	5.0	30000	0.43	35	1624
												Class-C Amp. Plate-Mod.	500	275	—	−50	75	9.0	3.3	25000	0.25	24	
2E22	30	6.3	1.5	750	250	10	13	0.2	8.0	5-pin M.	5J[14]	Class-C Amp.-Oscillator	500	250	—	—	60	4.5	—	—	—	—	2E22
RK66*	30	6.3	1.5!	600	300	3.5	12	0.25	10.5	5-pin M.	T-5C	Class-C Amp.-Oscillator	600	300	—	−60	90	11	5.0	—	0.5	40	RK66
												Class-C Amp. Plate-Mod.	500	—	—	−50	75	8.0	3.2	25000	0.23	25	
807*[7] 1625*[7]	30	6.3 12.6	0.9 0.45	750	300	3.5	11	0.2[11]	7.0	5-pin M. 7-pin M.	T-5BB T-9E	Class-C Amp. (Telegraphy)	750	250	—	−50	100	8.0	3.0	—	0.22	50	807 1625
												Class-C Amp. Plate-Mod.	600	275	—	−90	100	6.5	4.0	—	0.4	42.5	
RK20[5] RK20A RK46[5]	40	7.5 7.5 12.6	3.0 3.25 2.5	1250	300	15	14	0.01	12	5-pin M.	T-5C	Class-C Amp. (Telegraphy)	1250	300	45	−100	92	36	11.5	—	1.6	84	RK20 RK20A RK46
												Class-C Amp. (Telephony)	1000	300	0	−100	75	30	10	23000	1.3	52	
												Suppressor-Modulated Amp.	1250	300	−45	−100	48	44	11.5	—	1.5	21	
												Grid-Modulated Amp.	1250	300	45	−142	40	7.0	1.8	—	1.5	20	
HY69*[6]	40	6.3	1.5	600	300	5.0	15.4	0.23	6.5	5-pin M.	T-5D	Class-C Amp.-Oscillator	600	250	—	−60	100	12.5	4.0	30000	0.25	42	HY69
												Class-C Amp. Plate-Mod.	600	250	—	−60	100	12.5	5.0	30000	0.35	42	
												Modulated Doubler	600	200	—	−300	90	11.5	6.0	35000	2.8	27	
829**[10]	40	6.3 12.6	2.25 1.12	500	225	40	14.5	0.1[11]	7.0	Special	7BP[14]	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	83	829
												Class-C Amp. Plate-Mod.	425	200	—	−60	212	35	11	6400	0.8	63	
												Grid-Modulated Amp.	500	200	—	−38	120	10	2.0	—	0.5	23	
829A**[10]	40	6.3 12.6	2.25 1.12	750	240	7.0	14.4	0.1[11]	7.0	Special	7BP[14]	Class-C Amp.-Oscillator	750	200	—	−55	160	30	12	18300	0.8	87	829A
												Class-C Amp. Plate-Mod.	600	200	—	−70	150	30	12	13300	0.9	70	
												Grid-Modulated Amp.	750	200	—	−55	80	5.0	0	—	0.7	24	
829B**[10] 3E29**[10]	40	12.6 6.3	1.125 2.25	750	225	6	14.5	0.1	7.0	Special	7BP[14]	Class-C Amp. (Grid Mod.)	750	200	—	−55	80	5	0	—	0.7	24	829B 3E29
	28			600	225	7						Class-C Amp. (Plate-Mod.)	600	200	—	−70	150	30	12.0	13300	0.9	70	
	40			750	225	7						Class-C Amp. (Telegraphy)	750	200	—	−55	160	30	12.0	18300	0.8	87	
HY1269*[6]	40	6.3 12.6	3.5 1.75	750	300	5.0	16.0	0.25	7.5	5-pin M.	T-5DB[12]	Class-C Amp.-Oscillator	750	300	—	−70	120	15	4	—	0.25	63	HY1269
												Class-C Amp. Plate Mod.	600	250	—	−70	100	12.5	5	35000	0.5	42	
												Grid-Modulated Amp.	750	300	—	—	80	—	—	—	—	20	

TABLE XV — TETRODE AND PENTODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms [3]	Approx. Grid Driving Power Watts [4]	Approx. Carrier Output Power Watts	Type
RK47	50	10	3.25	1250	300	10	13	0.12	10	5-pin M.	T-5D	Class-C Amp. (Telegraphy)	1250	300	—	−70	138	14	7.0	—	1.0	120	RK47
												Class-C Amp. Plate-Mod.	900	300	—	−150	120	17.5	6.0	—	1.4	87	
												Grid-Modulated Amp.	1250	300	—	−30	60	2.0	0.9	—	4.0	25	
312A	50	10	2.8	1250	500	20	15.5	0.15	12.3	6-pin M.	T-6C	Class-C Amp. (Telegraphy)	1250	300	20	−55	100	36	5.5	—	0.7	90	312A
												Class-C Amp. Plate-Mod.	1000	—	40	−40	95	35	7.0	22000	1.0	65	
												Suppressor-Mod. Amp.	1250	—	−85	−50	50	42	5.0	22000	0.55	23	
804 [7]	50	7.5	3.0	1500	300	15	16	0.01 [11]	14.5	5-pin M.	T-5C	Class-C Amp. (Telegraphy)	1500	300	45	−100	100	35	7.0	34000	1.95	110	804
												Class-C Amp. Plate-Mod.	1250	250	50	−90	75	20	6.0	50000	0.75	65	
												Grid-Modulated Amp.	1500	300	45	−130	50	13.5	3.7	—	1.3	28	
												Suppressor-Mod. Amp.	1500	300	−50	−115	50	32	7.0	—	0.95	28	
305A	60	10	3.1	1000	200	6	10.5	0.14	5.4	4-pin M.	T-4CE	Class-C Amp. (Telegraphy)	1000	200	—	−200	125	—	—	—	—	85	305A
												Class-C Amp. (Telephony)	800	200	—	−270	125	—	—	—	—	70	
HY67	65	6.3 12.6	4.5 2.25	1250	300	10	—	0.19	14.5	5-pin M.	T-5DB	Class-C Amp. (Telegraphy)	1250	300	—	−80	175	22.5	10	—	1.5	152	HY67
												Class-C Amp. Plate-Mod.	1000	300	—	−150	145	17.5	14	—	2.0	101	
												Grid-Modulated Amp.	1250	300	—	—	78	—	—	—	—	32.5	
814 [7]	65	10	3.25	1500	300	10	13.5	0.1 [11]	13.5	5-pin M.	T-5D	Class-C Amp. (Telegraphy)	1500	300	—	−90	150	24	10	50000	1.5	160	814
												Class-C Amp. Plate-Mod.	1250	300	—	−150	145	20	10	48000	3.2	130	
												Grid-Modulated Amp.	1500	250	—	−120	60	3.0	2.5	—	4.2	35	
282A	70	10	3.0	1000	250	5.0	12.2	0.2	6.8	4-pin M.	T-4C	Class-C Amp. (Telegraphy)	1000	150	—	−160	100	—	—	—	—	33	282A
												Class-C Amp. Plate-Mod.	750	150	—	−180	100	—	50	—	—	50	
4E27/ 8001 [7]	75	5.0	7.5	2000	500	25	11	0.1 [11]	5.5	7-pin J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	—	−200	150	11	6.0	33000	1.4	230	4E27/ 8001
												Class-C Amp. Plate-Mod.	1800	400	—	−130	135	11	8.0	16000	1.7	178	
												Suppressor-Mod. Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	35	
HK257 * HK257B **	75	5.0	7.5	4000	500	25	13.8	0.04	6.7	7-pin J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6.0	—	1.4	230	HK257 HK257B
												Class-C Amp. Plate-Mod.	1800	400	60	−130	135	11	8.0	—	1.7	178	
												Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	35	
828 [7]	80	10	3.25	2000	750	23	13.5	0.05 [11]	14.5	5-pin M.	T-5C	Class-C Amp. (Telegraphy)	1500	400	75	−100	180	28	12	40000	2.2	200	828
												Class-C Amp. Plate-Mod.	1250	400	75	−140	160	28	12	30000	2.7	150	
												Grid-Modulated Amp.	1500	400	75	−150	80	4.0	1.3	—	1.3	41	
RK28 [5]	100	10	5.0	2000	400	35	15	0.02	15	5-pin J.	T-5C	Class-C Amp. (Telegraphy)	2000	400	45	−100	150	55	13	21000	2.0	210	RK28
												Class-C Amp. (Telephony)	1500	400	45	−100	135	52	13	21000	2.0	155	
												Suppressor-Modulated Amp.	2000	400	−45	−100	85	65	13	—	1.8	60	
												Grid-Modulated Amplifier	2000	400	45	−140	80	20	4.0	—	0.9	75	
RK48 [5] RK48A	100	10	5.0	2000	400	22	17	0.13	13	5-pin J.	T-5D	Class-C Amp. (Telegraphy)	2000	400	—	−100	180	40	6.5	—	1.0	250	RK48 RK48A
												Class-C Amp. (Telephony)	1500	400	—	−100	148	50	6.5	22000	1.0	165	
												Grid-Modulated Amplifier	1500	400	—	−145	77	10	1.5	—	1.6	40	
813	100	10	5.0	2000	400	22	16.3	0.2 [11]	14	7-pin J.	T-7DA	Class-C Amp. (Telegraphy)	2000	400	—	−90	180	15	3.0	107000	0.5	260	813
												Class-C Amp. (Telephony)	1600	400	—	−130	150	20	6.0	21600	1.2	175	
												Grid-Modulated Amplifier	2000	400	—	−120	75	3.0	—	—	—	50	

TABLE XV — TETRODE AND PENTODE TRANSMITTING TUBES — *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Base [1]	Socket Connections [2]	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen [3] Resistor Ohms	Approx. Grid Driving Power Watts [4]	Approx. Carrier Output Power Watts	Type
850	100	10	3.25	1250	175	10	17	0.25[11]	25	4-pin J.	T-3B	Class-C Amp. (Telegraphy)	1250	175	—	−150	160	—	35	—	10	130	850
												Class-C Amp. (Telephony)	1000	140	—	−100	125	—	40	—	10	65	
												Grid-Modulated Amplifier	1250	175	—	− 13	110	—	—	—	—	40	
860	100	10	3.25	3000	500	10	7.75	0.08[11]	7.5	4-pin M.	T-4CB	Class-C Amp.-Oscillator	3000	300	—	−150	85	25	15	—	7.0	165	860
												Class-C Amp. Plate-Mod.	2000	220	—	−200	85	25	38	100000	17	105	
RK28A	125	10	5.0	2000	400	35	15	0.02	15	5-pin J.	T-5C	Class-C Amp. (Telegraphy)	2000	400	45	−100	170	60	10	—	1.6	250	RK28A
												Class-C Amp. Plate-Mod.	1500	400	45	−100	135	54	10	18500	1.6	150	
												Grid-Modulated Amp.	2000	400	45	− 55	80	18	2.0	—	0.5	60	
												Suppressor-Mod. Amp.	2000	—	−45	−115[7]	90	52	11.5	30000	1.5	60	
803	125	10	5.0	2000	600	30	17.5	0.15[11]	29	5-pin J.	T-5C	Class-C Amp. (Telegraphy)	2000	500	40	− 90	160	45	12	—	2.0	210	803
												Class-C Amp. (Telephony)	1600	500	100	− 80	150	20	4.0	20000	4.0	155	
												Suppressor-Modulated Amp.	2000	—	−110	−100	80	48	15	35000	2.5	53	
												Grid-Modulated Amplifier	2000	600	40	− 80	80	20	4.0	—	2.0	53	
RK65	215	5.0	14	3000	500	35	10.5	0.24	4.75	4-pin J.	T-3BC	Class-C Amp. (Telegraphy)	3000	400	—	−100	240	70	24	—	6.0	510	RK65
												Class-C (Plate & Screen Mod.)	2500	—	—	−150	200	70	22	30000	6.3	380	
861	400	11	10	3500	750	35	14.5	0.1[11]	10.5	Special	T-1B	Class-C Amp. (Telegraphy)	3500	500	—	−250	300	40	40	—	30	700	861
												Class-C Amp. (Telephony)	3000	375	—	−200	200	—	55	70000	35	400	

[1] S. — small; M. — medium; O. — octal; J. — jumbo.
[2] See Transmitting Tube Base Diagrams.
[3] In plate-and-screen modulated Class-C amplifiers, connect screen-dropping resistor direct to r.f. B+ to mod., and by-pass for r.f. only. This does not apply to the 828.
[4] See Chapter 4–8 for discussion of grid driving power.
[5] Obsolete type. [6] Instant-heating filament for mobile operation.
[7] Intermittent commercial and amateur service ratings.
[8] Triode connection — screen-grid tied to plate.
[9] Calculated on basis of 33% efficiency at 100% modulation.
[10] Dual tube. Values for both sections, in push-pull.
[11] With external shielding.
[12] Terminals 3 and 6 must be connected together.
[13] Early tubes of this type do not have center-tapped filament.
[14] See Receiving Tube Base Diagrams.

Frequency limits:
* May be used at full ratings on 56–60 Mc. band and lower.
** May be used at full ratings on 112-Mc. band and lower.
*** May be used at full ratings on 224-Mc. band and lower.
**** May be used at full ratings above 300 Mc.

TABLE XVI — MAGNETRON AND VELOCITY-MODULATED TUBES

Type	Max. Plate or Collector Dissipation Watts	Cathode Volts	Cathode Amps.	Base	Socket Connections	Typical Operation	Plate or Collector Voltage	Grid No. 4 Voltage	Grid No. 3 Voltage	Grid No. 2 Voltage	Grid No. 1 Voltage	Grid No. 4 Current Ma.	Grid No. 3 Current Ma.	Grid No. 2 Current Ma.	Plate or Collector Current Ma.	Stabilizing Electrode Voltage	Stabilizing Electrode Current Ma.	Magnetic Field Gausses	Carrier Output Watts	Type
2J35[1]	4.0	1.8	2.0	4-pin M.	4AP	Magnetron Oscillator[6]	1000	—	—	—	—	—	—	—	4	650	10	1300	1.0	2J35
825[2,6]	50	6.3	0.75	Special	T-9C	Class-C Amp. (Grid-Mod.)	1500	800	3600	3600	−33	1.0	0.3	0.5	25	—	—	—	9.0	825
						Class-C Amp. (Telephony)	1500	800	3600	3600	−40	2.0	0.5	1.0	45	—	—	—	35	
410-R[3]		6.3		8-pin O.	T-9D	Oscillator-Amplifier or Frequency Multiplier	2500[4]	—	—	—[5]	—	—	—	—	—	—	—	—	20	410-R

[1] Transit-time split-plate type magnetron with internal circuit (approximate wavelength 10 cm.).
[2] Inductive-output amplifier (recommended for frequencies above 300 Mc.).
[3] Klystron (recommended for frequencies above 1000 Mc.).
[4] Collector anode voltage should be applied before applying grid voltage.
[5] Grid no. 2 (smoother grid) is electrically connected to collector anode.
[6] Focusing electromagnet (double lens) should be operated at approximately 1000 ampere turns.

TABLE XVII — TELEVISION TRANSMITTING TUBES

Type	Name	Socket Connections[1]	Heater Volts	Heater Amps.	Use	Collector Voltage	Pattern Electrode Voltage	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-off Grid Voltage[2]	Signal Plate Voltage	Collector Current µa.[3]	Beam Current µa.	Pattern Electrode Current[4]	Signal Plate Input[5]	Beam Resolution Capability[6]	Signal Output Volts	Type
1840	Orthicon	M	6.3	0.6	Direct and film pickup	—	—	300[7]	300	−40	—	—	1.0	—	—	—	0.03–0.15	1840
1847	Iconoscope	Fig. 1	6.3	0.6	Direct pickup	600	—	600	150	−120	—	—	—	—	—	—	—	1847
1848	Iconoscope	Fig. 25	6.3	0.6	Direct pickup	1200	—	1000[7]	300	−40	—	0.1	0.25	—	—	—	0.015–0.075	1848
1849[8]	Iconoscope	Fig. 21	6.3	0.6	Film pickup	1000	—	1000	360	−25	—	0.1	—	—	—	—	—	1849
1850	Iconoscope	Fig. 21	6.3	0.6	Direct pickup	Same as 1849												1850
1898	Monoscope	Fig. 20	2.5	2.1	Test pattern	—	950	1000	300	−60	—	—	2.0	2.0	—	—	—	1898
1899	Monoscope	Fig. 22	2.5	2.1	Test pattern	1700	1500	1500	390	−60	—	—	4.0	2.5	—	500	—	1899
2203	Monotron	Fig. 23	2.5	2.1	Test pattern	—	—	1000	400	−20	−150	—	—	—	5	300	0.1	2203

[1] Refer to Cathode Ray Tube Socket Connections.
[2] Adjust bias for minimum (most negative) value for satisfactory signal. Max. resistance in grid circuit should not exceed 1 meg.
[3] Collector current measurements made with mosaic not illuminated.
[4] Peak-to-peak signal value in µa.
[5] In mw./sq. cm. max.
[6] With full scanning.
[7] Accelerating electrode (Grid No. 2) voltage same as Anode No. 2 voltage.
[8] Obsolete.

Transmitting Tube Ratings

The ratings of transmitting tubes grouped in the transmitting tube tables are on the basis of the "absolute" system. This system enables the transmitter design engineer to choose his design values so as to obtain maximum performance within the tube ratings. Such design procedure has been considered practical for large transmitters where adequate controls are usually incorporated in the design and an experienced operator ordinarily is present to make any necessary adjustments.

In the absolute system, the maximum ratings shown for each type thus rated are limiting values above which the serviceability of the tube may be impaired from the viewpoint of life and satisfactory performance.

The maximum ratings given for each transmitting type on the foregoing data pages apply only when the tube is operated at frequencies lower than some specified value which depends on the design of the type. As the frequency is raised above the specified value, the radio-frequency current, dielectric losses, and heating effects increase rapidly. Most types can be operated above their specified maximum frequency provided the plate voltage and plate input are reduced in accordance with the information given by the asterisks on maximum operating frequencies for full ratings.

For certain air-cooled transmitting tubes, two sets of absolute maximum values are shown to meet diversified design requirements. One set is designated as CCS (Continuous Commercial Service) ratings, while the other is called ICAS (Intermittent Commercial and Amateur Service) ratings.

Continuous Commercial Service is defined as that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration. To meet these requirements, the CCS ratings have been established.

Intermittent Commercial and Amateur Service is defined to include the many applications where the transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life. These various factors have been taken into account in establishing the ICAS ratings.

Under the ICAS classification are such applications as the use of tubes in amateur transmitters, and the use of tubes in equipment where transmissions are of an intermittent nature. The term "intermittent" is used to identify operating conditions in all applications other than amateur in which no operating or "on" period exceeds 5 minutes and every "on" period is followed by an "off" or stand-by period of the same or greater duration.

ICAS ratings are considerably higher than CCS ratings. They permit the handling of greater power, but tube life under ICAS conditions, of course, is reduced. However, the transmitter designer may very properly decide that a small tube operated with ICAS ratings better meets his requirements than a larger tube operated with CCS ratings. Although such use involves some sacrifice in tube life, the period over which tubes will continue to give satisfactory performance in intermittent service can be extremely long depending on the exact nature of the service.

The equipment should be designed to operate the filament or heater of each tube type at rated normal value for full-load operating conditions under average voltage-supply conditions. Variations from this normal value due to voltage-supply fluctuation or other causes, should not exceed ± 5 per cent unless otherwise specified by the tube manufacturer.

The Catalog Section

In the following pages is a catalog-file of products of the principal manufacturers who serve the short-wave field. Appearance in these pages is by invitation—space has been sold only to those dependable firms whose established integrity and whose products have met with the approval of the American Radio Relay League.

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NATIONAL RADIO PRODUCTS

1945

DISTINGUISHED SERVICE IN TWO WORLD WARS

NATIONAL DIALS

The four-inch N Dial has an engine divided scale and vernier. The vernier is flush with the scale. The planetary drive has a ratio of 5 to 1, and is contained within the body of the dial. 2, 3, 4 or 5 scale. Fits ¼″ shaft. **Specify scale.**

N Dial — **List $7.50**

"Velvet Vernier" Dial, Type B, has a compact variable ratio 6 to 1 minimum, 20 to 1 maximum drive that is smooth and trouble free. An illuminator is available. The case is black bakelite. 1 or 5 scale. 4″ diam. Fits ¼″ shaft. **Specify scale.**

B Dial — **List $3.00**
Illuminator, extra — **List $.55**

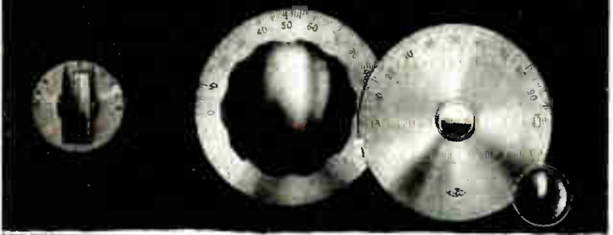
The original black bakelite "Velvet Vernier" Dial, Type A, is still an unchallenged favorite for general purpose use. The planetary drive has a ratio of 5 to 1. In 4 inch diameter with 2, 4 or 5 scale, and in 3⅜ inch diameter with 2 scale. Fits ¼″ shaft. **Specify scale.**

A Dial — **List $3.30**

The BM Dial is a smaller version of the B Dial (described in the opposite column) for use where space is limited. The drive ratio is fixed. Although small in size, the BM Dial has the same smooth action as the larger units. 1 or 5 scale. 3″ diam. Fits ¼″ shaft. **Specify scale.**

BM Dial — **List $2.75**

INEXPENSIVE DIALS



TYPE R
List $.85
1⅜″ Dia.
Etched Nickel Silver

TYPE O
List $1.65
3½″ Dia.

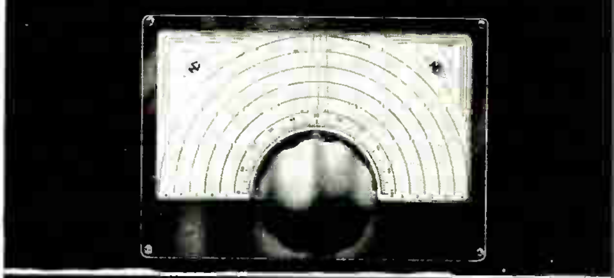
TYPE L
List $2.75
5″ Dia.

TYPE K
List $1.65
3½″ Dia.

TYPE M
List $2.75
5″ Dia.

R Dial scale 3 only but marked 10-0; O, K, L, M scale 2. All fit ¼″ shafts.

NEW! FOR INDIVIDUAL CALIBRATING



For experimenters who "build their own" and desire direct calibration. Fine for Freq. Monitors and ECO's.

- **Dial bezel size 5″ x 7¼″**
- **Five blank scales for direct calibration**
- **Employs Velvet Vernier Drive**
- **Easy to mount**

TYPE ACN — List $5.00

KNOBS

HRK (Fits ¼″ shaft) — **List $.95**
Black bakelite knob 2⅜″ diam.

HRP-P (Fits ¼″ shaft) — **List $.40**
Black bakelite knob 1¼″ long and ½″ wide. Equipped with pointer.

HRP — **List $.30**
The Type HRP knob has no pointer, but is otherwise the same as the knob above.

ACCESSORIES

ODL — **List $.55**
A locking device which clamps the rim of O, K, L and M Dials. Brass, nickel plated.

ODD — **List $.70**
Vernier drive for O, K, L, M or other plain dials.

SB (Fits ¼″ shaft) — **List $.30**
A nickel plated brass bushing ½″ diam.

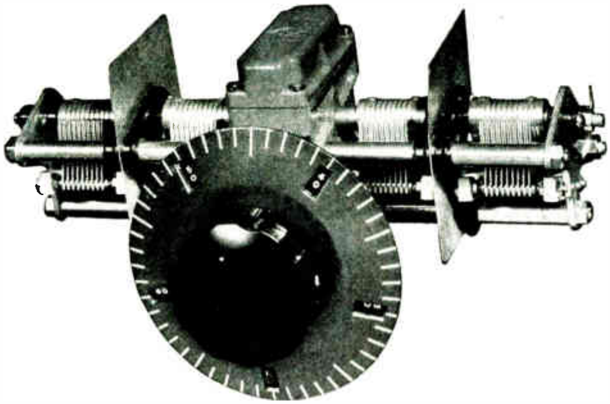
RSL (Fits ¼″ shaft) — **List $.95**
Rotor Shaft Lock for TMA, TMC and similar condensers.

DIAL SCALES

Scale	Divisions	Rotation	Direction of Condenser Rotation for Increase of dial reading
1	0-100-0	180°	Either
2	0-100	180°	Counter Clockwise
3	100-0	180°	Clockwise
4	150-0	270°	Clockwise
5	200-0	360°	Clockwise
6	0-150	270°	Counter Clockwise

• Priorities are required for all products in this catalog unless otherwise released by the War Production Board.

NATIONAL PRECISION CONDENSERS



The Micrometer dial reads direct to one part in 500. Division lines are approximately ¼″ apart. The dial revolves ten times in covering the tuning range, and the numbers visible through the small windows change every revolution to give consecutive numbering by tens from 0 to 500. The condenser is of extremely rigid construction, with four bearings on the rotor shaft. The drive, at the mid-point of the rotor, is through an enclosed preloaded worm gear with 20 to 1 ratio. Each rotor is individually insulated from the frame, and each has its own individual rotor contact. Stator insulation is Steatite. Plate shape is straight-line frequency when the frequency range is 2:1.

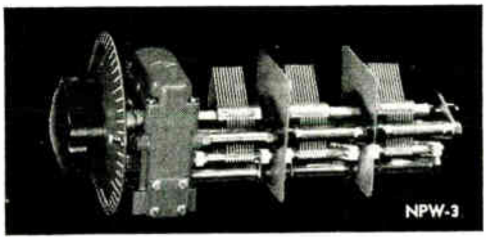
PW Condensers are available in 2, 3 or 4 sections, in either 160 or 225 mmf per section. Larger capacities cannot be supplied.

A single-section PW condenser with grounded rotor is supplied in capacities of 150, 200, 350 and 500 mmf single spaced, and capacities up to 125 mmf, double spaced.

PW condensers are all with rotor shaft parallel to the panel.

PW-1R	Single section right	**List $16.50**
PW-1L	Single section left	**List $16.50**
PW-2R	Double section right	**List $22.00**
PW-2L	Double section left	**List $22.00**
PW-2S	Single section each side	**List $22.00**
PW-3R	Double section right; single left	**List $26.50**
PW-3L	Double section left; single right	**List $26.50**
PW-4	Double section each side	**List $30.00**
PW-DO	Dial and knob only	**List $ 7.25**

NPW MODELS with micrometer dial

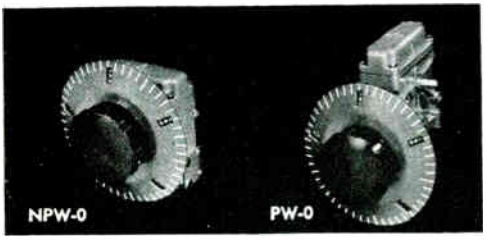


NPW-3. Three sections, each 225 mmf.
List $26.50

NPW-X. Three sections, each 25 mmf.
List $22.50

Both condensers are similar to PW models, except that rotor shaft is perpendicular to panel.

GEAR DRIVE UNITS with micrometer dial



NPW-O **List $12.00**

Uses parts similar to the NPW condenser. Drive shaft perpendicular to panel. One TX-9 coupling supplied.

PW-O **List $15.00**

Uses parts similar to the PW condenser. Drive shaft parallel to panel. Two TX-9 couplings supplied.

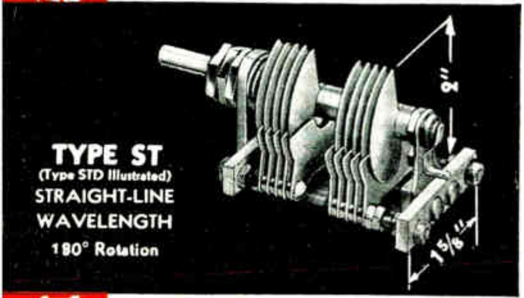
PW DIAL

PW dial, only, with eccentric **List Price $7.25**

PW dial is the same as that used on the condenser above. It revolves ten times in covering the complete range and as there is no gear reduction unit furnished, the driven shaft will revolve ten times also. The PW dial fits a shaft 5/16″ in diameter.

● Priorities are required for all products in this catalog until otherwise released by the War Production Board.

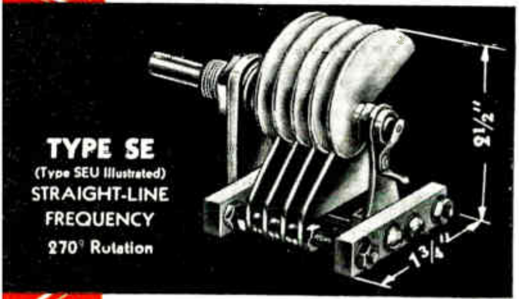
NATIONAL RECEIVING CONDENSERS



Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	List
SINGLE BEARING MODELS						
15 Mmf.	3 Mmf.	3	.018"	1 3/16"	STHS- 15	$1.50
25	3.25	4	.018"	1 3/16"	STHS- 25	1.65
50	3.5	7	.018"	1 3/16"	STHS- 50	1.75
DOUBLE BEARING MODELS						
35 Mmf.	6 Mmf.	8	.026"	2 1/4"	ST- 35	$1.65
50	7	11	.026"	2 1/4"	ST- 50	2.00
75	8	15	.026"	2 1/4"	ST- 75	2.25
100	9	20	.026"	2 1/4"	ST-100	2.50
140	10	27	.026"	2 3/4"	ST-140	2.75
150	10.5	29	.026"	2 3/4"	ST-150	2.75
200	12.0	27	.018"	2 1/4"	STH-200	3.00
250	13.5	32	.018"	2 3/4"	STH-250	3.30
300	15.0	39	.018"	2 3/4"	STH-300	3.50
335	17.0	43	.018"	2 3/4"	STH-335	4.00
SPLIT STATOR DOUBLE BEARING MODELS						
50–50	5–5	11–11	.026"	2 3/4"	STD- 50	$4.00
100–100	5.5–5.5	14–14	.018"	2 3/4"	STHD-100	5.00

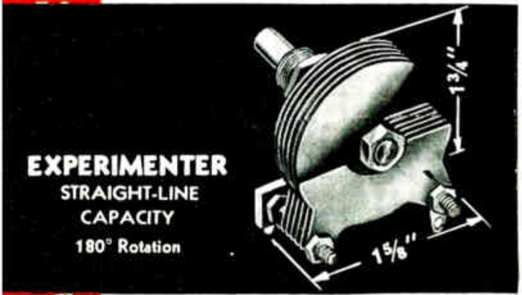
NOTE — Type SS Condensers, having straight-line-capacity plates but otherwise similar to the Type ST, are available. Capacities and Prices same as Type ST.

The ST Type condenser has Straight-Line Wavelength plates. All double-bearing models have the front bearing insulated to prevent noise. On special order a shaft extension at each end is available, for ganging. On double-bearing single shaft models, the rotor contact is through a constant impedance pigtail. Isolantite insulation.



Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	List
15 Mmf.	7 Mmf.	6	.055"	2 1/4"	SEU- 15	$2.75
20	7.5	8	.055"	2 1/4"	SEU- 20	3.00
25	8	9	.055"	2 1/4"	SEU- 25	3.00
50	9	11	.026"	2 1/4"	SE- 50	2.50
75	10	15	.026"	2 1/4"	SE- 75	2.75
100	11.5	20	.026"	2 1/4"	SE-100	3.00
150	13	29	.026"	2 3/4"	SE-150	3.25
200	12	27	.018"	2 1/4"	SEH-200	3.25
250	14	32	.018"	2 3/4"	SEH-250	3.50
300	16	39	.018"	2 3/4"	SEH-300	3.50
335	17	43	.018"	2 3/4"	SEH-335	3.85

TYPE SE — All models have two rotor bearings, the front bearing being insulated to prevent noise. A shaft extension at each end, for ganging, is available on special order. On models with single shaft extension, the rotor contact is through a constant impedance pigtail. The SEU models (illustrated) are suitable for high voltages as their plates are thick polished aluminum with rounded edges. Other SE condensers do not have polished edges on the plates. Isolantite insulation.



Capacity	Minimum Capacity	Length	Air Gap	No. of Plates	Catalog Symbol	List
15 Mmf.	3.5	1 5/16"	.045"	5	EX- 15	$ .95
25	3.75	1 5/16"	.045"	7	EX- 25	.95
35	3.75	1 5/16"	.045"	10	EX- 35	1.10
50	4	1 5/16"	.017"	6	EX- 50	1.00
100	4.75	1 5/16"	.017"	12	EX-100	1.10
140	5.5	1 5/16"	.017"	15	EX-140	1.40

The National "Experimenter" Type Condensers are low-priced models for general experimental work. They are of all-brass construction. The rotor has only one bearing. Plates can be removed without difficulty. Bakelite insulation.

• Priorities are required for all products in this catalog until otherwise released by the War Production Board.

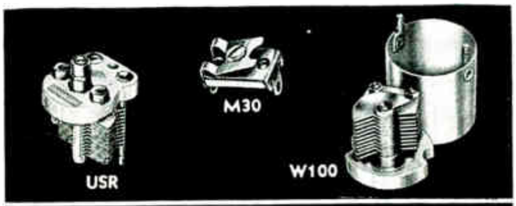
NATIONAL MINIATURE CONDENSERS

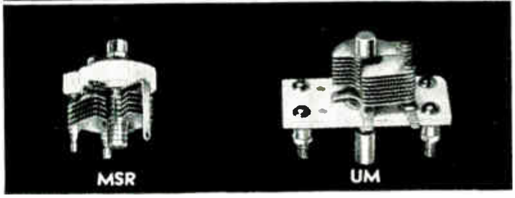
USR — *See table* — Type USR condensers are small, compact, low-loss units. Their soldered construction makes them particularly suitable for applications where vibration is present. Adjustment is made with a screw driver. Steatite base.

USE — *See table* — Type USE condensers are similar to Type USR, but are provided with a 1/4" diameter shaft extension at each end.

USL — *See table* — Type USL condensers are similar to Type USR, but are provided with a rotor shaft lock, so that the rotor can be clamped at any setting.

MSR, MSE, MSL — *See table* — Condensers of the MS series are similar in appearance to the US series described above, but they differ in making use of plates which are the same as those of the UM condenser. This and other small changes results in a more robust and rigid assembly. Other details of the MSR, MSE, and MSL are the same as the USR, USE, and USL respectively.





Capacity	Catalog Symbol			List
25 mmf.	USR-25	USE-25	USL-25	$1.45
50	USR-50	USE-50	USL-50	1.65
75	USR-75	USE-75	USL-75	1.90
100	USR-100	USE-100	USL-100	2.10
140	USR-140	USE-140	USL-140	2.50

Capacity	Catalog Symbol			List
25 mmf.	MSR-25	MSE-25	MSL-25	$1.45
50	MSR-50	MSE-50	MSL-50	1.65
75	MSR-75	MSE-75	MSL-75	1.90
100	MSR-100	MSE-100	MSL-100	2.10

Capacity	Minimum Capacity	No. of Plates	Air Gap	Catalog Symbol	List
15 mmf.	1.5	6	.017"	UM-15	$1.40
35	2.5	12	.017"	UM-35	1.65
50	3	16	.017"	UM-50	1.75
75	3.5	22	.017"	UM-75	1.90
100	4.5	28	.017"	UM-100	2.10
25	3.4	14	.042"	UMA-25	2.00
BALANCED STATOR MODEL					
25	2	4-4-4	.017"	UMB-25	$2.00

M-30 List $.35

Type M-30 is a small adjustable mica condenser with a maximum capacity of 30 mmf. Dimensions 13/16" x 9/16" x 1/2". Isolantite base.

W-75, 75 mmf. **List $2.50**
W-100, 100 mmf. **List $2.75**

Small padding condensers having very low temperature coefficient. Mounted in an aluminum shield 1 1/4" in diameter.

The **UM CONDENSER** is designed for ultra high frequency use and is small enough for convenient mounting in PB-10 and RO shield cans. They are particularly useful for tuning receivers, transmitters, and exciters. Shaft extensions at each end of the rotor permit easy ganging when used with one of our flexible couplings. The UMB-25 Condenser is a balanced stator model, two stators act on a single rotor. The UM can be mounted by the angle foot supplied or by bolts and spacers. See table for sizes.

Dimensions: Base 1" x 2 1/4", Mounting holes 5/8" x 1 23/32", Axial length 2 1/8" overall.

Plates: Straight line capacity, 180° rotation.

NATIONAL COMPANY, INC., MAL

NATIONAL NEUTRALIZING CONDENSERS

NC-600U List $.60
With standoff insulator

NC-600 List $.50
Without insulator

For neutralizing low power beam tubes requiring from .5 to 4 mmf, and 1500 max. total volts such as the 6L6. The NC-600U is supplied with a GS-10 standoff insulator screwed on one end, which may be removed for pigtail mounting.

STN List $2.00
The Type STN has a maximum capacity of 18 mmf (3000 V), making it suitable for such tubes as the 10 and 45. It is supplied with two standoff insulators.

TCN List $4.00
The Type TCN is similar to the TMC. It has a maximum capacity of 25 mmf (6000 V), making it suitable for the 203A, 211 and similar tubes.

NC-800 List $3.00
The NC-800 disk-type neutralizing condenser is suitable for the RCA-800, 35T, HK-54 and similar tubes. It is equipped with a micrometer thimble and clamp. The chart below gives capacity and air gap for different settings.

NC-75 List $4.50
For 75T, 808, 811, 812 & similar tubes.

NC-150 List $7.25
For HK354, RK36, 300T, 852, etc.

NC-500 List $13.75
For WE-251, 450TH, 450TL, 750TL, etc.

These larger disk type neutralizing condensers are for the higher powered tubes. Disks are aluminum, insulation steatite.

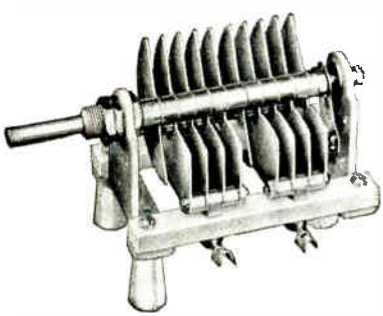
• Priorities are required for all products in this catalog until otherwise released by the War Production Board

NATIONAL TRANSMITTING CONDENSERS

TYPE TMS

is a condenser designed for transmitter use in low power stages. It is compact, rigid, and dependable. Provision has been made for mounting either on the panel, on the chassis, or on two stand-off insulators. Insulation is Isolantite. Voltage ratings listed are conservative.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
100 Mmf.	9.5	3″	.026″	1000v.	9	TMS-100	$2.75
150	11	3″	.026″	1000v.	14	TMS-150	3.00
250	13.5	3″	.026″	1000v.	22	TMS-250	3.30
300	15	3″	.026″	1000v.	27	TMS-300	4.00
35	8	3″	.065″	2000v.	7	TMSA-35	3.30
50	11	3″	.065″	2000v.	11	TMSA-50	3.60
DOUBLE STATOR MODELS							
50–50 Mmf.	6–6	3″	.026″	1000v.	5–5	TMS-50D	$4.25
100–100	7–7	3″	.026″	1000v.	9–9	TMS-100D	5.00
50–50	10.5–10.5	3″	.065″	2000v.	11–11	TMSA-50D	4.40



TYPE TMH

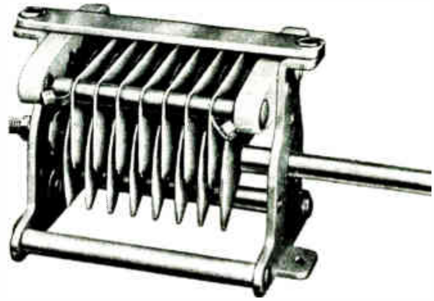
features very compact construction, excellent power factor, and aluminum plates .040″ thick with polished edges. It mounts on the panel or on removable stand-off insulators. Isolantite insulators have long leakage path. Stand-offs included in listed price.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List
SINGLE STATOR MODELS							
50 Mmf.	9	3¾″	.085″	3500v.	15	TMH-50	$3.85
75	11	3¾″	.085″	3500v.	19	TMH-75	4.40
100	12.5	5⅛″	.085″	3500v.	25	TMH-100	5.25
150	18	6½″	.085″	3500v.	37	TMH-150	6.60
35	11	5⅛″	.180″	6500v.	17	TMH-35A	5.75
DOUBLE STATOR MODELS							
35–35 Mmf.	6–6	3¾″	.085″	3500v.	9–9	TMH-35D	$6.00
50–50	8–8	5⅛″	.085″	3500v.	13–13	TMH-50D	6.60
75–75	11–11	6½″	.085″	3500v.	19–19	TMH-75D	8.00

• Priorities are required for all products in this catalog until otherwise released by the War Production Board.

NATIONAL TRANSMITTING CONDENSERS

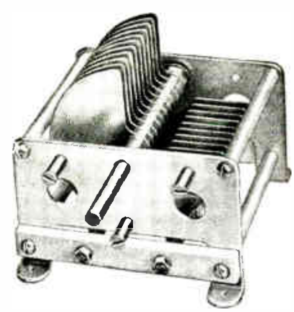
TYPE TMK



is a new condenser for exciters and low power transmitters. Special provision has been made for mounting AR-16 coils in a swivel plug-in mount on either the top or rear of the condenser, (see page 10). For panel or stand-off mounting. Isolantite insulation.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
35 Mmf.	7.5	2 7/32"	.047"	1500v.	7	TMK-35	$3.60
50	8	2 3/8"	.047"	1500v.	9	TMK-50	3.85
75	9	2 11/16"	.047"	1500v.	13	TMK-75	4.15
100	10	3"	.047"	1500v.	17	TMK-100	4.40
150	10.5	3 5/8"	.047"	1500v.	25	TMK-150	5.00
200	11	4 1/4"	.047"	1500v.	33	TMK-200	5.50
250	11.5	4 7/8"	.047"	1500v.	41	TMK-250	6.00
DOUBLE STATOR MODELS							
35–35 Mmf.	7.5–7.5	3"	.047"	1500v.	7–7	TMK-35D	$5.75
50–50	8–8	3 5/8"	.047"	1500v.	9–9	TMK-50D	6.50
100–100	10–10	4 1/4"	.047"	1500v.	17–17	TMK-100D	8.00
Swivel Mounting Hardware for AR 16 Coils						SMH	$ .15

TYPE TMC



is designed for use in the power stages of transmitters where peak voltages do not exceed 3000. The frame is extremely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are aluminum with buffed edges. Insulation is Isolantite. The stator in the split stator models is supported at both ends.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
50 Mmf.	10	3"	.077"	3000v.	7	TMC-50	$4.40
100	13	3 1/2"	.077"	3000v.	13	TMC-100	5.00
150	17	4 5/8"	.077"	3000v.	21	TMC-150	5.75
250	23	6"	.077"	3000v.	32	TMC-250	6.60
300	25	6 3/4"	.077"	3000v.	39	TMC-300	7.25
DOUBLE STATOR MODELS							
50–50 Mmf.	9–9	4 5/8"	.077"	3000v.	7–7	TMC-50D	$7.25
100–100	11–11	6 3/4"	.077"	3000v.	13–13	TMC-100D	8.25
200–200	18.5–18.5	9 1/4"	.077"	3000v.	25–25	TMC-200D	11.00

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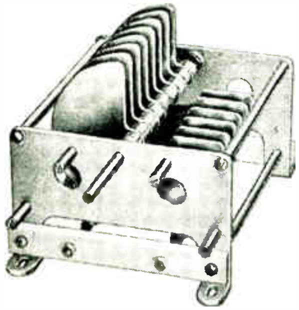
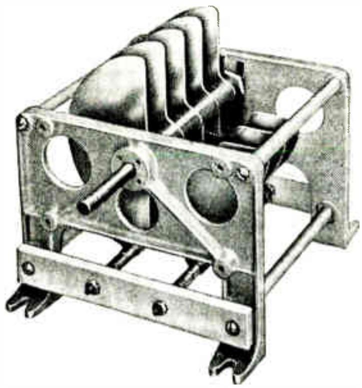
NATIONAL TRANSMITTING CONDENSERS



TYPE TMA

is a larger model of the popular TMC. The frame is extremely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are of heavy aluminum with rounded and buffed edges. Insulation is Isolantite, located outside of the concentrated field.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
			SINGLE STATOR MODELS				
300 Mmf.	19.5	4 9/16″	.077″	3000v.	23	TMA-300	$12.00
50	15	4 9/16″	.171″	6000v.	7	TMA-50A	6.50
100	19.5	6 7/8″	.171″	6000v.	15	TMA-100A	10.00
150	22.5	6 7/8″	.171″	6000v.	21	TMA-150A	12.00
230	33	9 5/16″	.171″	6000v.	33	TMA-230A	16.00
100	30	9 1/4″	.265″	9000v.	23	TMA-100B	13.50
150	40.5	12 1/2″	.265″	9000v.	33	TMA-150B	17.00
50	21	7 1/8″	.359″	12000v.	13	TMA-50C	8.00
100	37.5	12 7/8″	.359″	12000v.	25	TMA-100C	14.50
			DOUBLE STATOR MODELS				
200-200 Mmf.	15-15	6 7/8″	.077″	3000v.	16-16	TMA-200D	$15.00
50-50	12.5-12.5	6 7/8″	.171″	6000v.	8-8	TMA-50DA	11.00
100-100	17-17	9 5/16″	.171″	6000v.	14-14	TMA-100DA	17.50
60-60	19.5-19.5	12 1/2″	.265″	9000v.	15-15	TMA-60DB	18.50
40-40	18-18	12 7/8″	.359″	12000v.	11-11	TMA-40DC	13.50



TYPE TML

condenser is a 1 KW job throughout. Isolantite insulators, specially treated against moisture absorption, prevent flashovers. A large self-cleaning rotor contact provides high current capacity. Thick capacitor plates, with accurately rounded and polished edges, provide high voltage ratings. Sturdy cast aluminum end frames and dural tie bars permit an unusually rigid structure. Precision end bearings insure smooth turning and permanent alignment of the rotor. End frames are arranged for panel, chassis or stand-off mountings.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
			SINGLE STATOR MODELS				
75 Mmf.	25	18 1/16″	.719″	20,000v.	17	TML-75E	$28.75
150	60	18 1/16″	.469″	15,000v.	27	TML-150D	29.00
100	45	13 5/8″	.469″	15,000v.	19	TML-100D	26.00
50	22	8 5/16″	.469″	15,000v.	9	TML-50D	18.00
245	54	18 1/16″	.344″	10,000v.	35	TML-245B+	31.50
150	45	13 5/8″	.344″	10,000v.	21	TML-150B+	28.75
100	32	10 15/16″	.344″	10,000v.	15	TML-100B+	27.50
75	23.5	8 5/16″	.344″	10,000v.	11	TML-75B+	20.00
500	55	18 1/16″	.219″	7,500v.	49	TML-500A+	38.50
350	45	13 5/8″	.219″	7,500v.	33	TML-350A+	30.75
250	35	10 15/16″	.219″	7,500v.	25	TML-250A+	28.75
			DOUBLE STATOR MODELS				
30-30 Mmf.	12-12	18 1/16″	.719″	20,000v.	7-7	TML-30DE	$29.00
60-60	26-26	18 1/16″	.469″	15,000v.	11-11	TML-60DD	31.50
100-100	27-27	18 1/16″	.344″	10,000v.	15-15	TML-100DB+	35.00
60-60	20-20	13 5/8″	.344″	10,000v.	9-9	TML-60DB+	30.00
200-200	30-30	18 1/16″	.219″	7,500v.	21-21	TML-200DA+	38.50
100-100	17-17	10 15/16″	.219″	7,500v.	11-11	TML-100DA+	31.50

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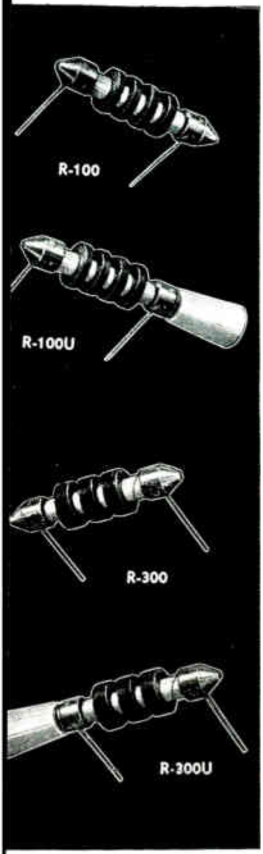
NATIONAL RF CHOKES



R-100 List $.50
Without standoff insulator

R-100U List $.60
With standoff insulator

R.F. chokes R-100 and R-100U are identical electrically, but the latter is provided with a removable standoff insulator screwed on one end. Both have Isolantite insulation and both have a continuous universal winding in four sections. Inductance 2½ m.h.; distributed capacity 1 mmf.; DC resistance 50 ohms; current rating 125 ma.

R-300 List $.50
Without insulator

R-300U List $.60
With insulator

R.F. chokes R-300 and R-300U are similar in size to R-100U but have higher current capacity. The R-300U is provided with a removable standoff insulator screwed on one end. Inductance 1 m.h.; distributed capacity 1 mf.; DC resistance 10 ohms; current rating 300 ma.

R.F. chokes are available in a variety of inductance values, ranging from 6 microhenries to 10 millihenries, in addition to those shown above. Various mounting arrangements are also available. Full information will be furnished on request.

R-152 List $2.50

For the 80 and 160 meter bands. Inductance 4 m.h., DC resistance 10 ohms, DC current 600 ma. Coils honeycomb wound on Isolantite core.

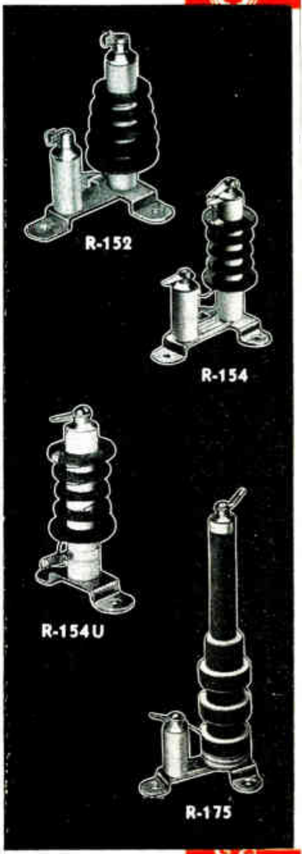
R-154 List $2.50

R-154U List $2.00

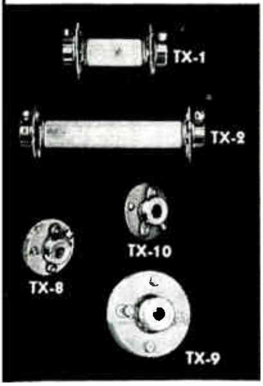
For the 20, 40 and 80 meter bands. Inductance 1 m.h., DC resistance 6 ohms, DC current 600 ma. Coils honeycomb wound on Isolantite core. The R-154U does not have the third mounting foot and the small insulator, but is otherwise the same as R-154. See illustration.

R-175 List $3.00

The R-175 Choke is suitable for parallel-feed as well as series-feed in transmitters with plate supply up to 3000 volts modulated or 4000 volts unmodulated. Unlike conventional chokes, the reactance of the R-175 is high throughout the 10 and 20 meter bands as well as the 40, 80 and 160 meter bands. Inductance 225 µh, distributed capacity 0.6 mmf., DC resistance 6 ohms, DC current 800 ma., voltage breakdown to base 12,500 volts.



NATIONAL SHAFT COUPLINGS



TX-1, Leakage path 1″ List $1.10

TX-2, Leakage path 2½″ List $1.25

Flexible couplings with glazed Isolantite insulation which fit ¼″ shafts.

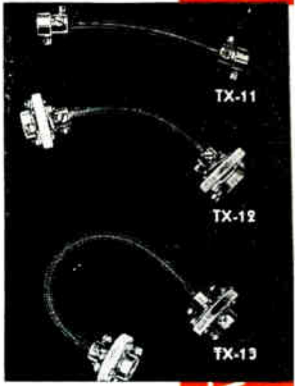
TX-8 List $.85
A non-flexible rigid coupling with Isolantite insulation. 1″ diam. Fits ¼″ shaft.

TX-9 List $1.25
This small insulated flexible coupling provides high electrical efficiency when used to isolate circuits. Insulation is Steatite. 1⅝″ diam. Fits ¼″ shaft.

TX-10 List $.60
A very compact insulated coupling free from backlash. Insulation is canvas Bakelite. 1 1/16″ diam. Fits ¼″ shaft.

TX-11 List $.70
The flexible shaft of this coupling connects shafts at angles up to 90 degrees, and eliminates misalignment problems. Fits ¼″ shafts. Length 4¼″.

TX-12, Length 4⅝″ List $1.40
TX-13, Length 7⅛″ List $1.65
These couplings use flexible shafting like the TX-11 above, but are also provided with Isolantite insulators at each end.

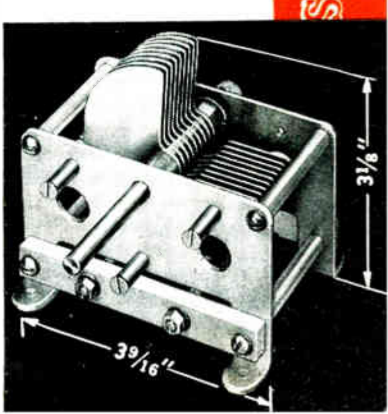


All prices subject to change without notice

NATIONAL GENERAL PURPOSE CONDENSERS

National EMC Condensers are made in large sizes for general purpose uses. They are similar in construction to the TMC Transmitting condenser, and have high efficiency and rugged frames. Insulation is Isolantite, and Peak Voltage Rating is 1000 Volts. Plate shape is Straight-Line Wavelength.

Capacity	Minimum Capacity	No. of Plates	Length	Catalog Symbol	List
150 Mmf.	9	9	4″	EMC-150	$4.50
250	11	15	2 15/16″	EMC-250	5.50
350	12	20	2 15/16″	EMC-350	6.75
500	16	29	4 3/8″	EMC-500	8.50
1000	22	56	6 1/4″	EMC-1000	12.50
SPLIT-STATOR MODEL					
350-350	12-12	20-20	6″	EMCD-350	$13.00



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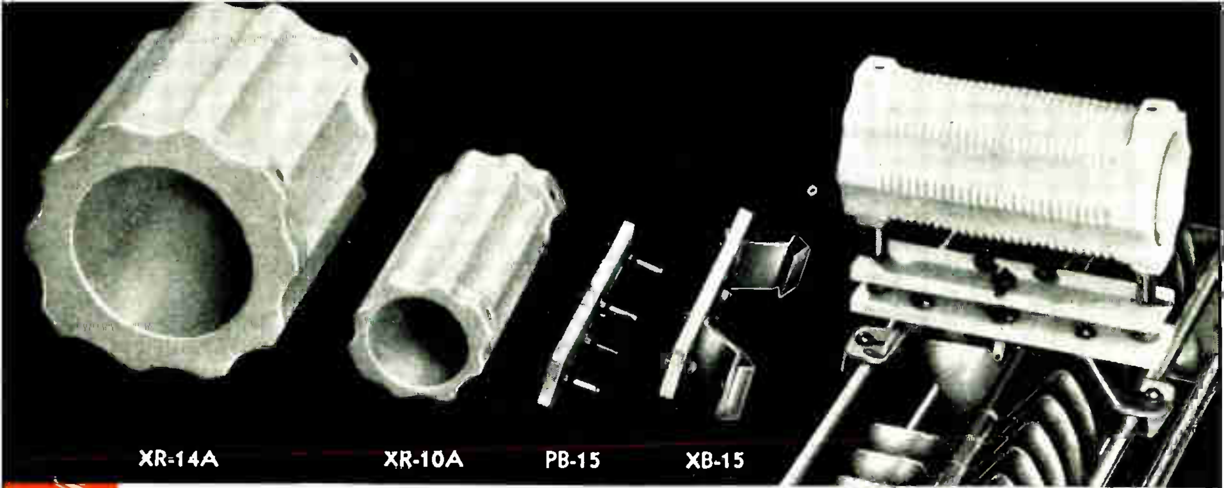


NATIONAL COM[illegible] MALDEN, M

TRANSMITTER COIL FORMS

The Transmitter Coil Forms and Mounting are designed as a group, and mount conveniently on the bars of a TMA condenser. The larger coil form, Type XR-14A, has a winding diameter of 5″, a winding length of 3¾″ (30 turns total) and is intended for the 80 meter band. The smaller form, Type XR-10A, has a winding length of 3¾″ and a winding diameter of 2½″ (26 turns total). It is intended for the 20 and 40 meter bands.

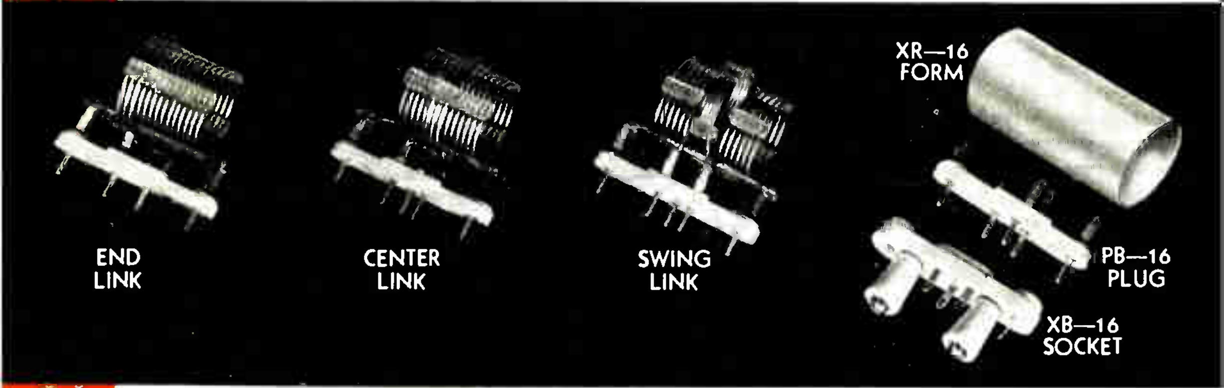
Either coil form fits the PB-15 plug. For higher frequencies, the plug may be used with a self-supporting coil of copper tubing. The XB-15 Socket may be mounted on breadboards or chassis, as well as on the TMA Condenser.

SINGLE UNITS

XR-10A, Coil Form only	List $1.65
XR-14A, Coil Form only	List $4.00
PB-15, Plug only	List $1.50
XB-15, Socket only	List $2.00

ASSEMBLIES

UR-10A, Assembly (including small Coil Form, Plug and Socket)	List $5.00
UR-14A, Assembly (including large Coil Form, Plug and Socket)	List $7.00



EXCITER COILS AND FORMS — TYPE AR-16 (Air Spaced)

TEMPORARILY DISCONTINUED

These air-spaced coils are suitable for use in stages where the plate input does not exceed 50 watts and are available in the sizes tabulated below. Capacities listed will resonate the coils at the low frequency end of the band and include all stray circuit capacities. All have separate link coupling coils and all fit the PB-16 Plug and XB-16 Socket.

The XR-16 Coil Form also fits the PB-16 Plug and XB-16 Socket. It has a winding diameter of 1¼″ and a winding length of 1¾″.

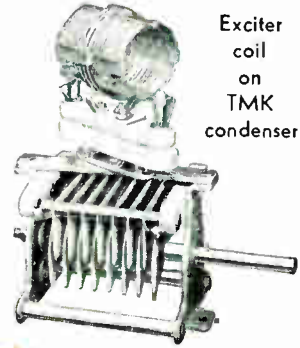
XR-16, Coil Form only	List $.70
PB-16, Plug-in Base only	List $.45
XB-16, Plug-in Socket only	List $.55
AR-16 Coils — Any type (see table) Including PB-16 Plug as illustrated. Each,	List $1.65

TEMPORARILY DISCONTINUED

Order by Catalog Symbol Shown in This Table

Band	End Link	Cap Mmf	Center Link	Cap Mmf	Swinging Link	Cap Mmf
5 meter	AR16-5E	20	AR16-5C	20	——	—
10 meter	AR16-10E	20	AR16-10C	20	AR16-10S	25
20 meter	AR16-20E	26	AR16-20C	26	AR16-20S	40
40 meter	AR16-40E	33	AR16-40C	33	AR16-40S	55
80 meter	AR16-80E	37	AR16-80C	37	AR16-80S	60
160 meter	AR16-160E	65	AR16-160C	65	——	—

TEMPORARILY DISCONTINUED



Exciter coil on TMK condenser

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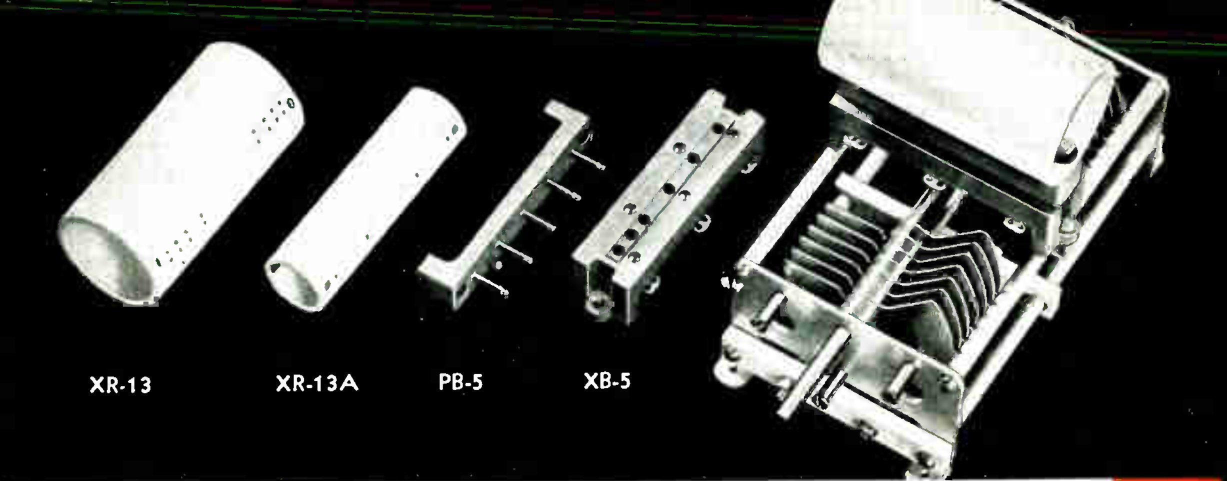


BUFFER COIL FORMS

National Buffer Coil Forms are designed to mount directly on the tie bars of a TMC condenser using the PB-5 Plug and XB-5 Socket. Plug and Socket are of molded R-39.

The two coil forms are of Isolantite, left unglazed to provide a tooth for coil dope. The larger form, Type XR-13, is 1¾" in diameter and has a winding length of 2¾". The smaller form, Type XR-13A, is 1" in diameter and provides a winding length of 2¾". Both forms have holes for mounting and for leads.

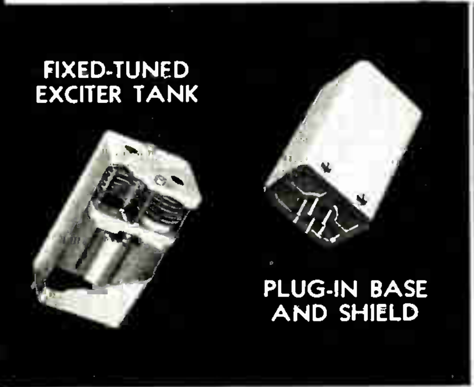
SINGLE UNITS

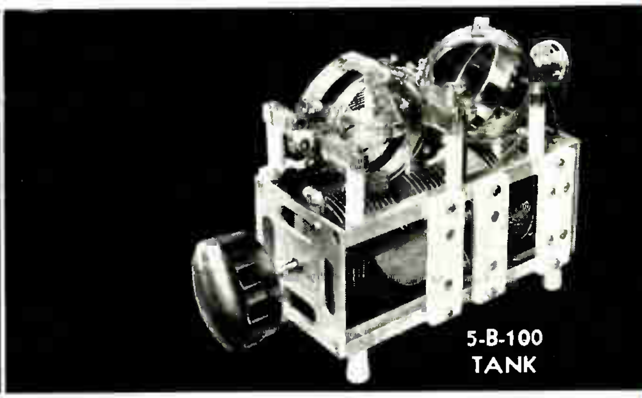
XR-13, Coil Form only	**List $1.25**
XR-13A, Coil Form only	**List $ .70**
PB-5, Plug only	**List $ .85**
XB-5, Socket only	**List $ .85**

ASSEMBLIES

UR-13A, Assembly (including small Coil Form, Plug and Socket) **List $2.25**

UR-13, Assembly (including large Coil Form, Plug and Socket) **List $2.75**





FIXED TUNED EXCITER TANK

Similar in general construction to National I.F. transformers, this unit has two 25 mmf., 2000 volt air condensers and an unwound XR-2 coil form.

FXT, without plug-in base	**List $5.00**
FXTB-5, with 5 prong base	**List $5.50**
FXTB-6, with 6 prong base	**List $5.50**

PLUG-IN BASE AND SHIELD

The low-loss R-39 base is ideal for mounting condensers and coils when it is desirable to have them shielded and easily removable. Shield can is 2" x 2⅞" x 4⅛".

PB-10-5, (5 Prong Base & Shield)	**List $.85**
PB-10-6, (6 Prong Base & Shield)	**List $.85**
PB-10A-5, (5 Prong Base only)	**List $.45**
PB-10A-6, (6 Prong Base only)	**List $.45**

5-B-100 TANK

TEMPORARILY DISCONTINUED

The National 5-B-100 is a complete tank circuit (including coils, condenser and R.F. choke), which tunes through five amateur bands with a single dial. The tank replaces the tuning condenser, set of five plug-in coils, plug-in coil socket and R.F. choke, without sacrificing efficiency or space, yet it costs no more.

The 5-B-100 is actually more compact than a tuning condenser and mounted plug-in coil for the same power capabilities. In addition to the compactness and wide tuning range advantages of the 5-B-100, the tank provides for the first time a real constant L/C ratio throughout the tuning range. Harmonics from the low-frequency bands are suppressed without sacrifice of efficiency on the high-frequency bands. Constant link loading or capacity coupling may be used.

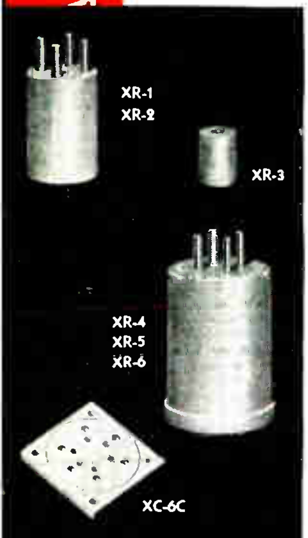
The 5-B-100 is an ideal plate tank for R.F. amplifiers using such tubes as 35T, 809, 811, 812, RK-11, RK-12, HK-24, HY-30Z, HY-51Z, etc. with input up to 1[illegible]0 watts (1250 volts unmodulated or 750 volts modulated maximum). Also ideal for grid tank of amplifiers up to 2 KW plate input.

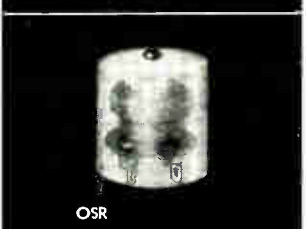
Four mounting insulators are supplied on the base. Overall dimensions are 4 inches wide, 6 inches high and 8 inches deep. Shipping weight, 5 lbs.

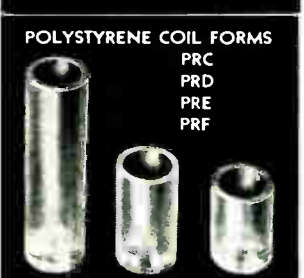
5-B-100 Tank, List $40.00

COMPANY
DEN, MASS., U.S.A.

NATIONAL PARTS







COIL FORMS

XR-1, Four prong, **List $.55**
XR-2, without prongs **List $.40**

Molded of R-39, permitting them to be grooved and drilled. Coil form diameter 1″, length 1½″.

XR-3 **List $.35**

Molded of R-39. Diameter 9⁄16″, length ¾″. Without prongs.

XR-4, Four prong, **List $.85**
XR-5, Five prong, **List $.85**
XR-6, Six prong, **List $.85**

Molded of R-39, permitting them to be grooved and drilled. Coil form diameter 1½″, length 2¼″. A special socket is required for the six-prong form.

XC6C, Special six-prong socket for XR-6 Coil Form, **List $.85**

IMPEDANCE COUPLER

S-101 **List $6.60**

A plate choke, coupling condenser and grid leak sealed in one case, for coupling the output of a regenerative detector to an audio stage. Used in SW-3U.

OSCILLATOR COIL

OSR **List $1.65**

A shielded oscillator coil which tunes to 100 KC with .00041 Mfd. Two separate inductances, closely coupled. Excellent for interruption-frequency oscillator in super-regenerative receivers.

H. F. COIL FORMS

Symbol	Outside Diameter	Length	List
PRC-1	⅜″	⅜″	$.20
PRC-2	⅜″	½″	.20
PRC-3	⅜″	¾″	.20
PRD-1	½″	½″	.20
PRD-2	½″	1″	.20
PRE-1	9⁄16″	¾″	.25
PRE-2	9⁄16″	1″	.25
PRE-3	9⁄16″	2″	.35
PRF-1	¾″	¾″	.35
PRF-2	¾″	1¼″	.45

COIL SHIELDS

RO, coil shield **List $.40**
2″ x 2⅜″ x 4⅛″ high

J30, coil shield **List $.40**
2½″ dia. x 3¾″ high

B30, coil shield **List $.40**
3″ dia. x 3¾″ high without mounting base.

B30-B, coil shield **List $.55**
Same as above, but with mounting base.

TUBE SHIELDS

TS, tube shield **List $.45**
With cap and base.

T58, tube shield **List $.45**
With cap and base, for 77, 78, etc. tubes.

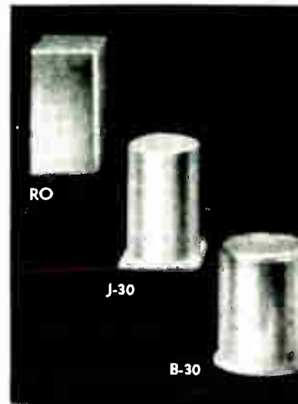
T78, tube shield **List $.45**
With cap and base, for 77, 78, etc. tubes.

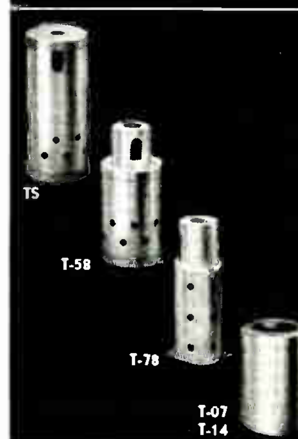
T14, tube shield **List $.45**
2⅛″ high, for 814, RK-20, etc.

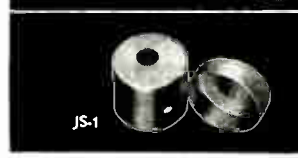
T07, tube shield **List $.45**
3″ high, for 807, RK-23, etc.

JACK SHIELD

JS-1, Jack shield **List $.40**
For shielding small standard jacks mounted behind a panel, or on the ends of extension cords.







NATIONAL CABINETS

The National Cabinets listed below are the same as those used in National Receivers, except that they are supplied in blank form. They are made of heavy gauge steel, and the paint is unusually well bonded to the metal. Sub-bases and bottom covers are included in the price.

	Width	Height	Depth	List Price
Type C-SW3	9¾″	7″	9″	$6.00
Type C-NC100	17¼″	8¾″	11¼″	9.50
Type C-HRO	16¾″	8¾″	10″	9.50
Type C-One-Ten	11″	7″	7¼″	5.00
Type C-SRR	7½″	7″	7½″	4.00

NATIONAL CABINETS

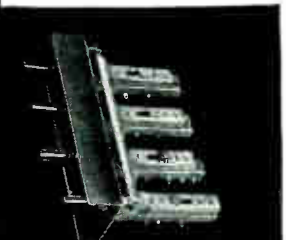

ACS-4


CHART FRAMES


COIL DOPE


NDC-8
NDC-10


NDC-2

PUSH SWITCH

ACS-4, Four gang, with trigger bar List $5.50

TEMPORARILY DISCONTINUED

ACS-1, Single section, less trigger bar List $1.40

The National Interlocking Push Switch has low losses, complete reliability and positive contacts. Insulation is R-39. The silverplated contacts are double pole, double throw.

CHART FRAME

The National Chart Frame is blanked from one piece of metal, and includes a celluloid sheet to cover the chart. Size 2¼″ x 3¼″, with sides ¼″ wide.

Type CFA List $.55

COIL DOPE

TEMPORARILY DISCONTINUED

CD-1 ... List $1.65

Liquid Polystyrene Cement — is ideal for winding coils, as it will not spoil the properties of the best coil form.

SPEAKER CABINETS

NDC-8 for 8″ speaker List $5.50

NDC-10 for 10″ speaker List $6.60

NDC-2 for 10″ speaker List $8.50

These metal speaker cabinets are acoustically correct. They are lined with acoustic felt, and are of welded construction to eliminate rattles. Finish is black wrinkle on NDC-8 and NDC-10. NDC-2 is finished in two-tone gray to match the NC-200 TG receiver.

I. F. TRANSFORMERS

IFC, Transformer, air core List $5.50

IFCO, Oscillator, air core List $5.50

Air dielectric condensers isolated from each other by an aluminum shield. Litz wound coils on a moisture proofed ceramic base. Shield can 4⅛″ x 2⅜″ x 2″. Available for either 175 KC or 450–550 KC. Specify frequency.

IFD, Diode Transformer, air core List $3.85

TEMPORARILY DISCONTINUED

Tuned primary and tuned, closely-coupled secondary for full-wave diode rectifier for noise silencing circuits, etc. 450–550 KC, air core only.

IFE, Transformer. Same as IFC but iron core, 450–550 KC only List $5.50

NATIONAL HIGH FIDELITY TRF UNITS

Each chassis provides a three-stage RF Amplifier tuned to one station only.

Each RF Transformer is tuned both primary and secondary (8 tuned circuits). The coupling is adjustable to include 10 KC with less than 1 db variation in the ... Sensitivity is adjustable from 5 microvolts to one volt. Three models cover ranges ... 250, and 1100–1700 KC. The chassis fits a standard 3½″ relay rack panel.

TEMPORARILY DISCONTINUED

DLUS, Tuner, wired and tested unit on ⅛″ steel, wrinkle finish, less tubes, List $82.50

DLUA, Tuner, same as DLUS but has 3/16″ aluminum panel, crackle finish, less tubes List $86.50

DLCA, Chassis as illustrated with sockets and terminals riveted in place List $5.00

DLPS, Steel ⅛″ panel List $1.65

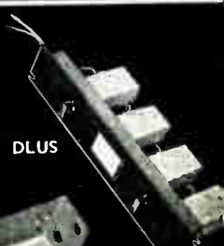
DLPA, Aluminum 3/16″ panel List $5.50

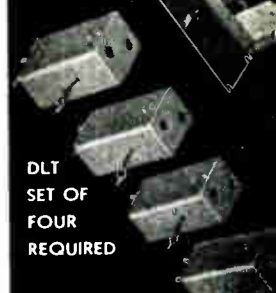
DLT, RF Transformer, set of four required List, each $7.25

(Specify operating frequency)


IFC


IFD


DLUS


DLT SET OF FOUR REQUIRED


DLCA

National Oscilloscopes have power supply and input controls built in. A panel switch permits use of the built-in 60-cycle sweep or external audio sweep for securing the familiar trapezoid pattern for modulation measurements.

CRM, less tubes List $21.00

1″ screen, using RCA-913 and 6X5 rectifier. Table model, 4¼″ x 6⅛″ x 8″.

CRR, less tubes List $35.00

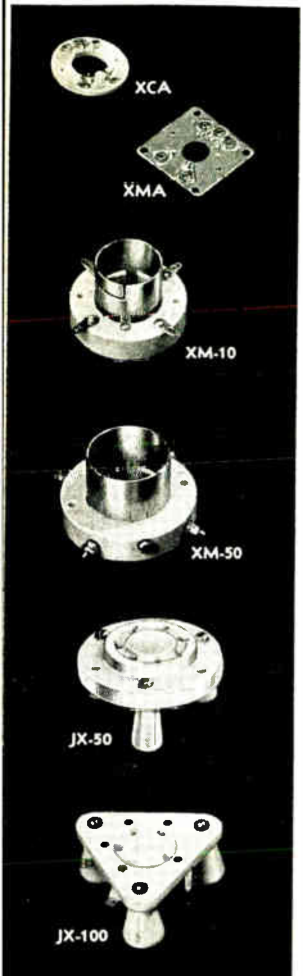
2″ screen, using RCA-902 and 6X5 rectifier. Relay rack mounting.

NATIONAL OSCILLOSCOPES


CRM


CRR

NATIONAL LOW-LOSS SOCKETS AND INSULATORS



XCA List $1.65

A low-loss socket for acorn triodes.

XMA List $2.20

For pentode acorn tubes, this socket has built-in by-pass condensers. The base is a copper plate.

XM-10 List $1.50

A heavy duty metal shell socket for tubes having the UX base.

XM-50 List $2.00

A heavy duty metal shell socket for tubes having the Jumbo 4-pin base ("fifty watters").

JX-50 List $1.35
Without Standoff Insulators

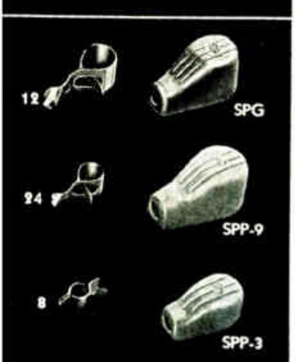
JX-50S List $1.65
With Standoff Insulators

A low-loss wafer socket for the 813 and other tubes having the Giant 7-pin base.

JX-100 List $3.30
Without Standoff Insulators

JX-100S List $4.00
With Standoff Insulators

A low-loss wafer socket for the 803, RK-28 and other tubes using the Giant 5-pin base.



SAFETY GRID & PLATE CAPS

SPG List $.40

9/16" Cap, R-39 L. L. insulation. These offer protection against accidental contact with High Voltage lobe caps.

SPP-9 List $.40

9/16" Cap L. L. ceramic insulation.

SPP-3 List $.35

3/8" Cap L. L. ceramic insulation

GRID & PLATE GRIPS

12, for 9/16" Caps List $.10
24, for 3/8" Caps List $.05
8, for 1/4" Cap List $.05
12 & **24** suitable for glass tubes
8 is for metal tubes

GS-1, 1/2" x 1 3/8" List $.40

GS-2, 1/2" x 2 7/8" List $.50

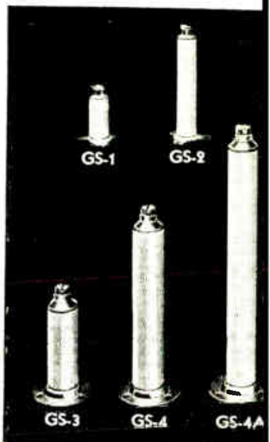
GS-3, 3/4" x 2 7/8" List $1.00

GS-4, 3/4" x 4 7/8" List $1.25

GS-4A, 3/4" x 6 7/8" List $1.75

Cylindrical low-loss steatite standoff insulators with nickel plated caps and bases.

GSJ, (not illustrated) List $.10
A special nickel plated jack top threaded to fit the 3/4" diameter insulators GS-3, GS-4 & GS-4A.



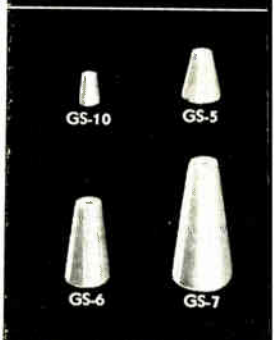
GS-5, 1 1/4" List, each $.40

GS-6, 2" List, each $.70

GS-7, 3" List, each $1.25

GS-10, 3/4", package of 10 List $1.20

These cone type standoff insulators are of low-loss steatite. They have a tapped hole at each end for mounting.



GS-8, with terminal List $.90

GS-9, with Jack List $1.25

These low-loss steatite standoff insulators are also useful as lead-through bushings.

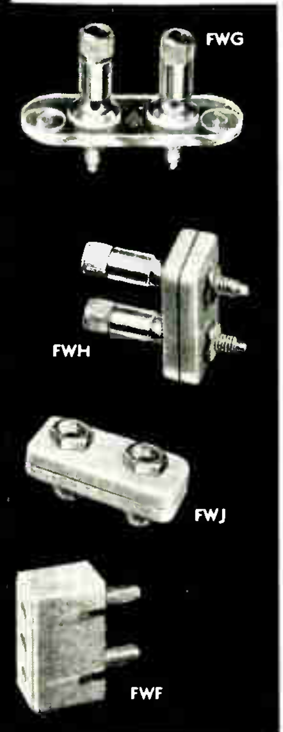
XC Series Sockets

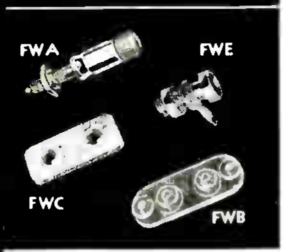
XC-4	List $.60
XC-5	List $.65
XC-6	List $.70
XC-7S	List $.75
XC-7L	List $.75
XC-8	List $.65

National wafer sockets have exceptionally good contacts with high current capacity together with low loss Isolantite insulation. All types have a locating groove to make tube insertion easy.

• *Priorities are required for all products in this catalog until otherwise released by the War Production Board*

NATIONAL LOW-LOSS SOCKETS AND INSULATORS





FWG — **List $.70**

A Victron terminal strip for high frequency use. The binding posts take banana plugs at the top, and grip wires through hole at the bottom, simultaneously, if desired.

FWH — **List $.95**

The insulators of this terminal assembly are molded R-39 and have serrated bosses that allow the thinnest panel to be gripped firmly, and yet have ample shoulders. Binding posts same as FWG above.

FWJ — **List $.75**

This assembly uses the same insulators as the FWH above, but has jacks. When used with the FWF plug (below), there is no exposed metal when the plug is in place.

FWF — **List $1.10**

This molded R-39 plug has two banana plugs on ¾" centers and fits FWH or FWJ above. Leads may be brought out through the top or side.

FWA, Post — **List, each $.30**
Brass Nickel Plated

FWE, Jack — **List, each $.20**
Brass Nickel Plated

FWC, Insulator — **List, per pair $.40**
R-39 Insulation

FWB, Insulator — **List, each $.10**
Polystyrene insulation

CIR Series Sockets

Any Type **List $.45**

Type CIR Sockets feature low-loss isolantite or steatite insulation, a contact that grips the tube prong for its entire length, and a metal ring for six position mounting. The sockets are supplied with two metal standoffs.

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AA-3 — **List $.60**

A low-loss steatite spreader for 6 inch line spacing. (600 ohms impedance with No. 12 wire.)

AA-5 — **List $.50**

A low-loss steatite aircraft-type strain insulator.

AA-6 — **List $.90**

A general purpose strain insulator of low-loss steatite.

XS-6 — **List, each $.20**

A low-loss isolantite bushing for ½" holes.

XP-6

Same as above but Victron. **List, box of ten $.85**

TPB — **List, per dozen $.85**

A threaded polystyrene bushing with removable .093 conductor moulded in, ¼" diam., 32 thread.

XS-7, (⅜" Hole) — **List $.55**
XS-8, (½" Hole) — **List $.75**

Steatite bushings. Prices include male and female bushings with metal fittings.

XS-1, (1" Hole) — **List $1.20**
XS-2, (1½" Hole) — **List $1.35**

Prices listed are per pair, including metal fittings. Insulation steatite.

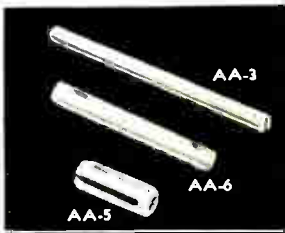
XS-3, (2¾" Hole) — **List $6.00**
XS-4, (3¾" Hole) — **List $7.25**

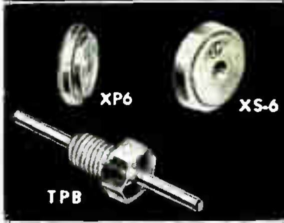
Prices are per pair, including metal fittings. These low-loss steatite bowls are ideal for lead-in purposes at high voltages.

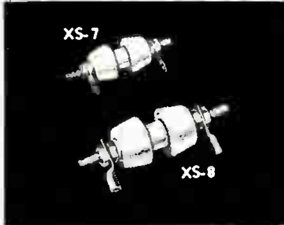
XS-5, Without Fittings — **List, each $8.25**

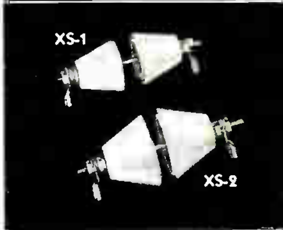
XS-5F, With Fittings — **List, per pair $17.00**

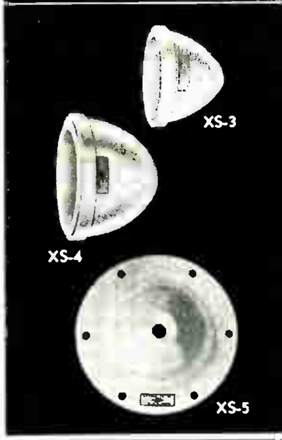
These big low-loss bowls have an extremely long leakage path and a 5¼" flange for bolting in place. Insulation steatite.

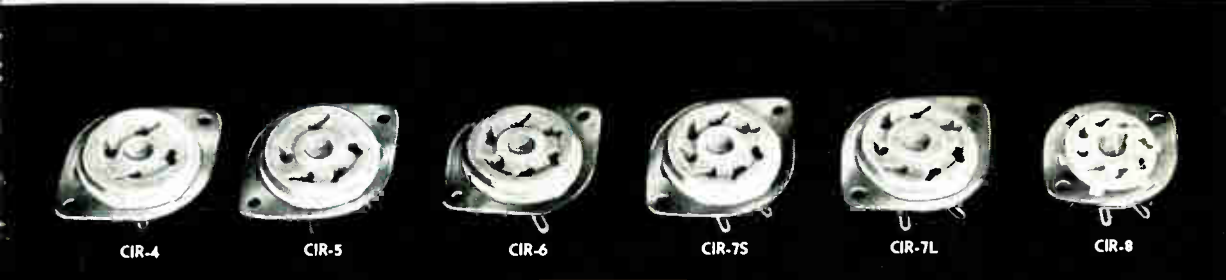




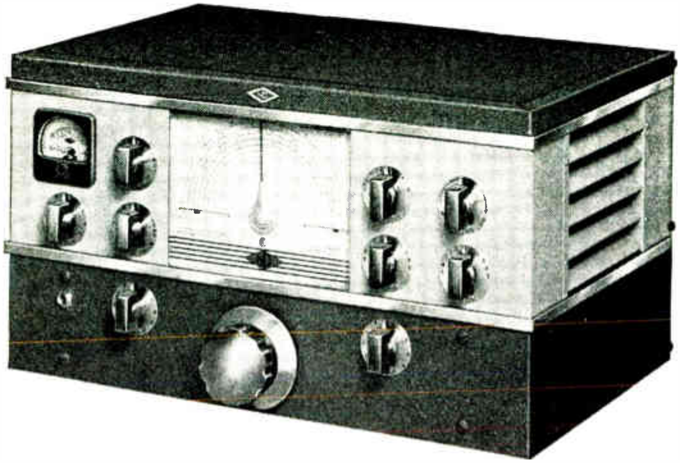








NATIONAL NC-200



The National NC-200 is a new communications receiver having a number of features not previously available. Twelve tubes are used in a highly perfected circuit that includes an extremely effective noise limiter. The crystal filter has an exceptionally wide selectivity range for use on both CW and phone, as well as a phasing circuit that makes rejection ratios as high as 10,000 to 1 available even when the interfering signal is only a few hundred cycles from the desired signal. The AVC holds the audio constant within 2 db for signals from 10 microvolts to 100,000 microvolts. The sensitivity of the NC-200 is particularly high, requiring only 1 microvolt input for 1 watt of audio output on the highest frequencies covered by the receiver. Signal-to-image ratio is better than 30 db at ten meters.

There are ten calibrated coil ranges, each with its own scale on the direct-reading dial. Six of these ranges provide continuous coverage from 490 KC to 30 MC. The remaining four ranges cover the 10, 20, 40 and 80 meter bands, each of which is spread over the major portion of the dial scale. Ranges are selected by a panel control knob. A movable-coil system similar to the NC-100 is used. The inertia-type dial drive has a ratio of about 20 to 1.

All models of the NC-200 are suitable for either AC or battery operation, having both a built-in AC power supply and a special detachable cable and plug for battery connection. Removal of the speaker plug disconnects both plate and screen circuits of the audio power stage thus providing maximum battery economy. The B supply filter and the standby switch are wired to the battery terminals, so that the filter is available for vibrator or dynamotor B supplies.

The ten-inch speaker is housed in a separate cabinet specially designed to harmonize with the trim lines of the receiver. The undistorted output is 8 watts.

All features expected in a fine communication receiver are provided. These include CW oscillator, Signal Strength Meter, B-supply switch, etc. A phonograph input jack is provided.

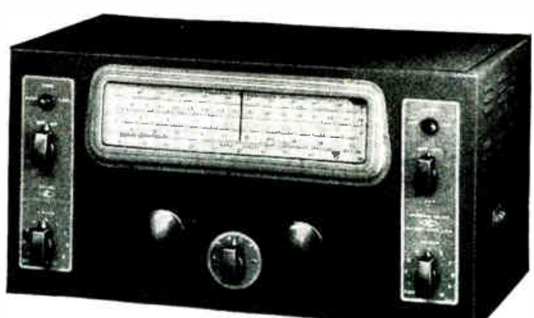
NC-200 TG, Table Model, two tone gray wrinkle receiver only. **List $265.83**

NC-2 TS, Table mounting 10" P.M. Loud Speaker in cabinet to match NC-200 TG above. **List $25.00**

NC-200 RG, Rack Model, gray wrinkle 3/16" aluminum panel receiver only. **List $289.33**

NC-2 RS, Rack Mounting 10" P.M. Loud Speaker on 10½" panel to match NC-200 RG above. **List $25.00**

NATIONAL NEW NC-45



The NC-45 receiver is an eight tube superheterodyne combining capable performance with low price. Features include a series valve noise limiter with automatic threshold control, tone control, CW oscillator, separate RF and AF gain controls, and AVC. Power supplies are self contained except for the battery model which must have an external source of heater and plate power, such as batteries or vibrapack.

A straight-line-frequency condenser is used in conjunction with a separate band spread condenser. This combination plus the full vision dial calibrated in frequency for each range covered and a separate linear scale for the band spread condenser, makes accurate tuning easy. Both condensers have inertia type drive. A coil switch with silver plated contacts selects the four ranges from 550 KC to 30 MC. Provision is made for either headphone or speaker.

Like all receivers which have no preselector stage, the NC-45 is not entirely free from images. However, where price is an important consideration, the NC-45 will be found a very satisfactory receiver.

NC-45 — Receiver only, complete with tubes, coils covering from 550 KC to 30 MC for 105–130 volts AC or DC operation — black finish. **List $84.17**
NC-45B — Receiver only, same as above but for battery operation, less batteries. **List $84.17**
NC-45A — Receiver only, same as above but for 105–130 volts AC only. **List $84.17**
NC-44TS — Loud Speaker in table mounting cabinet to match above receivers. **List $11.66**
RRA — Relay Rack Adapters designed for mounting these receivers in a standard relay rack. **List $2.75**
Shipping Weights: All models 45 pounds including speaker.

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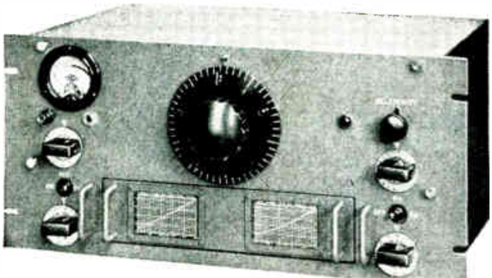


RACK MODEL

HRO table model, receiver only, complete with four sets of coils (1.7–4.0, 3.5–7.3, 7.0–14.4, 14.0–30.0 MCS). **List $329.50**

HRO Jr., table model, receiver only, with one set of 14 to 30 mc. coils. **List $198.00**

COILS

HRO Type	**E,** Range 900–2050 kc	**List $22.00**
HRO Type	**F,** Range 480–960 kc	**List $22.00**
HRO Type	**G,** Range 180–430 kc	**List $30.00**
HRO Type	**H,** Range 100–200 kc	**List $33.00**
HRO Type	**J,** Range 50–100 kc	**List $40.00**
HRO Jr. Type JA,	Range 14.0–30.0 mc	**List $18.25**
HRO Jr. Type JB,	Range 7.0–14.4 mc	**List $18.25**
HRO Jr. Type JC,	Range 3.5–7.3 mc	**List $18.25**
HRO Jr. Type JD,	Range 1.7–4.0 mc	**List $18.25**

MCS table model cabinet, 8" PM dynamic speaker and matching transformer **List $18.25**

697 Table power unit; 115 volt, 60 cycle input; 6.3 volt heater and 230 volt, 75 m.a. output, with tube **List $29.50**

See General Catalogue for relay rack mounting, coil containers and accessories

NATIONAL HRO

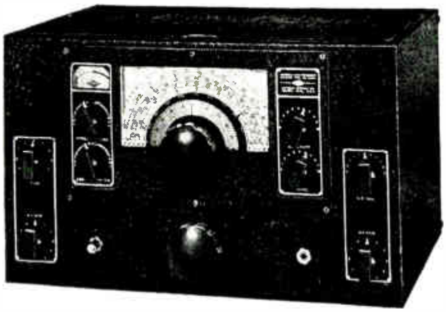
The HRO Receiver is a high-gain superheterodyne designed for communication service. Two preselector stages give remarkable image suppression, weak signal response and high signal-to-noise ratio. Air-dielectric tuning capacitors account, in part, for the high degree of operating stability. A crystal filter with both variable selectivity and phasing controls makes possible adjustment of selectivity over a wide range. Heterodynes and interfering c.w. signals may be "phased out" (attenuated) by correct setting of the phasing control. A signal strength meter, connected in a vacuum tube bridge circuit, is calibrated in S units from 1 to 9 and in db above S9 from 0 to 40. Also included are automatic and manual volume control features, a beat oscillator, a headphone jack and a B+ stand-by switch. Power supply is a separate unit. The standard model of HRO is supplied with four sets of coils covering the frequencies from 1.7 to 30 megacycles. Each coil set covers two amateur bands and the spectrum between. The higher frequency amateur band of each range, by a simple change-over operation, may be expanded to occupy 400 divisions of the 500 division PW instrument type dial.

For those who require the high performance of the HRO but do not need its extreme versatility, the HRO Jr. is offered. The fundamental circuit and mechanical details of both receivers are identical, but the HRO Jr. is simplified by omitting the crystal filter, signal strength meter and by supplying coils less the band-spread feature.

The frequency range of both the HRO and HRO Jr. may be extended to 50 kilocycles by using additional coil sets.

All models of the HRO are supplied with 6.3 volt heater type tubes. Table models and accessories are finished in black wrinkle enamel.

A technical bulletin covering completely all details will be supplied upon request.



NC-100A — complete with tubes. AC model — 10" speaker in cabinet. **List $220.00**

NC-100XA — complete with tubes and crystal filter. AC model — 10" speaker. **List $261.25**

NATIONAL NC-100A

These 11-tube superheterodyne receivers are self-contained (except for the speaker) in a table model cabinet that is readily adapted to relay rack mounting. One stage of R.F. and two stages of I.F. are used. Low loss insulation and high-Q coils give ample sensitivity and selectivity. Separate R.F. and Audio Gain Controls and a signal strength meter are mounted on the panel. Other controls are tone, CW Oscillator, AVC with amplified and delayed action, a B+ switch, and a phone jack. A self-contained power supply provides all necessary voltages including speaker field excitation. The range changing system is unique in that it combines the mechanical convenience of a coil switch with the electrical efficiency of plug-in coils.

All NC-100 series receivers are fitted with a noise limiter of truly remarkable effectiveness.

The NC-100A covers the range from 540KC to 30 MC. The large full vision dial is calibrated directly in megacycles and a separate high speed vernier scale provides high precision in logging.

NATIONAL POWER SUPPLIES

National Power Supplies are specially designed for high frequency receivers, and include efficient filters for RF disturbances as well as for hum frequencies. The various types for operation from an AC line are listed under the receivers with which they are used.

High voltage power supplies can be supplied for National Receivers for operation from batteries. These units are of the vibrator type.

686, Table model (165V., 50 MA.), for operation from 6.3 volts DC, with vibrator **List $49.50**

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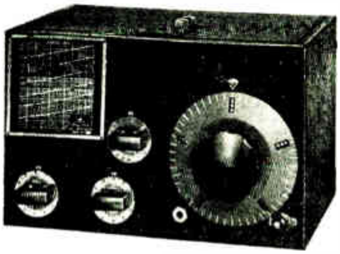

110 Receiver and 6 sets of coils, without tubes, speaker or power supply. List $93.50

5886 Power Supply for above receiver, with tube. List $32.50

NATIONAL ONE-TEN

The One-Ten Receiver fulfills the need for an adequate receiver to cover the field between one and ten meters.

A four-tube circuit is used, composed of one tuned R.F. stage, a self-quenching super-regenerative detector, transformer coupled to a first stage of audio which is resistance coupled to the power output stage. Tubes required: 954-R.F.; 955-Detector; 6C5-1st Audio, 6F6-2nd Audio.

NATIONAL SW-3

The SW-3U Receiver employs a circuit consisting of one R.F. stage transformer coupled to a regenerative detector and one stage of impedance coupled audio. This circuit provides maximum sensitivity and flexibility with the smallest number of tubes and the least auxiliary equipment. The single tuning dial operates a precisely adjusted two gang condenser; the regeneration control is smooth and noiseless, with no backlash or fringe howl; the volume control is calibrated from one to nine in steps corresponding to the R scale.

ONE UNIVERSAL MODEL — The circuit of the SW-3U is arranged for either battery or AC operation without coil substitution or circuit change. Battery operation utilizes two 1N5-G and one 1A5-G tubes. AC operation utilizes two 6J7-G and one 6C5-G tubes. Type 5886 AB power supply is recommended.

SW-3U, Universal model, without coils, phones, tubes or power supply. **List $38.50**

5886-AB, Power Supply, 113 V, 60 cycle, with 80 Rectifier. **List $32.50**

General Coverage Coils

Cat. No.	Range — Meters	List Per Pair
30	9 to 15	$3.85
31	13.5 to 25	3.85
32	23 to 41	3.85
33	40 to 70	3.85
34	65 to 115	3.85
35	115 to 200	3.85
36	200 to 360	4.40
37	350 to 550	4.40
38	500 to 850	5.50
39	850 to 1200	7.25
40	1200 to 1500	7.25
41	1500 to 2000	7.25
42	2000 to 3000	9.50

Band Spread Coils

30A	10 meter	$3.85
31A	20 meter	3.85
33A	40 meter	3.85
34A	80 meter	3.85
35A	160 meter	3.85

NATIONAL SCR-2

The SCR-2 is an extremely compact crystal controlled receiver for single channel reception mounted on a 3½" relay rack panel. It has two stages of tuned RF amplification, a dual purpose converter with crystal controlled oscillator, two stages of IF amplification, a detector and one audio stage. Auxiliary circuits are AVC, CW oscillator and noise limiter. Nine tubes are used, and the power supply is self-contained.

The SCR-2 is definitely a high performance receiver. Signal-to-noise ratio averages 10 db for an input of 2.5 microvolts. The AVC is flat within 4 db for inputs from 1 microvolt to well over 1 volt. Being crystal controlled, the frequency stability is excellent. The IF channel has a bandspread characteristic to allow for slight transmitter drift, etc.

As the SCR-2 receiver is intended for communication work, the audio channel has been deliberately made flat only from 100 to 1500 cycles, with increasing attenuation of higher frequencies, thus providing good intelligibility with maximum reduction of unwanted signals and noise.

SCR-2 receivers are available for use at fixed frequencies between 100 kcs and 18 mcs. A free booklet describing this receiver will be mailed on request.

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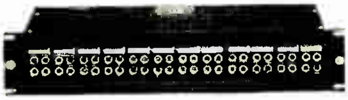
NATIONAL NRCM

MULTI-CHANNEL DIVERSITY RECEIVER

TYPE NRHA DUAL CHANNEL RECEIVER consists of two complete fixed frequency receivers, each entirely separate from the other except for a common power supply. The two channels of the NRHA are tuned to one transmitter to provide diversity reception. With four NRHA receivers, and four transmitters in simultaneous operation, eight separate channels with eight separate antennas are available for diversity reception of one signal.

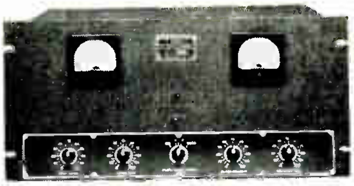
A separate monitor receiver is commonly used. This is often an HRO, but a special receiver such as the NRCL may be used



A patch cord panel is used for connection of the nine receiver channels to the following units. Jacks are provided for monitoring every channel.

A SIGNAL RECTIFIER UNIT receives the output from the nine channels and selects the output of whichever channel has the strongest signal. Since four frequencies are available, signal fading from such causes as changing skip-distance is largely eliminated, and the diversity action is very effective. As its name implies, the Signal Rectifier Unit also rectifies the signal, so that a CW signal is converted to a series of DC pulses.

A monitor panel provides a convenient means of control by the operator of levels into and out of the Signal Rectifier Unit. This panel is usually located at the operating position.



The **KEYER OSCILLATOR UNIT** has a self-contained audio oscillator which is keyed by the DC pulses from the signal rectifier. Low pass filters at the input of this unit remove beat notes, static and interference from the DC pulse, while high pass filters in the output remove any hum or other low frequency disturbance. Thus the final signal is a pure audio tone, free from interference and constant in pitch. It is remarkably easy to read.

The speaker panel is used to read the signal audibly. However, in a complete installation of this type, a tape recorder is commonly used to record the signal.

NEW EIMAC TETRODE

IS THE FORERUNNER OF MANY SENSATIONAL NEW EIMAC TUBES

Beginning now you can expect to hear a great deal about the New Eimac tubes which will soon be made available generally. The Eimac 4-250A Tetrode shown here is just the beginning.

Whether you are contemplating a new "ham rig" or highly specialized industrial electronic equipment—it is certain to be to your advantage to consult Eimac engineers early. While much technical information cannot be released publicly—considerable help can be given about your specific problem. Remember: Eimac is still first choice of leading electronic engineers throughout the world.

A pair of Eimac 4-250A Tetrodes, in the above photo, are operating at 100 MC with 4000 volts on the plates. Power input is 1.6 KW and power output 1.2 KW (75% efficiency). Plate current is 400 milliamperes, screen current is 20 milliamperes and grid current is 15 milliamperes. There are 300 volts on the screens and minus 250 volts on the grid.

Eimac Type 4-250A TETRODE

Filament	Thoriated Tungsten
Voltage	5.0 Volts
Current	10.5 Amperes
	Maximum Ratings
D. C. Plate Voltage	4000 Volts
D. C. Plate Current	350 Milliamperes
D. C. Screen Voltage	600 Volts
Plate Dissipation	250 Watts

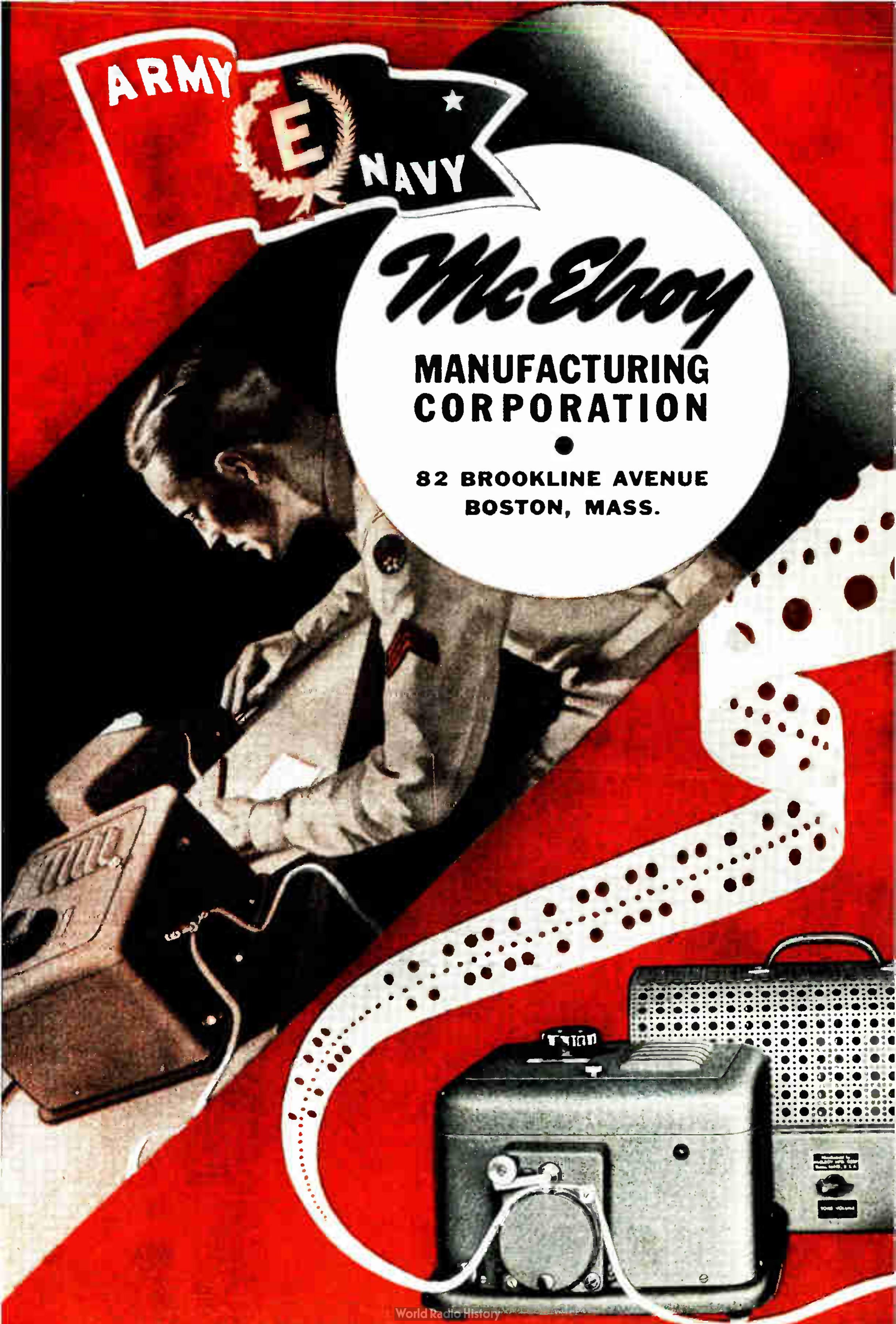
Follow the leaders to

Eimac TUBES

EITEL-McCULLOUGH, INC., 946 San Mateo Ave., SAN BRUNO, CALIF.

Plants located at: San Bruno, California and Salt Lake City, Utah

Export Agents: FRAZAR & HANSEN, 301 Clay Street, San Francisco 11, California, U. S. A.


ARMY
E
NAVY
McElroy
MANUFACTURING
CORPORATION
82 BROOKLINE AVENUE
BOSTON, MASS.

Within the past year, the men and women of McElroy have been honored with the award of both the Army-Navy "E" and the White Star. We worked hard for those citations . . . not for sake of the award . . . but because warfare for Victory needed the kind of equipment we produce to help form the vast network of radiotelegraph communications now in operation. After this war is over, we shall continue to progress. We shall continue to further the art of radiotelegraphy. We shall continue to give full support to the man who pounds the keys—ham or commercial. This is our pledge to the future.

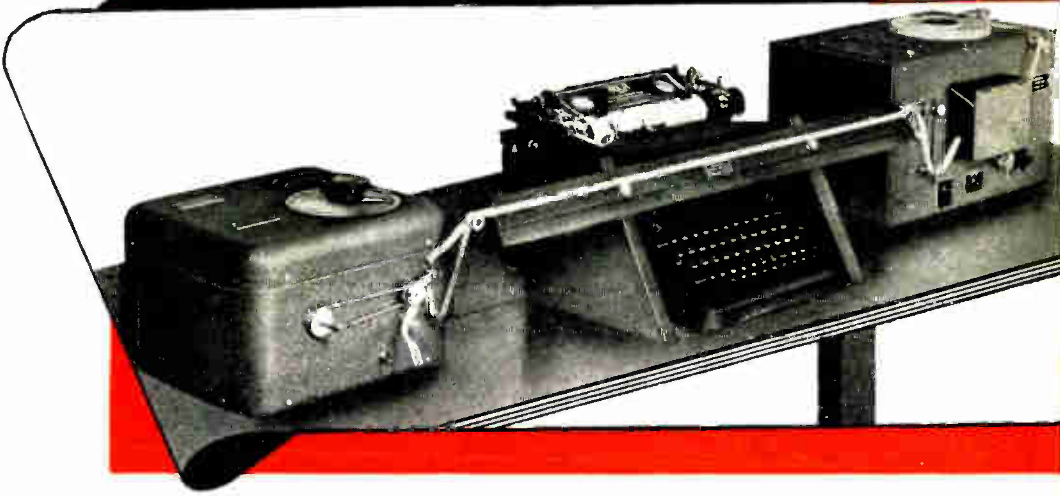
McElroy engineers never copy . . . never imitate. We create, design, build. We are never satisfied with mediocrity.

IMPORTANT NOTE: All McElroy Equipment is now being delivered promptly, for 110-120 volts, AC or DC, or any other voltages.

T. R. McElroy

President

McElroy MANUFACTURING CORP.
82 BROOKLINE AVE., BOSTON, MASS.



McELROY High-Speed Automatic Radiotelegraph Assemblies

The photograph in the top panel illustrates a complete McElroy automatic transmitting assembly . . . in the lower panel you see a McElroy automatic receiving assembly. These installations are typical of the high-speed radio telegraph equipment we supply to such international companies as R.C.A. Communications, Mackay Radio, Globe Wireless, Press Wireless . . . as well as to Army and Navy services everywhere.

On the following pages, each piece of equipment is individually illustrated and described. Technical manuals and operating instructions may be secured by writing direct to us.

ARMY E NAVY

McElroy MANUFACTURING CORP.
82 BROOKLINE AVE. BOSTON, MASS.

"NEVER BE SATISFIED WITH MEDIOCRITY"

From telegraph boy to head of the world's largest plant producing automatic radiotelegraph apparatus . . . is Ted McElroy's* own success saga. And the creed that drove him on—NEVER BE SATISFIED WITH MEDIOCRITY. This same spirit prevails throughout the McElroy organization where inquisitive engineers never copy and never imitate. They create, design, build . . .

Typical of the work they do is the new McELROY MODEL SR-900 SL-990 . . . a superior commercial recorder including an automatic noise limiter and signal leveller. Embodying new principles of design and operation, it will record clean, readable signals at speeds up to 350 words a minute . . . even under the most adverse conditions.

Your inquiries are invited. If a McElroy engineer can be of service to you, ask for one.

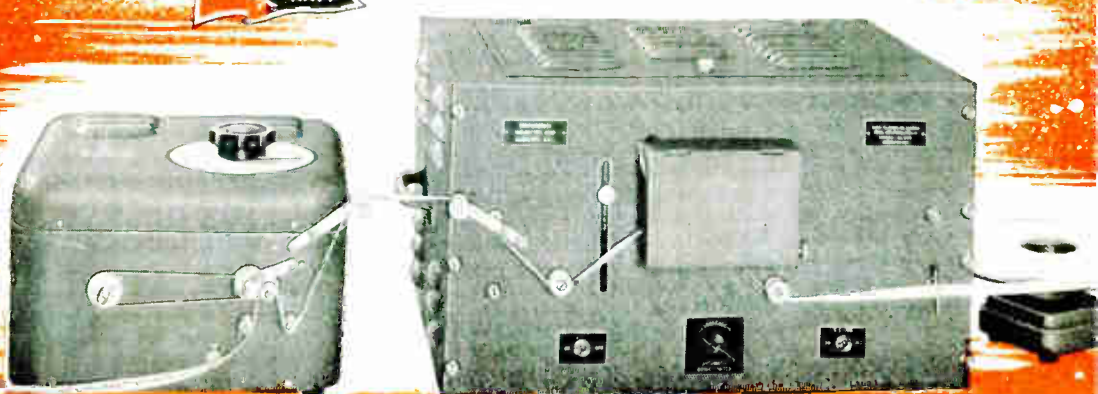
*WORLD CHAMPION RADIO TELEGRAPHER FOR MORE THAN 20 YEARS

McElroy MANUFACTURING CORP.
82 BROOKLINE AVE., BOSTON, MASS.

WORLD'S LARGEST MANUFACTURER OF AUTOMATIC RADIO TELEGRAPH APPARATUS

BUY MORE WAR BONDS

ARMY E NAVY



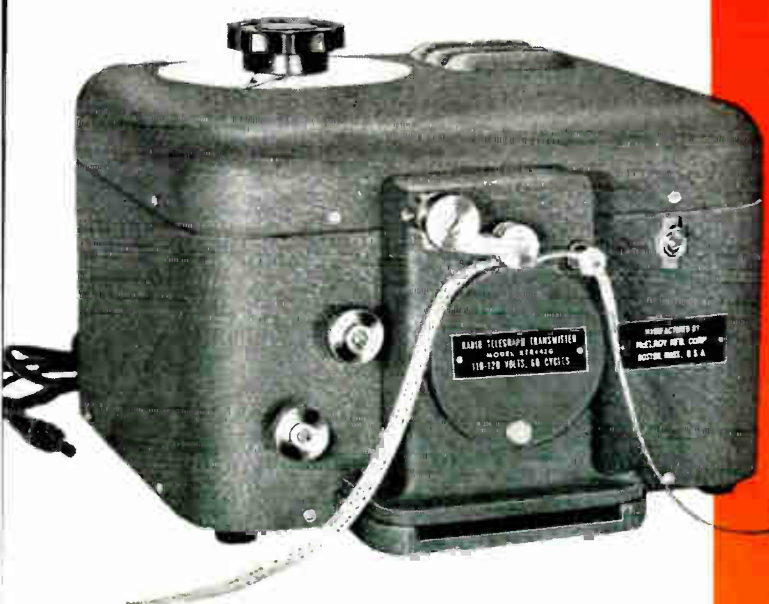
New!

McElroy Model XTR-442 G Automatic Transmitter

The motor drive of the XTR-442 G provides accurate calibration of words per minute which must be kept for indefinite periods at a fixed rate. It features, first, an entirely different type of head, with ball bearings and a newly developed plastic bearing and tool steel bell cranks. The speed range on the first set of gears is from 10 to 60 words a minute . . . an instantaneous gear shift provides speeds from 50 to 300 words per minute.

A second advantage of the motor is that the microswitch is operated by the lever extending throughout the front panel. This arrangement enables power to be removed from the motor while, simultaneously, the face of the drive pulley is withdrawn from the rim of the leather friction drive.

The XTR-442 G opens and closes any keying circuit to form mechanically precise signal elements, dots and dashes, in response to Wheatstone perforated tape. The unit will key either the intermediate relay of a radiotelegraph station or an audio oscillator for training radiotelegraph operators. The speed remains absolutely constant, unaffected by temperature rises or voltage fluctuations of the power. $490.00.



McElroy Model XTR-442 C High Speed Transmitter

Designed for speeds from 10 to 100 words per minute. This transmitter incorporates the newly developed McElroy microswitch that extends throughout the front panel, through the use of which power is removed from the motor and the face of the drive pulley is drawn away from the rim of the leather friction drive. A flat is thereby prevented from forming on the rim of the disc. The auto head is the same reliable McElroy model that has proved its worth in years of service . . . both in commercial and military applications. [illegible]

McElroy Model XTR-442 D

The same specifications as the XTR-442 C, excepting that this model is available for both AC and DC. [illegible]

McElroy MANUFACTURING CORP.
82 BROOKLINE AVE. BOSTON, MASS.

McELROY Medium Speed Recorder-Combination

Model SR-900 SL-990

Simply, yet ruggedly designed, the McElroy Model SR-900 SL-990 is an all-around Recorder suitable for speeds up to 300 words per minute. Except for the Tape Puller, this Recorder is completely self-contained. The circuit consists of an automatic noise limiter and signal leveller, terminating in a push-pull amplifier which drives a lightweight, low impedance moving coil through a heavy duty oxide rectifier.

Clean and readable signals may be recorded directly from the tone signal of any communications receiver having a 5000 or 5 to 50 ohm output. Because it is capable of rejecting static, background noise or weaker interfering signals, the Model SR-900 SL-990 operates efficiently under all conditions.

A unique method of controlling the flow of ink from the inkwell which is mounted in the back of the cabinet under the panel cover, gives this Recorder a distinct advantage over other models whose designs operate with the stylus approaching the tape from a horizontal position. The McElroy development has a control knob working through a slot on the front panel that raises and lowers the inkwell in a vertical position.

Built to provide enduring satisfaction, the Model SR-900 SL-990 is a requisite in any automatic receiving assembly . . . and for best results, we recommend that you use it in combination with the McElroy Tape Puller Model TP-890 C. Model SR-900 SL-990, $190.00.

McELROY New Tape Puller

Model TP-890 B

For use with high speed radiotelegraph recorders, school practice recorders and photo-tube keying units, the Model TP-890 C features the exclusive McElroy construction by which the power is removed from the motor while, at the same time, the face of the pulley is withdrawn from the rim of the friction drive, thus preventing a flat from developing on the rim of the disc.

The TP-890 C is used to give traction to standard ⅜-inch wide paper tape linked with a radiotelegraph line, and to wind tape on standard 16mm 400-ft. motion picture reels. The tape is drawn at an even speed per minute accurately controlled by a graduated dial. The knob, pointer and dial permit speeds to be varied from 0 to 100, but the rate is always the same when the pointer is returned to any given setting of the dial. Load differences, power line fluctuations or temperature variations have no effect on the efficiency of the instrument. It is also possible for the operator, without stopping the motor, to idle the tape puller so that no traction is applied to the tape and the reel does not revolve.

The TP-890 C (and the McElroy Tape Puller Model TP-890 D) are available in AC or DC, at $90.00. $10.00 additional for foot control on either AC or DC.

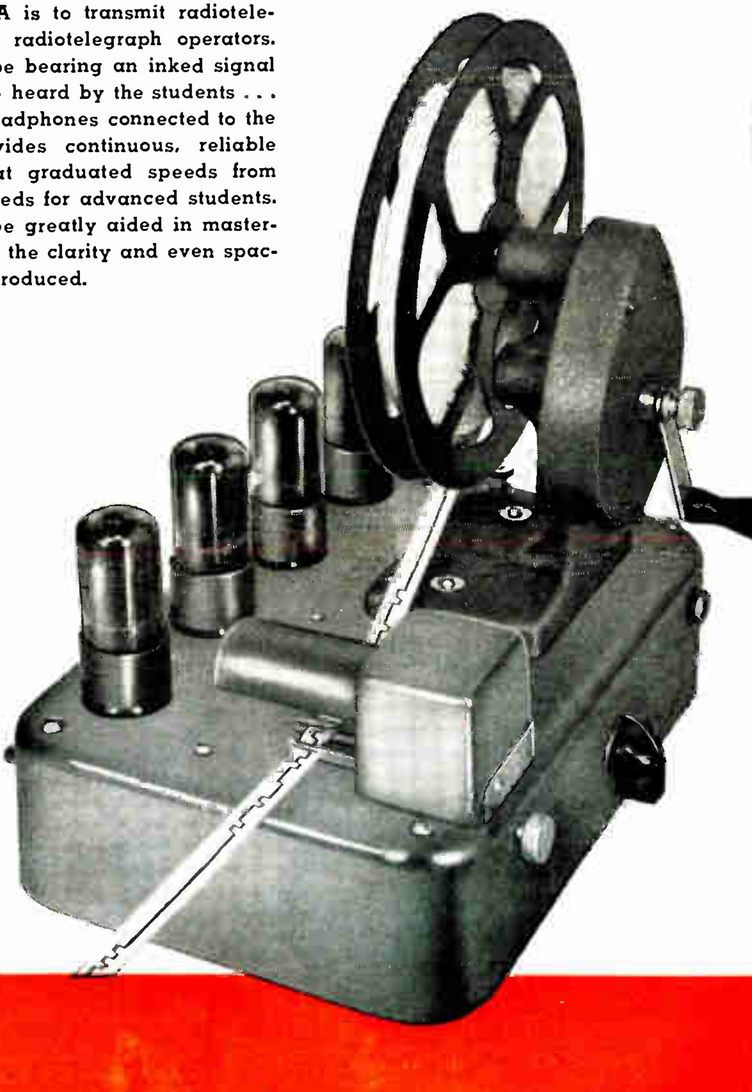
McElroy MANUFACTURING CORP.
[illegible] BROOKLINE AVE. BOSTON, MASS.

McELROY Model G-813-A Electronic Keyer

Regardless of the efforts of others to imitate the Model G-813 A Electronic Keyer, this McElroy development is still hailed as the outstanding job in the field. Used in conjunction with the McElroy Tape Puller Model TP-890 C, this instrument is the only one of its kind that has the practical advantage of keying only the signal line of the tape.

The purpose of the Model G-813 A is to transmit radiotelegraph signals for the training of radiotelegraph operators. In response to 3/8" wide paper tape bearing an inked signal line, it transmits signals which are heard by the students . . . as many as 50 at a time . . . in headphones connected to the keyer. The Model G-813 A provides continuous, reliable periods of signal transmission, at graduated speeds from slow for beginners to working speeds for advanced students. Students of radiotelegraphy will be greatly aided in mastering correctly sent code because of the clarity and even spacing with which the signals are reproduced.

ARMY E NAVY



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[illegible] BROOKLINE AVE., BOSTON, MASS.

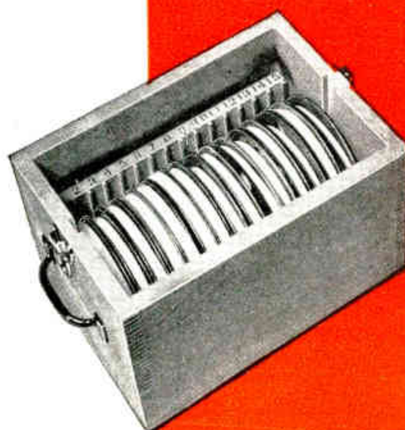
McELROY ACCESSORIES

McElroy Re-winds

Husky gears . . . both of them on shafts running through extra size oilite bearings...... **$20.00**

Practice Tapes

15-roll sets. G15L for sight reading, **$30.00** set. G15AA, U. S. Army tapes for photo-tube keying units, **$30.00** set. G15AM American Morse tapes for telegraph sounder practice, prices on application.



Tape Bridge

Provides a convenient channel across which standard 5/16" or 3/8" white paper tape bearing an inked radiotelegraph signal line is drawn for sight reading and typewriter transcriptions. **$20.00**

Operators Posture Chair

Designed for use with high speed automatic radiotelegraph apparatus **$20.00**

Operators Position Table

With liquid and cigarette-proof top...... **$120.00**

Ink Capsules

Each for one quart of ink...... **$1.00**

Quick Drying Fluid Prices on Application

Ink Recording Tape

20 roll, Prices on Large Quantities, Such as Tons, on Application.

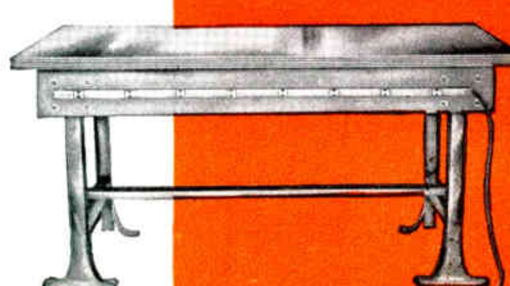
IMMEDIATE DELIVERY UP TO 10 TONS

Perforator Tape

Prices on Application.

IMMEDIATE DELIVERY UP TO 10 TONS

Tapes for All Kinds of Telegraph and Radio Work Can Be Supplied Promptly. Prices Upon Request.



Tape Stands

With warning signal so that the operator knows when roll of tape approaches the end. Designed for double rolls of tape...... **$45.00**

McElroy **MANUFACTURING CORP.**
82 BROOKLINE AVE. BOSTON, MASS.

ARMY E NAVY

McELROY CHART OF CODES AND SIGNALS

Respectfully Dedicated
to the
Radiomen of Our Armed Forces and Maritime Service

INTERNATIONAL FLAG SIGNALS

INTERNATIONAL MORSE
(CONTINENTAL CODE)

COMMERCIAL "Z" SIGNALS

AERONAUTICAL "Q" SIGNALS

AMERICAN MORSE

SEMAPHORE CODE

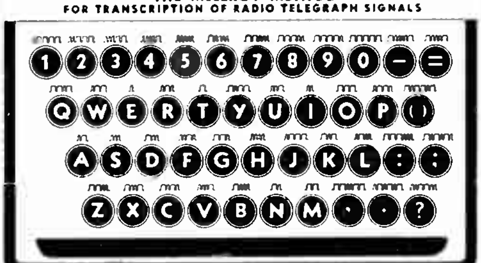
GREEK

A	Alpha
B	Beta
Γ	Gamma
Δ	Delta
E	Epsilon
Z	Zeta
H	Eta
Θ	Theta
I	Iota
K	Kappa
Λ	Lambda
M	Mu
N	Nu
Ξ	Xi
O	Omicron
Π	Pi
P	Rho
Σ	Sigma
T	Tau
Y	Ypsilon
Φ	Phi
X	Chi
Ψ	Psi
Ω	Omega
HY	Eta Ypsilon
YI	Ypsilon Iota
OY	Omicron Ypsilon
AI	Alpha Iota
AY	Alpha Ypsilon
EY	Epsilon Ypsilon
OI	Omicron Iota

RUSSIAN

А	A
Б	B
В	V
Г	G
Д	D
Е Э	E
Ж	J
З	Z
И	I
Й	Y
К	K
Л	L
М	M
Н	N
О	O
П	P
Р	R
С	S
Т	T
У	U
Ф	F
Х	KH
Ц	TS
Ч	CH
Ш	SH
Щ	SHCH
Ъ Ь	Mute
Ы	I
Ю	YU
Я	YA

THE McELROY METHOD
FOR TRANSCRIPTION OF RADIO TELEGRAPH SIGNALS



Typists can learn to transcribe inked slip at high speed in 4 to 6 weeks. Use GISPL set of practice tapes and associated equipment with this chart.

ARABIC

TURKISH

POSTAL TELEGRAPH — WESTERN UNION PHONETIC ALPHABET

WE'RE 100% IN THE VICTORY EFFORT AT PRESENT. PEACETIME WILL SEE AN EXPANDED VARIETY OF McELROY PRODUCTS TO MAKE EVEN MORE EFFICIENT THE RECEPTION AND TRANSMISSION OF DOTS AND DASHES.

U. S. ARMY AND U. S. NAVY PHONETIC ALPHABET

JAPANESE MORSE

A	KA	SA	TA	NA	HA	MA	YA	RA	WA
ア	カ	サ	タ	ナ	ハ	マ	ヤ	ラ	ワ
I	KI	SI	TI	NI	HI	MI	I	RI	(WI)
イ	キ	シ	チ	ニ	ヒ	ミ	イ	リ	ヰ
U	KU	SU	TU	NU	HU	MU	YU	RU	U
ウ	ク	ス	ツ	ヌ	フ	ム	ユ	ル	ウ
E	KE	SE	TE	NE	HE	ME	E	RE	(WE)
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GOVERNMENT PRESCRIBED PRINTING

A	J	S	1
B	K	T	2
C	L	U	3
D	M	V	4
E	N	W	5
F	O	X	6
G	P	Y	7
H	Q	Z	8
I	R		9
			Ø

"Q" SIGNALS

McELROY MANUFACTURING CORPORATION
82 BROOKLINE AVENUE, BOSTON, MASSACHUSETTS

In keeping with its policy of providing all services within its power, The American Radio Relay League makes available to amateurs and would-be amateurs literature properly prepared to present in the best form all available information pertaining to amateur radio. The fact that its offices are the national and international headquarters of radio amateurs, makes League publications authoritative, complete, up-to-the-minute, written from a thoroughly practical amateur's point-of-view. These publications are frequently revised and augmented to keep abreast of the fast-changing field. At this time they will be found particularly adaptable for radio training purposes. All are printed in the familiar *QST* format which permits thorough but economical presentation of the information. Most of the publications and supplies described in the following pages are handled by your dealer for your convenience.

•

THE AMERICAN RADIO RELAY LEAGUE, INC.

WEST HARTFORD 7
CONNECTICUT

THE
OFFICIAL MAGAZINE
OF THE
AMERICAN RADIO RELAY LEAGUE

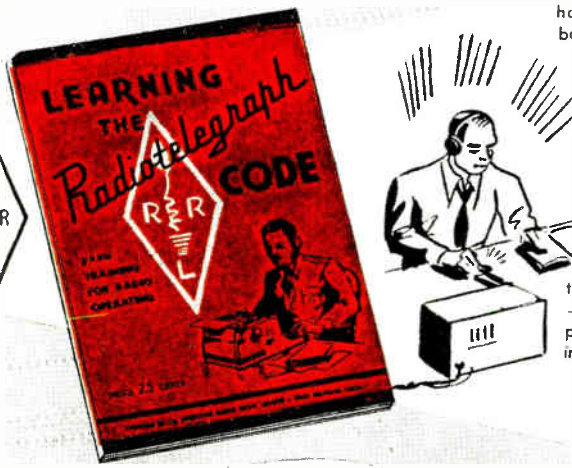
QST faithfully and adequately reports each month the rapid development which makes Amateur Radio so intriguing. Edited in the sole interests of the members of The American Radio Relay League, who are its owners, **QST** treats of equipment and practices and construction and design, and the romance which is part of Amateur Radio, in a direct and analytical style which has made **QST** famous all over the world. It is essential to the well-being of any radio amateur. **QST** goes to every member of The American Radio Relay League and membership costs $2.50 per year in the United States and Possessions. All other countries $3.00 per year. Elsewhere in this book will be found an application blank for A.R.R.L. membership.

For thirty years (and thereby the oldest American radio magazine) **QST** has been the "bible" of Amateur Radio.

The standard elementary guide for the prospective amateur. Features equipment which, although simple in construction, conforms in every detail to present practices. The apparatus is of a thoroughly practical type capable of giving long and satisfactory service—while at the same time it can be built at a minimum of expense. The design is such that a high degree of flexibility is secured, making the various units fit into the more elaborate station layouts which inevitably result as the amateur progresses. Complete operating instructions and references to sources of detailed information on licensing procedure are given.

25c

Postpaid Anywhere



Designed to train students to handle code skillfully and with precision, both in sending and in receiving, this booklet takes first rank among the League's publications which meet today's special training needs. Employing a novel system of code-learning based on the accepted method of sound conception, it is particularly excellent for the student who does not have the continuous help of an experienced operator or access to a code machine. It is similarly helpful home-study material for members of code classes. Adequate practice material is included for classwork as well as for home-study. There are also helpful data on high-speed operation, typewriter copy, general operating information — and an entire chapter on tone sources for code practice, including the description of a complete code instruction table with practice oscillator.

25c

Postpaid Anywhere

To obtain an amateur operator's license you must pass a government examination. The License Manual tells how to do that—tells what you must do and how to do it. It makes a simple and comparatively easy task of what otherwise might seem difficult. In addition to a large amount of general information, it contains questions and answers such as are asked in the government examinations. If you know the answers to the questions in this book, you can pass the examination without trouble.

25c

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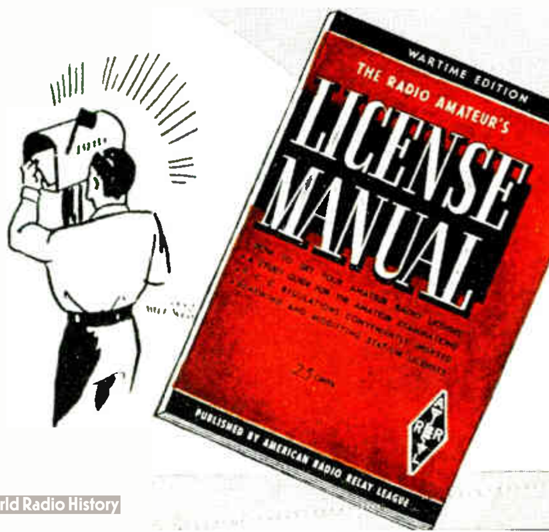
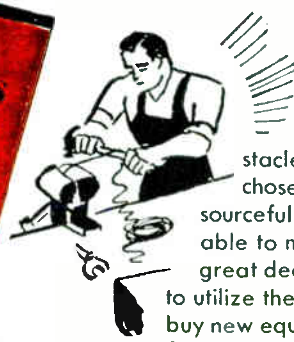


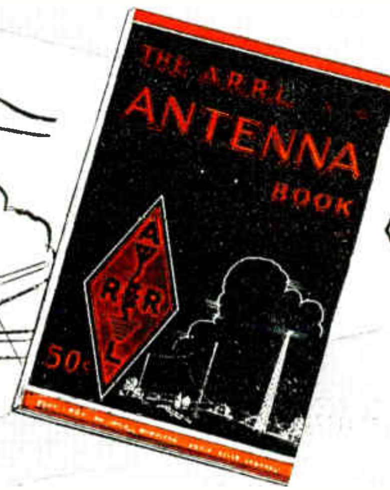


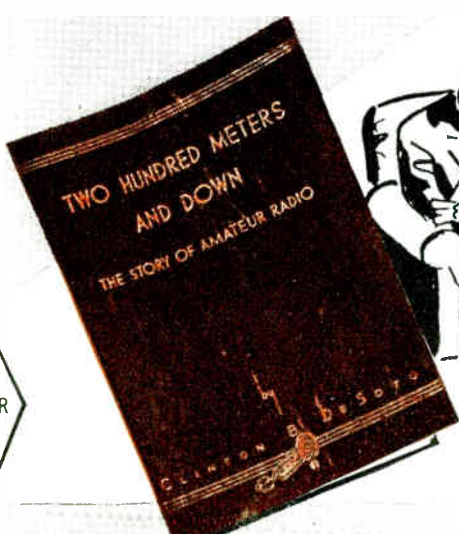
Amateurs are noted for their ingenuity in overcoming by clever means the minor and major obstacles they meet in their pursuit of their chosen hobby. An amateur must be resourceful and a good tinkerer. He must be able to make a small amount of money do a great deal for him. He must frequently be able to utilize the contents of the junk box rather than buy new equipment. Hints and Kinks is a compilation of hundreds of good ideas which amateurs have found helpful. It will return its cost many times in money savings—and it will save hours of time.

50c
Postpaid

A comprehensive manual of antenna design and construction, by the headquarters staff of the American Radio Relay League. Eighteen chapters, profusely illustrated. Both the theory and the practice of all types of antennas used by the amateur, from simple doublets to multi-element rotaries, including long wires, rhomboids, vees, phased systems, v.h.f. systems, etc. Feed systems and their adjustment. Construction of masts, lines and rotating mechanisms. The most comprehensive and reliable information ever published on the subject.

50c
Postpaid





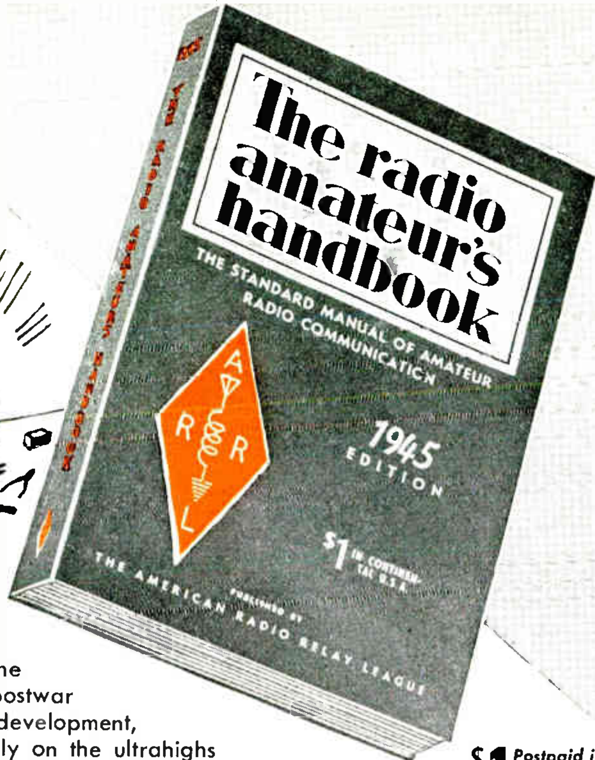
The Story of Amateur Radio, by Clinton B. DeSoto—a detailed, accurate presentation in full book length of all the elements that have served to develop the most unique institution of its kind in the history of the world. A book of history but not a history-book, TWO HUNDRED METERS AND DOWN: The Story of Amateur Radio tells in spirited, dramatic fashion the entire chain of significant events in the development of the art.

Approximately 200 pages, 90,000 words, with durable imitation leather red paper cover

$1.00 *Postpaid*

Deluxe edition bound in blue cloth

$2.00 *Postpaid*



Paving the way for postwar amateur development, particularly on the ultrahighs (microwaves) and other prospective new techniques, the 1945 Edition of the HANDBOOK includes diversified material new to its scope, while still retaining its time-proved treatment of the orthodox theory and practice of amateur radio—refined, modernized, reorganized for maximum convenience whether used as text, reference or constructional manual. . . . This Edition of the HANDBOOK contains more pages and more information per page than any HANDBOOK yet published. . . . Every subject encountered in practical radio communication is covered, arranged for maximum convenience to the reader, sectionalized by topics with abundant cross-referencing and fully indexed. . . . More than ever the ideal reference work, the 1945 edition also contains practical constructional information on tested and proved gear—always the outstanding feature of the HANDBOOK.

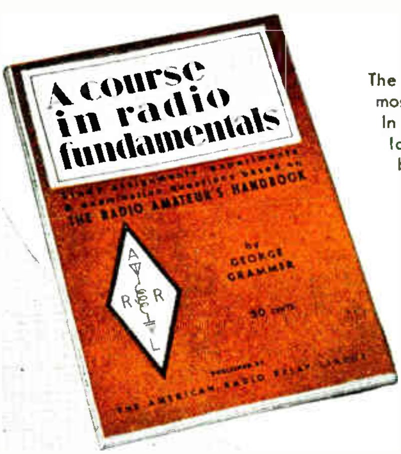
$1 *Postpaid in Continental U. S. A.*

$1.50 Postpaid Elsewhere

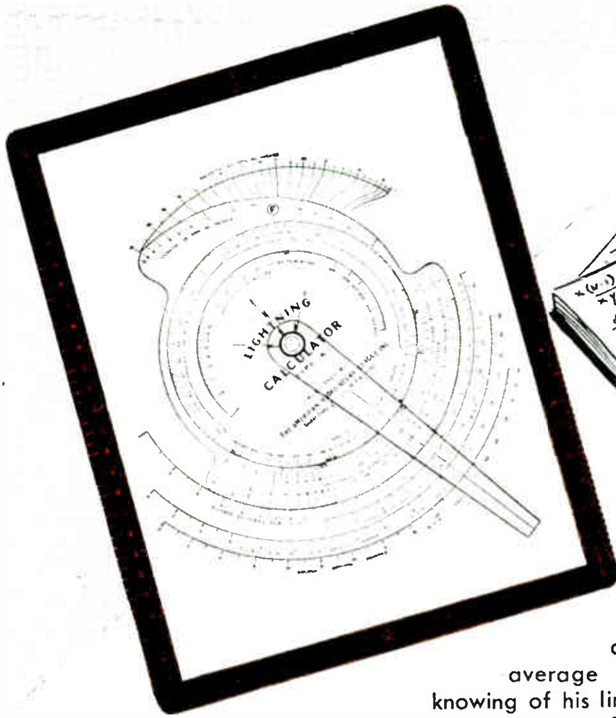
Buckram Bound $2.00

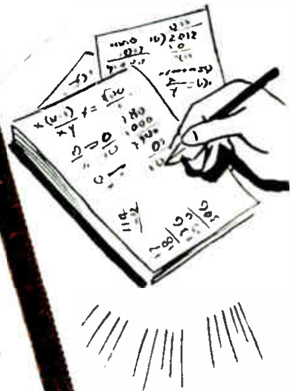
50c

Postpaid Anywhere



The objective in preparing this course was to accent those principles most frequently applied in actual radio communication. "A Course In Radio Fundamentals" is a study guide, examination book and laboratory manual. Its text is based on the "Radio Amateur's Handbook." Either the special edition for war training purposes or the Standard Edition may be used. References contained in the "Course" are identical in both editions. As a text, this book greatly smooths the way for the student of the technicalities of radio. It contains interesting study assignments, experiments and examination questions for either class or individual instruction. It describes in detail 40 experiments with simple apparatus giving a complete practical knowledge of radio theory and design.





Aware of the practical bent of the average amateur and knowing of his limited time, the League, under license of the designer, W. P. Koechel, has made available these calculators to obviate the tedious and sometimes difficult mathematical work involved in the design and construction of radio equipment. The lightning calculators are ingenious devices for rapid, certain and simple solution of the various mathematical problems which arise in radio and allied work. They make it possible to read direct answers without struggling with formulas and computations. They are tremendous time-savers for amateurs, engineers, servicemen and experimenters. Their accuracy is more than adequate for the solution of practical problems, and is well within the limits of measurement by ordinary means. Each calculator has on its reverse side detailed instructions for its use; the greatest mathematical ability required is that of dividing or multiplying simple numbers. They are printed in several colors. You will find lightning calculators the most useful gadgets you ever owned.

RADIO CALCULATOR

TYPE A

This calculator is useful for the problems that confront the amateur every time he builds a new rig or rebuilds an old one or winds a coil or designs a circuit. It has two scales for physical dimensions of coils from one-half inch to five and one-half inches in diameter and from one-quarter to ten inches in length, a frequency scale from 400 kilocycles through 150 megacycles; a wavelength scale from two to 600 meters; a capacity scale from 3 to 1,000 micro-microfarads; two inductance scales with a range of from one microhenry through 1,500; a turns-per-inch scale to cover enameled or single silk covered wire from 12 to 35 gauge, double silk or cotton covered from 0 to 36 and double cotton covered from 2 to 36. Using these scales in the simple manner outlined in the instructions on the back of the calculator, it is possible to solve problems involving frequency in kilocycles, wavelength in meters, inductance in microhenrys and capacity in microfarads, for practically all problems that the amateur will have in designing—from high-powered transmitters down to simple receivers. Gives the direct reading answers for these problems with accuracy well within the tolerances of practical construction.

$1.00

Postpaid

OHM'S LAW CALCULATOR

TYPE B

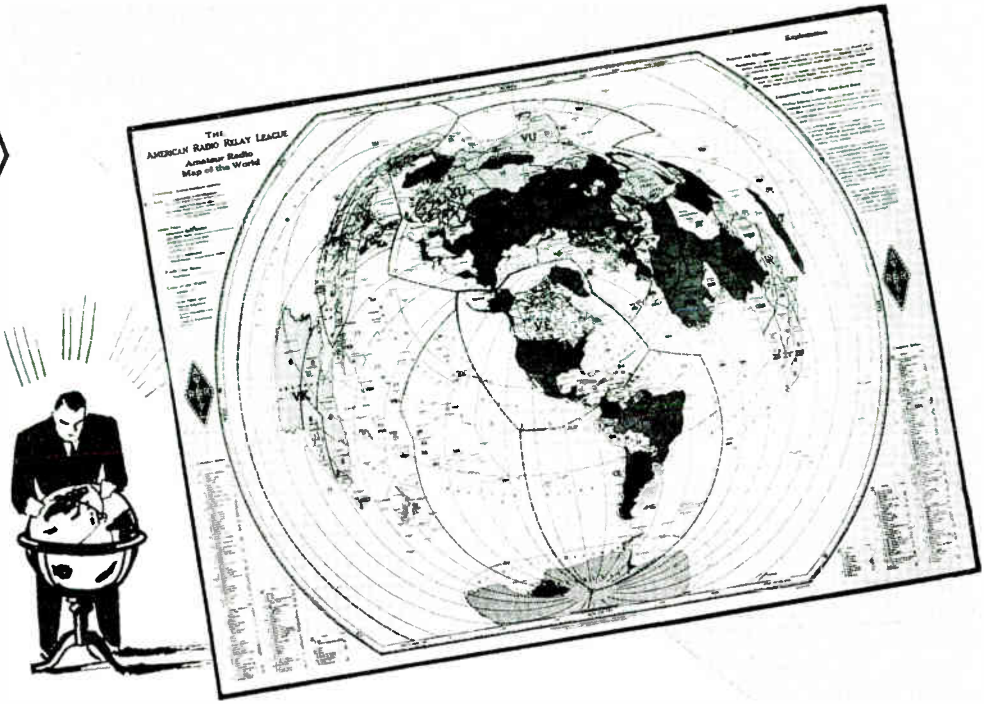
This calculator has four scales:

- A power scale from 10 microwatts through 10 kilowatts.
- A resistance scale from .01 ohms through 100 megohms.
- A current scale from 1 microampere through 100 amperes.
- A voltage scale from 10 microvolts through 10 kilovolts.

With this concentrated collection of scale, calculations may be made involving voltage, current, and resistance, and can be made with a single setting of a dial. The power or voltage or current or resistance in any circuit can be found easily if any two are known. This is a newly-designed Type B Calculator which is more accurate and simpler to use than the justly-famous original model. It will be found useful for many calculations which must be made frequently but which are often confusing if done by ordinary methods. All answers will be accurate within the tolerances of commercial equipment.

$1.00

Postpaid



The A.R.R.L. Amateur Radio **MAP** of the World is a special type of projection made by Rand, McNally to A.R.R.L. specifications. It gives great circle distance measurements in miles or kilometers within an accuracy of 2%. Shows all principal cities of the world; local time zones and Greenwich; WAC divisions; 230 countries, indexed; 180 prefixes, districts and subdivisions, where used; and U. S. examining points. Large enough to be usable, printed in six colors on heavy map paper, 30 x 40 inches.

$1.25
Postpaid Anywhere

MEMBERS STATIONERY

Members' stationery is standard 8½ x 11 bond paper which every member should be proud to use for his radio correspondence. Lithographed on 8½ x 11 heavy bond paper.

100 Sheets,
50c
250 Sheets,
$1.00
500 Sheets,
$1.75
Postpaid Anywhere

THE LEAGUE EMBLEM, with both gold border and lettering, and with black enamel background, is available in either pin (with safety clasp) or screw-back button type.

THE EMBLEM CUT: A mounted printing electrotype, ⅝″ high, for use by members on amateur printed matter, letterheads, cards, etc.

50c
Postpaid Anywhere

New G-E Lighthouse Tube...

FOR OPERATION IN THE ULTRA-HIGH-FREQUENCY SPECTRUM

THE REVOLUTIONARY G-E LIGHTHOUSE TUBE

Developed by G-E electronic engineers, this new disk-seal, parallel-plane, tube design extends the top frequency limits into the ultra-high-frequency spectrum. The generic name for tubes of this design is *megatron*.

The new Generol Electric lighthouse tube is one of the most important developments in electronic-tube history — paralleling many other G-E tube "firsts." The advent of this new tube makes possible improved electronic devices and dozens of new applications in radio and television. This is of special significance to the thousands of amateurs now in the armed forces. In the coming era of peace, they will be key figures in the expanded fields of electronics—FM and AM broadcasting, television, emergency and two-way communications, domestic and international commercial radio services, the fields of marine and aviation, commercial and industrial applications.

For over a quarter of a century, General Electric electronic research has produced many important developments. The new G-E lighthouse tube emphasizes again the fact that G.E. has contributed more basic electronic-tube developments to the communications and electronic industries than any other manufacturer.

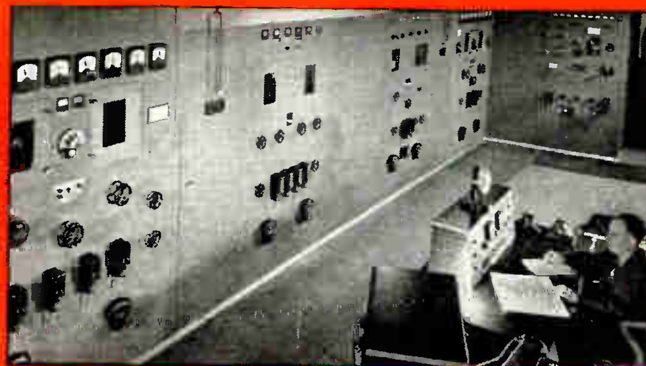
For information on receiving tubes, transmitting tubes, or industrial electronic tubes, write to Electronics Department, General Electric, Schenectady, N. Y.

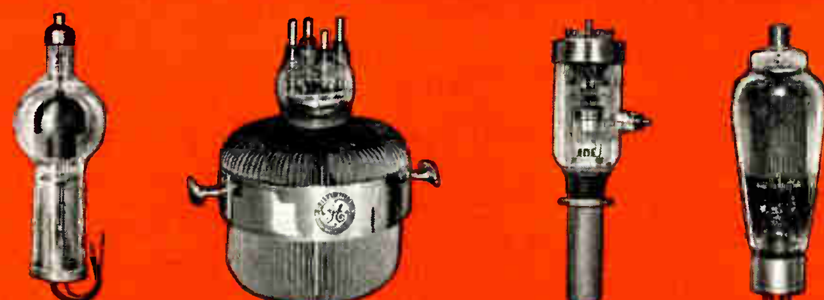
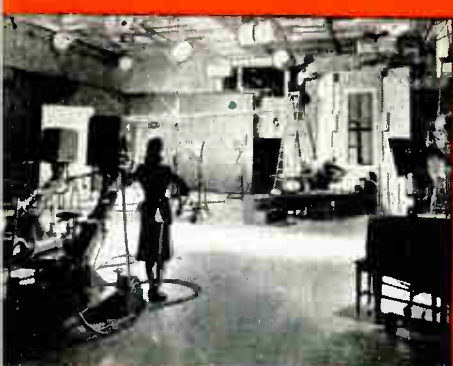
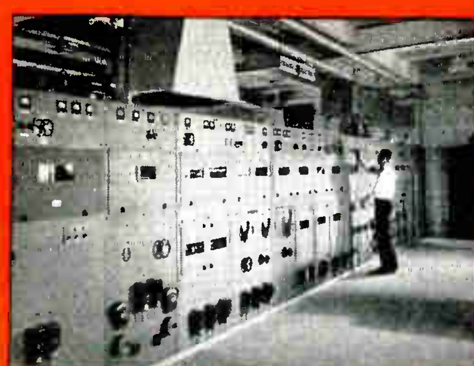
GENERAL ELECTRIC

SOME ELECTRONIC-TUBE "FIRSTS" CONTRIBUTED BY G. E.

- Thoriated tungsten filaments
- High-powered water-cooled transmitter tubes
- Hot-cathode, mercury-vapor rectifiers
- The screen-grid tube
- The steel-clad ignitron
- The thyratron
- The metal receiving tube
- Photoelectric tubes for commercial talking moving pictures.



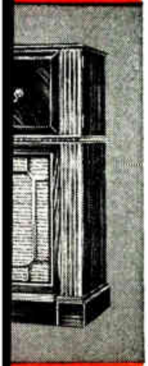




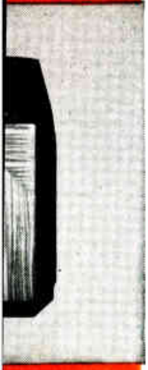
IMPORTANT GENERAL ELECTRIC ELECTRONIC PRODUCTS

INDUSTRIAL ELECTRONIC TUBES. General Electric is equipped to manufacture all types of electronic tubes — from tubes used in heating and welding, to tubes that control motors and machinery; to tubes that transmit, receive and amplify sound and signals; to tubes that measure light, sort, count, "see" through solids. Long experience, modern manufacturing equipment, rigid control and inspection contribute to the exceptionally high quality and dependability of G-E tubes. Shown (left to right) are the ignitron, thyratron, and kenotron.



FM AND AM TRANSMITTERS AND RECEIVERS. General Electric's unequaled experience in short-wave broadcasting is well known . . . all of America's 100 kw international broadcast transmitters have been built by G.E. G.E. has equipped more than a third of existing FM broadcast stations, and supplied a large portion of the 600,000 FM receivers now in use. G.E., in fact, is the only manufacturer with experience in building the complete FM system — including transmitter, antenna, and home receiver. Shown (left to right) are G-E FM and AM transmitters and G-E radio-phonograph combination incorporating AM and FM.

TRANSMITTING TUBES. General Electric has probably made more important contributions to the development of transmitting tubes than any other manufacturer. For example: G.E. developed tubes and circuits that produced the high-frequency oscillations that make broadcasting possible. G.E. developed the first water-cooled transmitting tube which made high-power broadcasting possible. G.E. developed the hot-cathode mercury-vapor tubes which cut broadcasting power losses tremendously. Shown are four typical G-E transmitting tubes.

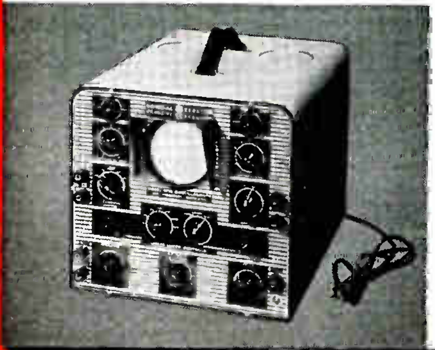


TELEVISION TRANSMITTER AND RECEIVERS. Evidence of G.E.'s leadership in studio planning and station equipping is Television Station WRGB, in Schenectady, New York. This televison "workshop" is one of the finest and most complete studios of its kind in the world. From WRGB will come much of programming knowledge and technical development which will bring the post-war expansion of television. Shown (left to right) are WRGB's studio, transmitter, and G-E AM and television receiver with FM for television sound.

EMERGENCY COMMUNICATION. It's coming! — two-way FM radio in every municipal police car. The G-E FM system for cities, towns, and public utilities provides amazing freedom from static and extremely low noise levels. General Electric AM police radio will be used in the wider areas covered by state and county public safety departments. Here, again, G.E.'s broad experience will provide unusually dependable emergency radio equipment.

CAPACITORS. General Electric has pioneered and developed a new, unique line of vacuum capacitors, rated from 7500 to 16,000 volts peak and from 25 to 100 mmfd. These circuits are common to military, aircraft, and amateur radio equipment.

The small size of the G-E vacuum capacitor is of especial importance in the design of high-frequency circuits. Only a tenth the size of similarly rated air capacitors, these capacitors also provide an internal voltage breakdown characteristic which is unaffected by altitude.



TESTING INSTRUMENTS. The new General Electric line of laboratory and testing equipment provides an extensive choice of portable, compact apparatus for accurate, rapid maintenance and testing of radio electronic circuits and parts. It includes G-E unimeters, tube checkers, bridges, signal generators, oscilloscopes, and other instruments — all planned for easy, error-free reading and long, dependable service. Shown at left is G-E oscillograph and frequency modulator for AM, FM, and television-receiver and transmitter trouble-shooting.

THE ELECTRONIC "BOOKS OF THE YEAR"

"How Electronic Tubes Work." Here's an electronics "first reader"—a simple explanation of the basic principles of electronic tubes . . . describes briefly important uses of tubes in industrial electronic equipment. *"How Electronic Tubes Work"* is *FREE*. Filing size—24 graphically illustrated pages.

"The ABC's of Radio." Here's a simple, easy-to-understand "Primer" on radio and its basic circuits . . . how they are designed and how they perform . . . what the fundamentals are . . . the various principles and theories of radio receivers and their service . . . 68 pages, clearly illustrated. *"The ABC's of Radio"* is offered at twenty-five cents in stamps or coin.

For either of these informative brochures . . . address Dept. 6-S, Electronics Department, General Electric, Schenectady, N. Y.

LEADER IN RADIO, TELEVISION, AND ELECTRONIC RESEARCH

GENERAL ELECTRIC

"ORANGE LEADER CALLING . . ." "***ROGER!***"

"Getting the message through" is the business of thousands of our ham friends these days.

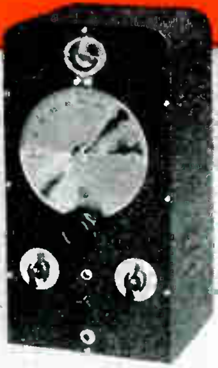
Some of them are running into their old friend, the Browning Frequency Meter (shown below), which is helping keep certain war rigs accurate.

Some of them will be interested to know that Browning has perfected and proved-in-the-field a balanced-capacitance signal system which has helped relieve the manpower situation in many plants by reducing the need for armed guard patrols. (Descriptive literature on request.)

All of them can look forward to returning to their own shacks in the sure knowledge that when peace comes Browning will be adding to ham operating pleasure with new, even better gear.

BROWNING

LABORATORIES, INCORPORATED
WINCHESTER, MASSACHUSETTS



HAMMARLUND

PRECISION PRODUCTS FOR BETTER RADIO

"HQ-120-X" AMATEUR RECEIVER

THE HAMMARLUND "HQ-120-X" meets the most critical demands of amateur and professional operators. Hammarlund engineers have gone beyond ordinary practice in designing this new and outstanding receiver. This ultra-modern 12-tube superheterodyne covers a continuous range of from 31 to .54 mc. (9.7 to 555 meters) in six bands, taking in all important amateur, communication, and broadcast channels. The "HQ-120-X" is not to be confused with modified broadcast sets. Two years were required to develop it. This is a special receiver with special parts throughout. Every wave range is individual—that is, each range has its own individual coil and a tuning condenser of proper value for maximum efficiency; thus, including the broadcast band does not decrease efficiency at high frequencies. Besides having all the necessary features for perfect short wave reception, such as A.V.C., beat oscillator, send-receive switch, phone jack and relay terminals, the "HQ-120-X" also includes a new and outstanding crystal filter circuit which is variable in 6 steps from full bandwidth to razor edge selectivity. This permits the use of the crystal filter for the reception of both voice and music. It is no longer necessary to contend with serious heterodyne interference. These annoying disturbances can be phased out with the phasing control on the panel. Other features include drift compensation for improved stability; a new and accurate "S" meter circuit for measuring incoming signal strength; antenna compensator to compensate for various antennas, and 310 degrees band spread for each amateur band from 80 to 10 meters. The band spread dial is calibrated in megacycles for each of these amateur bands. The main tuning dial is calibrated in megacycles throughout the entire range of the receiver. Gray finish. Rack adapter $6.00 extra.

Prices include Speaker and Tubes

Code	Type	Tuning Range	Speaker	Net Price
HQ-120-X	Crystal	31—.54 mc.	10" P.M. Dyn.	$168.00
Speaker cabinet (metal) 12½" x 12½" x 7 inches				3.90

Special model finished in black .... $168.00 Net
Speaker Cabinet, black to match .... 3.90 Net

Send for Descriptive Booklet

Code	Type	Spkr.	Tuning Range	Net Price
SP-210-X	Crystal	10"	15—560 meters	$318.00
SP-210-SX	Crystal	10"	7½—240 meters	318.00
SP-220-X	Crystal	12"	15—560 meters	330.00
SP-220-SX	Crystal	12"	7½—240 meters	330.00
PSC	10" speaker cabinet to match receiver			5.10

Special Models Covering Other Wave Ranges Available On Order

THE "SUPER-PRO"

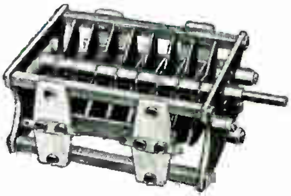
THIS 18-tube "SUPER-PRO" includes all the outstanding features which have made the "Super-Pro" famous, and in addition many recent developments have been added. The "Super-Pro" has a variable selectivity crystal filter. This crystal filter has five positions of selectivity—3 for phone and 2 for CW. The variable crystal filter, in addition to the variable band width I.F., provides a selectivity range of from less than 100 cycles to approximately 16 kc. The new "Super-Pro" also has an improved noise limiter designed to minimize interference caused by automobile ignition systems and disturbances of similar nature. Maximum image suppression is obtained with two stages of high selectivity tuned R.F. ahead of the first detector. Three stages of I.F. are employed and there are three stages of high fidelity audio amplification resulting in an output of approximately 14 watts. A new and improved "S" meter has been installed in the "Super-Pro" for accurately reporting relative signal strength. Other features include full band-spread on all bands; beat oscillator; send-receive switch; relay connections; phone connections; connections for phono-pickup; beautifully finished modernistic cabinet. The sensitivity of the "Super-Pro" is better than 1 microvolt. Available in rack mounting type at $10.50 extra.

Write for Circular

HAMMARLUND MANUFACTURING CO., INC. • 460 West 34th Street • New York City

HAMMARLUND

"TC" TRANSMITTING CONDENSER

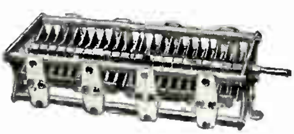


A moderately priced, heavy duty transmitting condenser, featuring aluminum end plates Isolantite insulation, non-inductive, self-cleaning silver plated beryllium contacts, full floating rotor bearing, non-magnetic rotor assembly, polished heavy aluminum plates accurately spaced. All, except type "L," have round edge plates of .040" thickness. Type "L" has .025" plates with plain edges. Type "F" has .230", 7500 V. air gap. Type "G," .200", 6750 V. Type "H," .171", 6000 V. Type "J," .100", 4250 V. Type "K," .084", 3750 V. Type "L," .070", 2000 V. air gap.

Available in a wide variety of capacities and working voltages, these condensers are ideal for modern up-to-date transmitters with power outputs ranging from 200 watts to 1 kw.

Type	Capacity	Overall Length	List
TC-220-L	220 mmf.	4 1/16	$ 6.30
TC-440-L	465 mmf.	5 7/8	9.10
TC-90-K	95 mmf.	2 15/16	5.70
TC-165-K	167 mmf.	4 1/16	6.50
TC-220-K	222 mmf.	4 5/8	8.00
TC-330-K	335 mmf.	6 1/2	10.00
TC-240-J	250 mmf.	6 1/2	10.20
TC-25-H	23.5 mmf.	2 15/16	5.10
TC-50-H	53 mmf.	4 1/16	6.00
TC-110-H	115 mmf.	6 1/2	9.00
TC-40-G	46 mmf.	4 1/16	7.00
TC-65-G	75 mmf.	5 7/8	8.80
TC-100-G	110 mmf.	7 1/2	11.20
TC-150-G	165 mmf.	10 5/8	14.80
TC-55-F	60 mmf.	5 7/8	8.00

"TCD" SPLIT STATOR TYPES



These split stator transmitting condensers are identical to the singles shown above, except that the stator sections are individual. Ideal for push-pull power amplifiers ranging in power up to 1 kw. They are of convenient size and lend themselves to construction of compact apparatus. Overall dimensions in back of panel are given in the accompanying table. The capacity values listed are for each section. The last letter in the code represents plate spacing and voltage rating. These are identical to those given above. Type "M"—plain plates, .030" air gap.

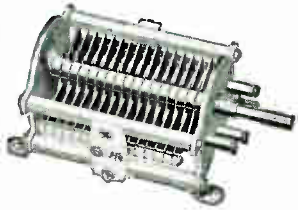
Type	Capacity	Overall Length	List
TCD-500-M	490 mmf.	4 1/16	$10.30
TCD-80-L	90 mmf.	4 1/16	8.30
TCD-210-L	215 mmf.	5 7/8	10.40
TCD-90-K	95 mmf.	4 5/8	9.40
TCD-165-K	167 mmf.	6 1/2	11.50
TCD-325-K	335 mmf.	11 1/16	20.50
TCD-240-J	250 mmf.	11 1/16	19.00
TCD-50-H	53 mmf.	6 1/2	9.80
TCD-110-H	115 mmf.	11 1/16	16.00
TCD-40-G	46 mmf.	7 1/2	10.50
TCD-75-G	85 mmf.	11 1/16	14.50
TCD-55-F	60 mmf.	11 1/16	13.70

"N" NEUTRALIZING CONDENSERS

Improved neutralizing condensers with heavy polished aluminum plates. Rounded edges. Isolantite. Fine adjusting screw. Positive lock. Horizontal adjustment. Type "N-10", 2 5/8" high x 1 1/16" deep. "N-15", 4 15/16" high x 3 1/2" deep. "N-20", 5 11/16" high x 4" deep.

Code	List
N-10—(2.1—10 mmf.)	$4.60
N-15—(3.2—14 mmf.)	8.70
N-20—(3.8—14 mmf.)	9.30

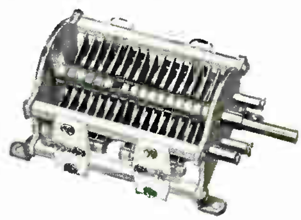
"MTC" TRANSMITTING CONDENSERS



Compact types, Isolantite insulation. Base or panel mounting. Polished aluminum plates. Stainless steel shaft. Size of 150 mmf. with .070" plates spacing only 4 5/8" behind panel. All type "B" condensers have round edge plates .025" in thickness. Type "C" has plain edge plates .025" thick. Self-cleaning wiping contact.

Code	Capacity	List
MTC-20-B	22 mmf.	$4.10
MTC-35-B	33 mmf.	4.30
MTC-50-B	50 mmf.	4.60
MTC-100-B	100 mmf.	5.30
MTC-150-B	150 mmf.	6.10
MTC-50-C	46 mmf.	4.10
MTC-100-C	105 mmf.	4.40
MTC-150-C	150 mmf.	4.80
MTC-250-C	255 mmf.	5.30
MTC-350-C	360 mmf.	5.80

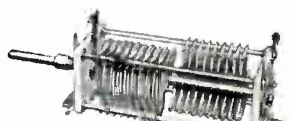
"MTCD" SPLIT-STATOR TYPES

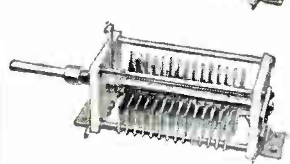


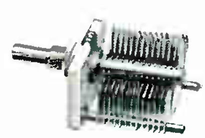
Same outstanding features as MTC singles except that stator sections are separate. Model 100-B with .070" plate spacing, only 5 3/4" behind panel. "B" models—rounded plates. "C" models—plain plate edges.

Code	Capacity	List
MTCD-20-B	22 mmf. per sect.	$5.60
MTCD-35-B	33 mmf. per sect.	6.00
MTCD-50-B	50 mmf. per sect.	6.50
MTCD-100-B	100 mmf. per sect.	8.75
MTCD-50-C	46 mmf. per sect.	5.50
MTCD-100-C	105 mmf. per sect.	6.00
MTCD-150-C	150 mmf. per sect.	6.50
MTCD-250-C	255 mmf. per sect.	7.50

A NEW LINE OF TRANSMITTING AND RECEIVING CONDENSERS







The new HFA and HFB receiving and transmitting condensers are the latest in condenser design. The HFB transmitting condenser, for example, has fully insulated rotor and control shaft permitting higher operating voltage for a given plate spacing. This new design results in more compact and efficient condenser construction and the insulated control shaft reduces the danger of electric shock to the operator. The HFB's are made in both dual and single stator types and in all important capacities.

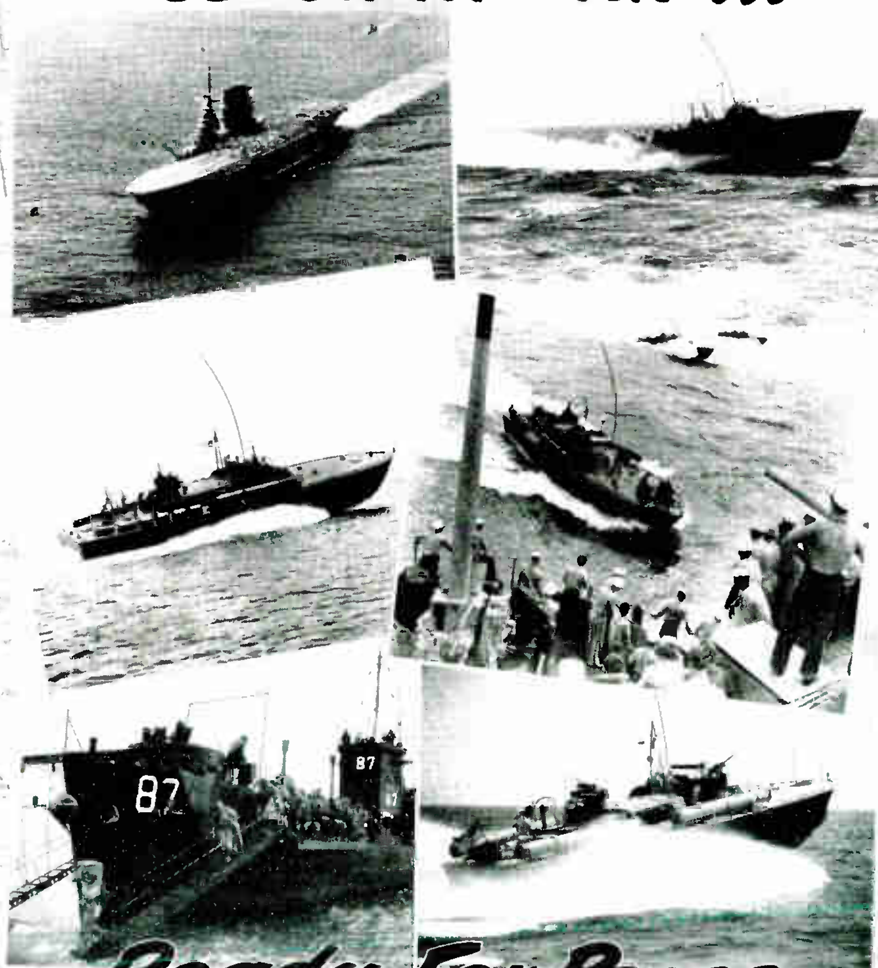
The HFA receiving condenser is a sturdier midget condenser intended for use in portable and aviation equipment where conditions of operations demand a better and more solid condenser. These, too, are available in a wide variety of size with both single and dual stators. All types, both HFA and HFB are of 100% soldered construction with brass plates, cadmium plated. Isolantite end plates.

Send for Latest Catalog!

HAMMARLUND MANUFACTURING CO., INC. • 460 West 34th Street • New York City

Premax Antennas

Tested in War...



Ready For Peace

Premax Products

DIV. CHISHOLM-RYDER CO., INC.

4519 Highland Ave., Niagara Falls, N.Y.

FOR LASTING SATISFACTION AND PRIDE IN AMATEUR RADIO ASSEMBLIES

AMPHENOL

Radio Cables and Parts

Amphenol Flexible Low-Loss Transmission Lines are a "must" in the new amateur-built sets. Also useful in television, frequency modulation, test equipment and similar applications. Plastic solid dielectric insures low losses under high frequency conditions. Coaxial and twinax types are illustrated. Impedances of from 50 to 72 ohms in the coaxial and 95 ohms in the twinax are basic.

When amateurs use Amphenol Parts in their equipment, they build in engineered units of dependable quality.

Amphenol Plastic Parts, known world-wide through their stalwart service in war and peacetime uses, have stood all tests of critical laboratory subjection and gruelling field service.

Some Amphenol products are featured on this page. There are many others available for the amateur who seeks the best in quality and long-life service. Send for Catalog No. 71 completely illustrative of Amphenol Radio Parts.

AVAILABLE SOON FROM YOUR DEALER

Low-Loss Connector Plug for use with U.H.F. Cables shown. Has mica filled bakelite insulation and silver plated metal parts.

Chassis or Box-Type Receptacle. One piece construction assures soldering ease. Conducts high-frequency current. Has low-loss dielectric. High conductivity socket contacts.

Amphenol Compact MIP Socket. World's strongest socket. Used extensively for the new, single-ended Octal tubes. Compact in size and low in capacity between contacts. Models available for high frequency. 1 5/16" mounting centers. An excellent "limited space" socket.

Standard Plug-In Coil Forms. Molded from Amphenol -912A polystyrene with low-loss characteristics. Designed for equipment which works the ultra-high frequencies.

Feed-Thru Bushing. Center hole accommodates wiring or rods up to 3/16" diameter. Made of No. 912A Amphenol polystyrene.

THE SET WORKS BETTER WHEN THESE MINIATURE SOCKETS ARE USED

Black Bakelite Miniature Socket has metal shield in center for grounding. Designed for 7 prong Miniature tubes which require a floating, low resistance contact.

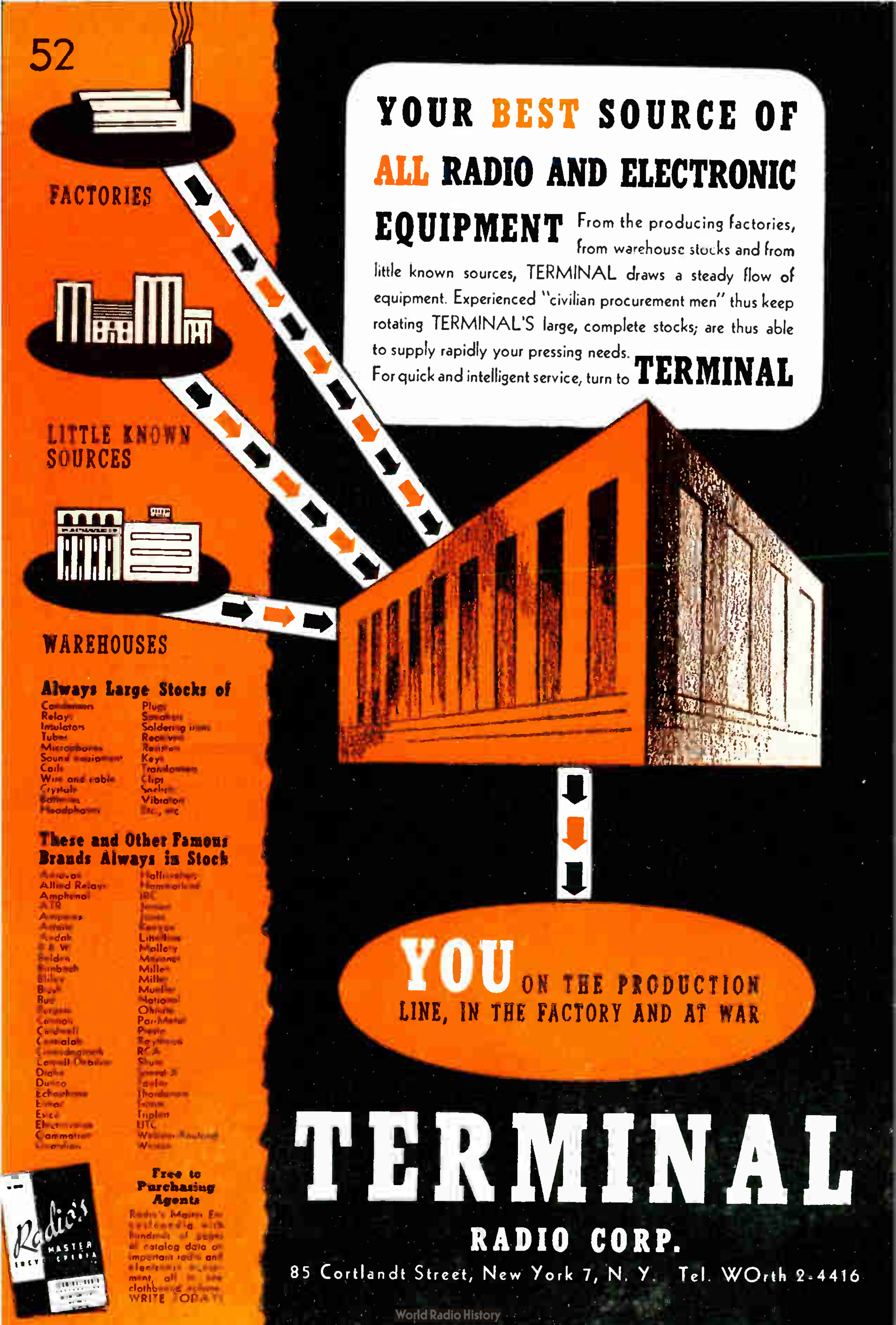
Polystyrene Miniature 5 Contact Socket for use with mating miniature coil form, to keep coil and circuit losses in U.H.F. equipment at a minimum.

Amphenol Socket for miniature tubes Especially molded for analyzers and checkers which have blank socket spaces, but do not have facilities for testing the new miniature tubes.

Depend upon AMPHENOL Quality

Connectors (AN, British, U.H.F.)
Fittings
Conduit
Cable Assemblies
U.H.F. Cable
Radio Parts
Plastics

AMERICAN PHENOLIC CORPORATION
CHICAGO 50, ILLINOIS
IN CANADA • AMPHENOL LIMITED • CANADA


FACTORIES
LITTLE KNOWN SOURCES
WAREHOUSES
YOUR BEST SOURCE OF ALL RADIO AND ELECTRONIC EQUIPMENT
From the producing factories, from warehouse stocks and from little known sources, TERMINAL draws a steady flow of equipment. Experienced "civilian procurement men" thus keep rotating TERMINAL'S large, complete stocks; are thus able to supply rapidly your pressing needs.
For quick and intelligent service, turn to TERMINAL
Always Large Stocks of
Relays
Insulators
Coils
Plugs
Keys
Clips
Vibrators
These and Other Famous Brands Always in Stock
Allied Relays
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Mallory
Miller
Par-Metal
RCA
Triplett
UTC
Free to Purchasing Agents
WRITE
YOU ON THE PRODUCTION LINE, IN THE FACTORY AND AT WAR
TERMINAL
RADIO CORP.
85 Cortlandt Street, New York 7, N. Y. Tel. WOrth 2-4416

THE AIRWAYS OF THE WORLD BECKON AGAIN TO THE WANDERER

To communicate freely with other amateurs around the world was once the happy avocation of "hams" in every part of the country. Through the war years, these hams turned to the grim business of battle and many heroic deeds are credited to them. Their training had been preparation for important wartime roles just as our background of thirty-five years' capacitor specialization enabled us to fulfill the enormous, war-quickened demands of the electronic industry in which capacitors were vital. It is good to turn to the exciting thoughts of tomorrow when the calls of friendship will go out again to far places. The great advances made in the design and building of C-D Capacitors during the war years will add much to the pleasures of peacetime amateur radio. Whatever your requirements, you will find C-D ready to supply you. Cornell-Dubilier Electric Corporation, South Plainfield, New Jersey.

SIX MODERN PLANTS

C-D has grown from one plant in 1940 to six great modern plants in 1944. Now better equipped to serve you with huge manufacturing facilities in central, convenient locations. C-D's six plants are located in South Plainfield, N. J., New Bedford, Worcester, and Brookline, Mass. and in Providence, R. I

CORNELL CAPAC

CORNELL-DUBILIER CAPACITORS

SIGMA *Instruments, Inc.*

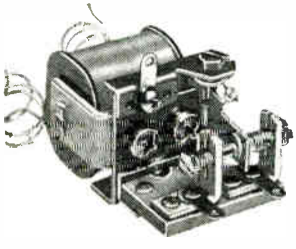
Specialists In Sensitive Relay Engineering and Manufacture

SIGMA SENSITIVE RELAYS FOR ELECTRONIC APPLICATIONS ARE FULLY GUARANTEED

SIGMA shares the responsibility for the proper operation of all Sigma relays . . . in other words, Sigma Instruments establishes the proper relay specifications in accordance with your problem requirements, then guarantees that the relays will perform with complete satisfaction.

Sigma Relays are not ordinarily sold as stock items, but upon definite recommendations as to the solution of particular problems. However, small quantities are available as quickly as stock items; because all components are in stock, and production is on the "short order" line.

Series 4 Recommended for: **Economy—Light weight—High Speed—Moderately severe environment (temperature and vibration)—Moderate sensitivity. *Aircraft performance on 30 to 50 milliwatt input.***

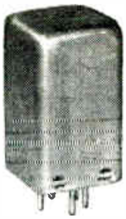


TYPE 4F
1⅜" x 1⅝" x 1 5/32"
Weight: 2 5/16 oz.

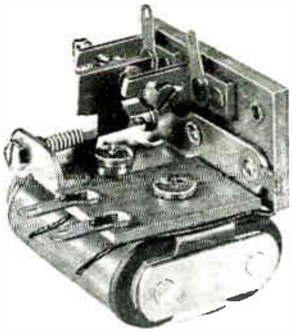
TYPE 4R
1½" x 1½" x 2⅝"
Mounted on 5-pin-tube base

TYPE 4A
2 5/16 " diam. x 2" high
Mounted on 5-pin-tube base

Series 5 are Best Adapted to: **Exceptionally severe environmental conditions, maintaining precise adjustment at extreme temperatures and after severe shocks (500 g's). Maximum sensitivity in small space and weight. *Aircraft performance on 5 milliwatt input.***



TYPE 5R
1½" x 1½" x 2¼"
Mounted on 5-pin-tube base



TYPE 5F
1⅜" x 1⅝" x 1⅜"

TYPE 5RH

Hermetic Seal: The 4R and 5R can be hermetically sealed (glass to metal) known as the 4RH and 5RH.

Exceptional Sensitivity Plus Performance in Difficult Environment

- Sigma Relays are available with coil resistances up to 18,000 Ohms.
- For A.C.—Both the types 4 and 5 are available with a built in full-wave rectifier giving D.C. sensitivity on A.C. input. The type 5 is available for time-delay service.

SIGMA can furnish specialized types for specific conditions and requirements . . . send us your problems.

SIGMA

Sigma Instruments, inc.

Sensitive RELAYS

78 CEYLON STREET BOSTON 21, MASS.


He'll Want the Finest!

JAMES KNIGHTS
Crystals

A Habit with us--
HERMETICALLY SEALED
ETCHED Crystals

The development of *Etched* quartz crystals was pioneered by engineers of the James Knights Company. Much of the original equipment now generally used in the crystal industry to produce these definitely superior type crystals was designed by James Knights men. Today, many thousands of J-K Etched Crystals in hermetically sealed holders are enabling our armed forces to better their communications.

When the Amateur returns to civilian radio operation he will find J-K Etched Crystals awaiting his most particular requirements. We look forward to cooperating with the Amateur and providing him with these better hermetically sealed Matched Crystals.

JK CRYSTALS

The JAMES KNIGHTS Co.
SANDWICH, ILLINOIS

CRYSTALS FOR THE CRITICAL

THERMADOR THERMATITE TREATING

Solves Heat  **Cold** 

Humidity  **Problems in**

Radio Transformers

Thermador Transformers are Thermatite treated to withstand extreme temperatures and humidity—arid or moist heat—dry or damp cold do not hamper their efficiency. Thermatite is the name of a process of accurate heat controlled vacuum impregnation developed and improved over a period of ten years.

Thermador also manufactures built-in Electric Heaters, Electric Ranges, Electric Water Heaters

THERMADOR TRANSFORMERS

THERMADOR ELECTRICAL MFG. CO.

5119 South Riverside Drive • Los Angeles 22, California

"Seven Leagues Ahead"

There's been a change of name!

Lafayette Radio Corporation

(CHICAGO AND ATLANTA)

is now known as

CONCORD RADIO CORPORATION

CONCORD • RADIO •

INDICATIVE OF THE CONCORD RADIO CORPORATION'S WARTIME AND POSTWAR DEVELOPMENTS IS THIS UNIQUE.

10-TUBE VERY HIGH FREQUENCY FM-AM RECEIVER KIT . . . WITH SEPARATE UFH TUNING UNIT

. . . for instruction in FM-AM receiver construction and high frequency technique

. . . when assembled, it becomes a perfect-working instrument

With the increased use of FM and Television in the postwar world, thousands of trained engineers, assemblers and maintenance men will be needed. This unique 10-tube FM-AM Receiver Kit, designed to approximate existing high frequency commercial and military units from a constructional standpoint, provides an inexpensive and practical way to lay the groundwork for your future operations.

Though this FM-AM Receiver Kit is primarily a "trainer," when assembled, it becomes a perfect-working instrument. The design is straightforward and simple . . . there are no frills or unnecessary components. All parts are made by nationally-known manufacturers, and all resistors and condensers are mounted on tie strips so that the kit may be easily torn down and reassembled. There are only two controls on the front panel—the tuning control and the volume control. In addition, there's a standard headphone jack for output, a switch for changeover from FM to AM, and a power switch in the AC line. Two models are available—one with a range from 88.6 to 107.6mc. . . . the other with a range from 115 to 140mc.

The Lafayette Radio Corporation

has, for 22 years, been one of the respected names in the field of radio and electronics. Our policies and our personnel, our reputation for integrity and outstanding service to our customers have enabled us to become one of the nation's great arsenals of radio and electronic equipment. Our customers include thousands of amateur radio operators as well as thousands of buyers in industry, government and military services, training schools, etc.

Although our organization is now known as the CONCORD RADIO CORPORATION we want to emphasize the fact that **nothing else besides the name has been changed.** Whether you order by mail, or visit us personally, you will find the same people to serve you, and the same fine merchandise and services as always. And, as usual, complete satisfaction is guaranteed in every transaction.

The Concord Radio Corporation

will, when the time is right, announce many new and unusual radio and electronic developments. Be sure that your name is on our list to receive CONCORD'S postwar literature and catalogs.

Free! Illustrated folder, completely describing this new FM-AM Receiver Kit. MAIL COUPON or send a penny postcard.

CONCORD RADIO CORPORATION
901 W. Jackson Blvd., Chicago 7, Ill., Dept. . . . .

Please send me your folder describing the new FM-AM Receiver Kit. Also put my name on your list for CONCORD'S literature and catalogs.

Name . . . .

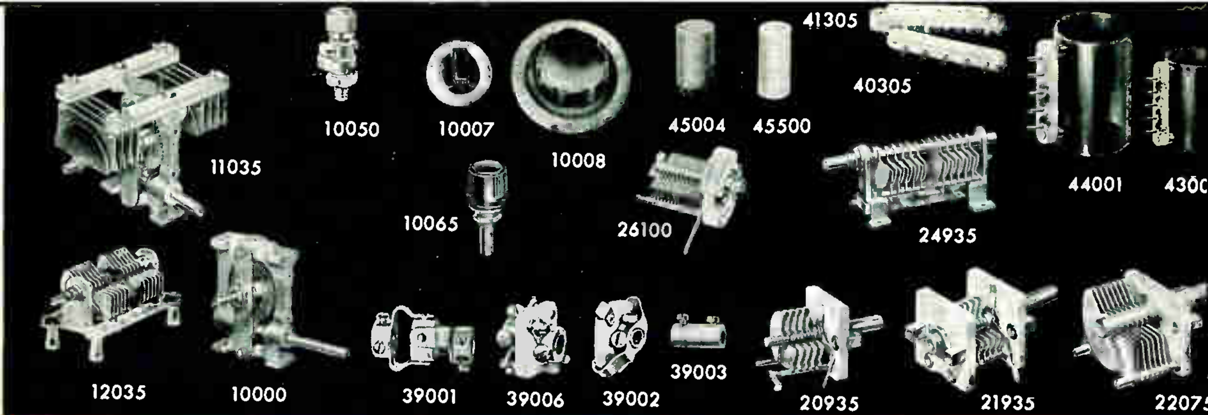
Address . . . .

City . . . . **State** . . . .

CONCORD RADIO CORPORATION

Lafayette Radio Corporation

901 W Jackson Blvd Chicago 7 [illegible] ★ [illegible] Street Atlanta 3 Georgia



MILLEN MODERN PARTS

MILLEN RADIO PRODUCTS are well designed MODERN PARTS for MODERN CIRCUITS, attractively packaged, moderately priced, and fully guaranteed. They have been designed with a view toward easy and practical application as well as efficient performance. For instance, the terminals are located so as to provide shortest possible leads, mounting feet are designed for easy insertion of screws and socket contacts, so that the solder won't run down inside them and make impossible the insertion of the tube, etc. Thus our slogan, "Designed for Application." Our general catalog is available for the asking either from your favorite parts supply house or direct from the factory.

11000, 12000, 13000, 14000 SERIES CONDENSERS
.077" air gaps for 3000 volt peak rating

MILLEN TYPE

Code	Capacity per side		Air Gap	Voltage Rating	Net Price
	Max.	Min.			
11035	36	4.6	.077"	3000	$6.90
11050	51	6.5	.077	3000	7.14
11070	74	9.5	.077	3000	7.80
13035	35	4.9	.077	3000	4.56
13050	49.5	6.3	.077	3000	5.20
13070	71	7.3	.077	3000	5.88
14200	204	10.7	.077	3000	7.80
14100	90.5	12.9	.171	6000	12.00
14050	50		.171	6000	7.20
14060	60		.265	9000	12.00

CONVENTIONAL SINGLE SECTION TYPE

Code	Capacity per section		Air Gap	Finish on Plates	Net Price
	Min.	Max.			
12935	9	37	.176"	Polished	$4.32
12936	9	37	.176	Plain	3.90
12536	6	43	.077	Plain	2.40
12551	7	55	.077	Plain	2.70
12576	9	76	.077	Plain	3.00
12510	12	101	.077	Plain	3.60
12515	18	151	.077	Plain	4.50

CONVENTIONAL DOUBLE SECTION TYPE

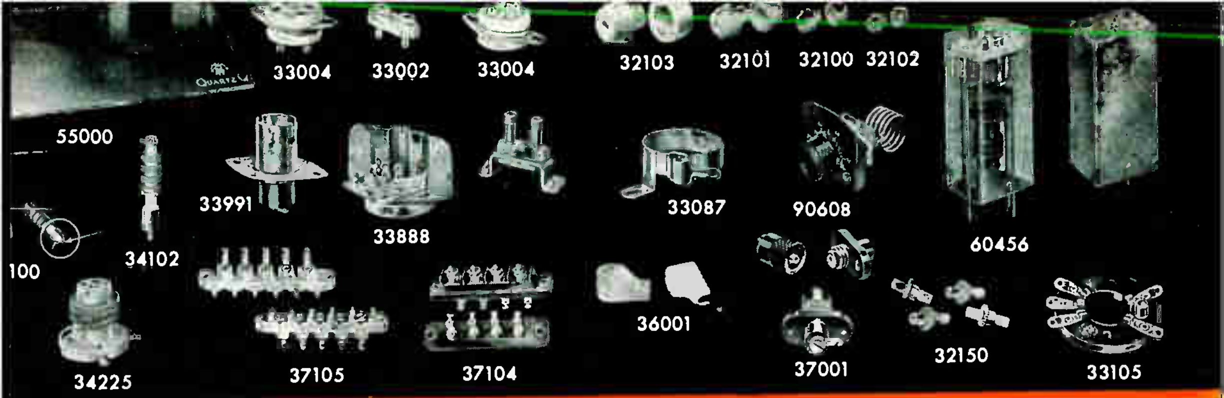
Code	Capacity per section		Air Gap	Finish on Plates	Net Price
	Min.	Max.			
12035	6	43	.077"	Polished	$4.32
12036	6	43	.077	Plain	3.90
12050	7	55	.077	Polished	5.10
12051	7	55	.077	Plain	4.32
12075	9	76	.077	Polished	5.61
12076	9	76	.077	Plain	5.40

Code	Description	Net Price
10000	Worm Drive Unit	$4.50
10001	Drum Meter Dial-0-100	1.85
10007	1 5/8" Nickel Silver Inst. Dial-0-100	.50
10008	3 1/2" Nickel Silver Inst. Dial-0-100	1.00
10050	Dial Lock	.45
10060	Shaft Lock for 1/4" Shafts	.36
10065	Vernier Drive Unit	.36
10067	Shaft Bearing, 1/4"	.21
15001	Neutral Condenser 0.7–4.3	.90
15002	Neutral Condenser 0.5–13.5	1.05
15003	Neutral Condenser 1.5–8.5	.90
15005	Neutral Condenser 3.4–14.6	2.00
15006	Neutral Condenser 2.8–9.1	3.00
20015	Steatite Ultra Midget 15 mmfd SS	.75
20035	Steatite Ultra Midget 35 mmfd SS	1.00
20050	Steatite Ultra Midget 50 mmfd SS	1.20
20100	Steatite Ultra Midget 100 mmfd SS	1.50
20140	Steatite Ultra Midget 140 mmfd SS	1.70
20920	Steatite Ultra Midget 20 mmfd DS	1.20
20935	Steatite Ultra Midget 35 mmfd DS	1.40
21050	Steatite Ultra Midget 50 mmfd SS	1.75
21100	Steatite Ultra Midget 100 mmfd SS	1.90
21140	Steatite Ultra Midget 140 mmfd SS	2.10
21935	Steatite Ultra Midget 35 mmfd DS	1.90
22075	Steatite Midget 75 mmfd SS	1.32
22100	Steatite Midget 100 mmfd SS	1.38
22140	Steatite Midget 140 mmfd SS	1.62
22915	Steatite Midget 15 mmfd DS	1.20
22935	Steatite Midget 35 mmfd DS	1.30
22950	Steatite Midget 50 mmfd DS	1.50
23075	Steatite Dual Midget 75 mmfd per section SS	2.60
23100	Steatite Dual Midget 100 mmfd per section SS	2.50
23925	Steatite Dual Midget 25 mmfd per section DS	2.25
23950	Steatite Dual Midget 50 mmfd per section DS	2.50
24100	100 mmfd per section, Single spaced	2.75
24935	35 mmfd per section, Double spaced	2.75
25130	93–130 Air Padder	1.50
26025	3.2–25 Air Padder	.96
26050	4–50 Air Padder	1.08
26075	4.3–76 Air Padder	1.20
26100	5–97 Air Padder	1.32
26140	6.5–140 Air Padder	1.60
26920	4.5–20 Air Padder	1.40
26935	5.5–36 Air Padder	1.50
27010	10 mmf Silver on Mica	.36
27025	25 mmf Silver on Mica	.36
27050	50 mmf Silver on Mica	.36
27100	100 mmf Silver on Mica	.36

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

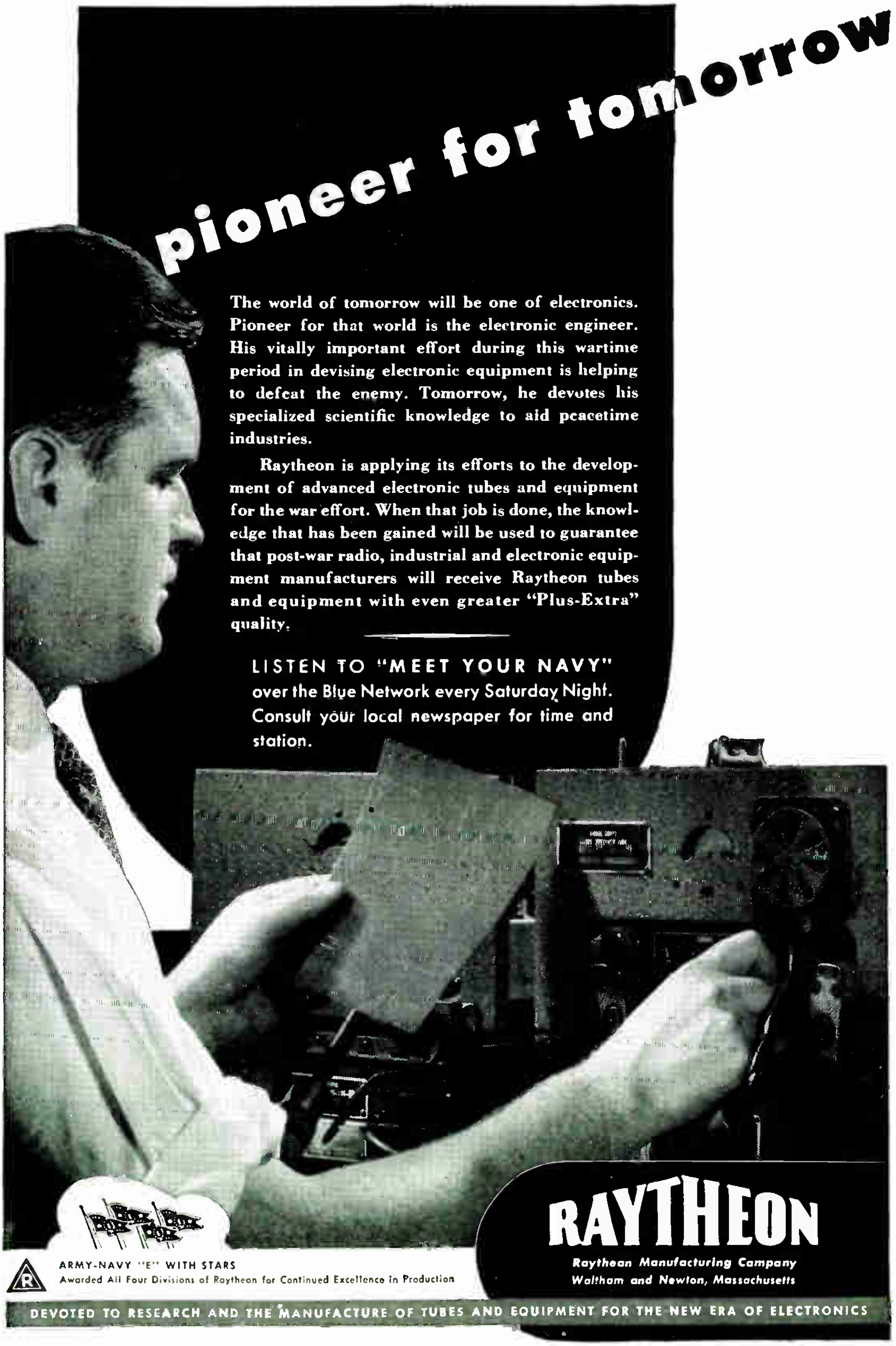
MALDEN, MASSACHUSETTS, U.S.A.



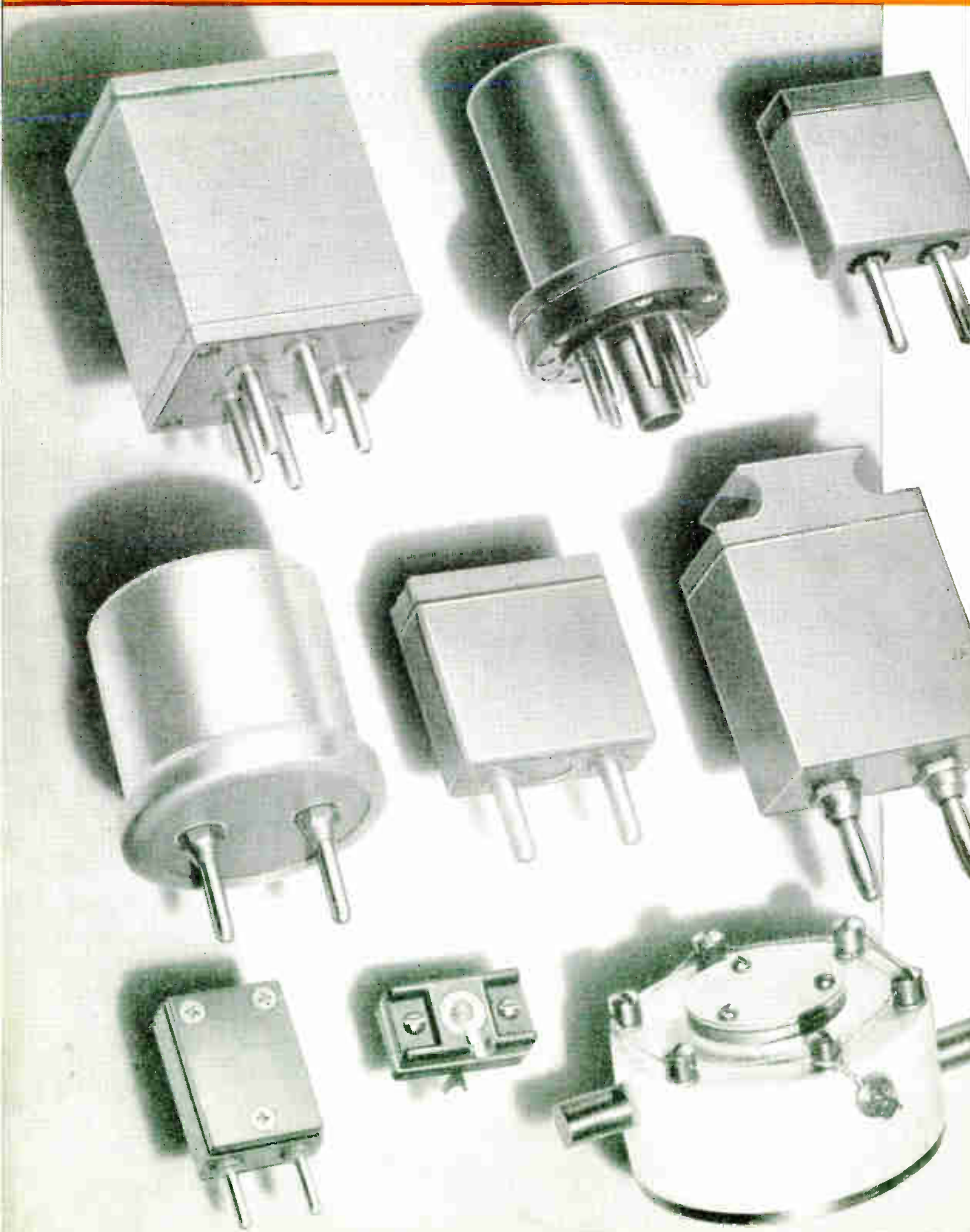
DESIGNED for APPLICATION

Code	Description	Net Price
27150	150 mmf Silver on Mica	$ .42
28030	30 mmfd Mica Padder	.15
30001	Standoff, ½ x 1⅜, QuartzQ	.15
30002	Standoff, ½ x 2⅞, QuartzQ	.21
30003	Standoff, ¾ x 2⅞, QuartzQ	.55
30004	Standoff, ¾ x 4⅞, QuartzQ	.65
31001	Standoff, ½ x 1, Isolantite	.20
31002	Standoff, ½ x 2½, Isolantite	.27
31003	Standoff, ¾ x 2, Isolantite	.30
31004	Standoff, ¾ x 3½, Isolantite	.42
31011	Cone, ¾ x ½, Steatite	.10
31012	Cone, 1 x 1, Steatite	.21
31013	Cone, 1½ x 1, Steatite	.27
31014	Cone, 2 x 1, Steatite	.75
31015	Cone, 3 x 1½, Steatite	.45
32100	Steatite Bushing for ⅜" hole	.30
32101	Steatite Bushing for ½" hole	.35
32102	Steatite Bushing for ¼" hole	.20
32103	Steatite Bushing for ¾" hole	.45
32150	Isolantite Thru-bushing, for ¼" hole	.05
32201	Steatite Bushing and Hardware	.75
32203	Steatite Bushing and Hardware	3.60
32300	Isolantite Bushing	1.80
33002	Crystal Socket	.25
33004	4 Prong Socket	.24
33005	5 Prong Socket	.24
33006	6 Prong Socket	.24
33007	7 Prong, Large, Socket	.24
33008	8 Prong, Octal, Socket	.24
33087	Base Clamp for 807 etc.	.30
33105	Acorn Socket, QuartzQ	.90
33888	Aluminum Shield for 33008	.18
33991	Socket for 991 etc.	.45
34010	Shielded 10 MH receiving	.75
34100	Universal 2.5 MH	.36
34101	Universal 2.5 MH, less Standoff	.30
34102	Commercial type 2.5 MH	.36
34140	Universal air core Transmitting	1.00
34150	Amateur Band Iron Core	1.75
34210	General Purpose RFC 10 MH	.60
34225	General Purpose RFC 25 MH	1.75
34240	General Purpose RFC 40 MH	.75
34285	General Purpose RFC 85 MH	1.25
34800	Interruption Frequency Oscillator Coil	1.20
36001	Ceramic Plate Cap, 9/16" for 866 etc.	.21
36002	Ceramic Plate Cap, ⅜" for 807 etc.	.21
37001	Black Bakelite Safety Terminal	.40
37104	Four Terminal, Black Bakelite	.60
37105	Five Terminal, Steatite	.75
37202	Steatite Plates, Pr.	.30
37211	Bracket	.15
37222	Terminal Posts, Pr.	.40
37501	Low Loss Mica Bakelite Safety Terminal	.55
38001	Isolantite 3/16" O.D. Beads (Pk of 50)	.30
38500	100 Beads, 5/16" dia., QuartzQ	.60
39001	Truly Flexible Isolantite	.36
39002	Conventional	.36
39003	Solid Brass N.P.	.21
39005	Universal Joint, Non-Insulated	.36
39006	Slide Action	.36
40205	Midget Plug	.24
40305	Intermediate size plug	.45
41205	Midget Socket	.30
41305	Intermediate size socket	.45
43001	QuartzQ blank form and plug	.90
43011		.90
43021	Midget coils for each	.90
43041	band. Mounted on No. 40205	.90

Code	Description	Net Price
43081	plug. No. 1 at end of code means	$ .90
43161	center link. No. 2, end link.	.90
44000	QuartzQ form 1¾" dia. x 3¾"	.75
44001	QuartzQ blank form and plug	1.20
44005		1.50
44010		1.50
44020		1.50
44040	"100 watt" coils	1.50
44080	for each band. Mounted on	1.90
44160	No. 40305 plug	2.10
44500	Swinging link and socket	1.75
45000	Coil Form, 1" dia. no p., low loss mica base Phenolic	.21
45004	Coil Form, 1" dia. 4 p., low loss mica base Phenolic	.30
45005	Coil Form, 1" dia. 5 p., low loss mica base Phenolic	.30
45500	Coil Form, ⅝" dia., Steatite	.45
46100	Coil Form, 1½" dia. no p., QuartzQ	.45
47001	Coil Form, ½" dia., QuartzQ	.10
47002	Coil Form, ½" dia., QuartzQ	.15
47003	Coil Form, ¾" dia., QuartzQ	.35
47004	Coil Form, ¾" dia., QuartzQ	.45
55001	Sheet, 3 x 8½ x .1, QuartzQ	.45
58000	Coil Dope, 2 oz., QuartzQ	.30
77083	"83" Hash Filter 250MA	1.00
77866	"866" Hash Filter 500MA	1.25 pr.
77872	"872" Hash Filter	1.40 pr.
79020	14mc Band Wave Trap	.90
79040	7mc Band Wave Trap	.90
79080	3.5mc Band Wave Trap	.90
79160	1.7mc Band Wave Trap	.90
	Air Trimmed	
60454	456 Diode Air Core	4.50
60455	456 Interstage (1) Air Core	4.50
60456	456 Interstage (2) Air Core	4.50
60501	5000 Interstage (2) Air Core	4.50
60502	5000 Diode Air Core	4.50
60503	5000 FM Interstage Air Core	4.50
60504	5000 FM Disc Air Core	4.50
62161	1600 Interstage Iron Core	2.50
62162	1600 Diode Iron Core	4.50
62454	456 Diode Iron Core	4.50
62456	456 Interstage Iron Core	4.50
63163	1600 BFO Air Core	4.50
63456	456 BFO Air Core	4.50
63503	5000 BFO Air Core	4.50
	Mica Trimmed	
67454	456 Diode Iron Core	1.25
67456	456 Interstage Iron Core	1.25
67503	5000 FM Interstage Air Core	1.50
67504	5000 FM Disc Air Core	1.50
	Permeability Tuned	
64454	456 Diode (2)	1.50
64456	456 Interstage (2)	1.50
65456	456 BFO	1.50
	Triple Tuned	
66454	456 Diode	1.75
66456	456 Interstage	1.75
90600	Complete set of four Wavemeters, in case	12.00
90605	Range 2.8 to 9.7 mc. Wavemeter	3.00
90606	Range 9.0 to 28 mc. Wavemeter	3.00
90607	Range 26 to 65 mc. Wavemeter	3.00
90608	Range 50 to 140 Wavemeter	3.00
90721	Hetrofil	4.00


pioneer for tomorrow
The world of tomorrow will be one of electronics. Pioneer for that world is the electronic engineer. His vitally important effort during this wartime period in devising electronic equipment is helping to defeat the enemy. Tomorrow, he devotes his specialized scientific knowledge to aid peacetime industries.
Raytheon is applying its efforts to the development of advanced electronic tubes and equipment for the war effort. When that job is done, the knowledge that has been gained will be used to guarantee that post-war radio, industrial and electronic equipment manufacturers will receive Raytheon tubes and equipment with even greater "Plus-Extra" quality.
LISTEN TO "MEET YOUR NAVY"
over the Blue Network every Saturday Night.
Consult your local newspaper for time and station.
ARMY-NAVY "E" WITH STARS
Awarded All Four Divisions of Raytheon for Continued Excellence in Production
RAYTHEON
Raytheon Manufacturing Company
Waltham and Newton, Massachusetts
DEVOTED TO RESEARCH AND THE MANUFACTURE OF TUBES AND EQUIPMENT FOR THE NEW ERA OF ELECTRONICS

In QUARTZ CRYSTALS for amateur or
significant advancements have



commercial radio, the most been introduced by... Bliley

This is a message from Bliley to the thousands of amateurs and professional engineers who are now serving their country in the armed forces and in essential communications industries. Bliley "grew up" with them.

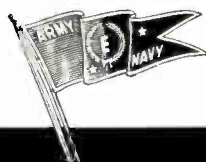
To these men and women Bliley crystals are still a familiar sight. They recognize, in the military crystal units used by our armed forces, many basic features that were pioneered by Bliley for application in peacetime services.

When tremendous production was demanded by our armed forces Bliley had the engineering background, the facilities and the production experience to provide a firm corner stone on which this volume production of radio crystals was successfully built. And, from the ranks of talented amateurs and radio engineers came a host of long-time friends who knew exactly how to use them.

But research has continued and experience has grown mightily to meet the challenge of war requirements. With the return to peace, and relaxation of wartime restrictions there will be better Bliley crystals for every application as well as new Bliley crystals for the new services that loom on the horizon. That's a promise.

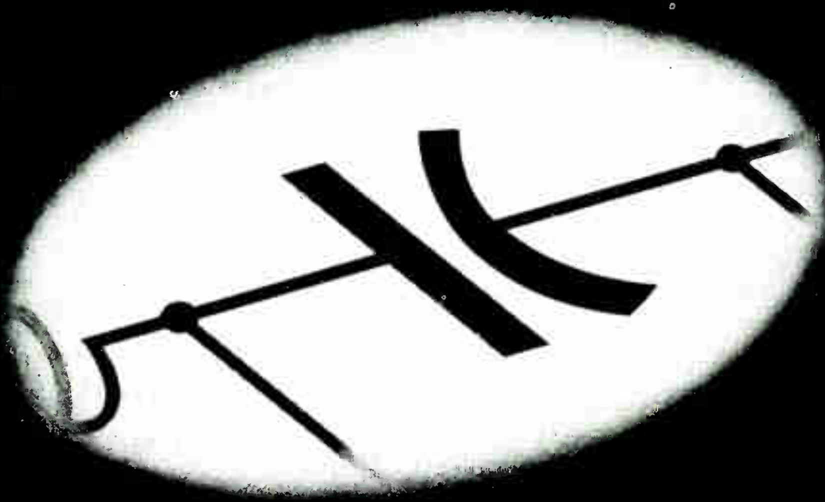
To our old friends, amateurs and professional engineers, we say, "Look to Bliley for crystal units that embody every advanced development."

DO MORE THAN BEFORE
BUY **EXTRA** WAR BONDS



Bliley
CRYSTALS

BLILEY ELECTRIC COMPANY, ERIE, PA.



ARMY
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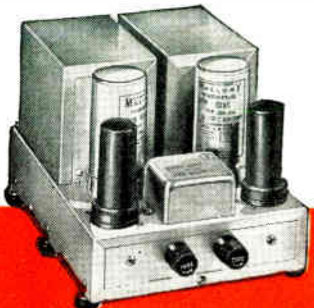
NAVY

P. R. MALLORY & CO. Inc.

MALLORY

Approved Precision Products



Mallory standard Vibrapacks include nominal input voltages of 6, 12, and 32 volts DC. Nominal output voltages from 125 to 400. Models available with switch for four output voltages in 25-volt stages. Hermetically sealed vibrators. High efficiency—low battery drain.

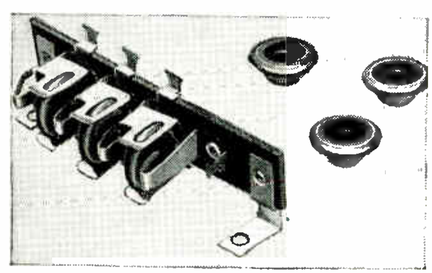
Vibrapacks* Provide Dependable Plate Power for Portable Equipment

Mallory Vibrapacks provide economical, efficient and dependable plate power for operating radio receivers, transmitters, PA systems, direction finders and other electronic equipment on vehicles, farms, portable equipment, or wherever commercial AC power is unavailable.

Write for Form E-555-D for detail information.

Mallory Transmitting Capacitors

Built with adequate safety factor for long life. Round and square can styles. Available in 20 stock sizes, working voltages from 600 to 6,000.



Better Amplifier Performance with Mallory Grid Bias Cells

Grid bias cells simplify circuits—eliminate parts—obtain wider frequency response, greater stability, and lower hum level. DC potential requires neither filtering nor interstage isolation. Available in 1 or 1¼ volt types. Holders available for one to four cells in series.

OTHER MALLORY PRODUCTS INCLUDE:

- Attenuators
- Battery Chargers
- Capacitors, Dry Electrolytic
- Capacitors, Paper
- Dial or Pilot Lights
- Ham Band Switches
- Jacks
- Jack Switches
- Knobs
- "L" Pad Attenuators
- Noise Filters
- Plugs
- Potentiometers
- Push Button Switches
- Rectifiers, Dry Disc
- Resistors
- Rheostats
- "T" Pad Attenuators
- Variohm* Resistors
- Replacements
- Vibrators
- Vitreous Resistors
- Volume Controls

*Reg. U. S. Pat. Off.

All Mallory standard parts available from Mallory Distributors

Write for Catalog M-773

FOR IMPROVED PERFORMANCE USE MALLORY APPROVED PRECISION PRODUCTS

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For complete, balanced, fully guaranteed instrumentation . . .

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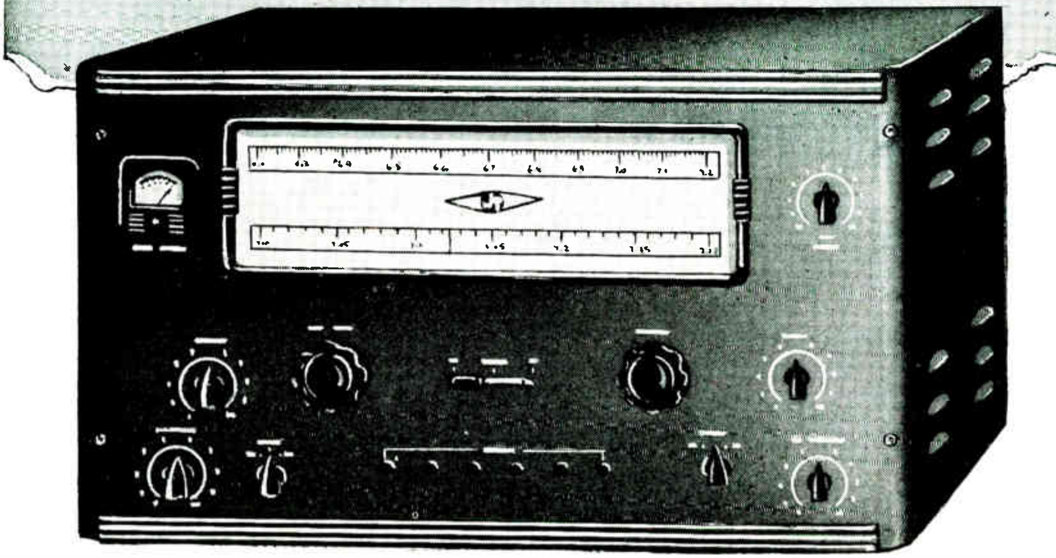

This new communication receiver, which meets the requirements of the advanced amateur and professional operator, is the latest development of Jefferson-Travis engineers. The Model CR-1 is a 15 tube superheterodyne which covers the frequency range of 540 Kc. to 32 Mc. in 5 bands. Tuning is simplified by the use of a large horizontal dial showing only the selected frequency band. A substantial amount of mechanical and electrical band spreading is available in the regular tuning system but, in addition to this, a supplementary tuning system is provided which permits the selection of any particular frequency band and spreading it out over the full length of the dial. A continuously variable selectivity system, *electronically controlled,* makes possible a variation of selectivity ranging from crystal filter sharpness to a broad flat top 16 Kc. wide for the reception of maximum fidelity consistent with broadcast conditions. The crystal filter and phasing control assure maximum selectivity on both phone and CW and has a high rejection ratio to separate closely spaced signals.

The pre-selector has two stages of selective tuned RF, eliminating image response, increasing signal strength and assuring better signal to noise ratio. The efficient new noise limiter incorporated in the Model CR-1 is superb in keeping interference caused by atmospheric and electrical conditions at a minimum.

All the oscillator circuits are designed for maximum stability over a wide range of supply voltage and temperature variations.

A 10" speaker, housed in a separate cabinet, is used with the high fidelity audio system. This system has push-pull output and provides an undistorted output of 10 watts. Other features built in this modern receiver include: Automatic Volume Control (providing a constant output for a very wide range of signal levels), RF and Audio Volume Controls, Push Button Band Selection, Beat Frequency Oscillator, Send-Receive Switch, Phone Jack and an S Meter for checking the relative signal strength of incoming signals.

The power supply operates with an extremely low hum level from 110 volts AC 60 cycles. The power supply is self contained in the set, and the entire Model CR-1 is constructed to provide maximum mechanical stability. The sturdy metal cabinet used has been designed so that the equipment can be used for either rack or cabinet mounting. Additional technical information available upon request. Specifications subject to change.

RCA-8005 TRANSMITTING TRIODE DE LUXE — Most powerful of the small triodes, a single RCA-8005 in plate-modulated service will take 240 watts (ICAS) with only 9 watts of grid drive at frequencies up to 60 Mc. The RCA-8005 has the same physical dimensions as the famous 809 and 812. *Net Price*, **$7.00**

RCA-816 HALF-WAVE MERCURY-VAPOR RECTIFIER — Designed and priced for real economy in transmitters of 400 watts output or less. Small as a receiving tube, but handles a peak inverse of 5000 volts, and a peak plate current of 0.5 amperes. Two RCA-816's in a full-wave circuit can deliver 1600 volts at 250 ma. with good regulation and exceptionally long life. *Net Price*, **$1.00**

RCA-815 PUSH-PULL R-F BEAM POWER AMPLIFIER — Providing push-pull beam power within one tube envelope, the RCA-815 will deliver an output of over 60 watts (class C telegraphy, ICAS) on all frequencies up to 125 Mc. It requires a plate voltage of only 400 to 500 volts, needs less than ½-watt of grid drive, and generally requires no neutralization on the lower frequencies, although it may at the higher frequencies. *Net Price*, **$4.50**

RCA-826 TRIODE FOR THE ULTRA HIGHS —Operates at maximum CCS ratings (60 watts plate dissipation) at frequencies as high as 250 Mc and at reduced ratings as high as 300 Mc. Specifically designed for use as an oscillator, r-f or frequency multiplier at the ultra-high frequencies. All terminals at one end of bulb permit use of short leads in neutralizing circuits. *Net Price*, **$19.00**

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YOU are undoubtedly planning ahead toward the day when you can resume your amateur operations. For your convenience in planning, we are listing here a part of the complete RCA line of tubes which covers the more important amateur types . . . even though they are not immediately available to you because of the War Emergency. We, too, look forward to the day when you can resume your place on the air . . . and when you can again rely on RCA for *performance-plus-economy* in tubes. Remember, *the Magic Brain of all electronic equipment is a Tube . . . and the fountain-head of modern Tube development is RCA.*

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802	Pentode	33 Watts*	3.50
803	Pentode	350 Watts	25.00
804	Pentode	150 Watts*	15.00
805	Triode	315 Watts	11.00
806	Triode	1000 Watts*	22.00
807	Beam	75 Watts*	2.25
808	Triode	200 Watts	7.75
809	Triode	100 Watts*	2.50
810	Triode	620 Watts*	13.50
811	Triode	225 Watts*	3.50
812	Triode	225 Watts*	3.50
813	Beam	360 Watts	22.00
815	UHF Twin Beam	75 Watts*	4.50
826	UHF Triode	125 Watts	19.00
828	Beam	270 Watts*	17.50
834	UHF Triode	125 Watts	12.50
1623	Triode	100 Watts*	2.50
1624	Beam	54 Watts	2.40
8000	Triode	620 Watts*	13.50
8003	Triode	330 Watts	12.00
8005	De Luxe Triode	300 Watts*	7.00
8012	UHF Triode	50 Watts	14.00
8025	UHF Triode	50 Watts*	14.50

*ICAS Rating

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816	5000 Volts	.125 amp	1.00
866-A/866	10000 Volts	.25 amp	1.50
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3AP1/906P1	3" Green Phosphor	13.50

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6C4	Miniature Triode, Amplifier, Oscillator	.90
6F4	Acorn Triode, Oscillator	10.25
6J4	Miniature Triode, UHF Amplifier	8.35
6J6	Miniature Twin Triode, Amplifier, Oscillator	1.85
954	Acorn Pentode, Amplifier, Detector	4.50
955	Acorn Triode, Detector, Oscillator	2.85
956	Acorn Pentode, Super-Control Amplifier	5.00
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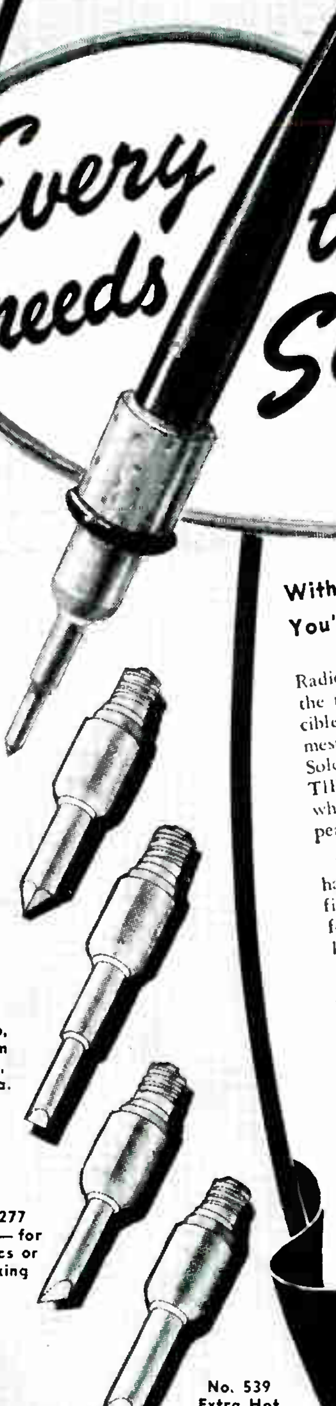

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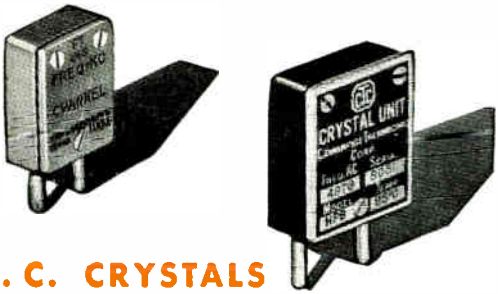
I-F TRANSFORMERS

TURRET TERMINAL LUGS

SPLIT LUGS

DOUBLE END TERMINAL LUGS



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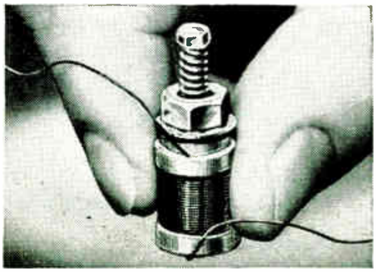

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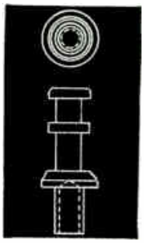
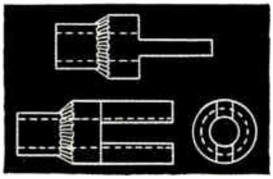
Just swage these heavily silver plated Turret Terminal Lugs to the board and in a jiffy you have a good, firm turret terminal. Quick soldering, too. Sufficient metal is used in the Lugs to give them strength but not enough to draw heat thus increasing soldering time.



C. T. C. Turret Terminal Lugs are stocked to meet 1/32", 2/32", 3/32", 4/32", 6/32", and 8/32" board thicknesses.

C. T. C. SPLIT LUGS

A .050 hole through the shaft permits wiring to these Split Lugs from either top or bottom without drilling or cutting. Just swage them to the board, then wire. Made of brass, heavily silver plated, C. T. C. Split Lugs are available in two sizes to fit 3/32" and 5/32" boards.



DOUBLE END TERMINAL LUGS

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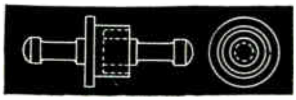

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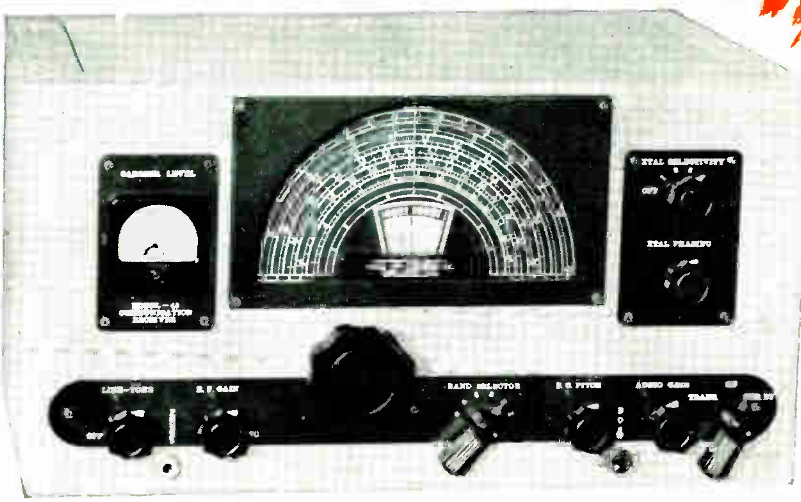

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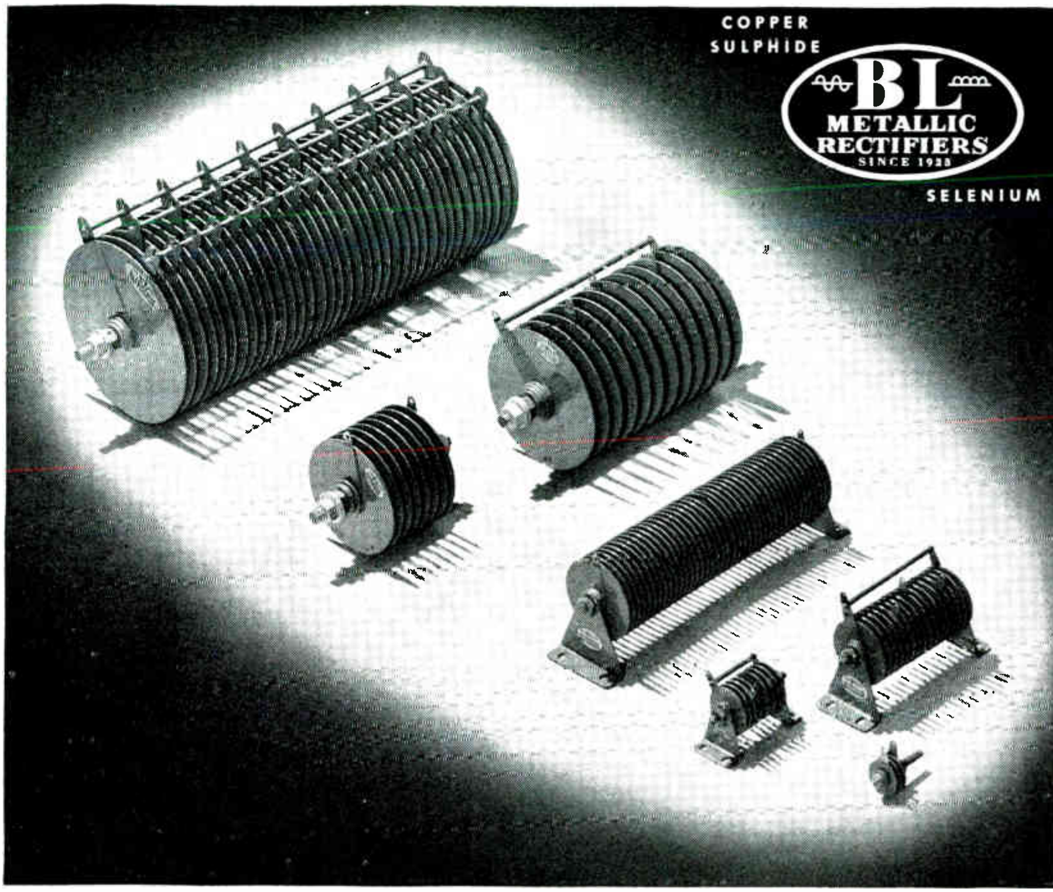

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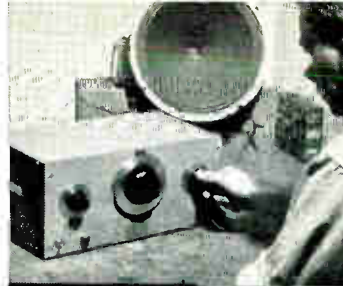

Audio frequency oscillators of the resistance-tuned type. Excellent frequency stability and freedom from wave form distortion. Model 200B Resistance-Tuned Oscillator above

Research into the fields of electrical engineering, physics, chemistry and all phases of science is greatly aided by the use of modern electronic measuring devices. The accuracy with which measurements are made is the yardstick of progress. Time is always an important element. Thus, electronic instruments should combine speed of operation with accuracy.

To gain speed of operation without sacrifice of accuracy, all *-hp-* instruments are designed to operate with a minimum of adjustment. For the first time a variable frequency audio frequency oscillator which does not require a zero adjustment is available. The *-hp-* vacuum tube voltmeter is as simple to operate as a multi-range d.c. instrument.

To accomplish certain measurements more quickly *-hp-* model 205AG incorporates a resistance-tuned audio oscillator, input meter, output meter, attenuator and impedance matching system . . . all in a single unit.

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Audio signal generators providing standardized voltage throughout the audio frequency range. Model 205AG Audio Signal Generator above

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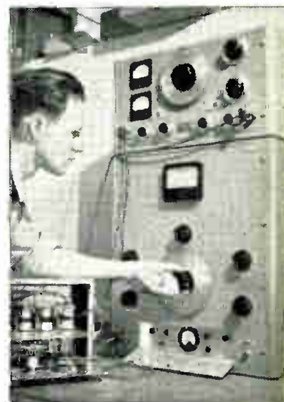
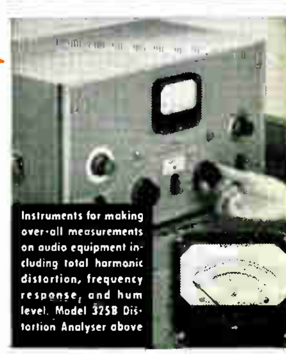


Harmonic wave analysers with many new features including variable band width. Model 300A wave analyser above



Instruments for making over-all measurements on audio equipment including total harmonic distortion, frequency response, and hum level. Model 325B Distortion Analyser above

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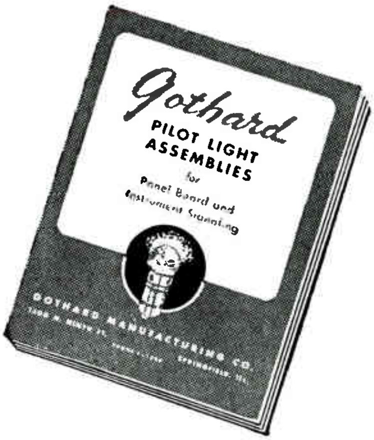

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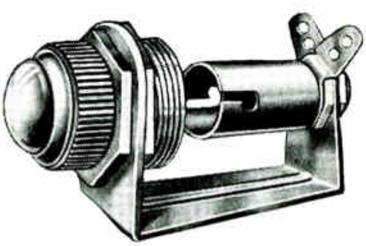
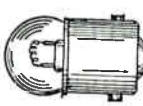
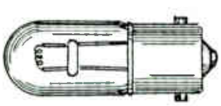
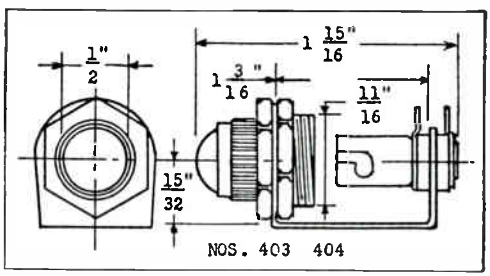




Mazda No. 51 .... 6-8 volt
Mazda No. 53 .... 12-16 volt
Mazda No. 356 ... 28 volt



Mazda No. 44 .... 6-8 volt
Mazda No. 1815 .. 12-16 volt
Mazda No. 313 ... 28 volt



No. 403—Faceted Jewel .... List $0.50
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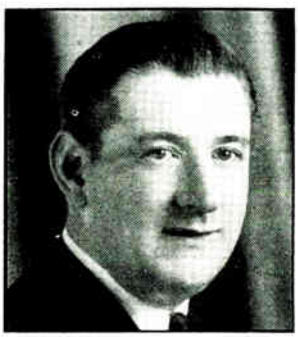

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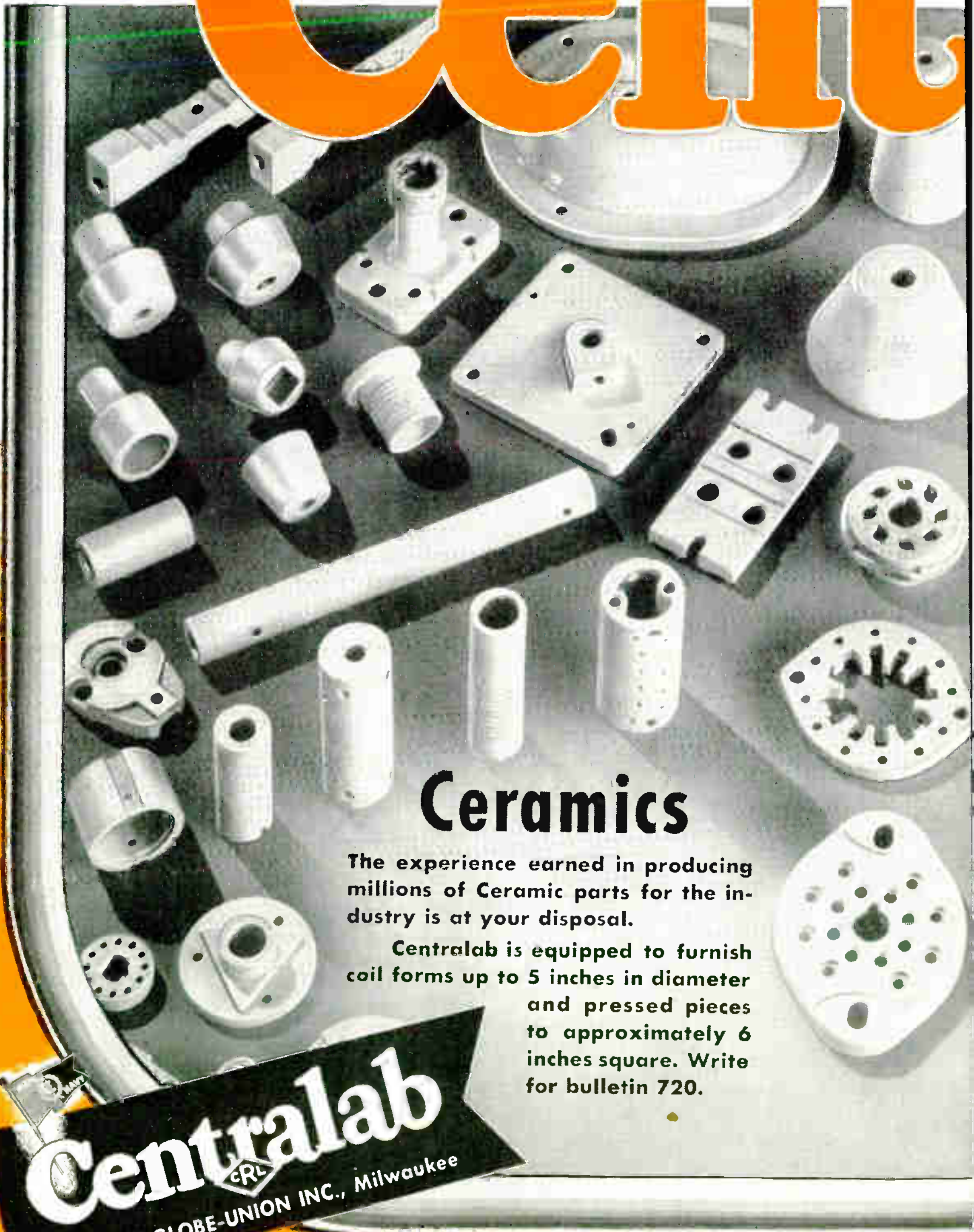

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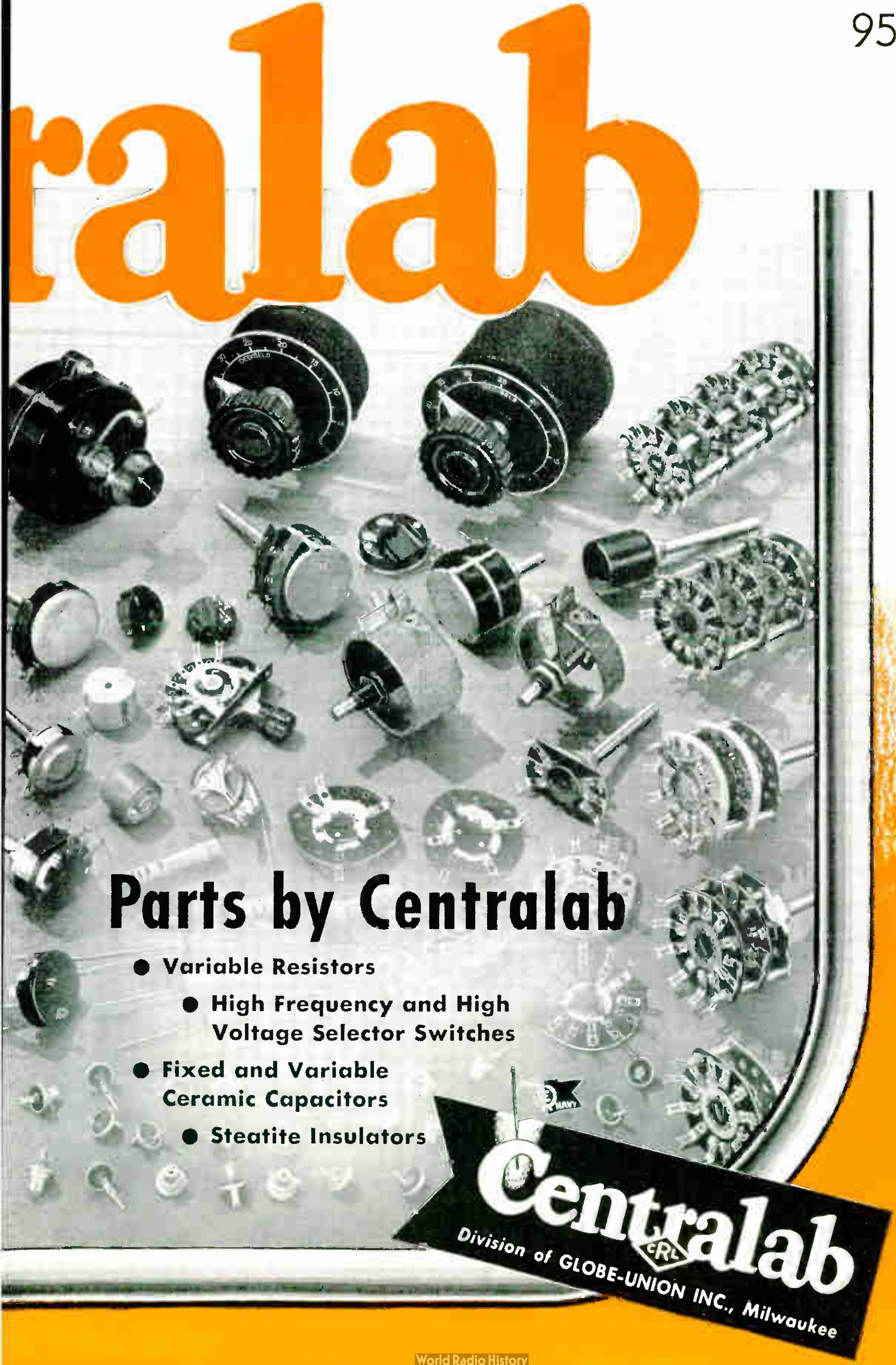
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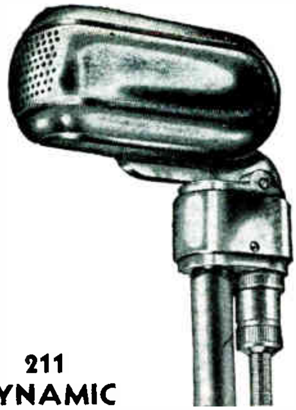

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No. 211 is a Rugged Dynamic utilizing a new type magnet structure and acoustic network. The high frequency range has been extended and the extreme lows have been raised 2 to 4 decibels to compensate for overall deficiencies in loud speaker systems. Unique diaphragm structure results in extremely low harmonic and phase distortion without sacrificing high output level. Tilting head, balanced line output connection. Chrome or gunmetal finish.

33D Dynamic
33X Crystal

The full satin chrome finish of 33D Dynamic adds class to any rig. 90° tilting head gives semi- or non-directional pick-up. Output level −54DB. Range 40–9,000 cycles. Ruggedly built for recorder or P.A. work. Built in transformer free from hum pick-up. Available in 200 or 500 ohms or high impedance. 33X Crystal same appearance as 33D. Level −52DB. Range 30–10,000 cycles.

22X
22D

22X Crystal is tops in performance. Reproduces clean and sharp. Smart engineering cuts feedback to minimum. Tilting head and removable 7-foot cable set. Built-in wind-gag permits outdoor operation. Crystal impregnated against moisture. Automatic barometric compensator. Chrome type finish. Level −52 DB. Range 30–7,000 cycles.

22D Dynamic is identical in appearance with 22X but has high level dynamic cartridge. Dependable indoors or out. Output −54 DB. Range 30–8,000 cycles. 200 or 500 ohms or high impedance.

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Plus Performance at Low Cost

Model CX

Crystal, in rich brushed chrome finish, with 7 foot removable cable set using Amphenol connectors. Level −55 DB. Range 50–7,000 cycles.

Model CD

Dynamic, same style and finish as CX, with removable 7 foot cable set. In 200–250 ohms, 500 ohms or hi-impedance. Level −52 DB. Range 50–7,000 cycles.

Model BX

Crystal mike for recording, P.A. and ham work. Bronze enamel finish. Level −55DB. Range 50–5,500 cycles. An excellent unit. With 7 foot cable.

Model BD

Dynamic, same finish as BX. Works indoors or out. Level −42 DB. Range 50–5,000 cycles. 200–250 ohms, 500 ohms or high impedance with 7 foot cable.

Han-D

Hang it, hold it, use it on desk or floor stands. Han-D does the job of several mikes. Available as 9X Crystal, in brushed chrome finish, Level −48DB, or 9D Dynamic in brushed chrome or gunmetal. Level −50DB, 200 or 500 ohms or hi-impedance.

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TRANSMITTING and RECTIFYING TUBES

AMPEREX TRANSMITTING TUBES

WATER-COOLED TYPES

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Mu	Gm	Capacitance Grid to Plate	*PLATE Max. Volts	*PLATE Max. Amps.	*PLATE Max. Dissipation Watts	*Nominal Output Watts	Max. Freq. MC. At Max. Plate Input	Max. Freq. MC. At 50% Max. Plate Input
207	22.0	52.0	20.0	6500	27.0	15000	2.00	10000	C20000	1.6	20
220C	21.5	41.0	35.0	5000	22.0	15000	1.00	10000	BR2500	3.0	30
228A	21.5	41.0	17.0	6500	23.4	6000	.75	3000	BR1000	1.5	20
232C	20.0	72.0	40.0	8000	22.0	20000	2.00	20000	BR8500	1.5	20
342A	20.0	67.0	40.0	6820	27.0	18000	1.40	25000	BR8500	4.0	16
343A	21.5	57.5	40.0	6750	23.5	15000	.70	10000	BR3500	4.0	8
520B	22.0	34.0	17.0	5000	27.0	10000	1.20	5000	C5000	2	......
846	11.0	51.0	40.0	2800	9.0	7500	1.00	1600	C2500	50	150
848	22.0	52.0	8.0	4200	27.0	12000	....	7500	A2000	1.6	20
858	22.0	52.0	42.0	4500	18.0	20000	2.00	20000	C20000	1.6	40
859	11.0‡	71.0	36.0	8000	15.0	20000	3.50	20000	C35000	1.5	40
863	22.0	52.0	50.0	7000	27.0	15000	2.00	7500	B22000	1.6	20
889	11.0	195.0	21.0	8000	17.5	7500	2.00	5000	C10000	50	150
891	11.0‡	60.0	8.0	4200	27.0	15000	2.00	5000	B22000	1.6	20
892	11.0‡	60.0	50.0	7000	32.0	10000	1.00	6600	CP6000	1.6	20
1652	14.5	52.0	14.0	....	27.0	7500	1.25	5000	C6000	1.5	10
HF50K	27.0** / 13.5**	100.0 / 200.0	36.0	1600	20.0	20000	5.00	30000	øC25000	10	50

‡ Single or two-phase filament (two units); voltage is per unit. ** Single or two-phase filament excitation. ø At upper frequency limit of 50 megacycles.

FORCED-AIR COOLED TYPES

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Mu	Gm	Capacitance Grid to Plate	*PLATE Max. Volts	*PLATE Max. Amps.	*PLATE Max. Dissipation Watts	*Nominal Output Watts	Max. Freq. MC. At Max. Plate Input	Max. Freq. MC. At 50% Max. Plate Input
220R†	21.5	41.0	35.0	5000	22.0	12500	1.00	6000	BR2500	4.0	30
232R†	20.0	72.0	40.0	8000	22.0	12500	2.00	7500	CP10000	3.0	20
343R†	21.5	57.5	40.0	6750	23.5	7500	1.50	5000	CP5000	4.0	30
889R‡	11.0	125.0	21.0	8000	19.0	6000	1.00	3000	CP4000	25	100
891R†	11.0‡	60.0	8.0	4200	28.0	10000	2.00	4500	B10000	1.6	20
892R†	11.0‡	60.0	50.0	7000	32.0	8500	1.00	4000	CP5000	1.6	20
HF3000°	21.5	40.5	16.0	6500	10.0	10000	1.35	3000	C7500	20	50
ZB3200°	21.5	40.5	85.0	5000	10.0	10000	1.50	2000	B8000	20	50

* $75.00 credit will be allowed against purchase of new tube if radiator and crate are returned in good condition.
‡ Single or two-phase filament (two units); voltage is per unit.
° All glass radiation and air-cooled transmitting tubes.
† $100.00 credit will be allowed against purchase of new tubes if radiator and crate are returned in good condition.

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Mu	Gm	Capacitance Grid to Plate	*PLATE Max. Volts	*PLATE Max. Ma.	*PLATE Max. Dissipation Watts	*Nominal Output Watts	Max. Freq. MC. At. Max. Plate Input	Max. Freq. MC. At. 50% Max. Plate Input
AB-150	10.0	3.25	5.3	3400	9.5	1500	......	100	AB150	......	......
HF- 60	10.0	2.50	20.0	5000	5.2	1600	150	60	C100	30	100
HF- 75	10.0	3.25	12.5	4000	2.0	2000	120	75	C150	75	200
HF-100	10.0	2.50	23.0	4200	4.5	1500	150	75	C150	30	150
HF-120	10.0	3.25	12.0	4500	10.5	1250	175	100	C150	20	80
HF-125	10.0	3.25	25.0	4500	11.5	1500	175	100	C200	30	90
HF-130	10.0	3.25	12.5	4300	9.0	1250	210	125	C170	20	90
HF-140	10.0	3.25	12.0	4500	12.5	1250	175	100	C150	15	60
HF-150	10.0	3.25	12.5	4300	7.2	1500	210	125	C200	30	100
HF-175	10.0	4.00	18.0	5000	6.3	2000	250	125	C300	25	100
HF-200	10.5	4.00	18.0	5000	5.8	2500	200	150	C350	20	100
HF-250	10.5	4.00	18.0	5000	5.8	3000	200	150	C375	20	100
HF-300	11.0	4.00	23.0	5600	6.5	3000	275	200	C600	20	100
ZB-120	10.0	2.50	30.0	5000	5.2	1500	160	75	B300	30	90
111H	10.0	2.50	23.0	4200	4.6	1500	160	75	C175	25	50
203A	10.0	3.25	25.0	4500	13.5	1250	175	100	C150	15	80
203H	10.0	3.25	25.0	4500	11.5	1500	175	100	C200	30	90
204A	11.0	3.85	23.0	4000	15.0	2500	275	250	C500	3	30
211	10.0	3.25	12.0	4500	12.5	1250	175	100	C150	15	80
211C	10.0	3.25	12.5	4300	9.0	1250	210	125	C175	20	90
211H	10.0	3.25	12.5	4300	7.2	1500	210	125	C200	30	100
212E	14.0	6.00	16.0	8000	19.0	2000	350	275	BR75	1.5	3.0
241B	14.0	6.00	16.0	8500	18.8	2000	350	275	C400	7.5	20
242A	10.0	3.25	12.5	3600	13.0	1250	150	85	A20	6	25
242B	10.0	3.25	12.5	3600	13.0	1250	150	100	A20	6	25
242C	10.0	3.25	12.5	3600	13.0	1250	150	100	A20	6	25
251A	10.0	16.00	10.5	3800	8.0	3000	600	1000	C1200	30	60
261A	10.0	3.25	12.0	4000	9.0	1250	210	125	C175	30	50
270A	10.0	9.75	16.0	5700	21.0	3000	375	350	C700	7.5	20
276A	10.0	3.25	12.0	4000	9.0	1250	210	125	C175	30	50
279A	10.0	21.00	10.0	5000	18.0	3000	800	1200	BR500	20	40
304B	7.5	3.25	11.0	2000	2.5	1250	100	50	C85	100	350
308B	14.0	6.00	8.0	7500	17.4	2250	325	250	A50	1.5	3
800	7.5	3.25	15.0	2000	2.5	1250	80	35	C65	60	190
801	7.5	1.25	8.0	1600	6.0	600	70	42	C25	60	120
805	10.0	3.25	50.0	4800	6.0	1500	210	125	B400	30	80
810	10.0	4.50	35.0	5000	4.8	2000	250	125	C375	30	100
830	10.0	2.50	8.0	2000	9.9	750	130	40	C60	6	50
830B	10.0	2.50	25.0	3080	11.0	1000	150	60	B175	15	65
833	10.0	10.00	35.0	8000	6.3	3000	500	300	C1000	30	100
834	7.5	3.25	11.0	2000	2.5	1250	100	50	C75	100	350
838	10.0	3.25	50.0	4800	8.0	1250	175	100	B275	30	120
841	7.5	1.25	30.0	750	7.0	425	60	15	B25	6	50
842	7.5	1.50	3.0	1250	7.0	425	......	12	A3	......	......
845	10.0	3.25	5.3	3400	11.5	1250	......	75	A25	......	......
849	11.0	5.00	19.0	6000	33.0	3000	350	300	B1225	3	30
849A	11.0	7.70	19.0	7600	11.5	4000	500	500	B1900	3	30
849H	11.0	7.70	19.0	7600	11.5	3500	500	500	C1180	20	40
851	11.0	15.50	20.5	15000	47.0	2500	1000	750	C1700	3	15
852	10.0	3.25	12.0	1200	2.6	3000	150	100	C165	30	120

* Ratings given are typical of the class of service in which the tube is most commonly used.

The letter preceding each rating identifies the particular class of service as follows:

A —power output per tube as Class A power amplifier and modulator

AB—power output per pair of tubes as Class AB power amplifier and modulator

B —power output per tube as Class B power amplifier and modulator

BR—power output per pair of tubes as Class B Radio Frequency power amplifier

C —power output per tube as Class C power amplifier or oscillator

CP—power output per tube as Class C plate modulated power amplifier

AMPEREX RECTIFYING TUBES

MERCURY VAPOR RECTIFIERS

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Peak Inverse Volts	Approx. ◆Ave. Plate Amps.	Peak Plate Amps.
249B	2.5	7.50	7500	0.50	1.5
258B	2.5	7.50	7500	0.50	1.5
266B	5.0	42.00	22000	7.00	20.0
267B	5.0	6.75	10000	1.25	5.0
315A	5.0	10.00	15000	1.50	6.0
575A	5.0	10.00	15000	1.50	6.0
857B	5.0	40.00	22000	10.00	40.0
866	2.5	5.00	7500	0.25	1.0
866A	2.5	5.00	10000	0.25	1.0
869B	5.0	20.00	20000	2.50	10.0
872A	5.0	6.75	10000	1.25	5.0

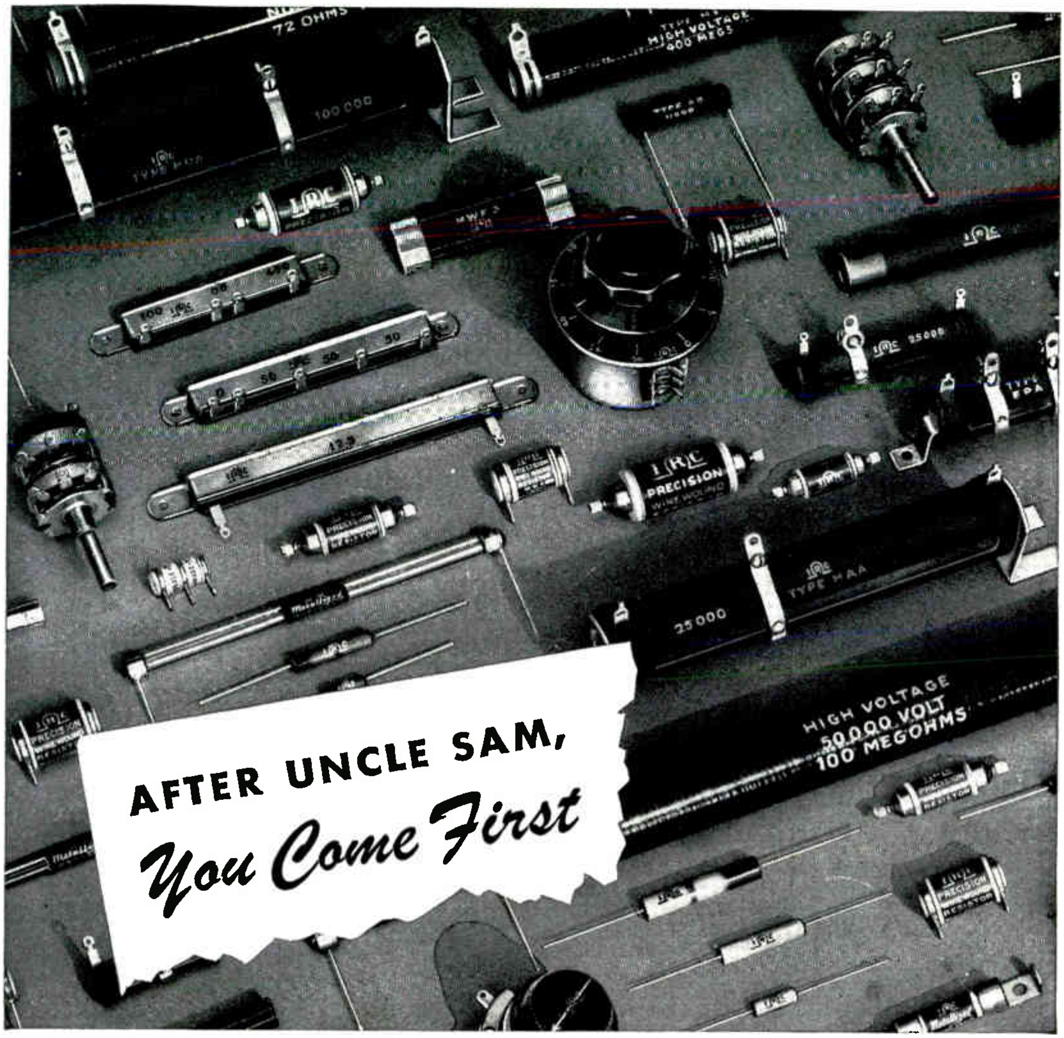
◆ Actual value will depend on wave-form resulting from load and filter circuit.

WATER-COOLED VACUUM RECTIFIERS

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Peak Inverse Volts	Ave. Emission Amps.	Peak Plate Amps.
222A	21.5	41.00	25000	7.0	5.5
237A	20.0	61.00	50000	10.0	8.0

RADIATION COOLED HIGH VACUUM RECTIFIERS

TYPE NO.	FILAMENT Volts	FILAMENT Amps.	Peak Inverse Volts	Ave. Emission Amps.	Peak Rate Amps.
8020	5	6	40000	0.100	0.750
221A	5	10	25000	0.300	1.5



We're extremely conscious of our obligation to the armed forces . . . and to you. During the war years we turned out millions upon millions of resistors used in equipment for ranging, control, detection, broadcasting and receiving. As a result we've added to our specialized knowledge a wealth of invaluable experience. This, plus operation of the world's largest resistor plants and a continuous study of postwar electronic products, assures availability to you of the resistances *preferred for performance* by both experts and amateurs.

IRC

INTERNATIONAL RESISTANCE CO.

401 N. Broad St., Philadelphia 8, Pa.

IRC makes more types of resistance units, in more shapes, for more applications than any other manufacturer in the world


STANCOR
Transformers
Battle-Tested!
Before a Stancor Transformer is shipped, it is "certified for service" by engineers whose tests simulate actual conditions in the field... Because "Stancor" is battle tested—right in our extensive laboratories—it has covered itself with glory on the battlefront. This is your assurance of the efficient performance of Stancor Products to which you may confidently look when the domestic market returns.
STANDARD TRANSFORMER CORPORATION
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SOME OF THE MANUFACTURERS WHOSE PRODUCTS WE DISTRIBUTE:

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American Phenolic Corp.
American Radio Hardware Co.
American Television & Radio Co.
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Harvey

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Service!

Technical "Know-how"

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Radio and Electronic Components and Equipment

More than just another distributor of radio and electronic components and equipment, HARVEY is a center of service and technical "know-how." This is our 18th year of buying, selling and merchandising the products of leading American manufacturers. We have built a fine reputation for delivering the goods on time, and for obtaining hard-to-find components. Our specially trained staff is prepared to answer questions on the effectiveness of substitutes, and to work out with you technical and priority problems. For present or postwar work, be sure to keep in touch with HARVEY!

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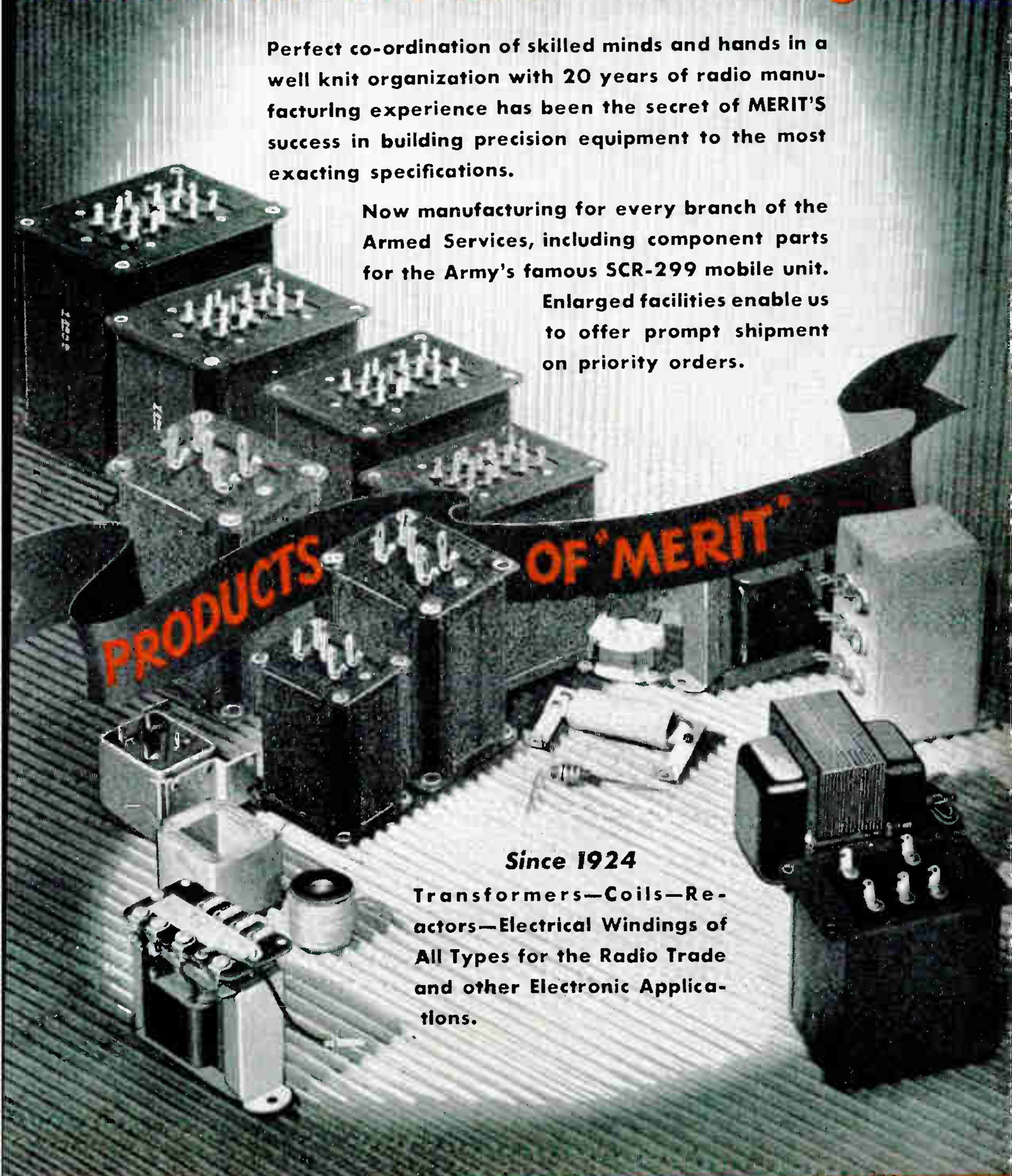
Remember! Our expediters assure you of the fastest possible deliveries. All merchandise sold with an ironclad guarantee of satisfaction.

PRECISION to the "Nth" Degree

Perfect co-ordination of skilled minds and hands in a well knit organization with 20 years of radio manufacturing experience has been the secret of MERIT'S success in building precision equipment to the most exacting specifications.

Now manufacturing for every branch of the Armed Services, including component parts for the Army's famous SCR-299 mobile unit. Enlarged facilities enable us to offer prompt shipment on priority orders.



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Transformers—Coils—Reactors—Electrical Windings of All Types for the Radio Trade and other Electronic Applications.

MERIT COIL & TRANSFORMER CORP.

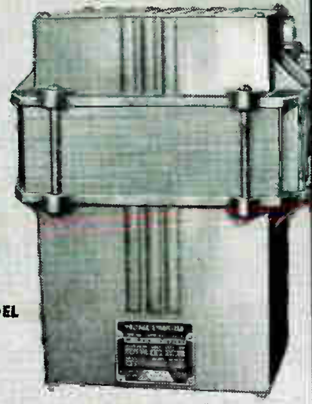
4427 NORTH CLARK ST. CHICAGO 40, U.S.A.

RAYTHEON VOLTAGE STABILIZERS

control fluctuating voltage to ±½%

- Stabilize at any load within their ratings.
- Hold constant varying A. C. input voltage to ± ½%.
- Quick action . . . fluctuating input voltage is stabilized within 2 cycles. Variations cannot be observed on an ordinary volt meter.
- Wide A.C. input voltage limits . . . 95 to 130 volts.
- Entirely automatic . . . no moving parts . . . requires no maintenance . . . connect it and forget it.

Bulletin DL48-537 is new — write for your copy.



ENDBELL MODEL

DIMENSIONS IN INCHES

Index Ref. No.	L 50~	L 60~	W	H
G-5	13 1/[illegible]	12 1/2	9 [illegible]	8 [illegible]
G-6	14 [illegible]/8	14	14 3/4	11 7/[illegible]
G-7	17 1/2	16 3/8	14 3/4	11 [illegible]

INPUT 95-130 V 60 CYCLES 1-PHASE OUTPUT 115 V PLUS OR MINUS ½%

WATTS	CASED Cat. No.	CASED Net Wt.	Index Ref. No.	UNCASED Cat. No.	UNCASED Net Wt.
30	VR-1	8 lbs.	G-1	VR-107	6 lbs.
30	VR-1-A†	8 "	G-1	VR-107-A†	6 "
60	VR-2	18 "	G-2	VR-207	16 "
120	VR-3	26 "	G-3	VR-307	22 "
250	VR-4	46 "	G-4	VR-407	36 "
500	VR-5	70 "	G-5		
1000	VR-6	140 "	G-6		
2000	VR-7	200 "	G-7		

†OUTPUT 6.0 or 7.5 VOLTS PLUS OR MINUS ½%.

INPUT 190-260 V 60 CYCLES 1-PHASE OUTPUT 220 230 V ± ½%

WATTS	CASED Cat. No.	CASED Net. Wt.	Index Ref. No.
2000	VR-7-A	200 lbs.	G-7

INPUT 95-130 V 50 CYCLES 1-PHASE OUTPUT 115 V PLUS OR MINUS ½%

WATTS	CASED Cat. No.	CASED Net Wt.	Index Ref. No.	UNCASED Cat. No.	UNCASED Net Wt.
25	VR-155	8 lbs.	G-1	VR-158	6 lbs.
60	VR-255	21 "	G-2	VR-258	19 "
120	VR-355	26 "	G-3	VR-358	23 "
250	VR-455	50 "	G-4	VR-458	42 "
500	VR-555	80 "	G-5		
1000	VR-655	150 "	G-6		
2000	VR-755	220 "	G-7		

INPUT 190-260 V 50 CYCLES 1-PHASE OUTPUT 220 230 V ± ½%

WATTS	CASED Cat. No.	CASED Net Wt.	Index Ref. No.	UNCASED Cat. No.	UNCASED Net Wt.
25	VR-521	8 lbs.	G-1	VR-510	6 lbs.
60	VR-522	21 "	G-2	VR-520	19 "
120	VR-523	26 "	G-3	VR-530	23 "
250	VR-524	50 "	G-4	VR-540	42 "
500	VR-525	80 "	G-5		
1000	VR-526	150 "	G-6		
2000	VR-527	220 "	G-7		

CASED MODEL

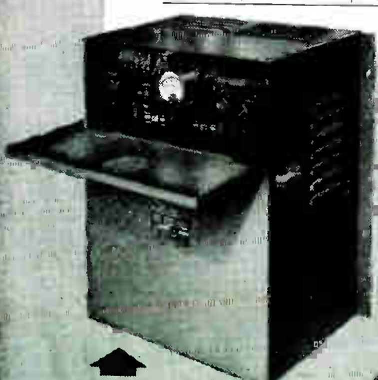
DIMENSIONS IN INCHES

Index Ref. No.	L	W	H
G-1	8 7/8	3 7/16	4 1/2
G-2	11 3/8	5 7/8	5 5/8
G-3	15	6	6 1/[illegible]
G-4	18 5/8	7	8 7/8

UNCASED MODEL

DIMENSIONS IN INCHES

Index Ref. No.	L	W	H Max.
G-1	7 9/16	3	3 9/16
G-2	7 7/8	5 11/16	5 [illegible]/16
G-3	12 7/32	5 11/16	5 3/8
G-4	15 9/16	6 3/4	7



Raytheon RectiChargeRs (battery chargers) . . . complete range of units from 1 to 12 amps . . . many models for 50 cycles input.

Raytheon RectiFilteRs (battery eliminators) . . . wide range of models available for varied operations . . . 50 or 60 cycles.

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MANUFACTURING COMPANY
190 WILLOW STREET, WALTHAM, MASS.

The coveted Army-Navy "E", for Excellence in the manufacture of war equipment and tubes, flies over all four Raytheon Plants where over 15,000 men and women are producing for VICTORY.

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Distributors of
RADIO and ELECTRONIC
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When victory comes — and the amateur can again apply his enthusiasm and energy in individual endeavor — advancement will be rapid in amateur radio. Then — as now — you can count on us for diversified stocks and prompt, friendly service in all your needs for components and equipment.
SIGNAL GENERATOR
Send for free Catalog
THE RADIO SHACK CORP.
"The RADIO SHACK
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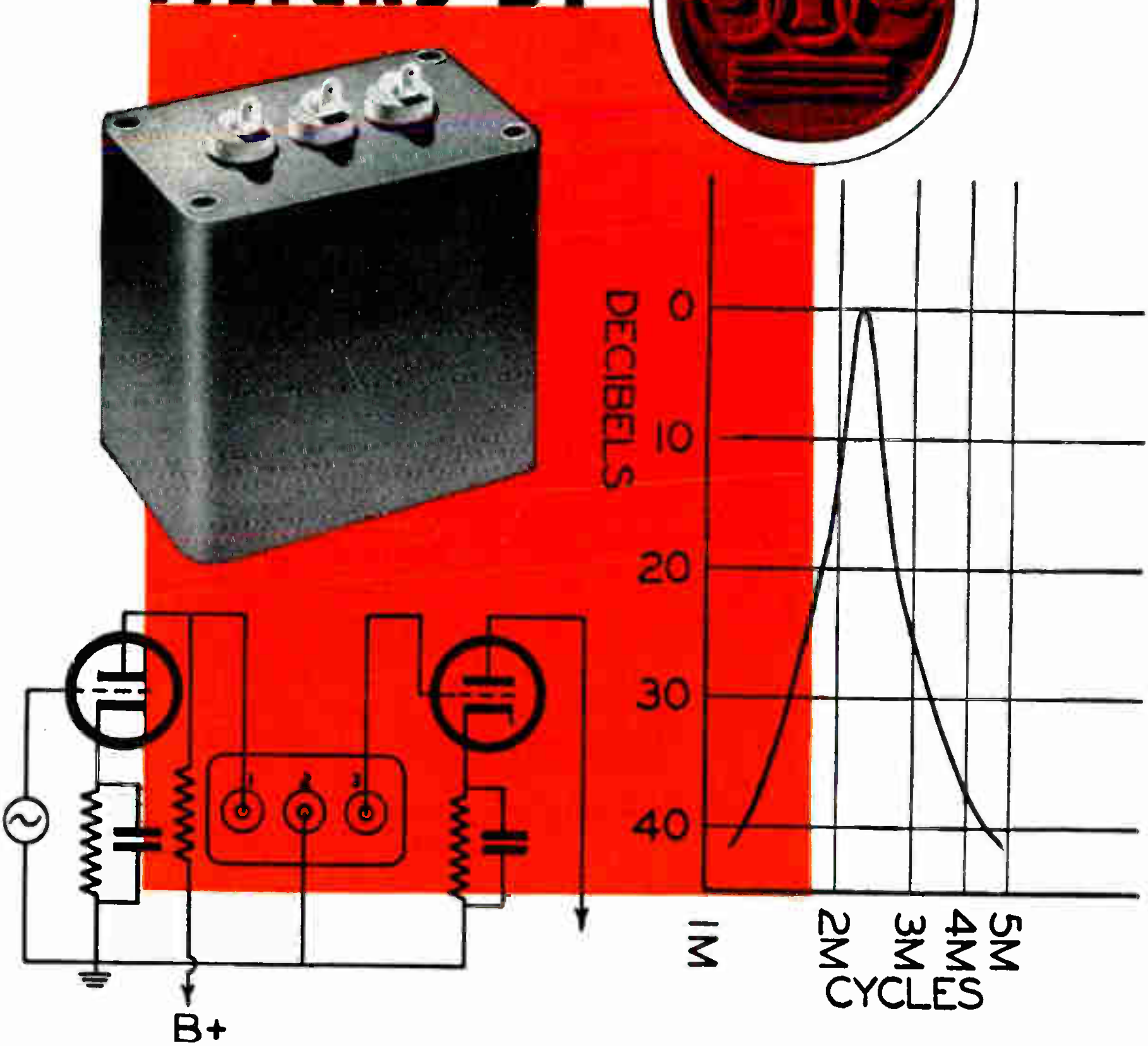
INTERSTAGE FILTERS BY UTC



Interstage filters lend themselves to effecting gain simultaneously with their frequency discrimination. The unit illustrated is a band pass unit which provides a 2:1 step-up ratio, with band pass attenuation of 40 DB per octave. This unit employs a dual alloy magnetic shield which reduces inductive pick-up to 150 Mv. per gauss. The dimensions in its hermetically sealed case are 1½ x 2½ x 2½. Filters of this type can be supplied for any band pass frequency from 200 to 10,000 cycles.

May we cooperate with you on design savings for your application . . . war or postwar?

United Transformer Co.

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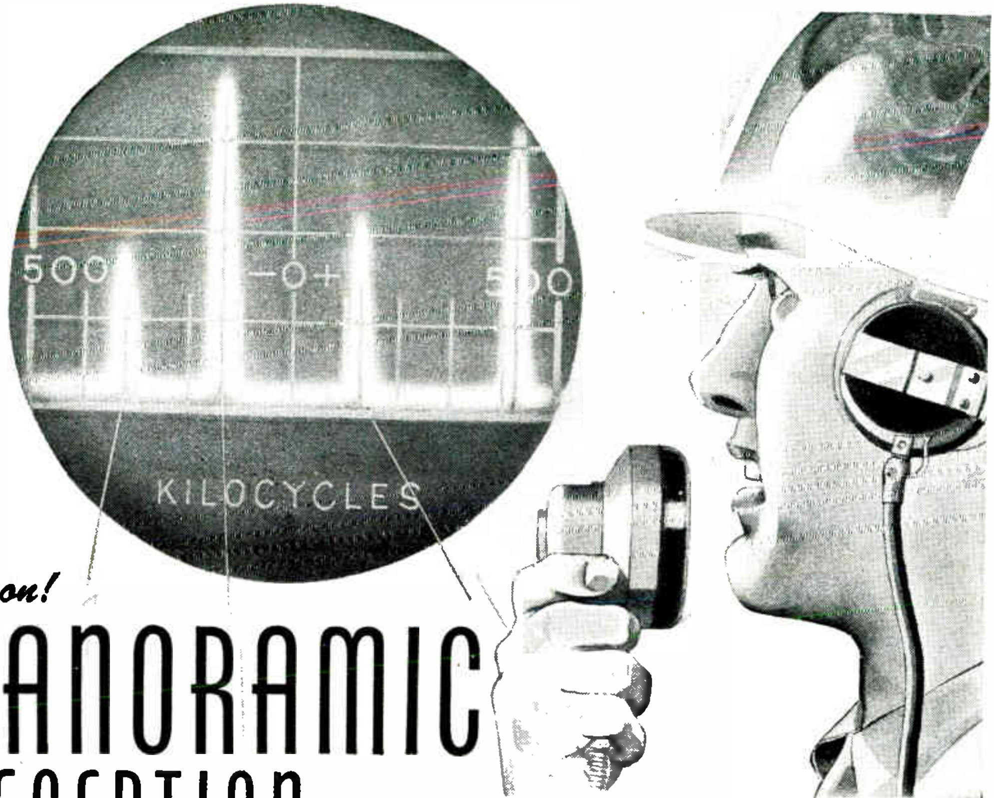
EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y., CABLES: "ARLAB"

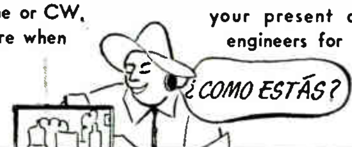


Soon!

PANORAMIC RECEPTION WILL BE USED BY GI JOE WITH HIS "HAM" RIG

When GI Joe takes off his helmet, he will still remember many of the things he is learning in the Army. As a radio operator, he uses ***Panoramic*** reception for effective monitoring and for catching tricks in enemy field communications. He recognizes its value for peacetime as well as for wartime. On the basis of military experience, he will want to make use of ***Panoramic*** reception for many more pleasant hours at his own rig. Because it ***SHOWS ALL SIGNALS ON A GIVEN BAND OF THE RADIO FREQUENCY SPECTRUM SIMULTANEOUSLY,*** GI Joe knows that ***Panoramic*** reception will tell him what stations are on the air, whether they are phone or CW, and what their signal strengths are when they reach him. Most important, he can be sure that he will miss very few calls in response to his CQ's.

Currently, ***Panoramic*** reception also is doing good work in laboratory development and industrial applications. Its ability to measure, interpret and compare variations in inductance, capacitance and resistance has created possibilities that are being utilized by far-sighted manufacturers. If ***Panoramic*** technique can be adapted to your present or future needs, ask our engineers for more detailed information.



PANORAMIC RADIO CORPORATION
242-250 WEST 55TH ST., New York 19, N.Y.

In Service on All Fronts

DUMMY ANTENNA RESISTORS

To check R. F. power, determine transmission line losses, check line to antenna impedance match. Helps tune up to peak efficiency. Non-inductive, non-capacitive, constant in resistance. 100 and 250 watt sizes in various resistances.

BROWN DEVIL RESISTORS

Small, extra sturdy, wirewound vitreous enameled resistors for voltage dropping, bias units, bleeders, etc. Proved right in vital installations the world over. 10 and 20 watt sizes in resistances up to 100,000 ohms.

PARASITIC SUPPRESSOR

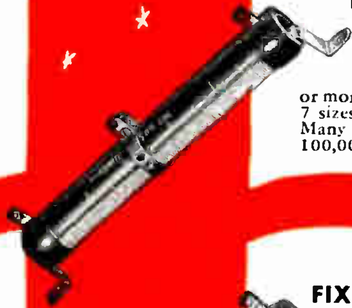
Small, light, compact non-inductive resistor and choke, designed to prevent u.h.f. parasitic oscillations which occur in the plate and grid leads of push-pull and parallel tube circuits. Only 1¾" long overall and ⅝" in diameter.

CENTER-TAPPED RESISTORS

For use across tube filaments to provide an electrical center for the grid and plate returns. Center tap accurate to plus or minus 1%. Wirewatt (1 watt) and Brown Devil (10 watt) units, in resistances from 10 to 200 ohms.

R. F. PLATE CHOKES

Single-layer wound on low power factor steatite cores, with moisture-proof coating. Built to carry 1000 M.A. 5 stock sizes from 2½ meters to 160 meters. 2½ and 5 meter chokes mount by wire leads. Larger sizes mount on brackets.



ADJUSTABLE DIVIDOHMS

You can quickly adjust these handy Dividohms to the exact resistance you want, or put on one or more taps wherever needed. 7 sizes from 10 to 200 watts. Many resistance values up to 100,000 ohms.

R. F. POWER LINE CHOKES

Keep R.F. currents from going out over the power line and causing interference with radio receivers. Also used at receivers to stop incoming R.F. interference. 3 stock sizes, rated at 5, 10 and 20 amperes.

FIXED RESISTORS

Resistance wire is wound over a porcelain core, permanently locked in place, insulated and protected by Ohmite vitreous enamel. Available in 25, 50, 100, 160 and 200 watt stock sizes, in resistances from 1 to 250,000 ohms.

Be Right with OHMITE

RHEOSTATS ★ RESISTORS ★ CHOKES ★ TAP SWITCHES

Ohmite Vitreous Enamel is unexcelled as a protective and bonding covering for resistors and rheostats.

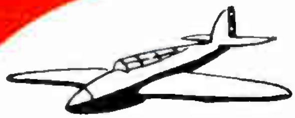
OHMITE Rheostats ★ Resistors ★ Chokes ★ Switches

CLOSE-CONTROL RHEOSTATS

Insure permanently smooth, close control of electronic devices, communications and electrical equipment. Widely used in industry and in planes, tanks, ships. All ceramic, vitreous enameled. 25, 50, 75, 100, 150, 225, 300, 500, 750 and 1000 watt sizes. Approved Army & Navy types.

HIGH-VOLTAGE SWITCH

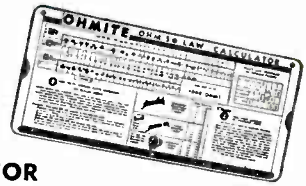
For general use where high voltage insulation is required. Suitable for circuits up to 1 K.W. rating. Used for band changing, meter switching, tapped transformer circuits, etc. Ceramic construction.



Many of you now engaged in vital war industries or in active service have long been familiar with the rugged dependability of Ohmite Products. Their wide use in planes, tanks and ships, in walkie-talkies and field units, in communications, electronic and electrical equipment, gives you added assurance in dealing with today's resistance-control problems. This is well worth remembering when you build original equipment or make vital replacements — *today and tomorrow.*

Besides the units shown here, there are Ohmite Non-Inductive Vitreous Enameled Resistors, Riteohm Precision Resistors, Hermetically-Glass-Sealed Resistors, Direction-Indicator Rheostats, Attenuators, and many others.

HANDY OHMITE OHM'S LAW CALCULATOR



Very useful in training schools, in laboratories and in industry. Figures ohms, watts, volts, amperes — quickly, easily. Solves any Ohm's Law problem with one setting of the slide. All values are direct reading. No slide rule knowledge is necessary. Scales on two sides cover the range of currents, resistances, wattages and voltages commonly used in radio and electronic applications. Size only 4 1/8″ x 9″. Send only 10¢ in coin to cover handling cost.

AUTHORIZED DISTRIBUTORS EVERYWHERE

HIGH-CURRENT TAP SWITCHES

Compact, all ceramic, multi-point rotary selectors for A.C. use. Silver to silver contacts. Rated at 10, 15, 25, 50 and 100 amperes with any number of taps up to 11, 12, 12, 12, and 8 respectively. Single or tandem assemblies.

LC-2 LINK CONTROL

Simplified, compact, convenient panel regulation of the transfer of R.F. energy thru the link or low impedance line used in many transmitters. Eliminates swinging coupling coils. All ceramic vitreous enameled construction.

SEND FOR FREE CATALOG 18 — Gives helpful information and data on Ohmite stock units for essential applications — lists hundreds of stock values. Very handy for quick reference.

OHMITE MANUFACTURING COMPANY

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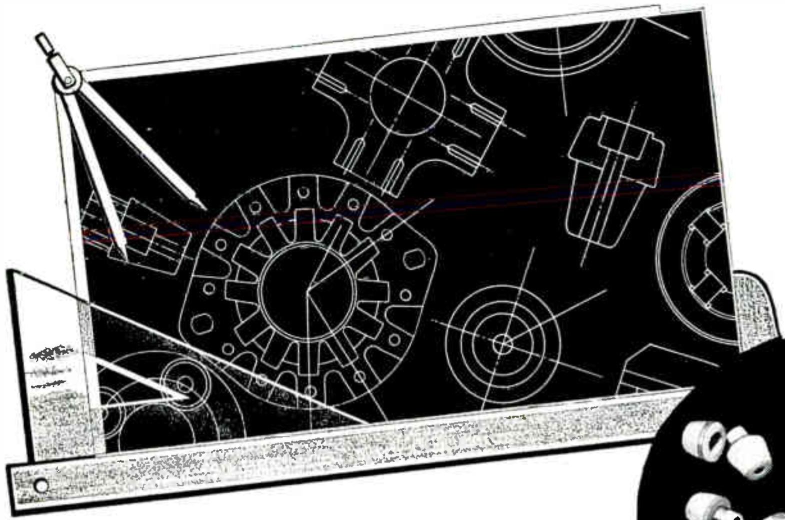
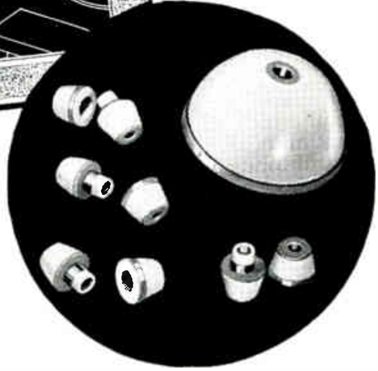




NEW INSULATOR DESIGN POSSIBILITIES FOR RADIO

General Ceramics' successful surfacing of steatite with thin films of silver, fired at a high temperature and then built up with an electroplated metal (silver, copper or tin), opens up new insulator design possibilities for very high frequency equipment, as well as for certain applications in the lower radio frequency field.

The metallic film can be applied to the surface of insulators to eliminate corona effect. The use of this combination permits improvement in the design of airplane strain and lead-through insulators.

The addition of a thin metallic surface film also permits soldering of metal parts directly to the steatite insulators. Water-tight seals may be made in this manner where temperature ranges encountered in service are limited.

Your inquiry regarding *Silver Surfaced Steatite* is invited.

ARMY E NAVY

General Ceramics
AND STEATITE CORP.
KEASBEY
NEW JERSEY

GAMMATRON TUBES

24G
PLATE DISS.
25 W.

257B
PLATE DISS.
75 W.

454
PLATE DISS.
250 W.

This complete line, covering a power range of 50 to 5,000 watts, embodies 18 years of pioneering and experience in the design and manufacture of tantalum tubes.

Special plate, grid, and filament design, and new metal-to-glass seals, give Gammatrons remarkable VHF performance. Other features: ability to withstand high plate voltages, complete protection against tube failure due to overloading, and long, efficient operating life.

The Gammatron engineers responsible for these developments will be glad to help you with your special problems.

HEINTZ AND KAUFMAN LTD.
SOUTH SAN FRANCISCO — CALIFORNIA U.S.A.

TYPE NO.	24	24G	54	254	257B	304L	304H	354C	354E	454L	454H	654	854L	854H	1054L	1554	2054A	3054
MAX. POWER OUTPUT: Class 'C' R.F.	90	90	250	500	230	1220	1220	615	615	900	900	1400	1800	1820	3000	3600	2000	5300
PLATE DISSIPATION: Watts	25	25	50	100	75	300	300	150	150	250	250	300	450	450	750	1000	1200	1500
AVERAGE AMPLIFICATION FACTOR	25	25	27	25		10	19	14	35	14	30	22	14	30	13.5	14.5	10	20
MAX. RATINGS: Plate Volts Plate M.A. Grid M.A.	2000 75 25	2000 75 25	3000 150 30	4000 225 40	4000 150 25	3000 1000 150	3000 1000 150	4000 300 60	4000 300 70	5000 375 60	5000 375 85	4000 600 100	6000 600 80	6000 600 110	6000 1000 125	5000 1000 250	3000 800 200	5000 2000 500
MAX. FREQUENCY, Mc.: Power Amplifier	200	300	200	175	150	175	175	50	50	150	150	50	125	125	100	30	20	30
INTERELECTRODE CAP: C g—p u.u.f. C g—f u.u.f. C p—f u.u.f.	1.7 2.5 0.4	1.6 1.8 0.2	1.8 2.1 0.5	3.6 3.3 1.0	0.08 10.5 In 4.6 Out	9 12 0.8	10.5 14 1.0	3.8 4.5 1.1	3.8 4.5 1.1	3.4 4.6 1.4	3.4 4.6 1.4	5.5 6.2 1.5	5 6 0.5	4 8 0.5	5 8 0.8	11 15.5 1.2	18 15 7	15 25 2.5
FILAMENT: Volts Amperes	6.3 3	6.3 3	5.0 5	5.0 7.5	5.0 7.5	5.10 26.13	5.10 26.13	5 10	5 10	5 11	5 11	7.5 15	7.5 12	7.5 12	7.5 21	11 17.5	10 22	14 45
PHYSICAL: Length, Inches Diameter, Inches Weight, Oz. Base	4 1/4 1 3/8 1 1/2 Small UX	4 1/4 1 3/8 1 1/2 Small UX	5 7/16 2 2 1/2 Std. UX	7 2 5/8 6 1/2 Std. 50 Watt	5 15/16 2 5/8 6 Giant 7 Pin	7 1/4 3 1/2 9 Johnson #213	7 3/4 3 1/2 9 Johnson #213	9 3 3/8 6 1/2 Std. 50 Watt	9 3 3/8 6 1/2 Std. 50 Watt	10 3 3/4 7 Std. 50 Watt	10 3 3/4 7 Std. 50 Watt	10 1/8 3 3/4 14 Std. 50 Watt	12 1/2 5 14 Std. 50 Watt	12 1/2 5 14 Std. 50 Watt	16 1/2 7 42 Johnson #214	18 6 56 HK 255	21 1/4 6 66 W.E. Co.	30 3/4 9 200 HK 255

*Beam Pentode.

WRITE FOR FULL DATA ON ALL **GAMMATRONS**

THE CHOICE OF GOOD OPERATORS EVERYWHERE

Because It's the World's Finest Semi-Automatic Radio Telegraph Key

GENUINE EASY-WORKING

VIBROPLEX

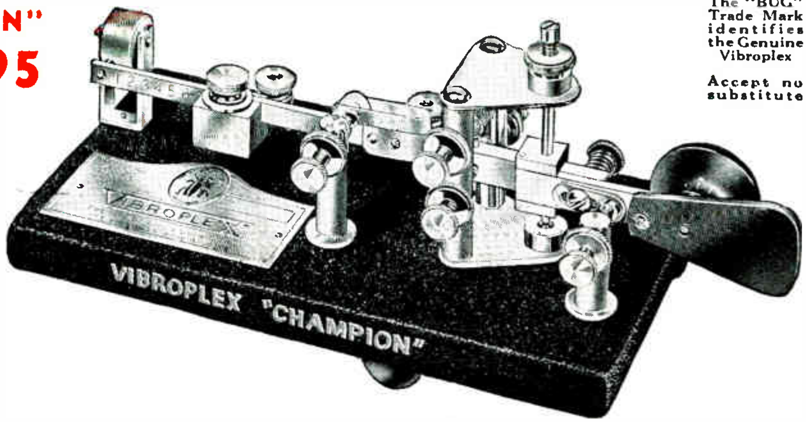
Reg. Trade Marks: Vibroplex, Lightning Bug, Bug

The "BUG" Trade Mark identifies the Genuine Vibroplex

Accept no substitute

THE "CHAMPION"

ONLY $9.95



A smart, efficient, full-size semi-automatic radio telegraph key. Easy to learn. Easy to operate. Black crackle base. Chrome-plated parts. 3/16th contacts. Without circuit closer, cord and wedge. *Radio use only.*

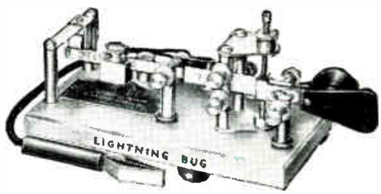
FAMOUS "ORIGINAL" MODEL

Here's a key that has won an international reputation for its uniformly smooth, accurate sending performance by many of the world's best operators. Effortless action . . . easy-working and easy to operate. Standard black crackle base. Bright machined parts. Precision construction including **DIE CUT** contact and main spring. 3/16th contacts. Complete with circuit closer, cord and wedge.................................**$15.95**

DeLuxe finish with Patent Jewel movement....**$19.50**

YES, the choice of good operators everywhere. These operators—many thousands of them have proved by actual telegraph experience that daily use of the Vibroplex semi-automatic key not only eliminates arm fatigue and prevents "glass arm" but also enables the operator to develop a higher degree of sending skill than is possible by any other means. Easy to learn. Easy to operate. Simply press lever—*Vibroplex does the rest.* Every Vibroplex key is built to exacting Vibroplex standards . . . precision machined . . . scientifically designed . . . improved in every possible detail.

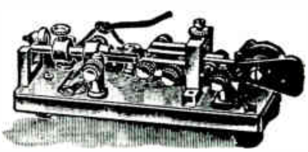
EASY-WORKING "LIGHTNING BUG" MODEL



A popular, low-priced key that does a remarkable sending job . . . responds to the lightest touch . . . easy-working . . . uniformly accurate and dependable. Many exclusive features contributing to easier and better sending. Standard black crackle base. Bright machined parts. Precision construction including **DIE CUT** contact and main spring. 3/16th contacts. Complete with circuit closer, cord and wedge...**$13.95**

DeLuxe finish with Patent Jewel movement......................**$17.50**

HANDY "BLUE RACER" MODEL



Patterned after the "Original" Vibroplex and capable of the same high-class sending performance that has made that key world famous. Smaller, handier. Suited to all classes of telegraph work where good sending is the prime essential. Standard black crackle base. Bright machined parts. Precision construction including **DIE CUT** contact and main spring. 3/16th contacts. Complete with circuit closer, cord and wedge..**$15.95**

DeLuxe finish with Patent Jewel movement................**$19.50**

DELUXE MODEL WITH PATENT JEWEL MOVEMENT

The aristocrat of Vibroplex keys made so by improvements and refinements, plus the sensational and exclusive **Patent Jewel** movement providing easier manipulation and higher quality signals than has ever before been attained. New, non-glare **Battle Ship Gray** Base. Bright machined parts. Colorful red switch knob, finger and thumb piece. Cord and wedge. Available in three designs: *Original, Blue Racer* and *Lightning Bug* models. Specify model when ordering. Remit by money order, or registered mail. Write for **FREE** illustrated catalog of Standard and DeLuxe Vibroplex keys.

The Genuine Vibroplex

CARRYING CASE

Keeps out dust, dirt and moisture. Prolongs life of key. Heavily reinforced. Flexible leather handle. Patent lock and key.....**$3.50**

THE VIBROPLEX CO., Inc.

833 Broadway, New York 3, N. Y.

J. E. Albright, President

Laboratory Standards

MODEL 84 U. H. F. STANDARD SIGNAL GENERATOR

Standard Signal Generators
U. H. F. Noise and Fieldstrength Meters
Square Wave Generators
Vacuum Tube Voltmeters
Pulse Generators
Moisture Meters
Megohm Meters

MODEL 62
VACUUM TUBE
VOLTMETER

MEASUREMENTS CORPORATION

BOONTON NEW JERSEY

PROVED TRUE BY TIME

TIME alone can prove how good capacitors are. And in the *test of time*, Tobe Capacitors have won a remarkable reputation for *long and satisfactory life*. It's the result of constant research, rigid inspections, specialized experience and conservative ratings. Whatever your condenser problems, let Tobe's experienced engineers help you to solve them!

LONG LIFE ASSURED

TOBE DEUTSCHMANN CORP.
TOBE
CANTON · MASSACHUSETTS

TRS 605, 5 mfd. 600 volts

SIZE— Overall height 5"

CONTAINER— 1-3/16"x 2-1/2"x 4"

Dimensions of other TRS models on request.

SPECIFICATIONS FOR TRS CAPACITORS

CAPACITY—.1 to 20 mfd.

WORKING VOLTAGE— 600 volts DC to 6,000 volts DC.

SHUNT RESISTANCE— 6,000 megohms per mfd.

RESISTANCE Terminal to Case— 10,000 megohms minimum:

POWER FACTOR—.002 to .005

VOLTAGE TEST Terminal to Case— 2,500 VDC for 600 volt condensers.

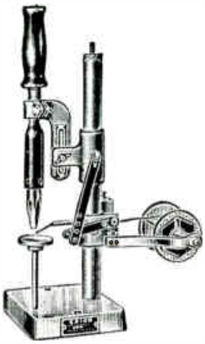
Capacitor unit tested at 2 times rated voltage.

Universal (wrap around) L or foot type and screw Spade-lug mounting brackets can be supplied.

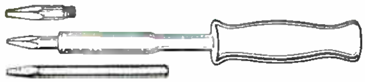
A small part in Victory today . . . **A BIG PART IN INDUSTRY TOMORROW**

ESICO

REG. U. S. PAT. OFF.



SOLDERING IRONS



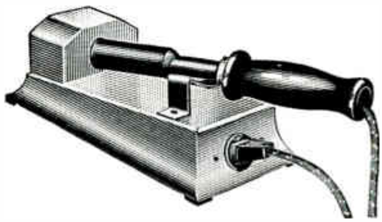
are widely used in industrial plants throughout the country. They are designed to withstand the strain of continuous service required of factory tools.

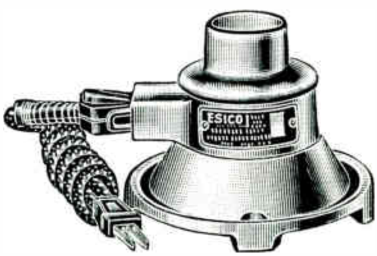
SPOT SOLDERING MACHINE

designed for treadle operation for advancement of iron and solder leaving operator's hands free for handling of product.

SOLDERING IRON TEMPERATURE CONTROLS

prevent overheating of soldering irons between soldering operations. Irons do not deteriorate when being used. The idle period causes oxidation and shortens life.





SOLDER POTS

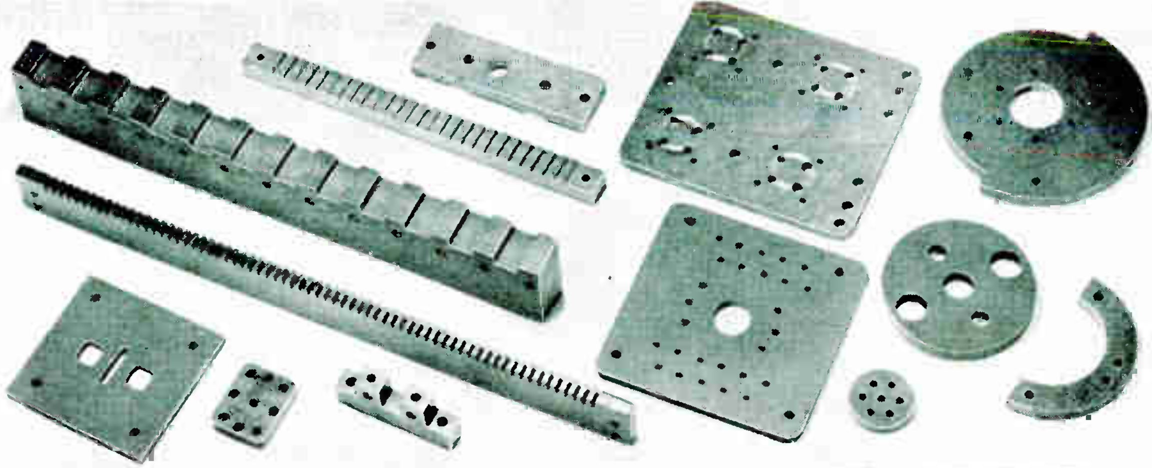
ruggedly constructed pots of various sizes designed for continuous operation and so constructed that they are easily and quickly serviced, should elements have to be replaced.

Write for Catalog

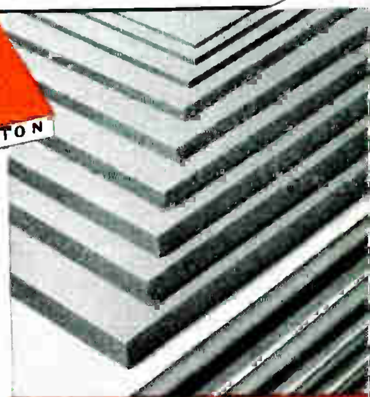
ELECTRIC SOLDERING IRON CO., INC.

2545 West Elm Street

DEEP RIVER, CONNECTICUT



Indispensable for h.f. and u.h.f. Insulation



MYKROY
PERFECTED MICA CERAMIC INSULATION

MYKROY is a perfected glass-bonded mica plastic-ceramic insulation regarded by engineers as the nearest approach to the text-book ideal of perfect insulation for all high frequency and ultra high frequency applications.

In MYKROY the combined special properties of several different types of dielectrics impart to it vastly superior characteristics:

MECHANICAL PROPERTIES

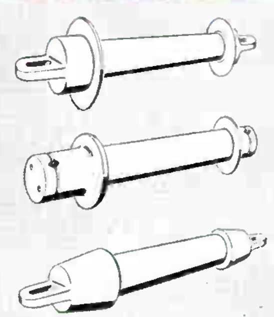
Modulus of Rupture	18000-21000 psi
Hardness	
Mohs Scale 3-4 BHN	BHN 500 Kg Load, 63-74
Impact Strength	ASTM Charpy .34-.41 ft. lbs.
Compression Strength	42000 psi
Specific Gravity	2.75-3.8
Thermal Expansion	.000006 per Degree Fahr.
Appearance	Brownish Grey to Light Tan

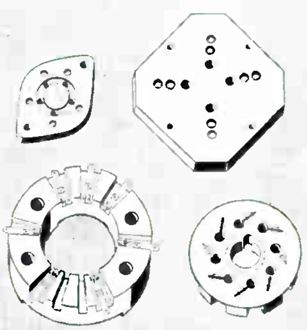
ELECTRICAL PROPERTIES

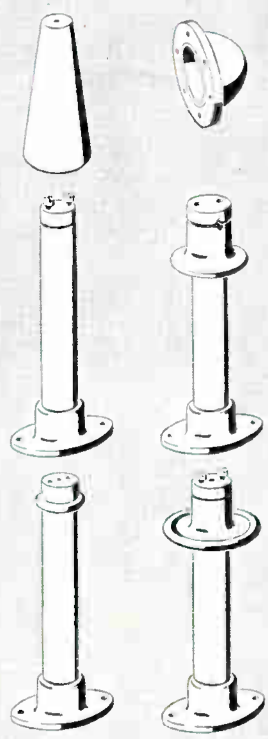
Dielectric Constant	6.5-7
Dielectric Strength (1/8")	630 Volts per Mil
Power Factor	.001-.002 (meets AWS L-4)

The illustrated radio components made of MYKROY are but a few of the many now available. MYKROY is also supplied in rods and sheets (up to 19¼ x 29¾) . . . machined or molded to your specifications.

Ask your dealer or write for
The MYKROY ENGINEERS MANUAL







ELECTRONIC MECHANICS INC.

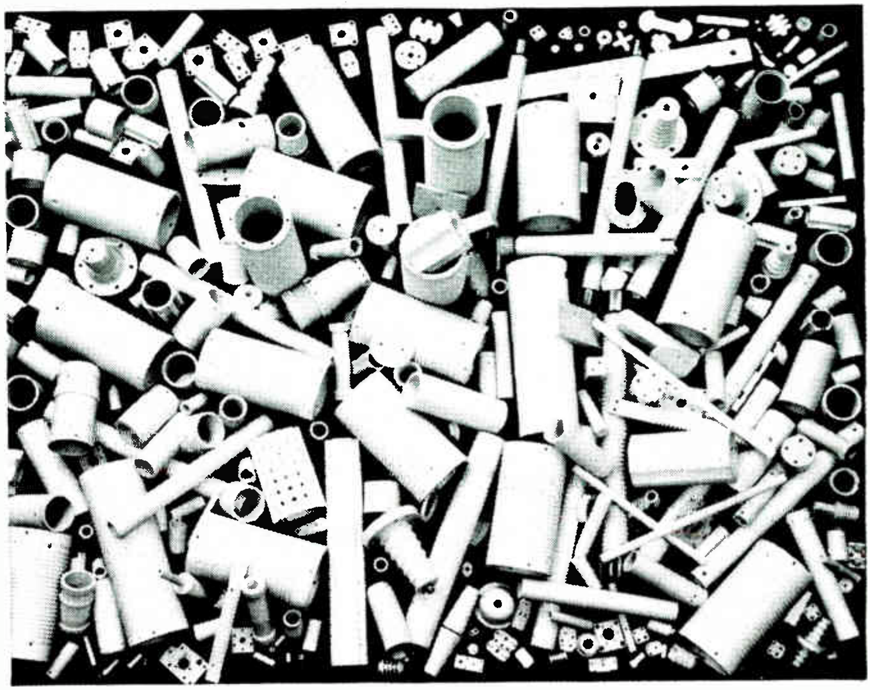
70 CLIFTON BLVD., CLIFTON, N. J.

Chicago (47), 1917 No. Springfield Ave.
Tel. Albany 4310
Export Office 89 Broad St., N. Y. C.

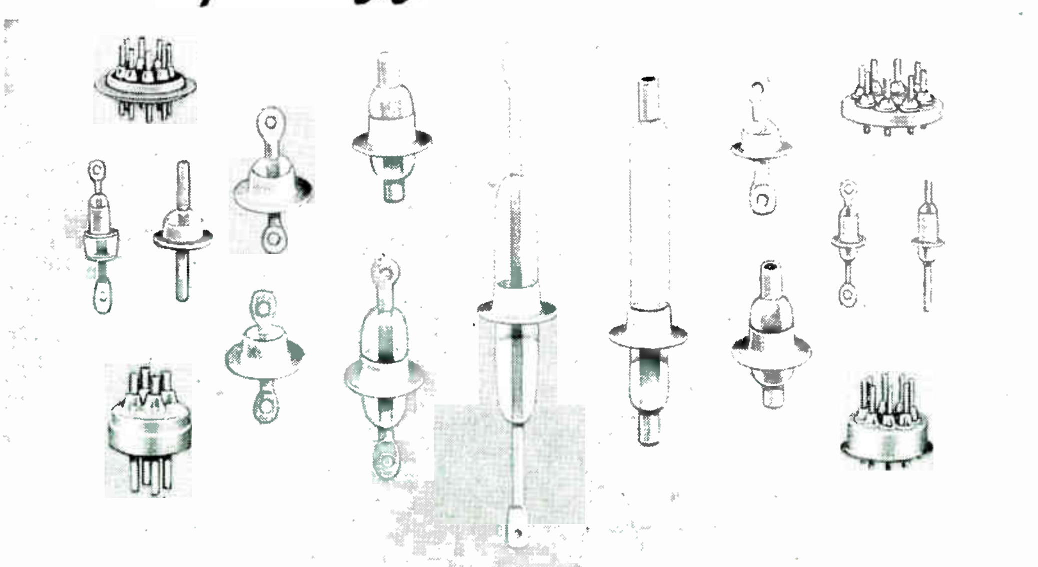
Stupakoff Steatite Insulators

PRECISION MADE . . UNLIMITED STYLES & SIZES

"FOR GREAT ACHIEVEMENT"



Stupakoff Hermetic Seals

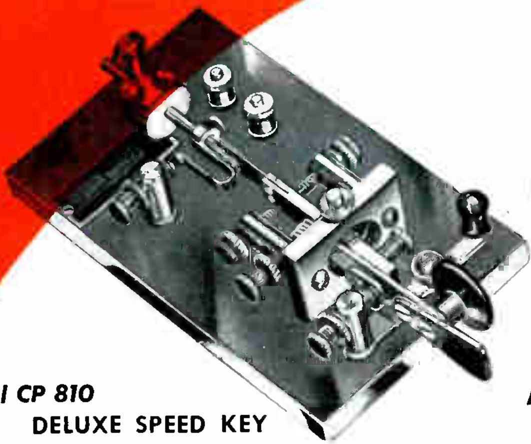


The metal, KOVAR*, a cobalt, nickel, iron alloy, has made possible a hermetically sealed terminal without the use of cements or gaskets. Illustrated are representative styles and sizes of single and multiple electrode type Kovar-glass terminals, manufactured by Stupakoff for the electronic industry

* TRADE MARK 337962 REGISTERED IN U.S. PATENT OFFICE

STUPAKOFF CERAMIC AND MANUFACTURING CO., LATROBE, PA.

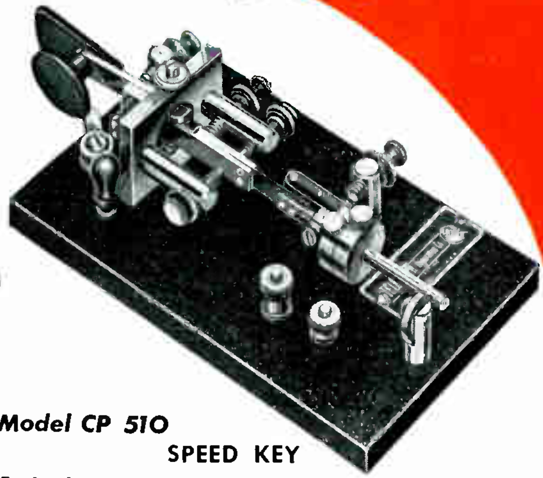
DESIGNED by the World's Champion Radio Telegrapher



Model CP 810

DELUXE SPEED KEY

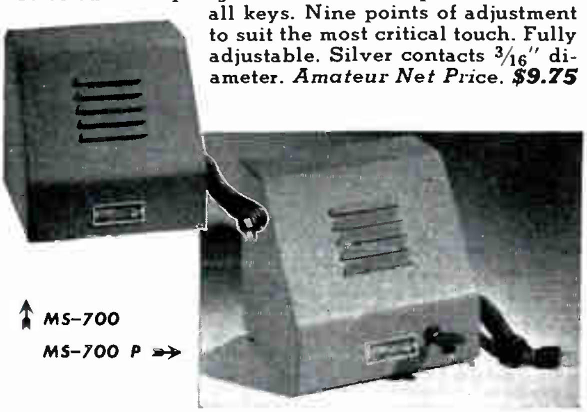
Finest commercial or amateur bug available. Chrome finished base and superstructure. Springs made of selected blue spring steel for uniform performance in all keys. Nine points of adjustment to suit the most critical touch. Fully adjustable. Silver contacts 3/16″ diameter. *Amateur Net Price.* ***$9.75***



Model CP 510

SPEED KEY

A high grade professional bug designed by World's Champion telegrapher. Battleship gray wrinkle finish and polished chrome. A masterpiece of precision craftsmanship. Silver contact points 3/16″ diameter. Springs, contacts and vibrating arm all fully adjustable.
Amateur Net Price ***$6.75***



↑ MS–700

MS–700 P →

MASTER OSCILLATONES

Designed for individual or class code-learning. Beautiful gray crackle finish steel cabinet. Continuously variable volume control, tone adjustable 500–1500 cycles. Terminals for head phones, and up to 300 phones may be used. Speaker may be on or off as desired. Operates on 110v. AC or DC. A 117N 7 tube is used.
MS-700. With Tube
Amateur Net Price. ***$11.85***
MS-700 P (*with key as shown*). With Tube
Amateur Net Price ***$15.90***

Model 300 **DELUXE HAND KEY**

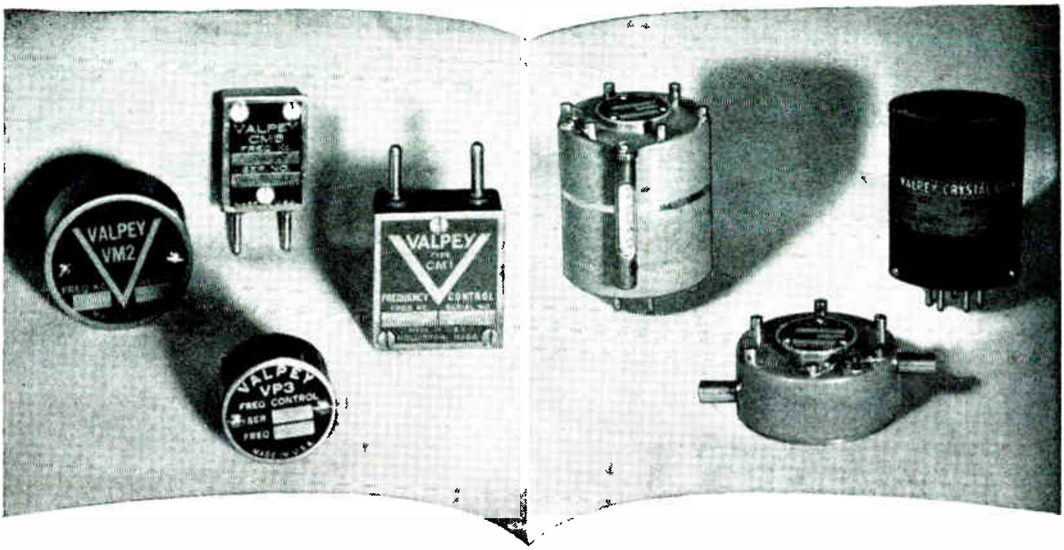
Beautifully finished in polished chrome and nickel. Sturdy, well-balanced construction for smooth, effortless keying. 3/16″ diameter contacts, adjustable for tension, spacing, and bearing position. Circuit closing switch is incorporated.
Amateur Net Price
$3.45

Model 200 **HAND KEY**

Right up to professional standard in performance and appearance. 3/16″ coin silver contacts. Accurately machined bearing screws. Black crackle finish metal base, with circuit-closing switch. *Amateur Net Price.*
$2.25

TELEGRAPH APPARATUS CO.

325 WEST HURON STREET • CHICAGO 10, ILL.



RELIABLE PERFORMANCE

Is Only One of the Many Reasons Why...

Valpey *Crystals*

Meet the Exact Specifications of Our Armed Forces

With peace will come again those products of industry such as radio, television, airplanes, refrigerators, washing machines, etc., which never really were luxuries so much as necessities . . . and to those who are thinking in terms of post-war, we are prepared to help build precision parts and accessories for such products required by the electronics industry again, as well as help design and produce many complex new ones that are destined to emerge from this war — plan ahead with — VALPEY.

QUARTZ CRYSTALS *for All Frequency Control Applications*

QUARTZ CRYSTAL MOUNTINGS

TEMPERATURE CONTROL OVENS

QUARTZ PRISMS, LENSES *and Special Optical Pieces*

WRITE NOW FOR "CRYSTONICS" INFORMATION

VALPEY CRYSTAL CORP.

"Craftsmanship in Crystals Since 1931"

HOLLISTON, MASS.

Ayer-Code

STREAMLINED METHOD

FOR RADIO-TELEGRAPH AMATEUR AND COMMERCIAL LICENSES.

LEARN YOUR CODE

Quickly **THE RIGHT WAY** *Easily*

by SOUND

TEACHES by rhythmic sound through skilled operation technique . . . No "sender" needed . . . Adjustable from 8 to 100 words per minute; high or low pitch, always crystal-clear . . . Developed principally for, and extensively used by, Army and Navy training schools . . . Unexcelled for economical and efficient instruction.

HOW IT WORKS — Photo-Electric Cell converts each message from paper tape to sound through electronic control of oscillator. Just place roll of tape on the keyer, plug into either AC or DC circuit, turn on the control switches. You can set the code speed just where you want it.

TAPES— Beginner, Ship to Shore Marine, Point to Point, Code Groups, Numbers, Mixed Code, "How To Become A Radio Operator", "A. R. R. L. License Manual", Plain English Press, Cypher, . . . Over 300 varieties from the simplest to the most intricate high-speed code.

For government and commercial uses: prices quoted immediately upon request.

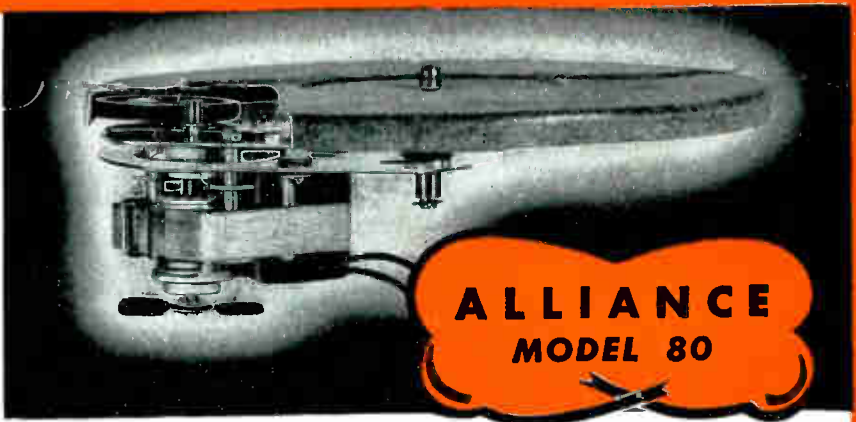
For individual use: Low rental prices. Write today.

Model #G-813-742. Keyer at front center, tape puller upper right. Unit supplied complete with ear phones and tape.

AYERS AUTOMATIC CODE MACHINES CO., INC.

711 BOYLSTON STREET
BOSTON 15, MASSACHUSETTS, U. S. A.

ALLIANCE "Even-Speed" PHONO-MOTORS



Simple and quiet in opera-
on — no gears — smooth
sitive friction rim drive.
od regulation characteristics
uniformity of table speed.
nply proportioned bearings
large oil reserves. Motor and
er plate shock mounted to
unting plate for low vibra-
n transfer to turntable and
tor board. Forced ventila-
n for cool operation. Slip
e fan precludes possibility
injury. Mounting plate main-
ns correct turntable height
gardless of mounting board
ckness. Maximum depth be-
v base mounting plate, 2 1/16
hes.

SE LIST PRICES: 110 V. 60 C.— 78 R.P.M. with 8" Table — $5.00; 9" Table — $5.28; 10" Table — $5.55.
TRAS ON BASE LIST PRICES: 220 V. 60 C.— $1.00; 110 V. 50 C.— $1.00; 220 V. 50 C.— $1.00; 110 V. 40 C.— $1.00; CESA Approved Type — 75c; Retractible Turntable Pin — 28c.

EVEN-SPEED" MODEL K PHONO-MOTOR

he 25-cycle Companion to the Model 80 Friction Drive Phonomotor. This phonomotor is designed specifically 25 cycle operation, having a motor of entirely new design, but employing the same efficient, positive, friction n drive as the popular Model 80. Interchange-ability in mounting is therefore obtained without sacrifice in per- mance. Maximum depth below base mounting plate, 2 1/4".

SE LIST PRICES: 110 Volt — 25 C.— 78 R.P.M. with 8" Table — $6.50; 9" Table — $6.75.
TRAS ON BASE LIST PRICES: Retractible Turntable Pin — 28c; CESA Approved Type — 75c.

LLIANCE PRECISION ASSURES EFFICIENCY AND DEPENDABILITY

ALLIANCE Fractional H. P. Motors for Amateur Needs

ondensed power for years of efficient service are built into these Alliance fractional . P. Motors "K" and "MS". They are ideal for operating fans, movie projectors, ght home appliances, toys, switches, motion displays, control systems and many her applications. Alliance Precision Built!

ODEL K SPECIFICATIONS: 117 V.— 60 Cycle — Cont. Open Rating — 50° C. Rise — No Fan.

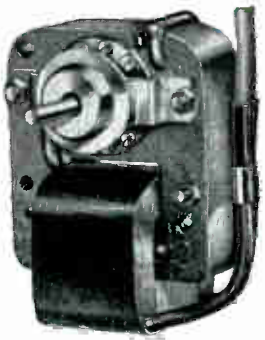
ack Thickness, Inches	.800	1.000	1.200
aring Torque — Oz. In. Cold	1.5	1.9	2.5
le Amps.— Hot	.52	.54	.56
le Watts — Hot	22.0	23.0	25.5
le R.P.M.— Hot	3450	3450	3450
ll Load Amps	.57	.60	.65

Stack Thickness, Inches	.800	1.000	1.200
Full Load Watts	28	32	36
Full Load Horsepower	.0068	.0085	.0100
Full Load Torque Oz. In.	2.4	2.9	3.5
Full Load — R.P.M.	2900	2900	2900

Rotor Shaft — Centerless Ground .171" Diameter
Bearings — Graphite Bronze Oilless Type, Self-Aligning, Amply Proportioned.

Motor — 110 Volt 60 C. Standard Version

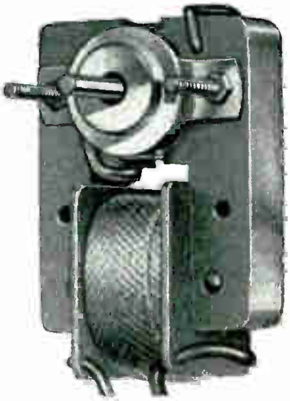
PRICES:
Spec. K828 — .8" Stack — List $4.00
Spec. K1002 — 1.0" Stack — List 4.75
Spec. K1201 — 1.2" Stack — List 5.50



New Model "K"
Full Size Motor Measures
2 1/8" x 2 3/8" x 3 1/16"

MODEL MS. SPECIFICATIONS: Consumes about 15 watts at 3000 R.P.M. without load. Speed 1500 to 3000 R.P.M., depending on frequency and load. Develops about .006 H.P. .171" Diameter Centerless ground steel shaft. Latest type graphite bronze oilless bearings, self-aligning and amply proportioned.

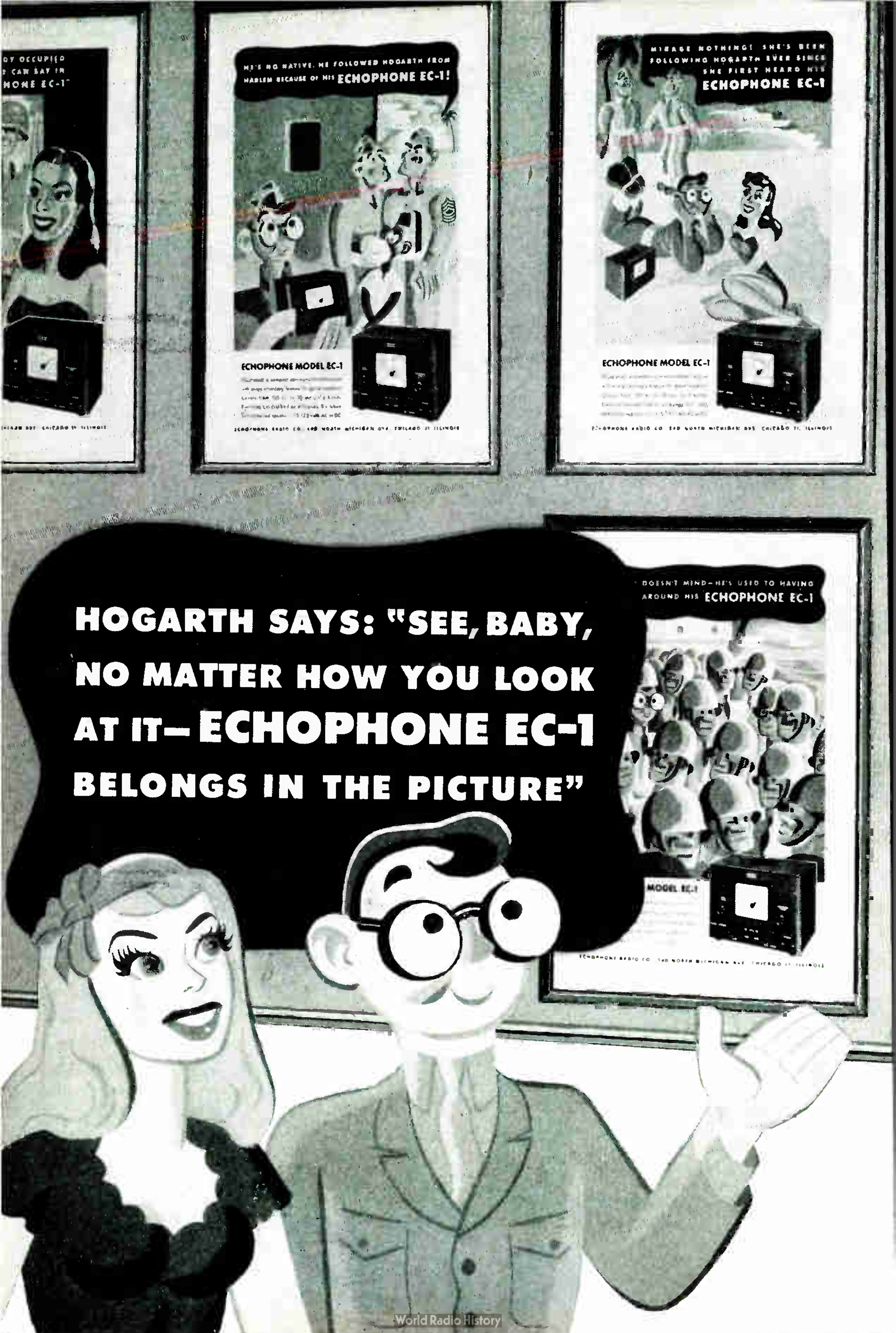
PRICES: Spec. 144 — 110 V. 60 C. Standard Version — List $3.00.

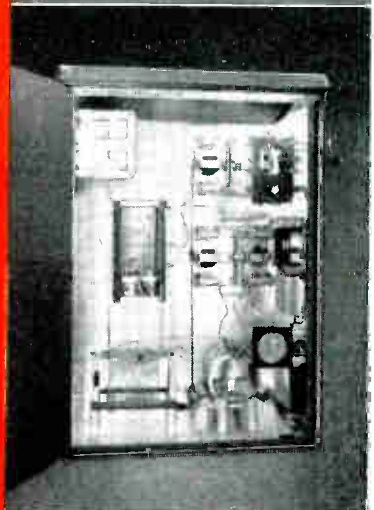


Model "MS"
Full Size Motor Measures
1 3/4" x 2 x 3 1/8"

ALLIANCE MANUFACTURING CO.

ALLIANCE . OHIO


HE'S NO NATIVE. HE FOLLOWED HOGARTH FROM HARLEM BECAUSE OF HIS ECHOPHONE EC-1!
ECHOPHONE MODEL EC-1
MIRAGE NOTHING! SHE'S BEEN FOLLOWING HOGARTH EVER SINCE SHE FIRST HEARD HIS ECHOPHONE EC-1
ECHOPHONE MODEL EC-1
HOGARTH SAYS: "SEE, BABY, NO MATTER HOW YOU LOOK AT IT—ECHOPHONE EC-1 BELONGS IN THE PICTURE"

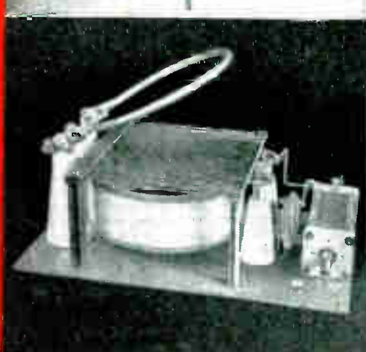


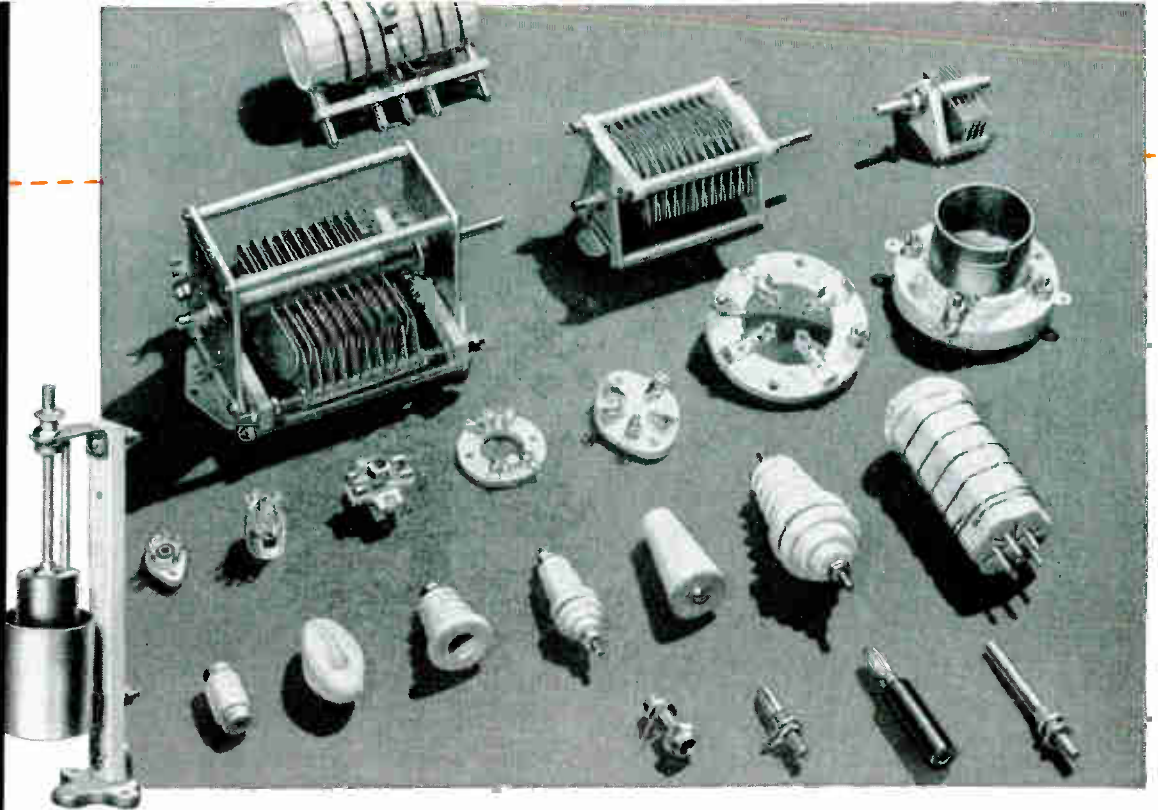
Illustrated down the side and across the bottom are just a few of the many commercial and broadcast items Johnson has been furnishing for many years. In most cases they are not suitable for Amateur use but we thought you might find them interesting.

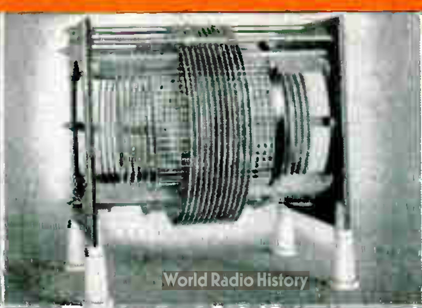
Phasing Equipment whereby the phase relationship and amount of power to each broadcast tower is controlled to produce a desired directional pattern; an Antenna Tower Coupling Unit; another Phasing Unit showing Johnson gas filled pressure condensers and inductors; and a Remote Meter Transformer are all shown along the left side.

Across the bottom will be seen an Open Wire Transmission Line Bracket; five sizes of Concentric Line fittings for use with gas filled lines; a higher power variable condenser; an Inductor; and Remote Control Tuning Equipment.

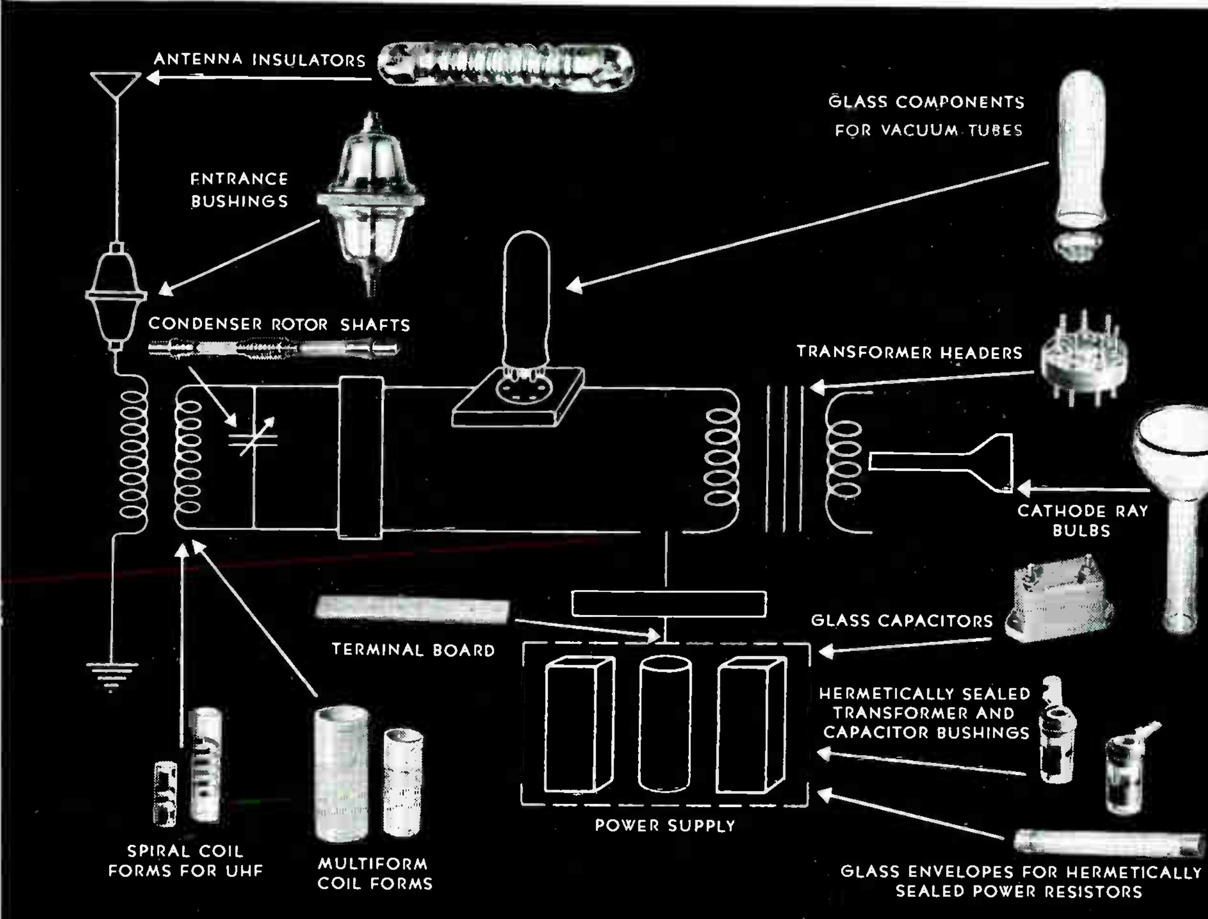
These items are representative of Johnson activities but by no means all-inclusive. Hardly a day passes without Johnson designing and building something entirely new — a piece of apparatus to do a special job or some variation of a previous design. This collective "know-how" of Johnson Engineers is your insurance of Johnson parts doing their job efficiently and economically.







ARE YOU LETTING GLASS HELP YOU ALL IT CAN?



IT WOULDN'T be surprising if you aren't familiar with everything glass is doing in electronic equipment today. Progress has been rapid. In the above *"circuit,"* for example, you'll find it on the job in (twelve) vital places. At Corning right now we're making a lot of other electronic glassware that we can't show. After the war we'll tell you all about it.

It's no accident that a major part of the electronic glassware in use got its start at Corning. We've dug in on some tough ones and ferreted out solutions. They told us we couldn't solder metal to glass — they needed glasses with a coefficient of expansion practically equal to that of fused quartz — they needed something to take the place of mica in capacitors — Corning Research found the answers to these and many other electronic problems.

Our 250 glass experts — the men behind "Corning Research" — our facilities and all our knowledge of glass are at your service. Write for a copy of an informative new booklet "There Will Be More Glass Parts in Postwar Electrical Products." Address Electronic Sales Dept. A-10, Bulb and Tubing Division, Corning Glass Works, Corning, N. Y.

CORNING *means* **Research in Glass**

Electronic Glassware

CORNING U.S. PAT. OFF. — PYREX U.S. PAT. OFF.

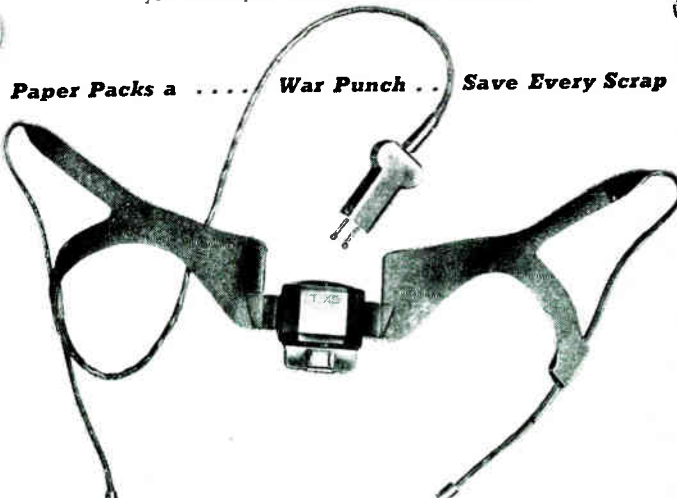
Electro-Voice
MICROPHONES

The extent of our line is but partially illustrated in this advertisement. Our current production is now being utilized in essential services. Soon, however, there will be Electro-Voice Microphones available for civilian use . . . and these will be described fully in subsequent advertisements.

In our South Bend laboratory, we have complete facilities for accurate frequency checking, harmonic wave analysis, measurement of ambient noise, etc. Electro-Voice Microphones reflect painstaking care in design and construction by superior performance in the field. They serve you better . . . for longer periods of time.

If your present limited quantity needs can be filled by any of our Standard Model Microphones, with or without minor modifications, we suggest that you contact your nearest radio parts distributor.

Paper Packs a . . . . War Punch . . Save Every Scrap



ELECTRO-VOICE CORPORATION • 1239 SOUTH BEND AVENUE • SOUTH BEND 24, INDIANA
Export Division: 13 East 40th Street, New York 16, N. Y., U. S. A. Cables: Arlab

CATHODE RAY TUBES

RECEIVING TUBES

TRANSMITTING TUBES

Condensers

Sav-a-shaft Volume Controls

Phototubes

Exciter Lamps

Panel Lamps

Flashlight Bulbs

Count on N.U.

FOR YOUR POST-WAR TUBES AND PARTS

Keep your eyes on N.U. for one of the most complete post-war lines of high grade tubes and accessories. Here you will find the products of advanced war-time research . . . the stepped-up performance you'll be wanting when amateur radio again comes into its own. Write for catalogs and name of nearest distributor.

National Union Radio Corporation, Newark 2, N. J.

Factories: Newark and Maplewood, N. J. Lansdale and Robesonia, Pa.

NATIONAL UNION

RADIO AND ELECTRONIC TUBES

Precision-Accuracy-Exactness IN CRYSTALS

BROADCAST

Type SR-2V. Adjustable mounting permits setting low drift crystal to exact frequency. Available with or without heating unit.

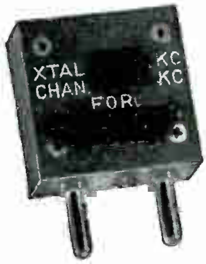
100 KC STANDARD

Type SR-100. 100 Kc crystal unit. New design—new stability. Designed for mounting in standard five-prong tube socket

AMATEUR

Type SR-H. Compact — pressure mounted. Mounts in standard eight-prong socket. Available in wide frequency range up to 10 Mc.

POLICE



Type SR-P. Standard three-quarter inch AT or BT, pressure mounted. Available in frequencies from 1 Mc to 5 Mc. Special cuts to order.

AIRCRAFT

Type SR-F. Typical of a wide variety of crystals carried in stock for civilian aircraft requirements. Available in all standard frequencies.

MULTI-CRYSTAL

Type SR-10S. Permits rapid selection of a wide choice of frequencies. This low loss holder accommodates ten metallized crystals.

SCIENTIFIC RADIO PRODUCTS CO.

SETS THE PACE!

There's always a trail-blazer with the birth of a new industry. At Scientific Radio Products we were among the first to pioneer and introduce mass production of faultless crystals to serve our nation at war. Latest development of our research engineers is a new plated crystal recently approved by the Signal Corps.

Today, in Scientific crystal units you have the UNBEATABLE COMBINATION of modern design, rugged construction, jewel-like precision and time-tested performance—advantages which have made our crystals first choice with those seeking the finest.

Your plans may include equipment in which crystals are used. Perhaps other developments of our engineers may be just what you're looking for. Call on us. We'll be glad to work with you on any problem.

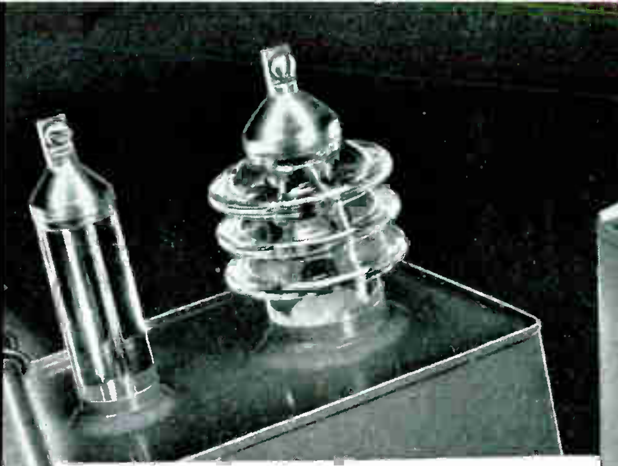
SCIENTIFIC RADIO PRODUCTS CO.

738 W. BROADWAY, COUNCIL BLUFFS, IA.

LEO MEYERSON W9GFQ   E. M. SHIDELER W9IFI

MANUFACTURERS OF PIEZO ELECTRIC CRYSTALS AND ASSOCIATED EQUIPMENT

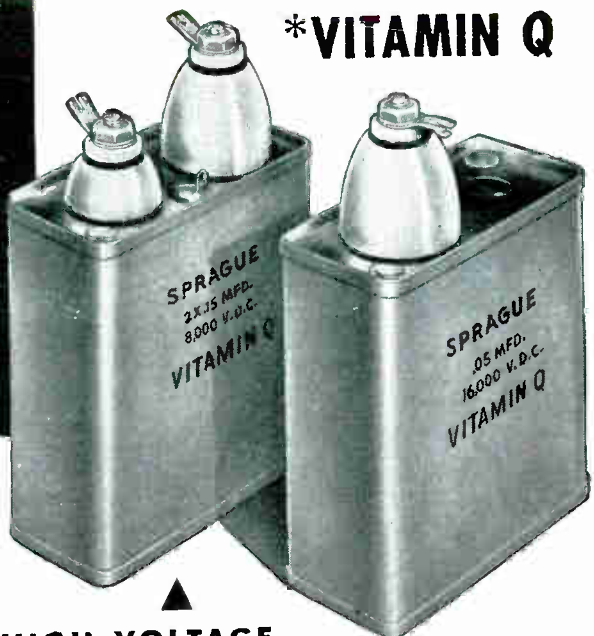
ARMY E NAVY



GLASS-TO-METAL SEALS

The old problem of protecting various capacitor and resistor types against leaks and moisture is solved by a unique glass-to-metal seal pioneered and perfected by Sprague. Glass capacitor bushings are ealed direct to the metal container and do not require djacent metal rings with "matched" coefficients of pansion. On Sprague *KOOLOHM Resistors, the nits are encased in glass tubes which are sealed irectly to the metal ends. The resulting seals are leak-roof, shock-proof, humidity-proof, and fungus-proof.

*VITAMIN Q



HIGH-VOLTAGE, HIGH-TEMPERATURE PROBLEMS SOLVED

When you've got both high voltage and high temperature to contend with in a capacitor application—well, ordinarily, you'd have a problem on your hands. Once again, however, Sprague engineering supplies the answer. Although extremely compact, Sprague Capacitors impregnated with *VITAMIN Q operate satisfactorily at thousands of volts at ambients as high as 105° C. Leakage resistance at room temperature is 20,000 megohms X microfarads—or at least five times better than previous types !

A Step Ahead!

Maybe the few Sprague developments illustrated here meet some immediate need of yours, maybe not. The point, however, is that, whatever you need in up-to-the-minute Capacitors or Wire Wound Resistors, Sprague engineers can supply it. From better by-pass or transmitting capacitors for your amateur rig, to miniature dry electrolytic replacements for more dependable radio service, or special capacitors or resistors for the most exacting engineering requirements, the name Sprague is your assurance of long, trouble-free performance.

SPRAGUE ELECTRIC COMPANY
(formerly Sprague Specialties Co.)
North Adams, Mass.

PERMITS 200° C. CONTINUOUS OPERATION

Many types of electrical equipment can now be designed for 200° C. continuous operation, thanks to the Sprague wartime development of *CEROC 200, a flexible ceramic (inorganic) insulation for copper, nickel, and other types of wire. Smaller equipment can be designed to do bigger jobs. *CEROC 200 dissipates heat rapidly and has an extremely good space factor. In most cases, wire insulated with it can be wound into coils, etc. on existing equipment. You'll be hearing a lot about *CEROC 200 in days to come !

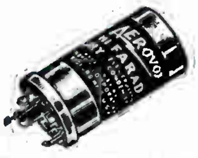
SPRAGUE

PIONEERS OF RADIO-ELECTRONIC PROGRESS

*Trademarks Reg. U. S. Pat. Off.

FOR THE WAR . . . *and after*

"Dandees" are meeting most wartime electrolytic requirements. PBS single-section units. 25 to 450 v. Also dual-section PRS-A concentrically-wound, three leads, and PRS-B separate-section, four leads. Polarity-indicating colored leads.



Aluminum-can electrolytics, still available in certain types, especially on high-priorities. Prong-base F type is typical of extensive Aerovox electrolytic line. Wherever possible, substitute Dandees or cardboard-case PBS.

Metal case paper condensers may still be available in some types, such as Type [illegible], [illegible] v. [illegible] to [illegible] mfd. Also the stamped metal-case '60, in [illegible] and [illegible] v., and particularly the uncased paper sections Type UC, [illegible] to 1000 v.

Tubular paper condensers Type '84 will be found highly satisfactory for most functions. Highly refined construction including extra-heavy-waxing insures excellent performance and life. 400 to 1600 v.

● Type '05 Hyvol oil-filled transmitting capacitors are typical of those veteran components emerging from the war. They are tougher than ever. They have proved that they "can take it" — and then some.

Whether on the fighting front that still remains, the production front, or the post-war "ham" front, you can continue to count on Aerovox for essential capacitors. There's a fully tried, tested and perfected type available for your every need — a real veteran of the war — and in due time, without high priorities.

Oil-filled capacitors still available against high priorities. Type '16 upright or inverted mounting, 400 to 1000 v. Also Type '30 "bathtub" for flat mounting with terminals on top, bottom or side, 400, 600 and 1000 v.

Mica capacitors are mighty scarce. Available only on highest priorities, whether it be the tiny molded-in-bakelite capacitors or the large bakelite-case medium-duty units.

Ask your Aerovox jobber for the latest Aerovox catalog. Ask for free subscription to the monthly Aerovox Research Worker. Or write direct.

Oil-filled transmitting capacitors are available on high priorities. In addition to large round-can '05, there is the inverted screw-mounting '10 type with new double-terminal feature. Also rectangular-can '09 in voltages up to 7500, and '20 series up to 50,000 v.

Heavy duty transmitting and electronic requirements are met by the stack-mounting 1550 series units, and also the cast aluminum case 1870. Ultra high-frequency requirements are met by the sulphur-filled 1860 series. Available on high priorities.

AEROVOX *Capacitors*

INDIVIDUALLY TESTED

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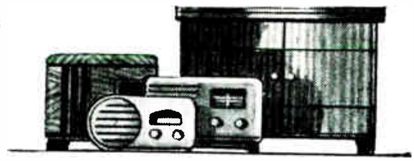
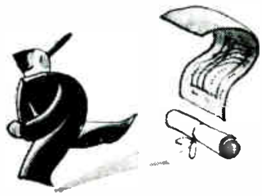
The ECA STORY

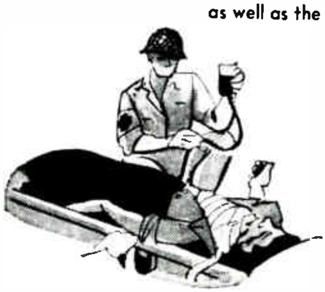


For almost a quarter of a century, most of the principals and personnel of ECA have had the opportunity to grow and expand with electronics. We've had experience producing many different types of highly specialized apparatus — including sound systems, test equipment and other electronic devices.

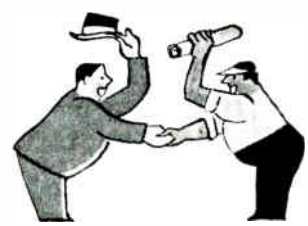


ECA engineers, designers and technicians are all accustomed to working to exacting laboratory standards. Rich in the fundamentals of radio and electronics, we can approach any problem with full confidence that the ultimate result will prove eminently satisfactory.

Naturally, during these crucial war years, our entire production is devoted to materials needed by the Armed Forces. Much of this equipment is of an extremely delicate and precise nature. All require maximum attention to design and construction to meet the standards of ECA as well as the government.



While devoting our working time 100% to war production, we have not forgotten home front activities. The Electronic Corporation of America is proud that each succeeding war bond drive has been over-subscribed, and we're equally proud of the blood donor award given to us by the American Red Cross.



Regularly at ECA, representatives of management meet with representatives of labor to discuss company policy, to fix production quotas, and to look after the needs of the individual worker. We have found that harmonious labor-management relations stimulate the output, efficiency and progress of our organization.

Under these splendid conditions will future ECA products be manufactured. Modern production techniques, trained personnel and precision laboratory and plant facilities will be utilized to produce superior ECA radios and electronic devices for home, industry and medical science. This, in effect, is the ECA story.

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It has been the problem of Sangamo engineers to design and produce capacitors that perform faithfully under these varying conditions, and so assure vitally needed communications at all times.
The wide variety of capacitors illustrated insures the availability of the proper unit for almost any mica capacitor requirement.
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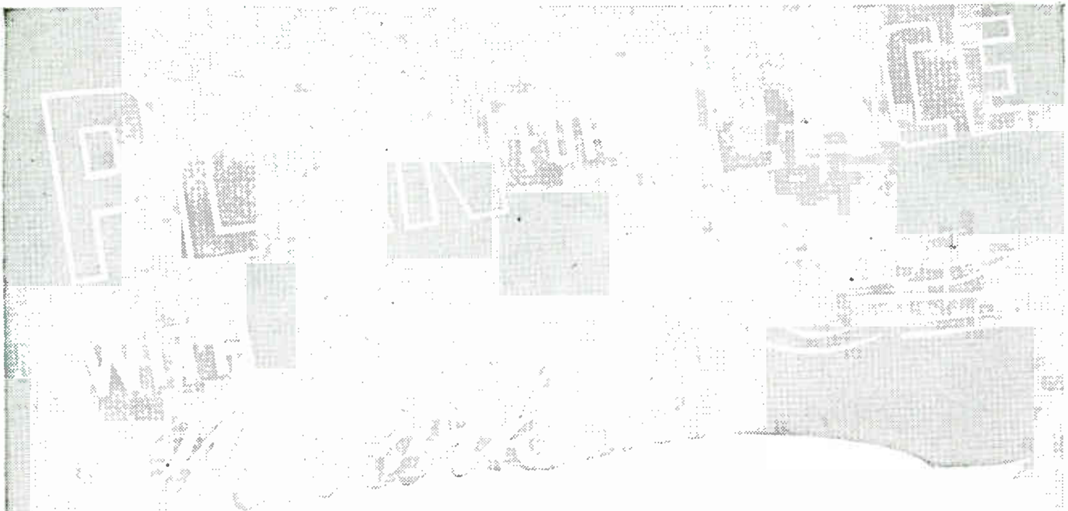

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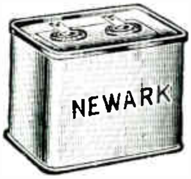

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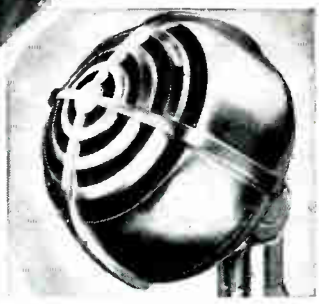
The Shure Super Cardioid, moving coil dynamic microphone is now being used by more than 700 radio stations throughout United States, including the major networks. It reduces background noise, room reflection, reverberation and feed back. Permits greater volume, simplifies pickup problems and allows greater freedom to performers. (556 and 55 Series)

ARMY E NAVY

SHURE
MICROPHONES



The Shure Uniplex is the only cardioid crystal microphone and gives the extra advantages of cardioid performance at low cost. (Model—730 B)

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Designers and Manufacturers of Microphones and Acoustic Devices

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Producers of Approved Precision Crystals for Radio Frequency Control

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Taylor Tubes is proud of its close and lengthy association with the advancement of Amateur Radio. The inception of our business was based on the idea of making *"More Watts per Dollar"* tubes for Amateurs. When Frank Hajek marketed the Taylor 866 at $1.65, the competitive price was $7.50. In the years following, Taylor Tubes carried on with its program of *"More Watts per Dollar"* tubes and in doing so made it possible for thousands of Amateurs to build better Rigs.

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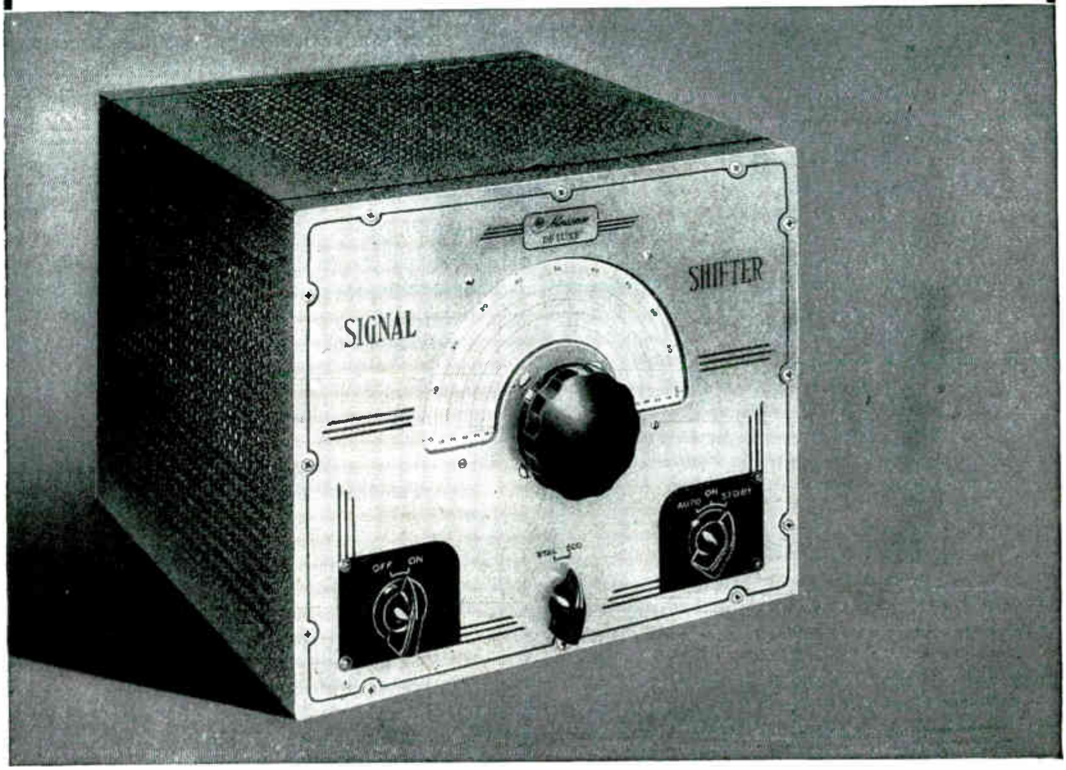

MEISSNER Signal Shifter praised for fine work in Africa, India and South Pacific

"Get them back! Spare nothing! Go to your jobbers, your dealers—the customers themselves. We want *every* Meissner Signal Shifter you can find!" Those were the military instructions to Meissner when war broke out.

And now, as American boys come home from afar with tales of adventure and achievement, they tell of the wonderful things accomplished under handicap by these selfsame Meissner Signal Shifters. For each is a variable-frequency exciter of remarkable stability, delivering a pure T9X output of 7½ watts on the 1.7-, 3.5-, 7-, and 14-mc bands. With these and many other advantages, it permits instant frequency change in any given band—*right from the operating position.* And soon you'll be sending for the *new* and *improved* Meissner Signal Shifter . . . finest product of its kind.

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"Ham" Radio and HYTRON

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THE radio amateur trained himself during peace to be invaluable to the Nation during war. Specializing on tubes exclusively designed for ham radio, Hytron when war began was prepared for immediate and direct conversion to war production. Hytron transmitting and special purpose tubes proved by the ham were ideally suited—with little or no changes—to military applications. Years of practical experience made Army and Navy specialists of radio amateurs overnight. Peacetime tools of these same hams, Hytron tubes joined immediately this new fighting team.

....PRESENT

HAMS with the Services in all parts of the world know the war job Hytron is doing. High-speed receiving tube techniques plus know-how derived from special purpose engineering of tubes for the amateur, make possible a flood of dependable Hytron radar and radio tubes to these fighting ex-hams and potential hams. Proud of winning the Army-Navy "E" for its performance on a huge production job, Hytron is also proud of its ham friends who are transforming innocent-appearing Hytron tubes into deadly weapons.

FUTURE....

THERE should be no concern about adequate post-war amateur frequencies. Excellent wartime performance on far-flung battle fronts has made for ham radio many enthusiastic and influential friends. The ARRL reports that it looks forward with absolute confidence to the opening of new frontiers in expanded frequency ranges to be made available to the post-war amateur. Hosts of hams will return to their old friend, Hytron. For the more familiar lower frequency bands—the very high frequencies—or the new superhighs—their choice will be Hytron.

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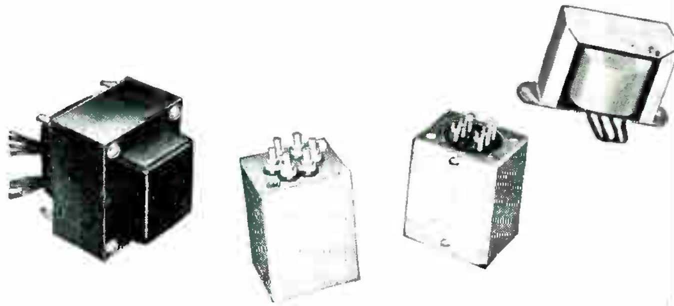
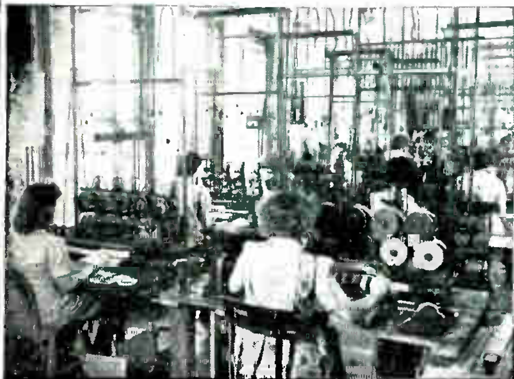
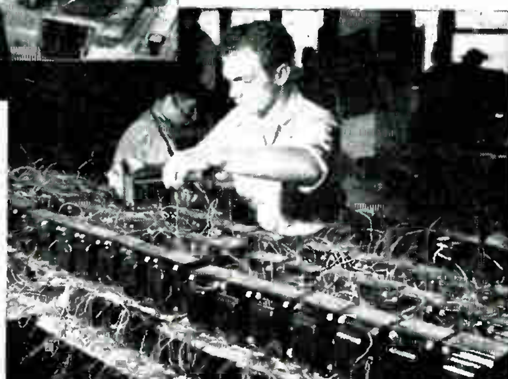
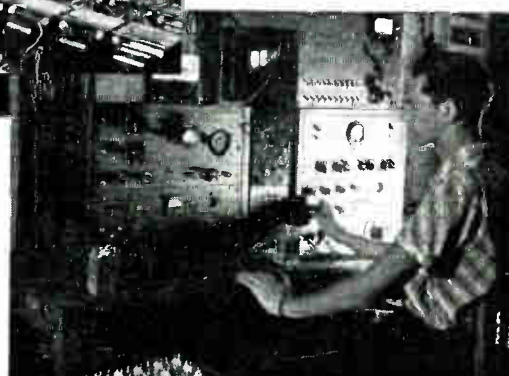




Section of a Freed winding room



Individual attention — quantity production



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INSULATION APPLIED BEFORE WIRE IS WOUND

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No power resistor can be one whit better than the insulation given its windings—and Koolohm ceramic insulation applied to the wire before it is wound gives you the maximum in this respect. Koolohms can be used safely up to their full rated wattage values. Their use of insulated wire permits larger wire sizes to be used, and guards against shorts and changed values. They give more resistance in smaller size, and are readily adaptable to almost any mounting style best suited to your production.

Standard Sprague Koolohms include 5- to 120-watt power types. Other Sprague Resistors include bobbin types, hermetically sealed power resistors, 5- to 150-watts, and meter multipliers. Write for new catalog—just off the press.

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24 G	575	866
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200	811	873

DR 24 G — a Better Triode Transmitter

This is an ideal tube for amateur applications. Its very small physical size (4⅜ high, weighing 1½ ounces) makes it desirable for new modern transmitters with a new high of efficiency. A product from our own laboratories.

A pair of 24G's will handle 300 watts input at 100 mc. in a Class C amplifier with exceptionally low driving power.

When used in a pulse oscillator, very high voltages may be applied in instantaneous pulses.

CHARACTERISTICS

Fil. Volts	6.3
Fil. Amp.	3.0
Amplification Factor	25

DIRECT INTER-ELECTRODE CAPACITANCE

Grid to Plate	1.7 uuf
Grid to Filament	2.5 uuf
Plate to Filament	.4 uuf

MAX. OPERATING CONDITIONS

Class C. Telegraphy

Plate Volt.	2000
Plate Current	75 ma.
Grid Current D.C.	25 ma.
Plate Dissipation	25

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To those who seek superior quality, we can only repeat what studio engineers in major broadcasting stations have told us over a period of years. They agree that ". . . no better instantaneous recording blanks were ever made . . ." than Gould-Moody "Black Seal" Blanks.
There are many reasons why this skeptical industry has been so positive in its choice of Gould-Moody Blanks. For the first time in the history of recording blanks, we have developed a formula that will not dry out and will be as faithful in reproduction after prolonged periods of storage as when originally cut. Careful design is responsible for such desirable features as engineered center thread action . . . four holes in every disc . . . and continued high fidelity through long months of play-backs. There are no wows, wheezes, or rumbles. Moreover, Gould-Moody "Black Seals" will not warp or deteriorate. Their true tonal reproductions, long life, and great economy place them head-and-shoulders above competition.
You can get Gould-Moody "Black Seal" Aluminum Instantaneous Recording Blanks without delay.
Old Aluminum Blanks recoated with the famous "Black Seal" formula on short notice
BUY AND HOLD MORE WAR BONDS UNTIL VICTORY
THE GOULD-MOODY COMPANY
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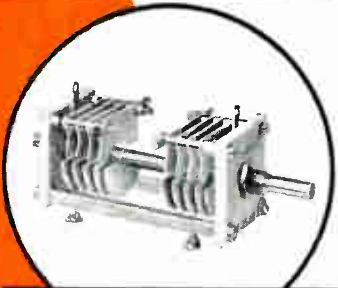
CARDWELL AIR CAPACITORS for Every Purpose

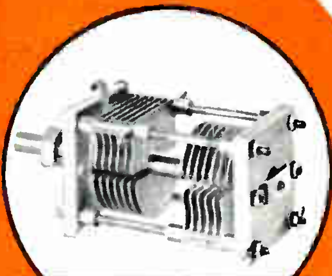
Cardwell condensers for amateur and commercial applications have a distinguished record of service in every hard won field of this war.

On the happy day when amateur radio returns to its own again, there will be an even wider selection of Cardwell components from which to choose.

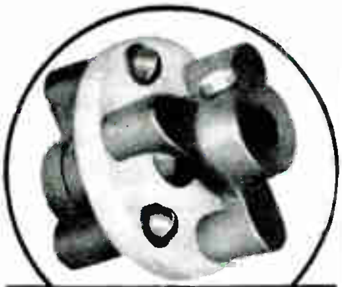
ZR-35-AS
PL-6003



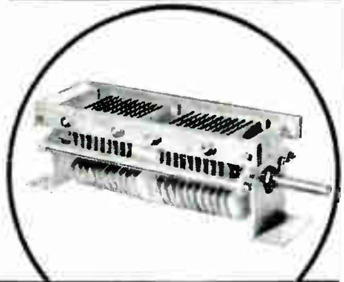
ET-15-AD
PL-6037



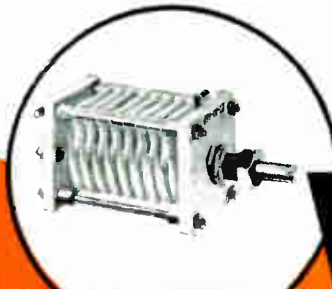
ER-50-ADP
PL-6051



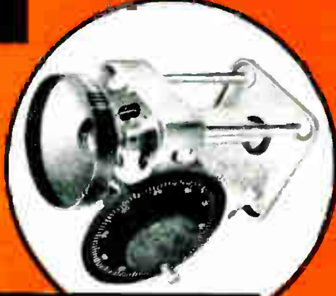
"ENF" RIGID COUPLING
PL-5201



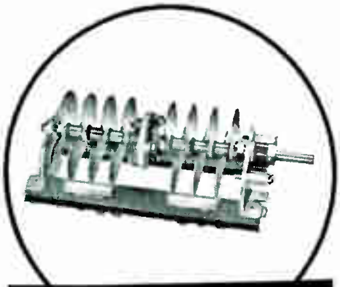
MT-100-GD PL-7030
with PL-5051 Mtg. Brackets



ET-30-ASP
PL-6052

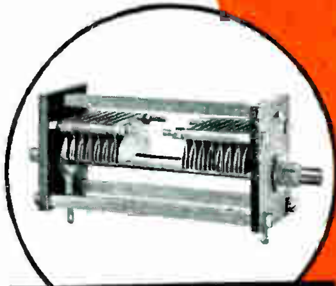


Frequency Meter Condenser
PT. No. 4.080

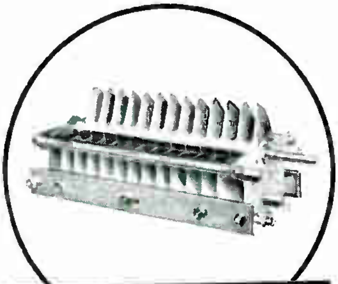


NA-12-NDI
PL-7115

Design engineers may choose freely from an extensive line of Cardwell capacitors and accessories. Models typical of Cardwell design and manufacture are illustrated.

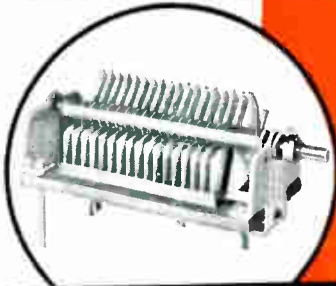


EO-35-AD
PL-6319



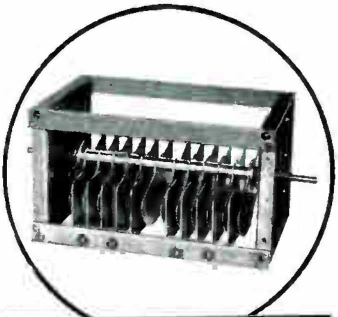
XC-100-XS
PL-8023

Special requirements demanded by the fast moving electronic art as a whole are being taken in their stride by Cardwell development engineers.

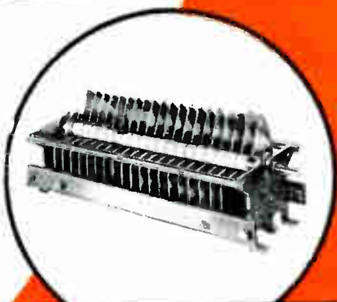


NUV-150-BS
PL-7261

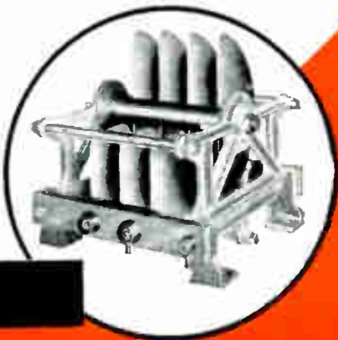
Cardwell's contract manufacturing resources for electronic equipment will be available, as always, for precision instrument production. Our war experiences will be translated into improved devices to implement the peace.



PK-125-QD
SPECIAL



TK-300-US
PL-9110



WX-95-VS
PL-9400

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THE ALLEN D. CARDWELL MANUFACTURING CORPORATION

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A balanced selection of good technical books, additional to the ARRL publications, should be on every amateur's bookshelf. We have arranged, for the convenience of our readers, to handle through the ARRL Book Department those works which we believe to be most useful. Make your selection from the following, add to it from time to time, and acquire the habit of study for improvement. ***Prices quoted include postage.*** Please remit with order.

Theory and Principles

BASIC RADIO PRINCIPLES, by *M. G. Suffern.* Elementary radio principles, circuits and components described in simple practical language, without mathematics. 271 pages, illustrated. 1943 . . . . **$3.00**

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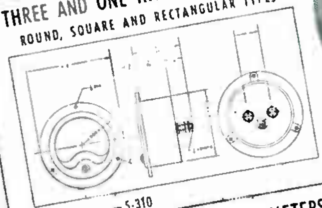
De JUR Electrical Indicating Instruments

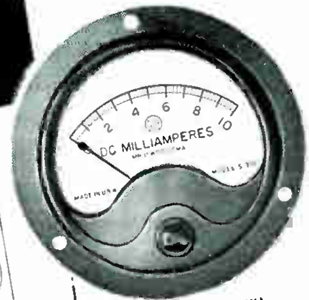
These instruments are normally calibrated for mounting on non-magnetic materials. If desired, instruments will be calibrated for use on magnetic panels. Thickness of the panel should be stated. Scales other than standard type can be supplied and prices will be sent upon request. Special divisions, markings and color combinations are available. Spade pointers are standard equipment. Knife-edge pointers can be supplied at additional cost. Should it be desired to shield the instrument from external magnetic fields, shields can be supplied. These shields increase the body diameter by 3/32 of an inch. Provisions can be made for rear-illumination. For this purpose, translucent scales are necessary. Instruments can be modified to special requirements on orders where the quantity permits such special work. Where these modifications are external, prices will depend upon the instrument sensitivity and range.

Models No. S-210 and No. S-310 are designed to comply with the standards adopted by the American Standards Association for electrical indicating instruments (2½" and 3½" round, flush mounting, panel type).

THREE AND ONE-HALF INCH METERS
ROUND, SQUARE AND RECTANGULAR TYPES

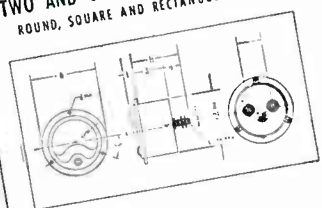


Cross-Section — S-310

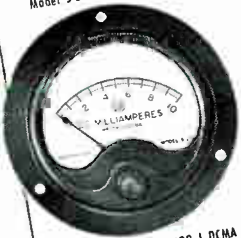


Model S-310 MR 25 I DCMA

TWO AND ONE-HALF INCH METERS
ROUND, SQUARE AND RECTANGULAR TYPES



Cross-Section — S-210



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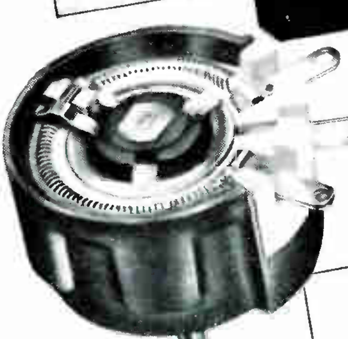
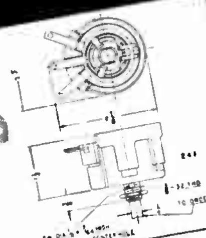
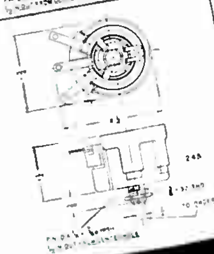




MODEL 241
SPECIFICATIONS

50 WATTS
RANGES—10 to 10,000 Ohms.
MECH. ROTATION—300°
ELEC. ROTATION—270°
WEIGHT—7 OZ.

RANGE IN OHMS	MODEL No.
0- 10	241
0- 50	241
0- 100	241
0- 500	241
0- 1,000	241
0- 5,000	241
0-10,000	241



MODEL 245
SPECIFICATIONS

25 WATTS
RANGES—10 to 10,000 Ohms.
MECH. ROTATION—300°
ELEC. ROTATION—270°
WEIGHT—7 OZ.

RANGE IN OHMS	MODEL No.
0- 10	245
0- 50	245
0- 100	245
0- 500	245
0- 1,000	245
0- 5,000	245
0-10,000	245

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GENERAL OFFICE: NORTHERN BLVD. AT 45th ST., LONG ISLAND CITY 1, N. Y.

hallicrafters

LIFTING THE CURTAIN ON A NEW WORLD OF AMATEUR RADIO

A special message to all Amateurs

hallicrafters RADIO

"WHAT ABOUT PEACETIME PRODUCTION AT hallicrafters?"

That's the subject of many inquiries we get every day. In partial reply we can and do make this promise: all of our attention and the best of our efforts will continue to be focused on the amateur — the ham, the fellow who actually helped us develop Hallicrafters equipment to the high pitch of perfection it enjoys today.

After all, it was the ham himself who helped us get that important thing we call "amateur technique" out of the shack and into the battle line. And it was the ham who went into the service and into the lab to keep working with short wave communications until it became what it has proved to be—a prime battle instrument, a life saver.

The ham is coming home from his war communications job bringing new enthusiasms, new appetites for the wider horizons that can be reached by the kind of radio he likes. All of us want to get back on the air again searching for new thrills, experiencing once again that old excitement.

For the expansion of amateur radio in all directions Hallicrafters will be ready — ready with new and finer equipment, a tougher kind of equipment that has been tried under fire and found to have what it takes.

At Hallicrafters you can be sure the ham will continue to be the key man in our post war plans and his wants will be the prime object of our peacetime production.

W. J. Halligan

The SUPER SKYRIDER,

THE new Hallicrafters Model SX-28A is a further refinement of the famous Model SX-28 that achieved such popularity with amateur and professional operators prior to Pearl Harbor. Embodying circuit refinements and constructional modifications necessitated by the arduous conditions of military service the new SX-28A offers the maximum in communications receiver performance to the discriminating buyer.

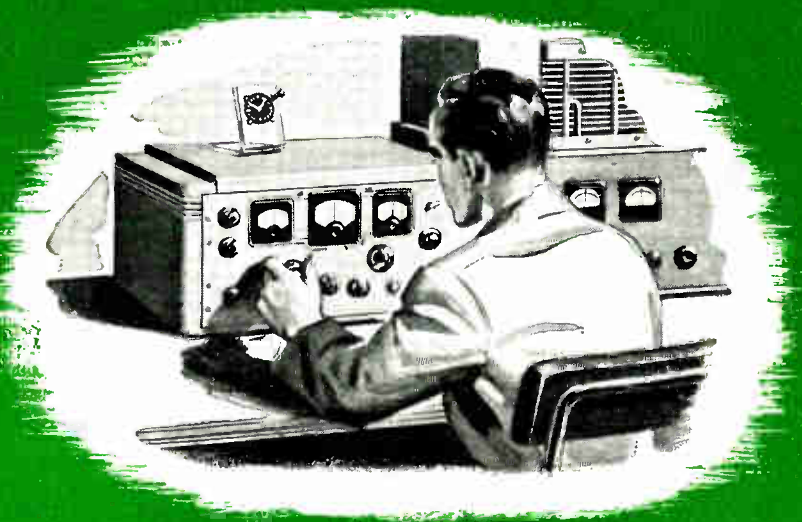
The traditional sensitivity and selectivity of the Model SX-28 have been further improved in this new Super Sky-Rider by the use of "micro-set" permeability-tuned inductances in the r.f. section. The inductances, trimmer capacitors, and associated components for each r.f. stage are mounted on small individual sub-chasses and may be removed as a unit for easy servicing.

Thousands of these fine receivers have seen service with the armed forces in all parts of the world and have maintained and enhanced Hallicrafters' reputation for outstanding quality and performance under the most difficult conditions.

hallicrafters RADIO

Model SX-28A

The Standard of Comparison



FEATURES

1. Frequency range 550 kc. to 42 Mc. continuous in 6 bands.

2. Main tuning dial accurately calibrated in megacycles.

3. Separate calibrated bandspread dial.

4. Two stages of radio frequency amplification.

5. Beat frequency oscillator, pitch variable from front panel.

6. Combination a.v.c.-b.f.o. switch.

7. Send-receive switch.

8. Lamb type 3-stage adjustable noise silencer.

9. Separate r.f. and a.f. gain controls.

10. Provision for battery or external power supply operation.

11. Push-pull 8-watt output stage.

12. Variable tone control, band-pass audio filter and bass boost switch.

13. Provision for break-in operation.

14. 500 or 5000 ohm output.

15. Six position i.f. and crystal filter selectivity switch.

16. Crystal phasing control.

17. "S" meter calibrated in S units and db. above S9.

18. Oscillator compensated for frequency drift.

19. Antenna compensator mounted on panel.

20. Separate a.v.c. amplifier.

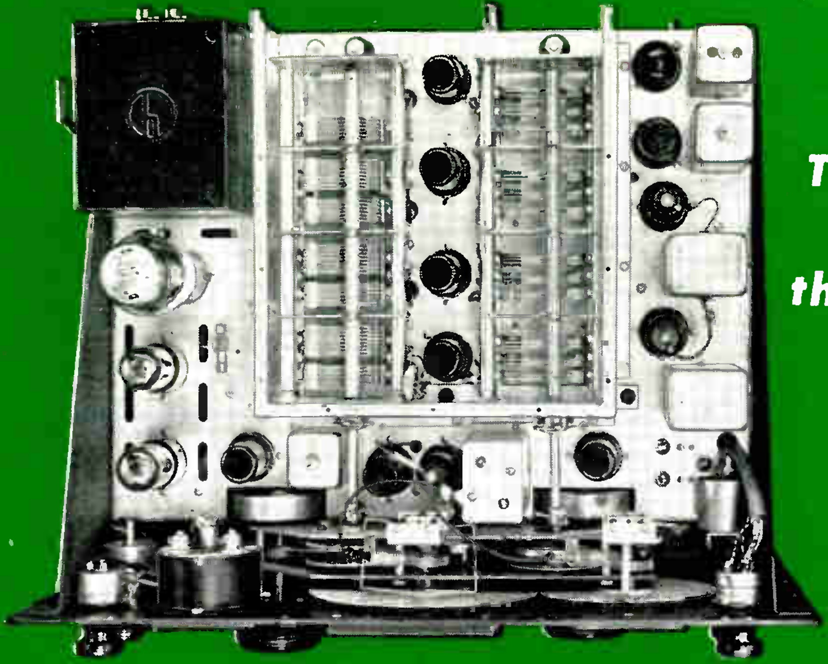
21. "Unit-style" r.f. sections for easy servicing.

22. "Micro-set" type coils in r.f. section permeability tuned.

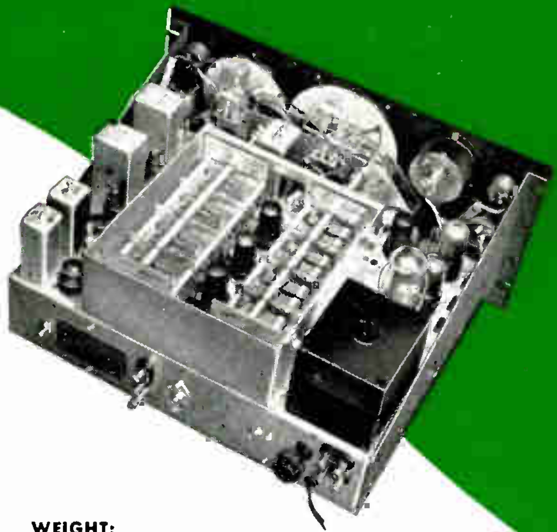
23. Dial lock on main tuning dial.

24. Inertia flywheel tuning and pre-loaded gear drive on main and bandspread dials.

25. Phonograph input jack.



The Inside Story of the Famous "28-A"



CONTROLS:

TONE and A.C. ON/OFF; B.F.O. (pitch control); BASS IN/OUT; A.F. GAIN; MAIN TUNING; R.F. GAIN; BAND SWITCH; ANT. TRIMMER; BANDSPREAD TUNING; A.V.C. and B.F.O. ON/OFF; SELECTIVITY; SEND/RECEIVE; CRYSTAL PHASING; A.N.L.; "S" meter adjustment on rear of chassis.

EXTERNAL CONNECTIONS:

Antenna-ground terminals arranged for single wire or doublet; speaker terminals for either 500 or 5000 ohm output; line cord and plug; line fuse; special socket, normally shorted by octal plug, provides for battery or external power supply operation and stand-by connection to transmitter; phonograph input jack. All connections are mounted on rear of chassis except headphone jack on panel.

PHYSICAL CHARACTERISTICS:

All components of the Super-Skyrider, Model SX-28A are mounted on a rugged steel chassis. Copper plated steel panel has etched black leatherette finish. Panel and chassis are joined by heavy side members. Cabinet is finished in gray wrinkle lacquer with chromium trim. Openings provided for cooling and ventilation.

DIMENSIONS:

Cabinet is 20½" long by 10" high by 14¾" deep. Panel is 19" long by 8¾" high. Clearance needed for relay rack mounting, 17⅜" long by 8¾" high by 14¾" deep.

WEIGHT:

Model SX-28A—75 pounds.
Packed for shipment—87 pounds.

FIFTEEN TUBES:

1—6AB7	1st r.f. Amplifier
1—6SK7	2nd r.f. Amplifier
1—6SA7	Mixer
1—6SA7	H.f. Oscillator
1—6L7	1st i.f. Amplifier Noise Limiter
1—6SK7	2nd i.f. Amplifier
1—6B8	A.v.c. Amplifier
1—6B8	2nd Detector and S Meter Tube
1—6AB7	Noise Amplifier
1—6H6	Noise Rectifier
1—6J5	Beat Oscillator
1—6SC7	1st Audio Amplifier
2—6V6GT	Push-Pull Output Amplifiers
1—5Z3	Rectifier

THE frequency range of the SX-28A extends from 550 kc. to 42 Mc. and is covered in six bands with suitable overlap at the band ends. In addition to the main tuning dial which is accurately calibrated in megacycles, there is a calibrated bandspread dial covering the frequency ranges of 3.5 to 4 Mc., 7 to 7.3 Mc., 14 to 14.4 Mc. and 28 to 30 Mc. Both dials are provided with flywheel inertia tuning.

One stage of r.f. amplification is used on frequencies below 3 Mc. and two stages on the higher frequency bands. These pre-selector stages using the new high-Q "micro-set" inductances assure a good signal-to-noise ratio and a high degree of selectivity. The Model SX-28A has an image ratio of 40 to 1 at 30 Mc.—350 to 1 at 14 Mc., and a proportionately increasing ratio as the frequency is decreased.

The three stage i.f. amplifier is designed to retain its adjustment under conditions of extreme change in temperature and humidity. The i.f. transformers are permeability tuned and are provided with small extra windings which can be connected to increase the coupling between circuits. These windings are used in conjunction with the crystal filter to furnish six different degrees of selectivity. Control is by means of a six-position panel switch. Any desired i.f. selectivity from wide-band high fidelity to razor-sharp c.w. reception is instantly available. In the medium and broad crystal positions the i.f. circuits function as a band-pass filter rather than as the more common broadly peaked resonant circuit and provide fully intelligible reception of radio telephone signals while holding interference and atmospherics to a minimum.

The SX-28A incorporates a double a.v.c. system. A.v.c. voltage for the r.f. and mixer tubes is taken from the broadly tuned carrier after it has passed through only three tuned i.f. circuits. A.v.c. for the i.f. stages, however, is taken from the carrier after it has passed through six tuned i.f.circuits. This arrangement provides a reduction in between-station noise and a more sharply defined aural tuning action. The "S" or signal intensity meter operates in conjunction with the a.v.c. and is calibrated in S units of approximately six db. each and in decibels above S9. A three position panel switch is provided for the control of a.v.c., "S" meter, and b.f.o. circuits.

Other features which contribute to the fine performance of the SX-28A are a Lamb three-stage noise silencer with panel adjustment; push-pull 6V6GT output stage with band-pass filter, bass boost, and tone control; antenna compensator; separate a.f. and r.f. volume controls; and panel stand-by switch with break-in control for transmitter. All controls and switches are conveniently arranged on the panel.

Model PM-23 SPEAKER

THIS Hallicrafters-Jensen speaker is designed for use with the larger Hallicrafters receivers. Of the permanent magnet type the Model PM-23 has a ten-inch cone and is mounted with its coupling transformer in a steel cabinet finished in gray wrinkle lacquer to match the receiver. Speaker opening is concealed by attractive metal grill. Transformer matches 5000 ohm output of receiver.

WEIGHT: packed for shipment, 22 pounds.

hallicrafters RADIO

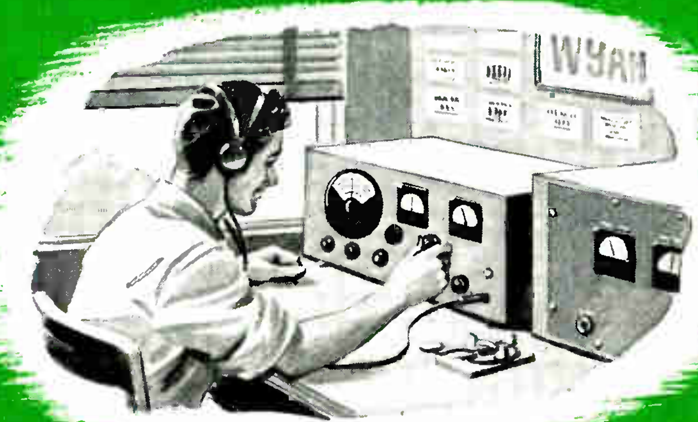
The SUPER

THE SUPER DEFIANT has long been one of Hallicrafters' most popular models. Incorporating every important feature for superb communications receiver performance, the Model SX-25 has achieved true economy without compromising quality.

The outbreak of war with its sudden demand for military communications receivers found Hallicrafters already in mass production of the Model SX-25 for amateur use. Production was immediately stepped up and tremendous quantities of these receivers were put into military communications work. Many minor modifications and improvements in quality of components to meet rigid military requirements were made but the basic design remains unchanged. The rugged construction, fine workmanship, and superb performance which proved so valuable in military service will continue to feature the Hallicrafters Model SX-25.

DEFIANT, *Model* SX-25

Every worthwhile feature at a moderate price



FEATURES

1. Frequency range 545 kc. to 42 Mc., continuous in 4 bands.
2. Main tuning dial accurately calibrated in megacycles.
3. Separate calibrated bandspread dial.
4. Two stages of radio frequency amplification.
5. Beat frequency oscillator, pitch variable from front panel.
6. A.v.c. switch.
7. B.f.o. switch.
8. Send-receive switch.
9. Automatic noise limiter.
10. Separate r.f. and a.f. gain controls.
11. Provision for battery or external power supply operation.
12. Push-pull 8-watt output stage.
13. High-low tone switch.
14. Provision for break-in operation.
15. 500 or 5000 ohm output.
16. Six position i.f. and crystal selectivity switch.
17. Crystal phasing control.
18. "S" meter calibrated in S units and db. above S9.
19. Oscillator compensated for frequency drift.
20. Inertia flywheel tuning on bandspread dial.

CONTROLS:

R.F. GAIN; BAND SWITCH; SELECTIVITY; MAIN TUNING; TONE HIGH-LOW; XTAL PHASING; BANDSPREAD; A.N.L. ON/OFF; A.F. GAIN; PITCH CONTROL; B.F.O. ON/OFF; SEND-REC.; S meter adjustment on rear of chassis.

EXTERNAL CONNECTIONS:

Antenna terminals arranged for doublet or single wire antenna. Speaker output for either 500 or 5000 ohms. Standby terminals for external control of receiver in conjunction with transmitter. Line cord and plug. Special socket, normally shorted by octal plug, for use of external power supply or batteries. All connections are mounted on rear of chassis except headphone jack on panel.

PHYSICAL CHARACTERISTICS:

The SUPER-DEFIANT, Model SX-25 is mounted in a steel cabinet finished in gray wrinkle lacquer. Ornamental metal grills in either end provide ventilation. Chassis is cadmium plated steel.

DIMENSIONS:

Receiver cabinet only—19½" long by 9½" high by 11⅛" deep.

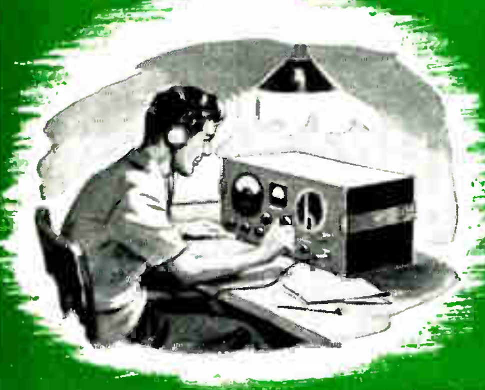
WEIGHT:

Model SX-25—38 pounds.
Packed for shipment—46 pounds.

TWELVE TUBES:

1—6SK7	1st r.f. Amplifier
1—6SK7	2nd r.f. Amplifier
1—6K8	1st Detector-Mixer, h.f. Oscillator
1—6SK7	1st i.f. Amplifier
1—6SK7	2nd i.f. Amplifier
1—6SQ7	2nd Detector, a.v.c. 1st Audio Amplifier
1—6SQ7	Phase Inverter
2—6F6	Push-pull Audio Output Stage
1—6H6	Automatic Noise Limiter
1—6J5GT	Beat Frequency Oscillator
1—80	Rectifier

The SKY CHAMPION, *Model* S-20R



COMPACT • RELIABLE

Top Performance in the Low Price Field

THE Hallicrafters SKY CHAMPION, Model S-20R is probably the greatest value ever offered in communications receivers. Its simplicity of design, excellent workmanship, and sturdy construction insure long, satisfactory service and make traditional Hallicrafters performance available to the purchaser of an economical receiver.

In common with its larger brothers, the Model S-20R has a distinguished war record, and like them, it has been strengthened and improved to cope with military requirements. Large quantities have been produced for the armed forces and have been used for training and communications purposes where performance was important, but the use of a complicated receiver was not justified. It is a compact, reliable receiver offering top performance in the low priced field.

FEATURES

1. Frequency range 550 kc. to 43 Mc., continuous in four bands.
2. Main tuning dial accurately calibrated in megacycles.
3. Separate electrical bandspread dial.
4. Beat frequency oscillator, pitch variable from front panel.
5. A.v.c. switch.
6. B.f.o. switch.
7. Send-receive switch.
8. Automatic noise limiter.
9. Separate r.f. and a.f. gain controls.
10. Provision for battery or external power supply operation.
11. 2½-watt output stage.
12. Three-position tone control.
13. Provision for break-in operation.
14. Provision for external S meter.
15. Inertia flywheel tuning on bandspread dial.
16. Internal rubber shock mounted 5" dynamic speaker.

CONTROLS:
R.F. GAIN; BAND SWITCH; AUDIO GAIN; MAIN TUNING; A.V.C. ON OFF; B.F.O. ON/OFF; BANDSPREAD TUNING; A.N.L. ON/OFF; TONE A.C. OFF HIGH/MED./LOW; PITCH CONTROL; SEND-REC.

EXTERNAL CONNECTIONS:
Antenna terminals for doublet or single wire antenna. Line cord and plug. Special socket for operation from external power supply. All connections except headphone jack are mounted on rear of chassis.

PHYSICAL CHARACTERISTICS:
Components of the Model S-20R are mounted on a strong cadmium-plated steel chassis. Cabinet is of steel finished in machine tool gray enamel with chrome trim. Internal five-inch dynamic speaker is held in rubber shock mounts.

DIMENSIONS:
Cabinet only—18½" long by 8½" high by 9⅜" deep.

WEIGHT:
Packed for shipment—32 pounds.

NINE TUBES:

1—6SK7	R.f. Amplifier
1—6K8	1st Detector-Mixer, h.f. Oscillator
1—6SK7	1st i.f. Amplifier
1—6SK7	2nd i.f. Amplifier
1—6SQ7	2nd Detector, a.v.c. and 1st Audio Amplifier
1—6F6G	2nd Audio Amplifier
1—6H6	Automatic Noise Limiter
1—6J5GT	Beat Frequency Oscillator
1—80	Rectifier

The SKYRIDER MARINE, *Model* S-22R

An Efficient Marine Receiver at a Moderate Price

THE Hallicrafters Model S-22R is specifically designed for marine service covering frequencies from 110 kc. to 18 Mc. Maximum convenience is assured through the use of a directly calibrated main tuning dial and the division of bands so that calling and working frequencies lie in the same band. An efficient mechanical bandspread with separate dial provides for easy logging. Excellent image rejection on the higher frequencies is achieved by the use of a 1600 kc. i.f. amplifier.

Special precautions have been taken to protect the Model S-22R against the hazards of salt sea atmosphere. Mica trimmer condensers are treated to maintain their adjustment, transformers are impregnated, and the chassis is nickel plated.

Many boats are provided with 110 volts d.c. and the SkyRider Marine is designed for a.c./d.c. operation. This feature makes the Model S-22R valuable for use ashore where a high performance receiver to operate from a d.c. line is desired.

FEATURES

1. Frequency range 110 kc. to 18 Mc. continuous in four bands.
2. Main tuning dial accurately calibrated in megacycles.
3. Mechanical bandspread with separate dial.
4. Beat frequency oscillator, pitch variable from front panel.
5. A.v.c. switch.
6. B.f.o. switch.
7. Send-receive switch.
8. Separate r.f. and a.f. gain controls.
9. Variable tone control.
10. Inertia flywheel tuning.
11. A.c./d.c. operation.
12. 1600 kc. iron core i.f. for maximum image rejection.
13. Internal rubber shock mounted 5" PM speaker.

CONTROLS:
R.F. GAIN; BAND SWITCH; AUDIO GAIN; A.V.C. ON OFF; MAIN TUNING; B.F.O. ON/OFF; TONE CONTROL; PITCH CONTROL; SEND-REC.

EXTERNAL CONNECTIONS:
Antenna terminals arranged for doublet or single wire. Line cord and plug. Phone jack on panel.

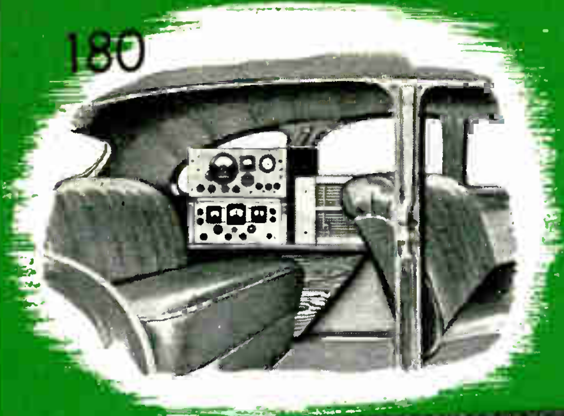
PHYSICAL CHARACTERISTICS:
The Model S-22R is mounted in a steel cabinet finished in black wrinkle lacquer with chrome trim. Steel chassis is nickel-plated. Five-inch PM dynamic speaker is built in.

DIMENSIONS:
Cabinet only—18½" long by 8½" high by 9⅜" deep.

WEIGHT:
Packed for shipment—31 pounds.

EIGHT TUBES:

1—6SK7	R.f. Amplifier
1—6K8	1st Detector-Mixer, r.f. Oscillator
1—6SK7	1st r.f. Amplifier
1—6SK7	2nd r.f. Amplifier
1—6SQ7	2nd Detector, a.v.c., 1st a.f. Amplifier
1—25L6	2nd a.f. Amplifier
1—6J5	Beat Frequency Oscillator
1—25Z5	Rectifier



Model S-36A

FM • AM • CW

THE new Hallicrafters a.m./f.m. receiver, Model S-36A, is designed for maximum performance on the very high frequencies. Using acorn tubes in the r.f. amplifier, first detector and high frequency oscillator circuits, the S-36A provides continuous frequency coverage from 27.8 to 143 megacycles. Either a limiter and discriminator for f.m. or a third i.f. amplifier, diode detector and noise limiter for a.m. may be switched into the circuit from the front panel. A beat frequency oscillator is provided for the reception of c.w. telegraph signals. The S-36A incorporates a new 3-watt audio system with a response curve which is essentially flat from 40 to 15,000 cycles. All components are of the highest quality and the entire receiver is designed for service in any climate. Combining f.m., a.m., and c.w. telegraph reception in one superbly engineered unit, the S-36A provides the utmost in very-high-frequency reception.

hallicrafters RADIO

Outstanding for Sensitivity—Stability High Fidelity—VHF Versatility

FEATURES

1. Frequency range 27.8 Mc. to 143 Mc. continuous in three bands.
2. Main tuning dial accurately calibrated in megacycles.
3. Mechanical bandspread dial.
4. R.f. stage with acorn tube.
5. Beat frequency oscillator, pitch variable from panel.
6. A.v.c. switch.
7. B.f.o. switch.
8. Send-receive switch.
9. Automatic noise limiter.
10. Separate r.f. and a.f. gain controls.
11. Push-pull high fidelity output stage.
12. 4-position tone control with bass boost.
13. Provision for break-in operation.
14. 500 or 5000 ohm output plus special balanced 600 ohm line.
15. Sharp-broad selectivity switch.
16. Dual purpose S and tuning meter.
17. Oscillator compensated for frequency drift.
18. Antenna compensator mounted on panel.
19. R.f. assembly easily removed for servicing.
20. Inertia flywheel tuning.
21. Hermetically sealed transformers and reactors.
22. All paper condensers oil impregnated and hermetically sealed.
23. Moisture proofed wiring.
24. F.m./a.m. switch.
25. Switch on chassis permits operation on 115 or 230 volts a.c.
26. Line fuse on panel.
27. Improved gear drive in dust proof housing.
28. S meter adjustable from front panel.

CONTROLS:

R.F. GAIN; A.V.C. ON/OFF; BAND SWITCH; ANTENNA; SEND-RECEIVE; SELECTIVITY; TONE; A.N.L. ON/OFF; TUNING; PITCH CONTROL; METER ADJUSTMENT; B.F.O. ON/OFF; A.M./F.M.; A.F. GAIN.

EXTERNAL CONNECTIONS:

Input terminals for single wire and doublet antennas. 500 ohm, balanced 600 ohm, and 5000 ohm terminals for speaker. Line cord and plug. Octal socket on rear of chassis permits operation from external power source such as batteries and makes provision for remote stand-by switch. This socket is normally shorted by octal plug. Line fuse is mounted on front panel.

PHYSICAL CHARACTERISTICS:

All components of the S-36A are mounted on a heavy steel chassis which is provided with special end plates for ease of maintenance. High frequency r.f. section is built in a separate chassis which may easily be removed for servicing. Cabinet is of steel finished in black wrinkle lacquer. Military type shock mounting is available if desired.

DIMENSIONS:

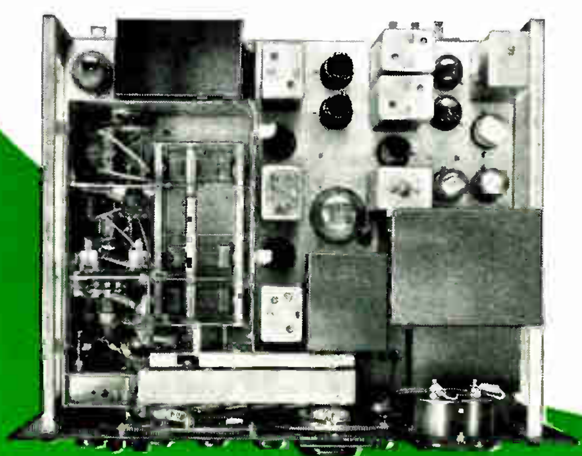
Model S-36A—19¼″ wide by 9½″ high by 15¾″ deep.
Model S-36A with military type shock mounting—21½″ wide by 11¼″ high by 15¾″ deep.

WEIGHT:

Packed for shipment—95 pounds.

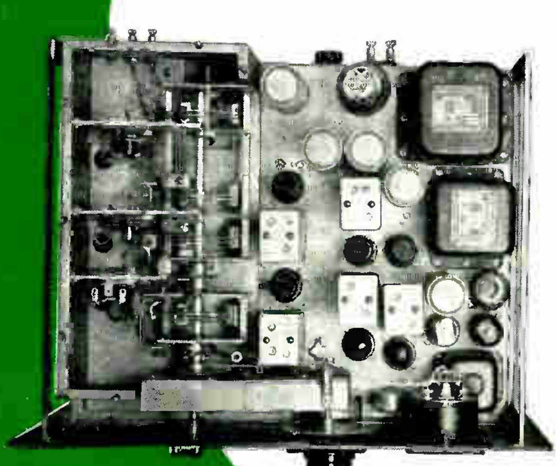
FIFTEEN TUBES:

1—956	(Acorn) Radio Frequency Amplifier
1—954	(Acorn) Converter-Mixer
1—6AC7 or 1852	First i.f. Amplifier
1—6AB7 or 1853	Second i.f. Amplifier
1—6SK7	Third i.f. Amplifier
1—6H6	A.m. Detector and Automatic Noise Limiter
1—6AC7 or 1852	F.m. Limiter
1—6H6	F.m. Discriminator
1—6SL7GT	Audio Amplifier
1—VR150	Voltage Regulator
2—6V6GT	Power Audio Amplifier
1—5U4G	Rectifier
1—6J5	Beat Frequency Oscillator
1—955	(Acorn) High Frequency Oscillator



Model S-37 FM • AM

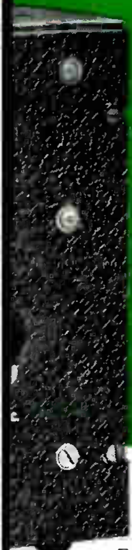
THE new Model S-37 has been designed to fill the need for very-high-frequency receiving equipment with the performance characteristics of Hallicrafters' top communications receivers and a frequency range extending above 200 Mc. Basically similar to the Model S-36A this new receiver incorporates the latest developments in v.h.f. circuit design and provides sensitivity and selectivity in the range from 130 to 210 Mc. that is in every way comparable to the performance of fine communications receivers on the standard frequencies.



A new pre-loaded gear drive with separate bandspread dial provides ease of tuning and the entire range of the receiver is covered without band switching. Two r.f. stages are used and, in conjunction with an intermediate frequency of 18 Mc., assure an amazingly high ratio of image rejection. Hermetically sealed transformers and capacitors make the Model S-37 suitable for use in any climate.

This new receiver again emphasizes Hallicrafters' pre-eminence in the commercial production of v.h.f. equipment.

The Highest Frequency Range of Any General Coverage Commercial Type Receiver



FEATURES

1. Frequency range continuous from 130 Mc. to 210 Mc.
2. Main tuning dial accurately calibrated in megacycles.
3. Mechanical bandspread dial.
4. Two r.f. stages with acorn tubes.
5. A.v.c. switch.
6. Send-receive switch.
7. Automatic noise limiter.
8. Separate r.f. and a.f. gain controls.
9. Variable tone control.
10. Provision for break-in operation.
11. 500 or 5000 ohm output.
12. Dual purpose S and tuning meter.
13. Oscillator compensated for frequency drift.
14. Antenna compensator mounted on panel.
15. R.f. assembly easily removed for servicing.
16. Inertia flywheel tuning.
17. Hermetically sealed transformers and reactors.
18. All paper condensers oil impregnated and hermetically sealed.
19. Moisture-proof wiring.
20. F.m./a.m. switch.
21. Provision for operation on 115 or 230 volts a.c.
22. Line fuse on rear of chassis.
23. Improved gear drive in dust proof housing.
24. "S" meter adjustable from front of panel.
25. 18 Mc. i.f. for maximum image rejection.

CONTROLS:

R.F. GAIN; A.V.C. ON/OFF; ANTENNA TRIMMER; SEND-RECEIVE; TONE; A.N.L. ON OFF; TUNING; METER ADJUSTMENT; A.M./F.M.; A.F. GAIN; POWER ON/OFF.

EXTERNAL CONNECTIONS:

Input terminals for single wire and doublet antennas. 500 ohm, and 5000 ohm terminals for speaker. Line cord and plug. Octal socket on rear of chassis permits operation from external power source such as batteries and makes provision for remote stand-by switch. This socket is normally shorted by octal plug. Line fuse is mounted on rear of chassis.

PHYSICAL CHARACTERISTICS:

All components of the S-37 are mounted on a heavy steel chassis which is provided with special end plates for ease of maintenance. High frequency R.F. section is built in a separate chassis which may easily be removed for servicing. Cabinet is of steel finished in black wrinkle lacquer. Military type shock mounting is available if desired.

DIMENSIONS:

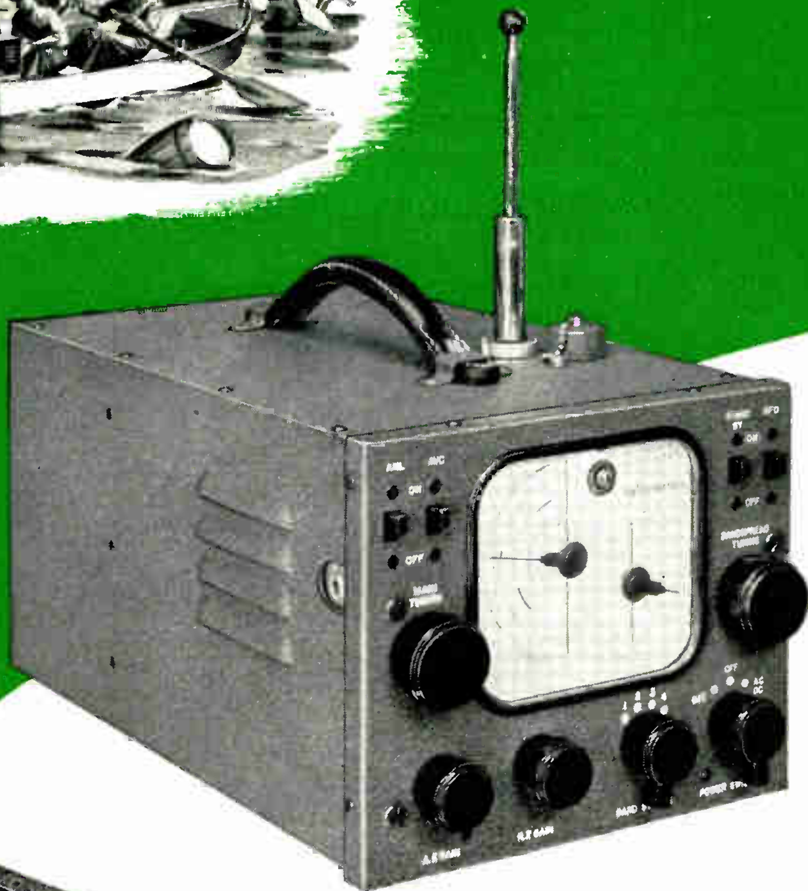
Model S-37—19¼" wide by 9½" high by 14-13/16" deep.

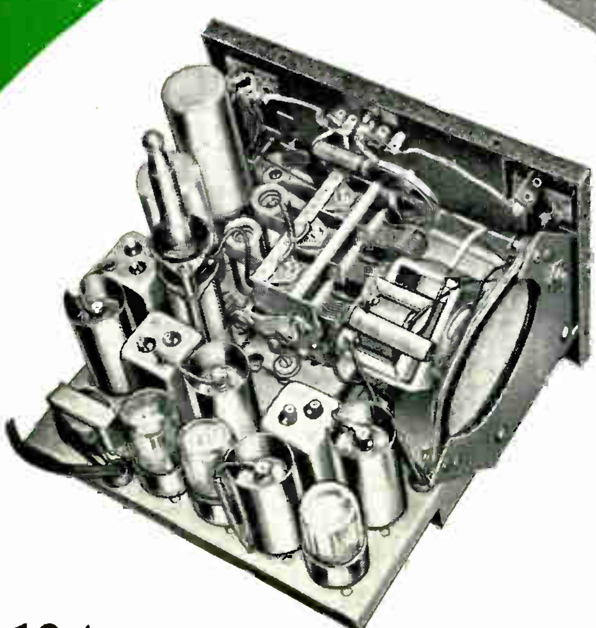
WEIGHT: Packed for shipment — 95 pounds.

FOURTEEN TUBES:

2—954	(Acorn) Radio Frequency Amplifiers
1—954	(Acorn) Converter-Mixer
1—6AC7	or 1852 First i.f. Amplifier
1—6AB7	or 1853 Second i.f. Amplifier
1—6SK7	Third i.f. Amplifier
1—6H6	A.M. Detector and Automatic Noise Limiter
1—6AC7	or 1852 F.M. Limiter
1—6H6	F.M. Discriminator
1—6SC7	Audio Amplifier
1—VR150	Voltage Regulator
1—6V6GT	Power Audio Amplifier
1—5U4G	Rectifier
1—955	(Acorn) High Frequency Oscillator

The





THE new Sky Ranger is a 9-tube communications receiver which combines the utmost in convenience with a high order of performance.

Incorporating electrical bandspread, b.f.o. for c.w. reception, automatic noise limiter, and the standard controls found on good communications receivers, the light weight, high sensitivity and collapsible built-in antenna of the Sky Ranger make it an ideal portable receiver. It may be operated on either 117 volts a.c. or d.c. or on its own self-contained batteries. Frequency coverage is contInous from .54 to 30.5 Mc. in 4 bands.

SKY RANGER, *Model S-39*

OUTSTANDING PERFORMANCE in its most compact form

FEATURES

1. Operates from its own self-contained batteries or 115 volts a.c. or d.c.
2. Frequency range 540 kc. to 30.5 Mc. continuous in four bands.
3. Main tuning dial accurately calibrated in megacycles.
4. Separate bandspread dial.
5. R.f. stage used on all bands.
6. Beat frequency oscillator.
7. A.v.c. switch.
8. B.f.o. switch.
9. Send-receive switch.
10. Automatic noise limiter.
11. Separate r.f. and a.f. gain controls.
12. Collapsible built-in antenna.
13. Moisture-proof wiring.
14. Components impregnated for use in tropical climates.
15. Neon on/off indicator to prevent waste of batteries.
16. Permeability tuned r.f. and i.f. stages.
17. Plug-in type filter capacitors.
18. Completely rainproofed for outdoor use.

CONTROLS:

MAIN TUNING; BANDSPREAD TUNING; A.F. GAIN; R.F. GAIN; BAND SWITCH; POWER SWITCH; A.N.L. ON/OFF; A.V.C. ON/OFF; STANDBY ON/OFF; B.F.O. ON/OFF.

EXTERNAL CONNECTIONS:

Socket and plug are provided to connect doublet or single wire antenna. A.c./d.c. power cord is carried in a closed compartment at rear of set. Phone jack permits use of headphones and shuts off loud speaker automatically.

PHYSICAL CHARACTERISTICS:

The S-39 is housed in a strong steel cabinet finished in olive drab wrinkle lacquer. All components are mounted on a pressed steel chassis and the entire receiver is designed for hard service. Particular care has been taken to make all components easily accessible for servicing.

DIMENSIONS:

Cabinet alone—7¼" high by 8¾" wide by 13½" deep.

Over all: 8½" high by 8¾" wide by 15¼" deep.

WEIGHT:

Model S-39—28 pounds, with batteries.

NINE TUBES:

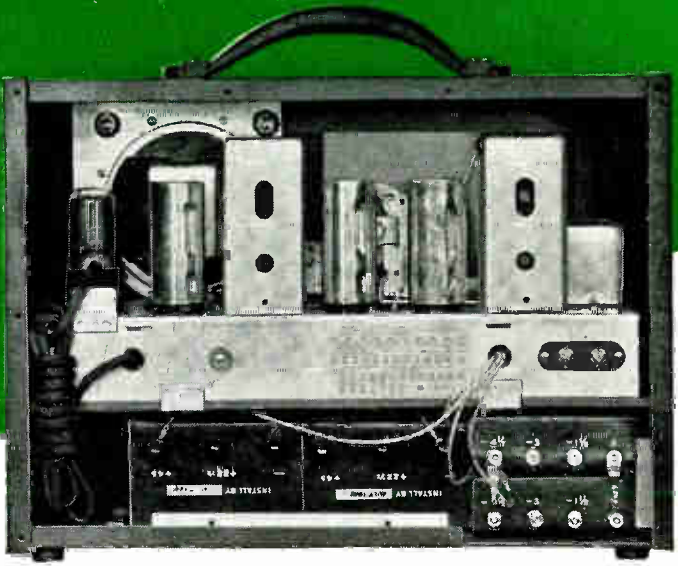
1—1T4	R.f. Amplifier
1—1R5	Mixer
1—1P5GT	First i.f. Amplifier
1—1P5GT	Second i.f. Amplifier
1—1H5GT	Second Detector, First Audio Amplifier and a.v.c.
1—1H5GT	Beat Frequency Oscillator, Automatic Noise Limiter
1—3Q5GT	Second Audio Amplifier
2—35Z5GT	Rectifiers

hallicrafters RADIO

The SKY COURIER *Model* RE-1

THE Hallicrafters SKY COURIER, Model RE-1, sets a new standard for portable broadcast receivers. Built to the same exacting specifications as Hallicrafters' communications receivers, the RE-1 combines unusual sensitivity and selectivity on the standard and short-wave broadcast bands with ruggedness, convenience, and portability. The SKY COURIER employs high-Q iron core inductances in all r.f. and i.f. circuits, operates from its own self-contained batteries or from 115 volts a.c. or d.c., and is mounted in a sturdy metal cabinet with recessed control panel affording maximum protection against accidental damage. This new Hallicrafters portable is truly an ideal receiver for those who demand the utmost in long distance performance.

A portable with real short wave performance



FEATURES

1. Operates from its own self-contained batteries or 115 volts a.c. or d.c.
2. Frequency range 550 kc. to 1600 kc., 2.8 Mc. to 7.8 Mc., 7 Mc. to 19 Mc.
3. Main tuning dial accurately calibrated in megacycles.
4. Separate bandspread dial.
5. High-Q iron core inductances throughout assure maximum sensitivity.
6. Dual audio output stages, combining economy on battery operation and maximum output on power line operation.
7. Components impregnated for use in tropical climates.
8. Self-contained 5" PM speaker, moisture resistant cone.
9. Recessed control panel for protection and convenience.
10. Antenna carried on reel on rear of cabinet.
11. Plug-in. type filter capacitors.
12. Special socket for line cord provides automatic changeover from a.c./d.c. to battery operation.

CONTROLS:

BANDSPREAD TUNING; BAND SWITCH; MAIN TUNING; VOLUME.

EXTERNAL CONNECTIONS:

Line cord for a.c./d.c. operation carried in compartment in back of cabinet. When used with self-contained batteries, line cord is plugged into socket at rear of chassis. Antenna and ground terminals on back of chassis.

BATTERY REQUIREMENTS:

Two Burgess No. 5308 or equivalent and four Burgess No. 2370 standard terminal type or two Burgess G3 plug type or equivalent.

PHYSICAL CHARACTERISTICS:

The SKY COURIER is mounted in a sturdy steel cabinet finished in machine tool gray wrinkle lacquer. Control panel is recessed for protection of knobs, dial, etc. Speaker mounted on front panel. Battery compartment in bottom of cabinet separated from cadmium plated chassis by non-corroding shelf. Cabinet rests on four rubber feet, carrying handle is of leather.

DIMENSIONS:

Cabinet: 10½" high by 15" wide by 8" deep. Allow 1¼" extra height for handle and feet.

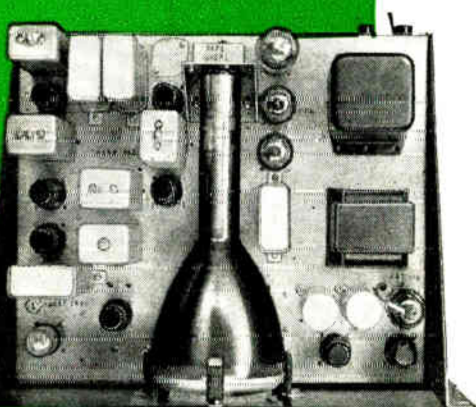
WEIGHT:

Model RE-1: 30 pounds (with batteries).

SEVEN TUBES:

1—1R5	Mixer
1—1N5GT	First i.f. Amplifier
1—1N5GT	Second i.f. Amplifier
1—1H5GT	Second Detector, a.v.c., and First Audio Amplifier
1—50L6GT	Audio Power Amplifier for a.c./d.c. operation
1—3Q5GT	Audio Power Amplifier for Battery Operation
1—35Z5GT	Rectifier for a.c./d.c. operation

hallicrafters RADIO



Model S-35

PANORAMIC RECEIVER

THE Hallicrafters PANORAMIC RECEIVER, Model S-35 is one of the newest and most interesting applications of the cathode-ray tube. This equipment, a special adapter mounted complete with an SX-28A receiver, makes possible the visual monitoring of whole sections of the frequency spectrum up to 100 kc. in width. All stations on the air in the portion of the spectrum being monitored are visible on the screen of the S-35. The station which is audible in the speaker or headphones always appears in the center of the oscilloscope screen. As the receiver is tuned the entire picture shifts across the screen.

The panoramic adapter unit consists of a chassis and panel of approximately the same dimensions as the SX-28A. Only one electrical connection is made between the adapter and the SX-28A and it does not interfere in any way with the normal operation of the receiver.

NOTE: All of the following are in addition to the normal equipment of the standard Model SX-28A receiver.

CONTROLS:

R.F. GAIN; SWEEP WIDTH; A.V.C.; VERTICAL (gain); HORIZONTAL (gain); POWER ON/OFF.

EXTERNAL CONNECTIONS:

Line cord and plug. Input lead from mixer stage of SX-28A.

PHYSICAL CHARACTERISTICS:

Panoramic adapter components are mounted on a steel chassis. Panel is of same dimensions as the SX-28A. Panel has etched black leatherette finish. Panel and chassis are joined by rugged end braces, and adapter unit and SX-28A are mounted together in sturdy metal cabinet finished in gray wrinkle lacquer.

DIMENSIONS:

Cabinet only, 20½" wide by 18⅝" high by 18" deep.

WEIGHT:

Model S-35—105 pounds.

Packed for shipment—166 pounds.

FOURTEEN TUBES:

1—6SG7	455 kc. Input Amplifier
1—6SA7	1st Detector
1—6SK7	100 kc. i.f. Amplifier
1—6SQ7	2nd Detector and Vertical Amplifier
1—6SN7GT	Sawtooth Oscillator
1—6SJ7	Return Trace Blanking Tube
1—6AC7	Reactance Modulator
1—6J5	R.f. Oscillator
1—6SC7	Horizontal Amplifier
1—2X2/879	High Voltage Rectifier
1—80	Low Voltage Rectifier
1—VR105	Voltage Regulator
1—VR150	Voltage Regulator
1—5AP1	Cathode-ray Tube

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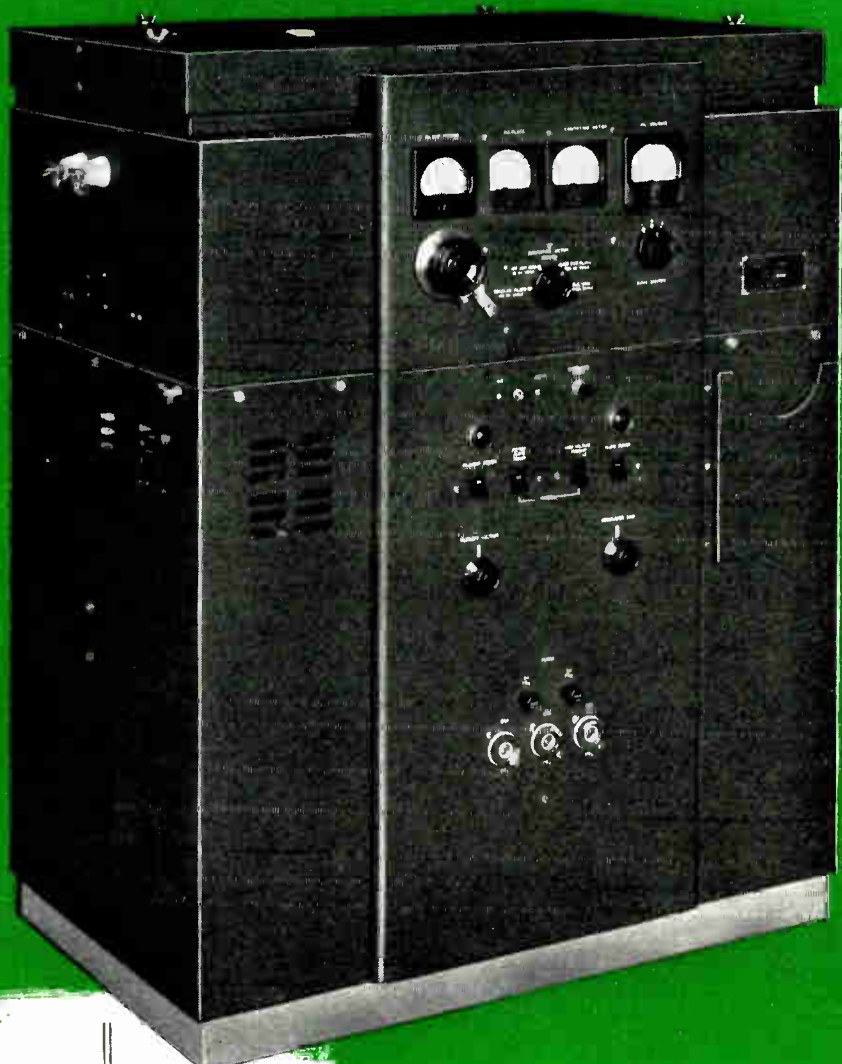


Model HT-4E

TELEPHONE AND TELEGRAPH TRANSMITTER

"The Voice of Victory"

189

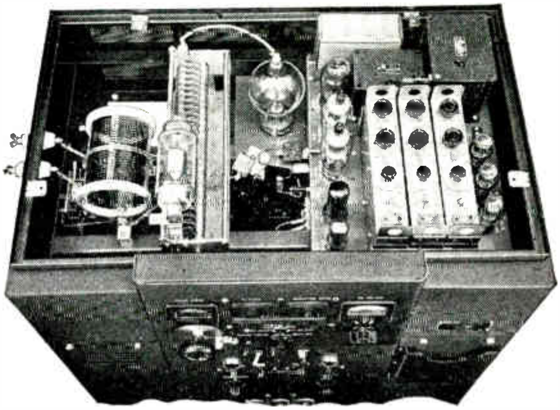
HALLICRAFTERS' Model HT-4E transmitter has the most distinguished war record of any piece of radio communications equipment. First produced several years before Pearl Harbor and designed to meet the requirements of the most exacting amateur operators, the Model HT-4 was selected as the transmitter for the SCR-299 mobile radio station. This famous Signal Corps unit, built by Hallicrafters, has been acclaimed by high military authorities as "the best piece of radio equipment in any army."

The performance of this superb transmitter on every battle front and under the most adverse conditions has become one of the great legends of the war. Originally intended for use as a mobile unit over ranges of a few hundred miles, the SCR-299 so far surpassed expectations that it was soon operating in long distance service over thousands of miles. Commanding officers in the field diverted many of them to use as fixed headquarters stations. SCR-299's were set up as permanent broadcast transmitters in the far corners of the earth, and, dismounted from their trucks, they have been flown into the most remote outposts, there to establish vital communications. All of these outstanding accomplishments were made possible by the sterling performance and rugged construction of the HT-4 and its successors.

Radio operators who were acquainted with the pre-war Model HT-4 are not surprised at its wartime achievements but they will be more than pleased with the many refinements and conveniences now available in the new Model HT-4E. Like other Hallicrafters products, this transmitter has undergone a continuous series of modifications and improvements to cope with the hazards of war and most of these refinements will prove as valuable to the amateur operator as they have to the Signal Corps. Among these wartime changes are: adoption of vacuum padding capacitors for low frequency operation, redesign of exciter tuning units to permit v.f.o. as well as crystal-controlled operation, addition of guide channels to make the insertion of tuning units easy and positive, addition of a remote control relay to switch from phone to c.w. and vice-versa, use of a side-tone oscillator in the speech amplifier to permit monitoring of c.w. transmissions, addition of locking rings to hold tubes firmly in position, slight redesign of cabinet for greater rigidity, and many others.

Refined and strengthened, battle tested under every conceivable hardship, and built by the thousands for service on every continent, the Hallicrafters HT-4E is ready for the reopening of amateur radio.

With the return of peace, this high-power transmitter will again take its place in the leading amateur stations and will once more be heard around the world. The proud owner of a new Hallicrafters HT-4E can rest secure in the knowledge that he has "the best piece of radio equipment" in any amateur station.



FEATURES

1. Coils available for frequency range 1.5 Mc. to 18 Mc.
2. Power output 450 watts c.w., 325 watts phone (continuous operation).
3. Oscillator and buffer stages may be pre-tuned for any three operating frequencies and selected by a panel switch.
4. High level class B modulation.
5. Plug-in pre-tuned r.f. exciter units.
6. Transmitter may be remotely controlled and keyed from speech amplifier.
7. Crystal or v.f.o. operation.
8. All operating controls on front panel.
9. Phone-c.w. operation controlled by single switch.
10. Break-in operation provided for.
11. Metering of all exciter stages and power amplifier grid current through meter switching.
12. All components plainly identified.
13. Voltage regulated oscillator power supply.
14. Optimum LC ratio on all bands due to plug-in vacuum padding condenser.
15. Heavy duty components used throughout.
16. Compact, unit style construction for maximum efficiency.
17. Filament voltage adjustment.
18. Modulator bias adjustment.
19. Filament power switch.
20. Exciter power switch.
21. Plate power switch.
22. High voltage protect switch.
23. Overload reset button.
24. Phone-c.w. switch.
25. Four power supplies.
26. Dual overload relays in high voltage supply.
27. Phone-c.w. relay.
28. Plate power relay.
29. Filament voltmeter on power amplifier.
30. Power amplifier plate current meter.
31. All fuses on front panel.
32. Dial lock on power amplifier tuning.
33. Guides for easy insertion of r.f. exciter units.
34. Tuning chart pocket on panel.
35. Overmodulation limiter on speech amplifier.
36. Modulator plate meter in speech unit for monitoring.
37. Sidetone oscillator (keying monitor).

"With the return of peace, this high-power transmitter will again take its place in the leading amateur stations and will once more be heard around the world. The proud owner of a new Hallicrafters HT-4E can rest secure in the knowledge that he has 'the best piece of radio equipment' in any amateur station."

CONTROLS:

PLATE TUNING; EXCITATION METER SWITCH; BAND SWITCH; CW/PHONE; OVERLOAD RESET; FILAMENT POWER; EXCITER PLATE POWER; HIGH VOLTAGE PROTECT; PLATE POWER; FILAMENT VOLTAGE; MODULATOR BIAS. On speech amplifier: GAIN; MOD. LIMITER; SIDETONE ON/OFF; TRANS. ON; TRANS. OFF.

Note: Three tuning units may be plugged into exciter unit at one time. Each unit has controls for OSCILLATOR, DOUB., and INT. AMP. These are pretuned and the desired channel is selected by the BANDSWITCH.

METERS:

P.A. PLATE; EXCITATION METER; FIL. VOLTAGE; MODULATOR PLATE METER (on speech amplifier.)

EXTERNAL CONNECTIONS:

A.c. plug and cord, antenna terminals, socket for speech amplifier input and power; key and microphone inputs on speech amplifier panel.

PHYSICAL CHARACTERISTICS:

All components of the HT-4E are mounted on heavy steel chasses, finished in gray lacquer. Cabinet is of heavy gauge steel, finished in black wrinkle. Speech amplifier is in its own table model cabinet, finished in black wrinkle.

DIMENSIONS:

Model HT-4E overall: 32⅝" wide by 39⅞" high by 21⅜" deep.

WEIGHTS:

Model HT-4E: 412 pounds.
Packed for shipment: 500 pounds.

TWENTY-THREE TUBES:

1—6V6GT	Crystal or v.f.o. Oscillator
1—6L6	Intermediate Amplifier
2—807	Buffer Amplifiers
1—250TH	R.f. Power Amplifier
3—VR150	Voltage Regulators
2—5Z3	Rectifiers
2—100TH	Class B Modulators
2—2A3	Class B Drivers
2—866	High Voltage Rectifiers
1—6SQ7	Microphone Amplifier
1—6J5	Speech Amplifier
1—6SN7GT	Phase Inverter
1—6SN7GT	Push-pull Output
1—6SR7	Modulation Limiter
1—6SN7GT	Sidetone Generator
1—80	Speech Amplifier Power Supply Rectifier

Speech Amplifier
Model HT-5E

hallicrafters RADIO

ANTENNA TUNING UNITS

THE antenna tuning units shown on this page were designed for use with the Model HT-4E transmitter. With these two units the transmitter can be matched to any type or size of antenna with the maximum possible transfer of energy.

MODEL AT-2 ANTENNA TUNING UNIT

Designed for use with a two wire transmission line, this unit employs the well known pi-section network. Has heavy duty capacitors and ceramic insulated plug-in inductances and is equipped with an antenna changeover relay to permit the use of one antenna for transmitting and receiving.

DIMENSIONS:
Model AT-2 overall: 22″ wide by 9″ high by 16½″ deep.

WEIGHT:
Model AT-2: 35 pounds.
Packed for shipment: 39 pounds.

MODEL AT-3 ANTENNA TUNING UNIT

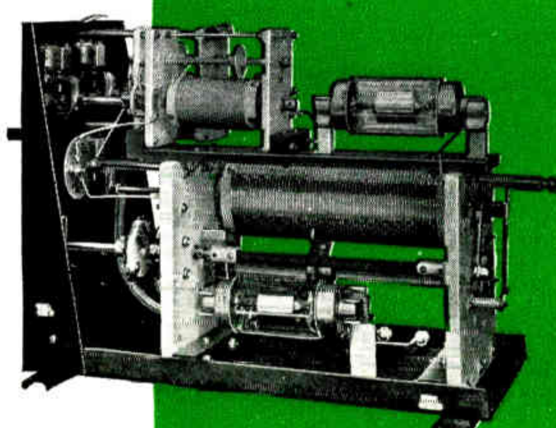
This unit which was used in recent versions of the SCR-299 represents an outstanding achievement in high-frequency design. Covering all frequencies between 1.5 and 18 Mc. without the use of plug-in inductances, the Model AT-3 will tune any single wire antenna from a fifteen foot whip to a long wire. This unit is ceramic insulated to withstand the high r.f. voltages which are generated when antennas are operated far below their fundamental frequencies and will prove invaluable to the operator who is compelled to use an antenna of inadequate size.

DIMENSIONS:
Model AT-3 overall: 10¼″ wide by 14¼″ high by 24″ deep.

WEIGHT:
Model AT-3: 48 pounds.
Packed for shipment: 56 pounds.

AT-2

AT-3



HALLICRAFTERS' Model HT-9 is an ideal medium power transmitter. Designed for maximum flexibility and convenience, it is completely self-contained, requiring only a microphone or key, antenna, and source of 115-volt a.c. power to go on the air.

Five individual plug-in tuning units and crystals may be accommodated in the exciter section simultaneously. Band switching is easily accomplished by changing one coil in the final amplifier and selecting the desired exciter frequency by means of a panel switch. Exciter units are pre-tuned and the only additional operation needed is a slight adjustment of the final tank tuning capacitor.

Separate meters are provided for the power amplifier plate and grid circuits and a third meter may be switched into either the exciter or modulator cathode circuit. All controls are conveniently arranged on the panel and a safety interlock switch is provided for protection against accidental shock when the cabinet is open.

FEATURES

1. Frequency range 1500 kc. to 18 Mc. and amateur 28 Mc. band.
2. Power output 100 watts on c.w., 75 watts on phone.
3. Five operating frequencies may be pre-set in the oscillator and buffer doubler stages and selected at will by means of the bandswitch.
4. 100 percent modulation with low distortion.
5. All operating controls on front panel.
6. Metering of cathode current of exciter or modulator, power amplifier grid, and power amplifier plate.
7. Input for any medium level, high impedance microphone.
8. Carrier hum more than 40 db. below 100% modulation.
9. Frequency response flat within 3 db. from 100 to 5000 cylces.
10. Antenna coil will match any resistive load from 10 to 600 ohms.
11. Line fuses mounted on rear of chassis.
12. Convenient table mounting.
13. Rugged construction and oversize components assure dependable operation.

DRAFTED INTO SERVICE

Model HT-9

TELEPHONE AND TELEGRAPH TRANSMITTER

CONTROLS:
AUDIO GAIN, (speech amplifier) OFF; CATHODE CURRENT EXC.-MOD.; PLATE PWR. ON/OFF; FIL. PWR. ON/OFF; C.W.-PHONE; BAND SWITCH; TRANSMIT-STANDBY; PLATE TUNING.

METERS:
CATHODE CURRENT; P.A. GRID; P.A. PLATE.

EXTERNAL CONNECTIONS:
Antenna terminals. Terminal strip for key, antenna relay, and remote control of receiver. Line cord and plug. Two line fuses. Microphone input connector (on left end of cabinet). All connections except microphone are located on rear of chassis.

PHYSICAL CHARACTERISTICS:
The Model HT-9 is constructed on a heavy cadmium plated steel chassis. Cabinet is of steel finished in black wrinkle lacquer and is provided with heavy rubber mounting feet. Ventilating openings in top and sides assure adequate cooling. Interlock switch under lid cuts high voltage supply when cabinet is opened.

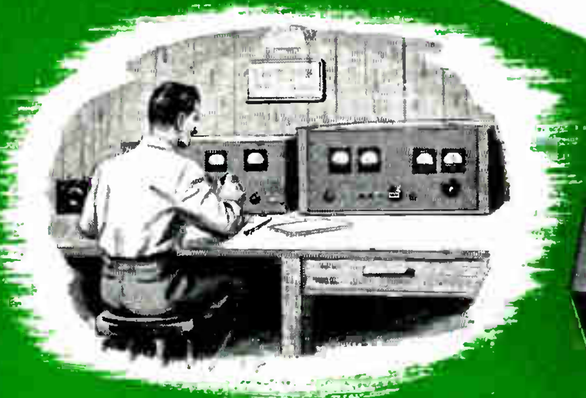
DIMENSIONS:
Model HT-9, overall clearance: 29⅛" wide by 12½" high by 20½" deep.

WEIGHT:
Model HT-9 transmitter: 120 pounds.
Packed for shipment: 125 pounds.

TUNING UNITS:
Final amplifier coils and exciter tuning units are available for the 1.75, 3.5, 7, 14 and 28 Mc. amateur bands. General coverage coils and units for all frequencies between 1.5 and 18 Mc. may be obtained on special order.

FOURTEEN TUBES:

1—6L6	Crystal Oscillator (used above 8 Mc. only)
1—6L6	Crystal Oscillator or Doubler
1—814	Final r.f. Amplifier
1—6SJ7	1st Speech Amplifier
1—6J5	2nd Speech Amplifier
4—6L6	Push-pull Parallel Modulator Stage
2—5Z3	Rectifiers
1—80	Rectifier
2—866	Rectifiers



At the outbreak of war the Army bought up as many HT-9 transmitters as could be located. They were rushed by air to a remote outpost where reliable communications were urgently needed. At last accounts these veteran transmitters were still on the job, helping to extend the lines of victory.

hallicrafters RADIO

Model HT-6

TELEPHONE AND TELEGRAPH TRANSMITTER

FILLING a long felt need for a low cost high performance transmitter, the HT-6 offers most of the desirable features found in Hallicrafters' larger units. Complete sets of coils and crystals for any three bands may be plugged in, pre-tuned, and selected at will by means of a panel switch. All operating controls are conveniently arranged on the front panel. Metering of all circuits is provided by a switch which places the meter in the proper circuit. E.c.o. operation is available at any point in the amateur bands if desired.

A high quality audio system assures complete modulation, and is designed for use with any medium level microphone. Sockets are provided on the rear of the chassis to permit emergency operation from external power supplies.

FEATURES

1. Frequency range amateur bands from 1.7 Mc. to 60 Mc.
2. Normal power output 25 watts, phone or c.w.
3. Three operating frequencies may be pre-set in the transmitter and selected by means of the bandswitch.
4. 100 percent modulation with low distortion.
5. All operating controls on front panel.
6. Metering of all circuits through use of multi-range meter and switch.
7. Input for any medium level, high impedance microphone.
8. Carrier hum more than 40 db. below 100% modulation.
9. Frequency response flat within 3 db. between 125 and 5000 cycles.
10. Antenna coil to match all common resistive loads.
11. Line fuse mounted on rear of chassis.
12. Convenient table mounting.

CONTROLS:

AUDIO GAIN; METER SWITCH; CW/PLATE OFF/PHONE SWITCH; BAND SWITCH; PLATE CIRCUIT TUNING; ON/OFF; TRANSMIT/STANDBY.

EXTERNAL CONNECTIONS:

Antenna terminals. Line cord and plug. Microphone input socket. Remote control socket. Two external power supply sockets.

PHYSICAL CHARACTERISTICS:

All components of the HT-6 are mounted on a rugged gray lacquered steel chassis, and housed in an attractive steel cabinet finished in machine tool gray.

DIMENSIONS:

Model HT-9—20″ wide by 9″ high by 15″ deep.

SHIPPING WEIGHT:

67 pounds.

TUNING UNITS:

Final amplifier coils and exciter tuning units are available for the 1.75, 3.5, 7, 14, 28 and 56 Mc. amateur bands.

NINE TUBES:

1—6J5	Oscillator (56 Mc. band only)
1—6L6	Crystal or e.c.o. Oscillator
1—807	Power Amplifier
1—6SQ7	Speech Amplifier
1—6SC7	Phase Inverter
2—6L6G	Modulators
2—5Z3	Rectifiers

The ENSIGN, *Model* HT-11

MARINE RADIOTELEPHONE—Safety—Convenience—Economy

THE Hallicrafters ENSIGN, Model HT-11 marine radiotelephone provides the safety and convenience of radio communication at a price within the reach of all boat owners. Comprising a 12-watt crystal-controlled transmitter and a five-tube superheterodyne receiver mounted in a single small cabinet, the ENSIGN provides instantaneous ship-to-shore and ship-to-ship radiotelephone communication and excellent broadcast reception. Small enough to find room in boats of any size, the Model HT-11 is designed for complete reliability combined with utmost simplicity of operation.

TRANSMITTER FEATURES

1. Instant selection of any 3 transmitter frequencies, crystal-controlled.
2. Twelve watts output.
3. Transmitter may be used in the range 2000 kc. to 3000 kc.
4. Can be used with any length antenna.
5. Convenient "push to talk" operation.
6. Separate economical low drain power supply.
7. Rust and corrosion protected throughout.
8. Small size for ease of installation.
9. Can be supplied for use with any power source.
10. Panel mounted chart for recording of operating frequencies.

RECEIVER FEATURES

1. Two bands; broadcast 550 kc. to 1700 kc. and marine 2000 kc. to 3000 kc.
2. Receiver output may be switched to speaker or handset.
3. Built in moisture resistant PM speaker.
4. Illuminated, easily read tuning dial.

CONTROLS:

SPEAKER-PHONES; TRANSMITTER FREQUENCY; RECEIVER TUNING; BAND-SWITCH; VOLUME; TRANS. FILS. ON/OFF.
Push-to-talk button on hand-set.

EXTERNAL CONNECTIONS:

Antenna terminal on top of cabinet. Power cable plugs into receptacle at left end of cabinet.

PHYSICAL CHARACTERISTICS:

Both transmitter and receiver components are mounted on a single nickel-plated chassis. Cabinet is finished in gray wrinkle lacquer. Speaker grill and controls are on front of cabinet and handset is permanently connected and carried in cradle at left end of cabinet.

POWER SUPPLY:

Power supplies are available for the following voltages: 6 volts d.c.; 12 volts d.c.; 32 volts d.c.; 115 volts d.c.; 115 volts a.c.

Power supplies are mounted separately and are connected to the Model HT-11 by a cable.

DIMENSIONS:

Cabinet only, 14⅛″ wide by 9⅛″ high by 9¼″ deep.
Overall including handset on cradle 16½″ wide by 10⅛″ high by 10″ deep.
D.c. Power Supply with Cover, 13″ wide by 9½″ high by 8⅜″ deep.
A.c. Power Supply with Cover, 9¼″ wide by 7⅜″ high by 7¾″ deep.

WEIGHT:

Model HT-11—31 pounds.
D.c. power supply—21 pounds.
A.c. power supply—21 pounds.
Add 3 pounds to any of above for shipping weight.

NINE TUBES:

Receiver	
1—6SK7	R.f. Amplifier
1—6K8	1st Detector, Mixer, h.f. Oscillator
1—6SK7	I.f. Amplifier
1—6SQ7	2nd Detector, a.v.c., 1st a.f. Amplifier
1—6K6G	2nd a.f. Amplifier
Transmitter	
1—6V6	Crystal-controlled Oscillator
1—807	R.f. Amplifier Output Stage
2—6V6	Push-pull Modulator Stage

The COMMODORE, *Model* HT-14

MARINE RADIOTELEPHONE

Dependable Communications on the High Seas

THE new Hallicrafters COMMODORE, Model HT-14 Marine Radiotelephone incorporates every feature experience has shown desirable for ship-to-shore and ship-to-ship telephone service. A commercial adaptation of the famous Hallicrafters-built SCR-543, the HT-14 basic design has been literally "battle tested." With 6 crystal-controlled channels selected simultaneously in both transmitter and receiver and an output of 45 watts capable of 100 percent amplitude modulation, the HT-14 is an ideal medium power marine radiotelephone.

TRANSMITTER FEATURES

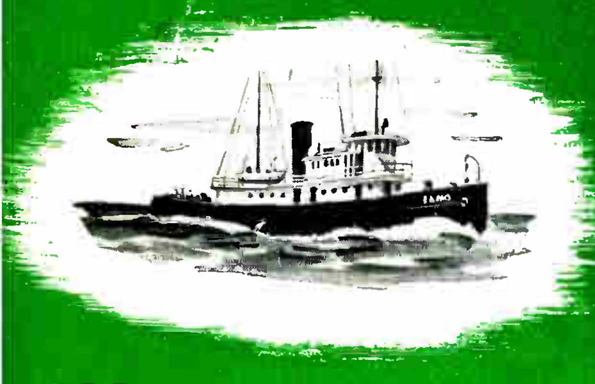
1. Instant selection of any 6 operating frequencies, crystal-controlled in both transmitter and receiver.
2. 45 watts output.
3. Frequency range, 1680 to 4450 kc.
4. Any antenna from 15 feet to a long wire may be used.
5. "Push-to-talk" switch on handset.
6. All components rust and corrosion resistant.
7. Metering of antenna current, final amplifier grid and plate, and modulator plate provided.
8. Entire unit easily removable for ease of servicing.
9. May be operated from 115 volts a.c., 12, 32 or 115 volts d.c.
10. Chart mounted on panel for recording of operating frequencies.
11. All operating adjustments may be made at front of unit.

RECEIVER FEATURES

1. Two ranges; 1680 kc. to 2750 kc. and 2750 to 4450 kc., either crystal controlled or manually tuned.
2. Crystal receiver frequencies switched simultaneously with those of the transmitter.
3. Iron core, high-Q inductances used in the r.f., detector, and oscillator circuits provide maximum gain.
4. Exceptionally flat automatic volume control.
5. Newly developed diode noise limiter and audio filter circuit.
6. Receiver output may be used on handset or speaker.
7. 5" PM speaker with moisture resistant cone.

CONTROLS:

OPERATING CHANNEL SWITCH (6 positions); TRANSMITTER ANTENNA TUNING; RECEIVER TUNING; RECEIVER BAND SWITCH; A.F. GAIN; NOISE CONTROL; STATIC FILTER ON/OFF; SPEAKER ON/OFF; METER SWITCH; RECEIVER POWER ON/OFF; TRANSMITTER POWER ON/OFF; SEND-RECEIVE SWITCH (located on hand set, thumb operated).



METERS:
Antenna current ammeter is of the thermocouple type and is flush mounted on the upper panel. Range, 0 to 2½ amperes. A dual range d.c. milliammeter 0-15-300 m.a. is mounted on the lower panel and can be connected to read final amplifier plate current, final amplifier grid current, and modulator plate current.

CONNECTIONS:
The antenna connector is mounted on a standoff insulator on top of cabinet. Handset plugs into a receptacle at lower left corner of cabinet. The cable to the power supply unit plugs into a socket at lower right corner of cabinet. The power supply has a line cord for connection to the 115-volt a.c. supply line. The steel cabinet should be connected to a good ground.

ANTENNA REQUIREMENTS:
The Model HT-14 Radiotelephone is designed to operate with any antenna from a 15-foot whip to a long wire. For maximum transmitting range, the antenna should be large and as high above water as possible. With single-masted boats, an insulated forestay makes a satisfactory antenna. On boats having two masts, the antenna should be supported between the mast-heads and may consist of one or more wires.

INSTALLATION:
A universal type of shock mounting is furnished with the HT-14 permitting installation either on a bulkhead or table. Special screw type fasteners hold the HT-14 to the shock mounting and permit its easy removal for servicing.

PHYSICAL CHARACTERISTICS:
The HT-14 Radiotelephone is mounted in a steel cabinet. The cabinet is divided into 2 sections which are held together by heavy clamps. The upper section contains the radio frequency components of both transmitter and receiver. The lower section holds the speech amplifier and modulator. The loud speaker is mounted in the center of the lower panel and the handset is hung in a bracket at the left. All operating controls and switches are conveniently placed. The power supply unit is mounted in a separate cabinet.

POWER SUPPLY:
Power supply combinations for use on four different voltages are available. The 115-volt a.c. power supply unit is mounted in a separate cabinet. The 32-volt (or 110-volt) d.c. models include a 32-volt (or 110-volt) d.c. rotary converter which supplies power to the 115-volt a.c. power supply unit. The 12-volt d.c. model includes a 12-volt dynamotor type power supply unit, instead of the 115-volt a.c. power supply unit, in a cabinet of the same dimensions.

DIMENSIONS (overall):
Main cabinet: 23" high by 21" wide by 16¼" deep.
Power supply cabinet. 9⅜" high by 16" wide by 15" deep.
These measurements include protruding parts.
Note: Shock mounts add 2¾" to height or depth according to type of installation.

WEIGHTS:
Main cabinet—110 lbs.
115-volt a.c. Power Supply—67 lbs.
Combined shipping weight—275 lbs.
For d.c. operated models, add 55 lbs. to shipping weight.

TWENTY TUBES:

Transmitter

1—6L6G	Crystal Oscillator
2—807	R.f. Amplifier
1—12J5GT	Speech Amplifier
4—6L6G	Modulator

Receiver

1—6SK7-GT G	R.f. Amplifier
1—6SA7-GT G	First Detector
1—6SK7-GT G	I.f. Amplifier
1—6H6-GT G	Second Detector, a.v.c. and Noise Limiter
1—6SK7GT G	First Audio Amplifier
1—6K6-GT G	Second Audio Amplifier
1—6J5-GT G	High Frequency Oscillator

Power Supply

1—80	Rectifier (for receiver)
4—5Z3	Rectifiers (for transmitter)

In War...

Men and women of Hallicrafters can lay claim to a distinguished production record in the service of their country during the war emergency. Up to the time this handbook went to press in the Fall of '44, Hallicrafters workers were privileged to receive four separate Army-Navy "E" awards. Another star was added to this service record when the U. S. Navy in one of the first recognitions of its kind awarded Hallicrafters workers a Certificate of Achievement for untiring efforts in helping the electronics industry speed the production of vital war material.

And in Peace...

You can be assured that the same degree of perfection and high quality of workmanship that earned these numerous citations will be carried over into peacetime production.

hallicrafters RADIO

THE HALLICRAFTERS COMPANY, MANUFACTURERS OF RADIO AND ELECTRONIC EQUIPMENT, CHICAGO 16, U.S.A.

Available to radio clubs or other interested parties, Hallicrafters sound film, "The Voice of Victory," showing how the HT-4 went to war. In either 16 mm. or 35 mm. size. Loaned without charge. Write the advertising department.

The American Radio Relay League

THE American Radio Relay League, Inc., is a non-commercial association of radio amateurs, bonded for the promotion of interest in amateur radio communication and experimentation, for the relaying of messages by radio, for the advancement of the radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

It is an incorporated association without capital stock, chartered under the laws of Connecticut. Its affairs are governed by a Board of Directors, elected every two years by the general membership. The officers are elected or appointed by the Directors. The League is non-commercial and no one commercially engaged in the manufacture, sale or rental of radio apparatus is eligible to membership on its board.

"Of, by and for the amateur," it numbers within its ranks practically every worth-while amateur in the nation and has a history of glorious achievement as the standard-bearer in amateur affairs.

Inquiries regarding membership are solicited. A bona fide interest in amateur radio is the only essential qualification; ownership of a transmitting station and knowledge of the code are not prerequisite.

Membership Application Blank WILL BE FOUND ON BACK OF THIS PAGE →

Application for Membership

AMERICAN RADIO RELAY LEAGUE

***Administrative Headquarters:* West Hartford 7, Conn., U. S. A.**

..............................194...

AMERICAN RADIO RELAY LEAGUE,
West Hartford, Conn., U. S. A.

Being genuinely interested in Amateur Radio, I hereby apply for membership in the American Radio Relay League, and enclose $2.50 ($3.00 in foreign countries) in payment of one year's dues*, $1.25 of which is for a subscription to QST for the same period. Please begin my subscription with the............... issue.

The call of my station is........................

The class of my operator's license is..............

I belong to the following radio societies...........

....

Send my Certificate of Membership ☐ or Membership Card ☐ (indicate which) to the address below:

Name..

....

....

A bona fide interest in amateur radio is the only essential requirement but full voting membership is granted only to licensed radio amateurs of the United States and Canada. Therefore, if you have a license, please be sure to indicate it above.

[*The dues are $2.50 per year in the United States and Possessions. All other countries $3.00 per year.]

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