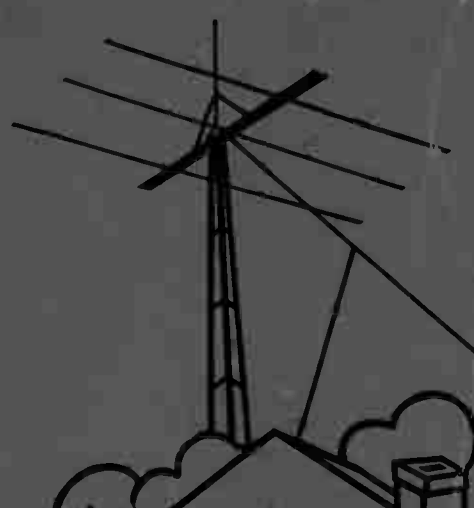
26TH EDITION · 1949

The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION



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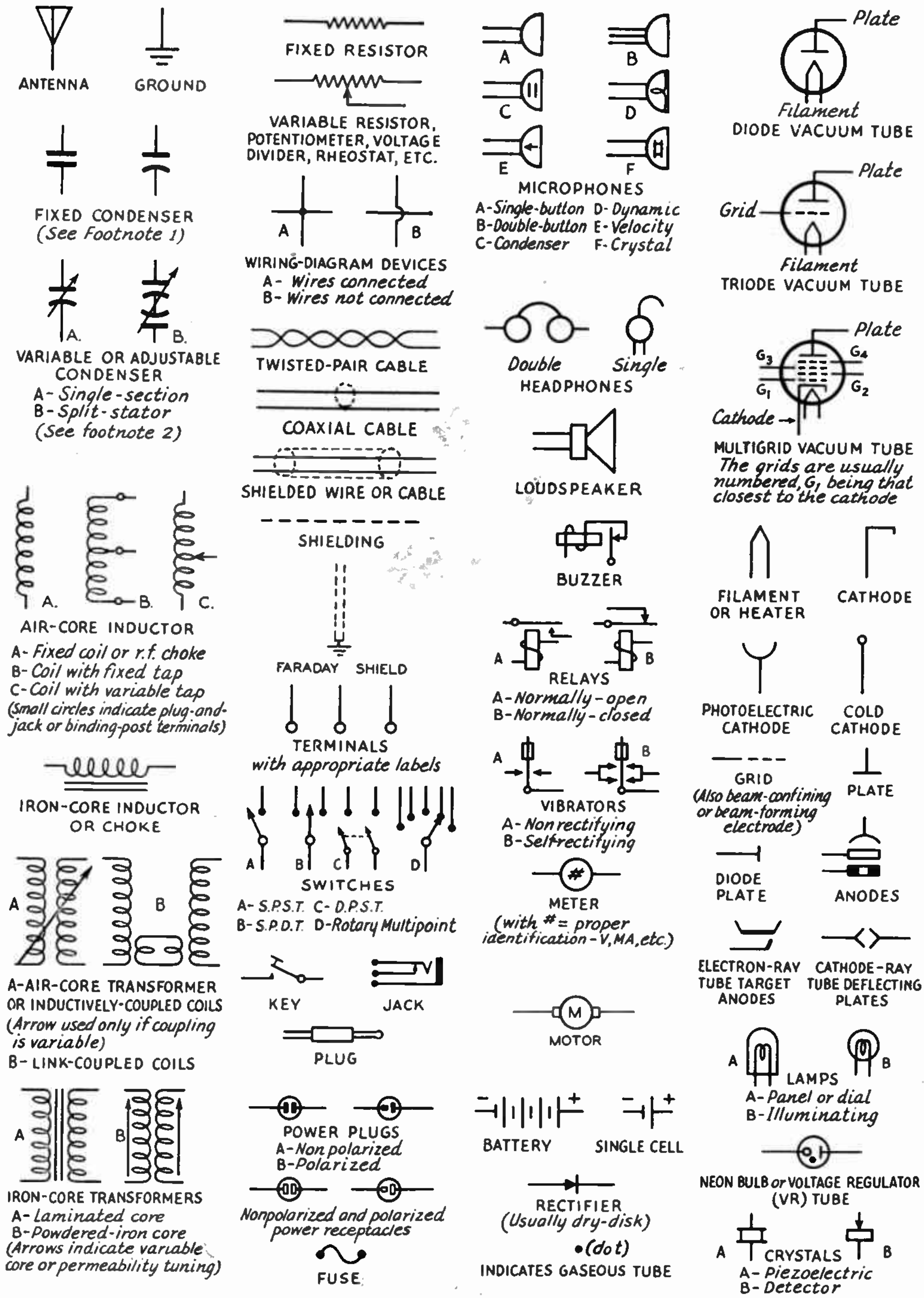
PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

THE RADIO AMATEUR'S HANDBOOK

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



[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 modern symbol, the curved line indicates the moving element (rotor plates) in variable and adjustable air- or mica-dielectric condensers.

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.

THE RADIO AMATEUR'S HANDBOOK

by the

HEADQUARTERS

STAFF

of the

AMERICAN

RADIO

RELAY

LEAGUE

West Hartford

Connecticut, U.S.A.

ARRL

1949 Twenty-Sixth

Edition

Twenty-Sixth Edition

(Of the previous twenty-five editions, 1,928,250 copies were published.)

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FOREWORD

With the appearance of this twenty-sixth edition *The Radio Amateur's Handbook* attains an all-time circulation figure over the two million mark, exemplifying the book's wide acceptance during its twenty-three years of continuous publication. Produced primarily for amateurs by the staff of the amateur's own organization, The American Radio Relay League, the *Handbook* has also won universal acceptance in other segments of the technical radio world — engineering, servicing, operating and educating. This wide dependence on the *Handbook* is founded on its practical utility, its treatment of radio communication problems in terms of how-to-do-it rather than by abstract discussion and abstruse formulas.

Synchronized to the needs of a fast-moving and progressive science, continuous revision has always been a feature of the *Handbook* — always with the objective of presenting the soundest and best aspects of current amateur practice rather than merely the new and novel.

In contrast to most publications of a comparable nature, the *Handbook* is printed in the convenient format and understandable language of the League's monthly magazine, *QST*. This, together with extensive and usefully-appropriate catalog advertising by reputable manufacturers producing equipment for radio amateurs, makes it possible to distribute for a very modest charge a work which in volume of subject matter and profusion of illustration surpasses most available radio texts selling for several times its price.

This twenty-sixth edition of the *Handbook* carries on the plan so successfully inaugurated in the preceding edition. Emphasis is again placed not on textbook style but instead on simpler and more understandable discussion of the facts that an amateur should know to get the most out of designing and using his apparatus; related theory and practice are arranged to complement each other. A large assortment of new equipment has been added to the receiver and transmitter chapters, gear attuned to the technical requirements imposed by operation in heavily-populated amateur bands and in neighborhoods sprouting television lookers-in. Technical progress on the ultrahighs has not been neglected either, a large number of new units being described in the u.h.f. chapters. As a final touch, the Communications Department sections have been thoroughly revised to conform to present approved standards of station operation.

It is sincerely hoped that this new edition will succeed in bringing as much assistance and inspiration to amateurs and newcomers to the hobby as have its many predecessors.

A. L. BUDLONG
Acting Secretary, A.R.R.L.

West Hartford, Conn.
December, 1948

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THE AMATEUR'S CODE

ONE 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.

TWO THE AMATEUR IS LOYAL . . . He owes his amateur radio to the American Radio Relay League, and he offers it his unswerving loyalty.

THREE THE AMATEUR IS PROGRESSIVE . . . He keeps his station abreast of science. It is built well and efficiently. His operating practice is clean and regular.

FOUR THE AMATEUR IS FRIENDLY . . . Slow and patient sending when requested, friendly advice and counsel to the beginner, kindly assistance and cooperation for the broadcast listener; these are marks of the amateur spirit.

FIVE 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.

SIX THE AMATEUR IS PATRIOTIC . . . His knowledge and his station are always ready for the service of his country and his community.

Electrical Laws and Circuits

Everyone knows that radio is electrical in nature, and it is taken for granted that to know anything about the operation of radio equipment you have first to know something about electricity and electrical circuits. The amount of electrical knowledge you need in amateur radio depends on how far you delve into the technicalities of the various types of transmitters, receivers and measuring equipment that amateurs use. If you're just getting started you do not need very much, but as you progress you will find that you will acquire, more or less unconsciously, a great deal of basic information. That is, you will if you make a conscientious effort to understand and analyze the things that you observe in using radio gear.

The purpose of this chapter is to provide the answers to many questions about circuits that will come up in the course of building and operating an amateur station. It is intended as a *practical* reference section rather than a course in "theory." You can study it consecutively if you wish, of course. However, it should be even more valuable to you in showing how everyday problems can be solved when the occasion to solve them arises.

Fundamentals

ELECTRIC AND MAGNETIC FIELDS

At the bottom of everything in electricity and radio is a **field.** Although a field is not too easy to visualize, we need to have some appreciation of what it is if electrical effects are to be understood. When something occurs at one point in space because something else happened at another point, with no visible means by which the "cause" can be related to the "effect," we say the two events are connected by a "field." It does not matter whether or not the field is "real" — that is, whether it is something physical although, like air, invisible. The important point is that the distant effects are *predictable*, and it is convenient to attribute them to properties of a field. The fields with which we are concerned are the **electric** and **magnetic,** and the combination of the two called the **electromagnetic** field.

A field has two important properties, *intensity* (*magnitude*) and *direction*. That is, the field exerts a *force* on an object immersed in it; intensity measures the amount of force exerted while direction tells the direction in which the object on which the force is exerted will tend to move. An electrically-charged object in an electric field will be acted on by a force that will tend to move it in a direction determined by the direction of the field. Similarly, a magnet in a magnetic field will be subject to a force. Everyone has seen demonstrations of magnetic fields with pocket magnets, so intensity and direction are not hard to grasp.

A "static" field is one that is fixed in space. Such a field can be set up by a stationary electric charge **(electrostatic field)** or by a stationary magnet **(magnetostatic field).** But if either an electric or magnetic field is moving in space or changing in intensity, the motion or change sets up the other kind of field. That is, a changing electric field sets up a magnetic field, and a changing magnetic field generates an electric field. This interrelationship between magnetic and electric fields makes possible such things as the electromagnet and the electric motor. It also makes possible the **electromagnetic waves** by which radio communication is carried on, for such waves are simply traveling fields in which the energy is alternately handed back and forth between the electric and magnetic fields.

Lines of Force

We need, obviously, some way to compare the intensity and direction of different fields. This is done by picturing the field as made up of **lines of force,** or **flux lines.** These are purely imaginary threads that show, by the direction in which they lie, the direction the object on which the force is exerted will move. The *number* of lines in a chosen cross section of the field is a measure of the *intensity* of the force. The number of lines per square inch, or per square centimeter, is called the **flux density.**

ELECTRICITY AND THE ELECTRIC CURRENT

Electrical effects are caused by extremely small particles of electricity called **electrons.** Everything physical is built up of atoms, particles so small that they cannot be seen even through the most powerful microscope. But the atom in turn consists of still smaller particles — several different kinds of them. One type of particle is the electron. An ordinary atom consists of a central core, called the **nucleus,** around which one or more electrons circulate somewhat as the earth and other planets circulate around the sun. Both the nucleus and the electrons are electrical, but the kind of electricity associated with the nucleus is called **positive** and that associated with the electrons is called **negative.**

The important fact about these two "opposite" kinds of electricity is that they are strongly attracted to each other. Also, there is a strong force of repulsion between two charges (a collection of electrified particles is called a **charge**) of the *same* kind. The positive nucleus and the negative electrons are attracted to each other, but two electrons will be repelled from each other and so will two nuclei. The fact that an atom contains both positive and negative charges makes it tend to stay together as a unit; in a normal atom the positive charge on the nucleus is exactly balanced by the total of the negative charges on the electrons. It *is* possible, though, for an atom to lose one of its electrons; when that happens the atom has a little less negative charge than it should — or, to put it another way, it has a net positive charge. Such an atom is said to be **ionized,** and in this case the atom is a **positive ion.** If an atom picks up an extra electron, as it sometimes does, it has a net negative charge and is called a **negative ion.** A positive ion will attract any stray electron in the vicinity, including the extra one that may be attached to a nearby negative ion. In this way it is conveniently possible for electrons to travel from atom to atom, and when such movement occurs on a measurable scale (millions or billions of electrons moving) we have a detectable **electric current.**

Conductors and Insulators

The movement of electrons can take place in a solid, a liquid, or a gas. In liquids and gases, positive and negative ions, as well, are free to move when attracted electrically, but in solids only the electrons move. However, movement of electrons or ions is not possible in all substances. Atoms of some materials, notably metals and acids, will give up an electron readily, but atoms of other materials will not part with any of their electrons even when the electric force is extremely strong. Materials in which electrons or ions can be moved with relative ease are called **conductors,** while those that refuse to permit such movement are called **nonconductors** or **insulators.** The following listing shows how some common materials divide between the conductor and insulator classifications:

Conductors	*Insulators*
Metals	Dry Air
Carbon	Wood
Acids	Porcelain
	Textiles
	Glass
	Rubber
	Resins

Electromotive Force

The electric force (called **electromotive force,** and abbreviated **e.m.f.**) that causes current flow may be developed in several ways. The action of certain chemical solutions on dissimilar metals sets up an e.m.f.; such a combination is called a **cell,** and a group of cells forms an electric **battery.** The amount of current that such cells can carry is limited, and in the course of current flow one of the metals is eaten away. The amount of electrical energy that can be taken from a battery consequently is rather small. Where a large amount of energy is needed it is usually furnished by an electric **generator,** which develops its e.m.f. by a combination of magnetic and mechanical means. Large generators in power houses supply the energy that is distributed to homes and factories.

In picturing current flow it is natural to think of a single, constant force causing the electrons to move. When this is so, the elec-

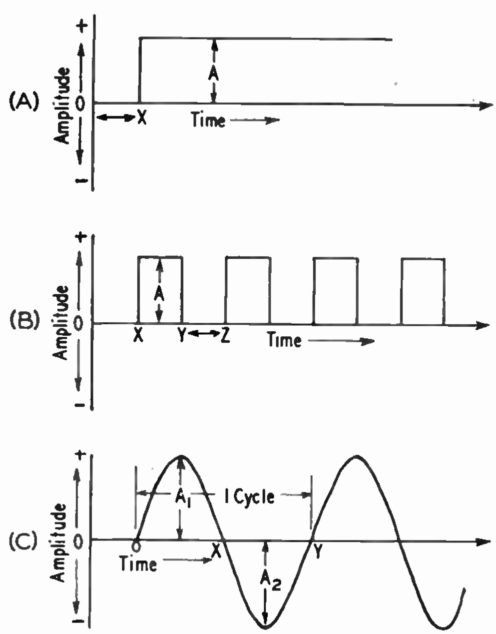


Fig. 2-1 — Three types of current flow. A — direct current; B — intermittent direct current; C — alternating current.

trons always move in the same direction through a path or circuit made up of conductors connected together in a continuous chain. Such a current is called a **direct current**, abbreviated **d.c.** It is the type of current furnished by batteries and by certain types of generators. However, it is also possible — and desirable as well — to have an e.m.f. that periodically reverses. With this kind of e.m.f. the current flows first in one direction through the circuit and then in the other. Such an e.m.f. is called an **alternating** e.m.f., and the current is called an **alternating current** (abbreviated **a.c.**). The reversals **(alternations)** may occur at any rate from a few per second up to several billion per second. Two reversals make a **cycle**; in one cycle the force acts first in one direction, then in the other, and then returns to the first direction. The number of cycles in one second is called the **frequency** of the alternating current.

Direct and Alternating Currents

The difference between direct current and alternating current is shown in Fig. 2-1. In these graphs the horizontal axis measures time, increasing toward the right away from the vertical axis. The vertical axis represents the amplitude or size of the current, increasing in either the up or down direction away from the horizontal axis. If the graph is *above* the horizontal axis the current is flowing in one direction through the circuit (indicated by the + sign) and if it is *below* the horizontal axis the current is flowing in the reverse direction through the circuit (indicated by the − sign). Fig. 2-1A shows that, if we close the circuit — that is, make the path for the current complete — at the time indicated by X, the current instantly takes the amplitude indicated by the height A. After that, the current continues at the same amplitude as time goes on. This is an ordinary *direct* current.

In Fig. 2-1B, the current starts flowing with the amplitude A at time X, continues at that amplitude until time Y and then instantly ceases. After an interval YZ the current again begins to flow and the same sort of start-and-stop performance is repeated. This is an *intermittent* direct current. We could get it by alternately closing and opening a switch in the circuit. It is a *direct* current because the *direction* of current flow does not change; the graph is always on the + side of the horizontal axis.

In Fig. 2-1C the current starts at zero, increases in amplitude as time goes on until it reaches the amplitude A_1 while flowing in the + direction, then decreases until it drops to zero amplitude once more. At that time (X) the *direction* of the current flow reverses; this is indicated by the fact that the next part of the graph is below the axis. As time goes on the amplitude increases, with the current now flowing in the − direction, until it reaches amplitude A_2. Then the amplitude decreases until finally it drops to zero (Y) and the direction reverses once more. This is an *alternating* current.

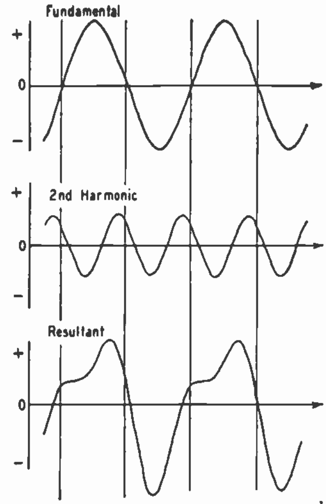


Fig. 2-2 — A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have opposite polarities, the resultant is the *difference*; if the negative-polarity component is larger, the resultant is negative at that instant.

Waveforms

The graph of the alternating current is what is known as a **sine wave**. Sine-wave alternating current is the simplest — but not the only — kind. Notice that the variations in amplitude are quite regular and that the "negative" half-cycle or alternation is exactly like the "positive" half-cycle except for the reversal of direction. The variations in many a.c. waves are not so smooth, nor is one half-cycle necessarily just like the preceding one in shape. However, these more complex waves actually can be shown to be the sum of two or more sine waves of frequencies that are exact integral (whole-number) multiples of some lower frequency. The lowest frequency is called the **fundamental** frequency, and the higher frequencies (2 times, 3 times the fundamental frequency, and so on) are called **harmonics.**

Fig. 2-2 shows how a fundamental and a second harmonic (twice the fundamental) might add to form a complex wave. A little thought will show that simply by changing the relative amplitudes of the two waves, as well as the times at which they pass through zero amplitude, an infinite number of waveshapes can be constructed from just a fundamental and second harmonic. Waves that are still more complex can be constructed if more than two harmonics are used.

Electrical Units

The unit of electromotive force is called the **volt**. An ordinary flashlight cell generates an e.m.f. of about 1.5 volts. The e.m.f. commonly supplied for domestic lighting and power is 115 volts, usually a.c. having a frequency of 60 cycles per second. The voltages used in radio receiving and transmitting circuits range from a few volts (usually a.c.) for filament heating to as high as a few thousand d.c. volts for the operation of power tubes.

The flow of electric current is measured in **amperes**. One ampere is equivalent to the movement of many billions of electrons past a point in the circuit in one second. Currents in the neighborhood of an ampere are required for heating the filaments of small power tubes. The *direct* currents used in amateur radio equipment usually are not so large, and it is customary to measure such currents in **milliamperes**. One milliampere is equal to one one-thousandth of an ampere, or 1000 milliamperes equals one ampere.

In assigning a value to an alternating current or voltage, it is necessary to take into account the difference between direct and alternating currents. A "d.c. ampere" is a measure of a *steady* current, but the "a.c. ampere" must measure a current that is continually varying in amplitude and periodically reversing direction. To put the two on the same basis, an a.c. ampere is defined as the amount of current that will cause the same heating effect (see later section) as one ampere of steady direct current. For a sine-wave alternating current, this **effective** (or **r.m.s.**) value is equal to the *maximum* amplitude of the current (A_1 or A_2 in Fig. 2-1C) multiplied by 0.707. The **instantaneous** value of an alternating current is the value that the current measures at any selected instant in the cycle.

If all the instantaneous values in a sine-wave alternating current are averaged over a *half*-cycle, the resulting figure is the **average** value of the alternating current. It is equal to 0.636 times the maximum amplitude. The average value is useful in connection with rectifier systems, as described in a later chapter.

These definitions of units apply to a.c. voltage as well as to current.

FREQUENCY AND WAVELENGTH

Frequency Spectrum

The electrical energy supplied for household use usually has a frequency of 60 cycles per second. Frequencies ranging from about 15 to 15,000 cycles per second are called **audio frequencies**, because the vibrations of air particles that our ears recognize as sounds occur at the same rate. Audio frequencies (abbreviated a.f.) are used to actuate loudspeakers and thus create sound waves.

Frequencies above about 15,000 cycles are called **radio frequencies (r.f.)** because they are useful in radio transmission. Frequencies all the way up to and beyond 10,000,000,000 cycles have been used for radio purposes. At radio frequencies the numbers become so large that it becomes convenient to use a larger unit than the cycle. Two such units in everyday use are the **kilocycle**, which is equal to 1000 cycles and is abbreviated **kc.**, and the **megacycle**, which is equal to 1,000,000 cycles or 1,000 kilocycles and is abbreviated **Mc.** The accompanying table shows how to convert frequencies expressed in one unit into frequencies in another unit.

The various radio frequencies are divided off into classifications for ready identification. These classifications, listed below, constitute the **frequency spectrum** so far as it extends for radio purposes at the present time.

Frequency	*Classification*	*Abbreviation*
10 to 30 kc.	Very-low frequencies	v.l.f.
30 to 300 kc.	Low frequencies	l.f.
300 to 3000 kc.	Medium frequencies	m.f.
3 to 30 Mc.	High frequencies	h.f.
30 to 300 Mc.	Very-high frequencies	v.h.f.
300 to 3000 Mc.	Ultrahigh frequencies	u.h.f.
3000 to 30,000 Mc.	Superhigh frequencies	s.h.f.

Wavelength

We said earlier that radio waves are traveling fields of electric and magnetic force. These fields travel at great speed — so great that, so far as we can observe, "cause" and "effect" are simultaneous. Nevertheless, it does take a definite amount of time for the effect of a field set up at one point to be felt at a point some distance away.

Radio waves travel at the same speed as light — 300,000,000 meters or about 186,000 miles a second. They are always set up by a radio-frequency current flowing in a circuit, because the rapidly-changing current sets up a magnetic field that changes in the same way, and the varying magnetic field in turn sets up a varying electric field. And whenever this happens, the two fields move outward at the speed of light.

Suppose our r.f. current has a frequency of 3,000,000 cycles per second. The fields, then, will go through complete reversals (one cycle) in 1/3,000,000 second. In that same period of time the fields — that is, the wave — will move 300,000,000/3,000,000 meters, or 100 meters. (The meter is the unit of length commonly used in all sciences. We could use miles, feet, or inches, though, if those units were more convenient.) By the time the wave has moved that distance the next cycle has begun and a new wave has started out. The first wave, in other words, covers a distance of 100 meters before the beginning of the next, and so on. This distance is the "length" of the wave, or **wavelength**.

The longer the time of one cycle — that is, the lower the frequency — the greater the distance occupied by each wave and hence the longer the wavelength. The relationship between wavelength and frequency is shown by the formula

$$\lambda = \frac{300{,}000}{f}$$

where λ = Wavelength in meters
f = Frequency in kilocycles

or

$$\lambda = \frac{300}{f}$$

where λ = Wavelength in meters
f = Frequency in megacycles

Example: The wavelength corresponding to a frequency of 3650 kilocycles is

$$\lambda = \frac{300{,}000}{3650} = 82.2 \text{ meters}$$

Most of our dealings are with frequency, if for no other reason than that it can be measured much more accurately than wavelength. However, we cannot ignore wavelength; it enters into the calculation of the size of "linear" circuits such as antennas.

Resistance

The ease with which we can force an electric current through a conductor varies with the material, shape and dimensions of the conductor. Given two conductors of the same size and shape, but of different materials, the amount of current that will flow when a given e.m.f. is applied to the conductor will be found to vary with what is called the **resistance** of the material. The lower the resistance, the greater the current for a given value of e.m.f.

Resistance is measured in **ohms.** A circuit has a resistance of one ohm when an applied e.m.f. of one volt causes a current of one ampere to flow. The **resistivity** of a material is the resistance, in ohms, of a cube of the material measuring one centimeter on each edge. One of the best conductors is copper, which is why this metal is so widely used in electrical circuits. It is frequently convenient, in making resistance calculations, to compare the resistance of the material under consideration with that of a copper conductor of the same size and shape; Table 2-I gives the ratio of the resistivity of the material to that of copper.

The longer the path through which the current flows the higher the resistance of that conductor. For direct current and low-frequency alternating currents (up to a few thousand cycles per second) the resistance is *inversely* proportional to the cross-sectional area of the path the current must travel; that is, given two conductors of the same material and having the same length, but differing in cross-sectional area, the one with the larger area will have the lower resistance.

Resistance of Wires

It is readily possible to combine all these statements about resistance in a single formula that would enable us to calculate the resistance of conductors of any size, shape and material. However, in most practical cases the problem will be to determine the resistance of a round wire of given diameter and length — or its opposite: finding a suitable size and length of wire to supply a desired amount of resistance. Such problems can be easily solved with the help of the information in the copper-wire table in Chapter Twenty-Four. This table gives the resistance, in ohms per thousand feet, of each standard wire size.

Example: Suppose a resistance of 3.5 ohms is needed and some No. 28 wire is on hand. The wire table in Chapter 24 shows that No. 28 has a resistance of 66.17 ohms per thousand feet. Since the desired resistance is 3.5 ohms, the length of wire required will be

$$\frac{3.5}{66.17} \times 1000 = 52.89 \text{ feet.}$$

Or, suppose that the resistance of the wire in the circuit must not exceed 0.05 ohm and that the length of wire required for making the connections totals 14 feet. Then

$$\frac{14}{1000} \times R = 0.05 \text{ ohm}$$

where R is the maximum allowable resistance in ohms per thousand feet. Rearranging the formula gives

$$R = \frac{0.05 \times 1000}{14} = 3.57 \text{ ohms/1000 ft.}$$

Reference to the wire table shows that No. 15 is the smallest size having a resistance less than this value.

When the wire is not copper, the resistance values given in the wire table in Chapter Twenty-Four should be multiplied by the ratios given in Table 2-1 to obtain the resistance.

Example: If the wire in the first example were iron instead of copper the length required for 3.5 ohms would be

$$\frac{3.5}{66.17 \times 5.65} \times 1000 = 9.35 \text{ feet.}$$

Temperature Effects

The resistance of a conductor changes with its temperature. Although it is seldom necessary to consider temperature in making the

TABLE 2-I
Relative Resistivity of Metals

Material	*Resistivity Compared to Copper*
Aluminum (pure)	1.70
Brass	3.57
Cadmium	5.26
Chromium	1.82
Copper (hard-drawn)	1.12
Copper (annealed)	1.00
Iron (pure)	5.65
Lead	14.3
Nickel	6.25 to 8.33
Phosphor Bronze	2.78
Silver	0.94
Tin	7.70
Zinc	3.54

resistance calculations required in amateur work, it is well to know that the resistance of practically all metallic conductors increases with increasing temperature. Carbon, however, acts in the opposite way; its resistance *decreases* when its temperature rises. The temperature effect is important when it is necessary to maintain a constant resistance under all conditions. Special materials that have little or no change in resistance over a wide temperature range are used in that case.

Resistors

Resistance has important uses in electrical and radio circuits. A "package" of resistance made up into a single unit is called a **resistor.** Resistors having the same resistance value may be considerably different in size and construction. The flow of current through resistance causes the conductor to become heated; the higher the resistance and the larger the current, the greater the amount of heat developed. Consequently, high-resistance resistors intended for carrying large currents must be physically large so the heat can be radiated quickly to the surrounding air. If the resistor does not get rid of its heat quickly it might reach a temperature that would cause it to melt or burn. Types of resistors used in radio circuits are shown in the photograph.

Conductance

The reciprocal of resistance (that is, $1/R$) is called **conductance.** It is usually represented by the symbol G, and the higher its value the greater the conductivity of the circuit. A circuit having large conductance has low resistance, and vice versa. In radio work the term is used chiefly in connection with vacuum-tube characteristics. The unit of conductance is the **mho.** A resistance of one ohm has a conductance of one mho, a resistance of 1000 ohms has a conductance of 0.001 mho, and so on. A unit frequently used in connection with vacuum tubes is the **micromho,** or one-millionth of a mho. It is the conductance of a resistance of one megohm.

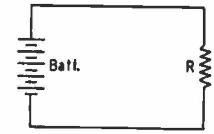


Fig. 2-3 — A simple circuit consisting of a battery and resistor.

OHM'S LAW

The simplest form of electric circuit is a battery with a resistance connected to its terminals, as shown by the symbols in Fig. 2-3. A complete circuit must have an unbroken path so current can flow out of the battery, through the apparatus connected to it, and back into the battery. The circuit is **broken,** or **open,** if a connection is removed at any point. A **switch** is a device for making and breaking connections and thereby closing or opening the circuit, either allowing current to flow or preventing it from flowing.

The values of current, voltage and resistance in a circuit are by no means independent of each other. The relationship between them is known as **Ohm's Law.** It can be stated as follows: The current flowing in a circuit is directly proportional to the applied e.m.f. and inversely proportional to the resistance. Expressed as an equation, it is

$$I \text{ (amperes)} = \frac{E \text{ (volts)}}{R \text{ (ohms)}}$$

The equation above gives the value of current when the voltage and resistance are known. It may be transposed so that any of the three quantities may be found when the other two are known:

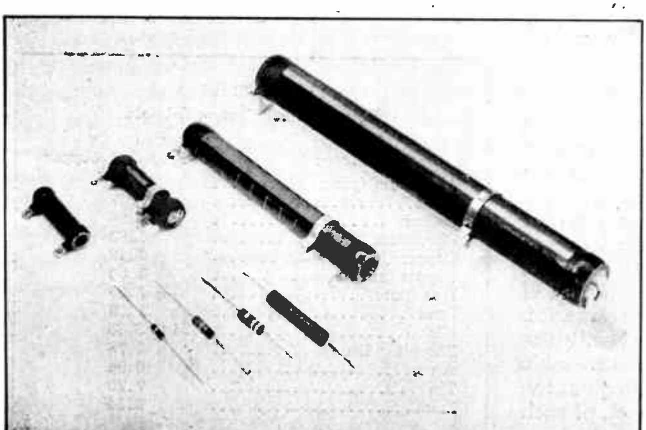
$$E = IR$$

(that is, the voltage acting is equal to the current in amperes multiplied by the resistance in ohms) and

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

(or, the resistance of the circuit is equal to the applied voltage divided by the current).

All three forms of the equation are used almost constantly in radio work. It must be



Types of resistors used in radio equipment. Those in the foreground with wire leads are carbon types, ranging in size from ½ watt at the left to 2 watts at the right. The larger resistors use resistance wire wound on ceramic tubes; sizes shown range from 5 watts to 100 watts. Three are the adjustable type, using a sliding contact on an exposed section of the resistance winding.

remembered that the quantities are in *volts*, *ohms* and *amperes*; other units cannot be used in the equations without first being converted. For example, if the current is in milliamperes it must be changed to the equivalent fraction of an ampere before the value can be substituted in the equations.

Table 2-II shows how to convert between the various units in common use The prefixes attached to the basic-unit name indicate the nature of the unit. These prefixes are:

micro — one-millionth (abbreviated μ)
milli — one-thousandth (abbreviated m)
kilo — one thousand (abbreviated k)
mega — one million (abbreviated M)

For example, one microvolt is one-millionth of a volt, and one megohm is 1,000,000 ohms. There are therefore 1,000,000 microvolts in one volt, and 0.000001 megohm in one ohm.

The following examples illustrate the use of Ohm's Law:

The current flowing in a resistance of 20,000 ohms is 150 milliamperes. What is the voltage? Since the voltage is to be found, the equation to use is $E = IR$. The current must first be converted from milliamperes to amperes, and reference to the table shows that to do so it is necessary to divide by 1000. Therefore,

$$E = \frac{150}{1000} \times 20{,}000 = 3000 \text{ volts}$$

When a voltage of 150 is applied to a circuit the current is measured at 2.5 amperes. What is the resistance of the circuit? In this case R is the unknown, so

$$R = \frac{E}{I} = \frac{150}{2.5} = 60 \text{ ohms}$$

No conversion was necessary because the voltage and current were given in volts and amperes.

How much current will flow if 250 volts is applied to a 5000-ohm resistor? Since I is unknown,

$$I = \frac{E}{R} = \frac{250}{5000} = 0.05 \text{ ampere}$$

Milliampere units would be more convenient for the current, and 0.05 amp. $\times$ 1000 = 50 milliamperes.

SERIES AND PARALLEL RESISTANCES

Very few actual electric circuits are as simple as the illustration in the preceding section. Commonly, resistances are found connected in a variety of ways. The two fundamental methods of connecting resistances are shown in Fig. 2-4. In the upper drawing, the current flows from the source of e.m.f. (in the direction shown by the arrow, let us say) down through the first resistance, R_1, then through the second, R_2, and then back to the source. These resistors are connected in series. The current everywhere in the circuit has the same value.

In the lower drawing the current flows to the common connection point at the top of the two resistors and then divides, one part of it flowing through R_1 and the other through R_2. At the lower connection point these two currents again combine; the total is the same as the current that flowed into the upper common connection. In this case the two resistors are connected in parallel.

TABLE 2-II
Conversion Values for Fractional and Multiple Units

To change from	*To*	*Divide by*	*Multiply by*
Units	Micro-units Milli-units Kilo-units Mega-units	 1000 1,000,000	1,000,000 1000
Micro-units	Milli-units Units	1000 1,000,000	
Milli-units	Micro-units Units	 1000	1000
Kilo-units	Units Mega-units	 1000	1000
Mega-units	Units Milli-units		1,000,000 1000

Resistors in Series

When a circuit has a number of resistances connected in series, the total resistance of the circuit is the sum of the individual resistances. If these are numbered R_1, R_2, R_3, etc., then

$$R \text{ (total)} = R_1 + R_2 + R_3 + R_4 + \ldots\ldots$$

where the dots indicate that as many resistors as necessary may be added.

Example: Suppose that three resistors are connected to a source of e.m.f. as shown in Fig. 2-5. The e.m.f. is 250 volts, R_1 is 5000 ohms, R_2 is 20,000 ohms, and R_3 is 8000 ohms. The total resistance is then

$$R = R_1 + R_2 + R_3 = 5000 + 20{,}000 + 8000 = 33{,}000 \text{ ohms}$$

The current flowing in the circuit is then

$$I = \frac{E}{R} = \frac{250}{33{,}000} = 0.00757 \text{ amp.} = 7.57 \text{ ma.}$$

(We need not carry calculations beyond three significant figures, and often two will suffice because the accuracy of measurements is seldom better than a few per cent.)

Voltage Drop

Ohm's Law applies to *any part* of a circuit as well as to the whole circuit. Although the current is the same in all three of the resistances in the example, the total voltage divides

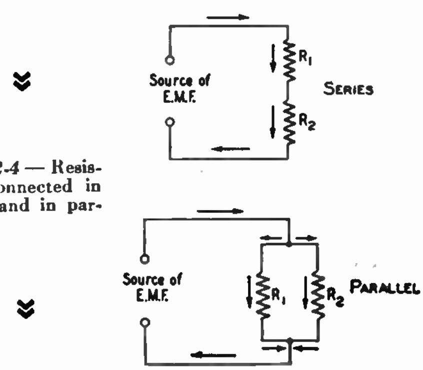


Fig. 2-4 — Resistors connected in series and in parallel.

among them. The voltage appearing across each resistor can be found from Ohm's Law.

Example: If the voltage across R_1 (Fig. 2-5) is called E_1, that across R_2 is called E_2, and that across R_3 is called E_3, then

$$E_1 = IR_1 = 0.00757 \times 5000 = 37.9 \text{ volts}$$
$$E_2 = IR_2 = 0.00757 \times 20{,}000 = 151.4 \text{ volts}$$
$$E_3 = IR_3 = 0.00757 \times 8000 = 60.6 \text{ volts}$$

The total voltage must equal the sum of the individual voltage drops:

$$E = E_1 + E_2 + E_3 = 37.9 + 151.4 + 60.6 = 249.9 \text{ volts}$$

The answer would have been more nearly exact if the current had been calculated to more decimal places, but as explained above a very high order of accuracy is not necessary.

In a simple series circuit like that in Fig. 2-5, the voltage drop across each resistance can be calculated very simply, if only the drop and not the current is wanted. The drop across each resistor is proportional to the ratio of the individual resistance to the *total* resistance. Thus

$$E_1 = \frac{R_1}{R_1 + R_2 + R_3} \times 250$$
$$= \frac{5000}{5000 + 20{,}000 + 8000} = \frac{5000}{33{,}000} \times 250$$
$$= 37.8 \text{ volts}$$
$$E_2 = \frac{20{,}000}{33{,}000} \times 250 = 151.5 \text{ volts}$$
$$E_3 = \frac{8000}{33{,}000} \times 250 = 60.5 \text{ volts}$$

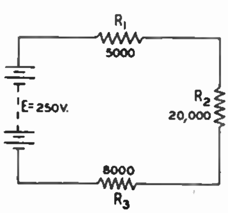


Fig. 2-5 — An example of resistors in series. The solution of the circuit is worked out in the text.

In problems such as this considerable time and trouble can be saved, when the current is small enough to be expressed in milliamperes, if the resistance is expressed in kilohms rather than ohms. When resistance in kilohms is substituted directly in Ohm's Law the current will be in milliamperes if the e.m.f. is in volts.

Example: Since 5000 ohms = 5 kilohms, 20,000 ohms = 20 kilohms, and 8000 ohms = 8 kilohms, the equations above become

$$I = \frac{E}{R} = \frac{250}{33} = 7.57 \text{ ma.}$$

$$E_1 = IR_1 = 7.57 \times 5 = 37.9 \text{ volts}$$
$$E_2 = IR_2 = 7.57 \times 20 = 151.4 \text{ volts}$$
$$E_3 = IR_3 = 7.57 \times 8 = 60.6 \text{ volts}$$

Resistors in Parallel

In a circuit with resistances in parallel, the total resistance is *less* than that of the *lowest* value of resistance present. This is because the total current is always greater than the current in any individual resistor. The formula for finding the total resistance of resistances in parallel is

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \frac{1}{R_4} + \cdots}$$

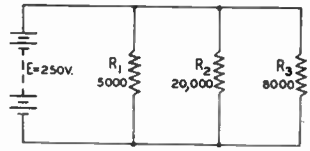


Fig. 2-6 — An example of resistors in parallel. The solution is worked out in the text.

where the dots again indicate that any number of resistors can be combined by the same method. For only two resistances in parallel (a very common case) the formula is

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

Example: If a 500-ohm resistor is paralleled with one of 1200 ohms, the total resistance is

$$R = \frac{R_1 R_2}{R_1 + R_2} = \frac{500 \times 1200}{500 + 1200} = \frac{600{,}000}{1700}$$
$$= 353 \text{ ohms}$$

It is probably easier to solve practical problems by a different method than the "reciprocal of reciprocals" formula. Suppose the three resistors of the previous example are connected in parallel as shown in Fig. 2-6. The same e.m.f., 250 volts, is applied to all three of the resistors. The current in each can be found from Ohm's Law as shown below, I_1 being the current through R_1, I_2 the current through R_2 and I_3 the current through R_3.

For convenience, the resistance will be expressed in kilohms so the current will be in milliamperes.

$$I_1 = \frac{E}{R_1} = \frac{250}{5} = 50 \text{ ma.}$$
$$I_2 = \frac{E}{R_2} = \frac{250}{20} = 12.5 \text{ ma.}$$
$$I_3 = \frac{E}{R_3} = \frac{250}{8} = 31.25 \text{ ma.}$$

The total current is

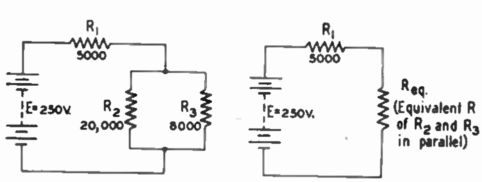
$$I = I_1 + I_2 + I_3 = 50 + 12.5 + 31.25$$
$$= 93.75 \text{ ma.}$$

The total resistance of the circuit is therefore

$$R = \frac{E}{I} = \frac{250}{93.75} = 2.66 \text{ kilohms } (= 2660 \text{ ohms})$$

Resistors in Series-Parallel

An actual circuit may have resistances both in parallel and in series. To illustrate, we use the same three resistances again, but now connected as in Fig. 2-7. The method of solving such a circuit is as follows: Consider R_2 and R_3 in parallel as though they formed a single resistor. Find their equivalent resistance. Then this resistance in series with R_1 forms a simple



EQUIVALENT CIRCUIT

Fig. 2-7 — An example of resistors in series-parallel. The solution is worked out in the text.

series circuit, as shown at the right in Fig. 2-7.

Example: The first step is to find the equivalent resistance of R_2 and R_3. From the formula for two resistances in parallel,

$$R_{eq.} = \frac{R_2R_3}{R_2 + R_3} = \frac{20 \times 8}{20 + 8} = \frac{160}{28} = 5.71 \text{ kilohms}$$

The total resistance in the circuit is then

$$R = R_1 + R_{eq.} = 5 + 5.71 \text{ kilohms} = 10.71 \text{ kilohms}$$

The current is

$$I = \frac{E}{R} = \frac{250}{10.71} = 23.4 \text{ ma.}$$

The voltage drops across R_1 and $R_{eq.}$ are

$$E_1 = IR_1 = 23.4 \times 5 = 117 \text{ volts}$$
$$E_2 = IR_{eq.} = 23.4 \times 5.71 = 133 \text{ volts}$$

with sufficient accuracy. These total 250 volts, thus checking the calculations so far, because the sum of the voltage drops must equal the total voltage. Since E_2 appears across both R_2 and R_3,

$$I_2 = \frac{E_2}{R_2} = \frac{133}{20} = 6.75 \text{ ma.}$$

$$I_3 = \frac{E_2}{R_3} = \frac{133}{8} = 16.6 \text{ ma.}$$

where I_2 = Current through R_2
I_3 = Current through R_3

The total is 23.35 ma., which checks sufficiently close with 23.4 ma., the current through the whole circuit.

There is a general rule for handling such complex circuits: Reduce the various resistances in parallel or series in *parts* of the circuit to *equivalent* resistances that then can be handled as *single* resistances in a simpler circuit. Eventually this process will lead to a simple series or parallel circuit from which the current and voltage drops can be calculated. Once these are known, Ohm's Law can be applied to each part of the circuit to determine currents and voltage drops in individual resistances.

POWER AND ENERGY

Power — the rate of doing work — is equal to voltage multiplied by current. The unit of electrical power, called the *watt*, is equal to one volt multiplied by one ampere. The equation for power therefore is

$$P = EI$$

where P = Power in watts
E = E.m.f. in volts
I = Current in amperes

Common fractional and multiple units for power are the *milliwatt*, one one-thousandth of a watt, and the *kilowatt*, or one thousand watts.

Example: The plate voltage on a transmitting vacuum tube is 2000 volts and the plate current is 350 milliamperes. (The current must be changed to amperes before substitution in the formula, and so is 0.35 amp.) Then

$$P = EI = 2000 \times 0.35 = 700 \text{ watts}$$

By substituting the Ohm's Law equivalents for E and I, the following formulas are obtained for power:

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

$$P = I^2R$$

These formulas are useful in power calculations when the resistance and either the current or voltage (but not both) are known.

Example: How much power will be used up in a 4000-ohm resistor if the voltage applied to it is 200 volts? From the equation

$$P = \frac{E^2}{R} = \frac{(200)^2}{4000} = \frac{40{,}000}{4000} = 10 \text{ watts}$$

Or, suppose a current of 20 milliamperes flows through a 300-ohm resistor. Then

$$P = I^2R = (0.02)^2 \times 300 = 0.0004 \times 300 = 0.12 \text{ watt}$$

Note that the current was changed from milliamperes to amperes before substitution in the formula.

Electrical power in a resistance is turned into heat. The greater the power the more rapidly the heat is generated. We said earlier that if a resistor is to handle considerable power it must be large in size and must be constructed in such a way that the heat will be carried off rapidly by the surrounding air. This prevents the temperature of the resistor from rising to a dangerous point. Resistors for radio work are made in many sizes, the smallest being rated to "dissipate" (or carry safely) about ¼ watt. The largest resistors used in amateur equipment will dissipate about 100 watts.

However, electrical power is not always turned into heat. The power used in running a motor, for example, is converted to mechanical motion. The power supplied to a radio transmitter is largely converted into radio waves. Power applied to a loudspeaker is changed into sound waves. Nevertheless, every electrical device has some resistance, so a part of the power supplied to it is dissipated in that resistance and hence appears as heat even though the major part of the power may be converted to another form.

Efficiency

In devices such as motors and vacuum tubes, the object is to obtain power in some other form than heat. Therefore power used in heating is considered to be a loss, because it is not the *useful* power. The **efficiency** of a device is the *useful* power output (in its converted form) divided by the power input to the device. In a vacuum-tube transmitter, for example, the object is to convert power from a d.c. source into a.c. power at some radio frequency. The ratio of the r.f. power output to the d.c. input is the efficiency of the tube. That is,

$$Eff. = \frac{P_o}{P_i}$$

where $Eff.$ = Efficiency (as a decimal)
P_o = Power output (watts)
P_i = Power input (watts)

Example: If the d.c. input to the tube is 100 watts and the r.f. power output is 60 watts, the efficiency is

$$Eff. = \frac{P_o}{P_i} = \frac{60}{100} = 0.6$$

Efficiency is usually expressed as a percentage; that is, it tells what per cent of the input power will be available as useful output. The efficiency in the above example is 60 per cent.

If a resistor is used purely for generating heat — as in an electric heater or cooker — its efficiency is practically 100 per cent, because all of the power input is converted into the desired form of power output. However, generating heat is usually not the desired end when resistors are used in radio equipment. The power losses in them are tolerated because very often a resistor performs a function that could not be conveniently or economically performed by any other device.

Energy

In residences, the power company's bill is for electric **energy,** not for power. What you pay for is the *work* that electricity does for you, not the *rate* at which that work is done. Electrical work is equal to power multiplied by time; the common unit is the **watt-hour,** which means that a power of one watt has been used for one hour. That is,

$$W = PT$$

where W = Energy in watt-hours
P = Power in watts
T = Time in hours

Other energy units are the **kilowatt-hour** and the **watt-second.** These units should be self-explanatory.

Energy units are seldom used in amateur practice, but it is obvious that a small amount of power used for a long time can eventually result in a "power" bill that is just as large as though a large amount of power had been used for a very short time.

Capacitance and Condensers

Suppose two flat metal plates are placed close to each other (but not touching) as shown in Fig. 2-8. Normally, the plates will be electrically "neutral"; that is, the number of electrons in each plate will just balance the number of atomic nuclei and there will be no electric charge.

Now suppose that the plates are connected to a battery through a switch, as shown. At the instant the switch is closed, electrons will be attracted from the upper plate to the positive terminal of the battery, and the same number will be repelled into the lower plate from the negative battery terminal. This electron movement will continue until enough electrons move into one plate and out of the other to make the e.m.f. between them the same as the e.m.f. of the battery. (That this must be so should be fairly obvious. The plates are conductors, and when they are connected to the battery, the battery voltage must appear between them.)

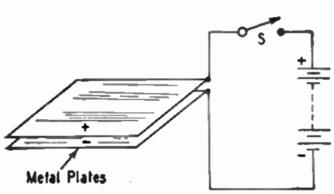


Fig. 2-8 — **A simple condenser.**

If the switch is opened after the plates have been charged, the top plate is left with a deficiency of electrons and the bottom plate with an excess. In other words, the plates remain charged despite the fact that the battery no longer is connected. They remain charged because with the switch open there is nowhere for the electrons to go. However, if a wire is touched between the two plates (**short-circuiting** them) the excess electrons on the bottom plate will flow through the wire to the upper plate, thus restoring electrical neutrality to both plates. The plates have then been **discharged.**

The two plates constitute an electrical **condenser,** and from the discussion above it should be clear that a condenser possesses the property of storing electricity. It should also be clear that during the time the electrons are moving — that is, while the condenser is being charged or discharged — a *current* is flowing in the circuit even though the circuit is "broken" by the gap between the condenser plates. However, the current flows *only* during the time of charge and discharge, and this time is usually very short. There can be no *continuous* flow of direct current through a condenser.

The charge or quantity of electricity that can be placed on a condenser when a given voltage is applied depends on its **capacitance** or **capacity.** The larger the plate area and the smaller the spacing between the plates the

TABLE 2-III
Dielectric Constants and Breakdown Voltages

Material	*Dielectric Constant*	*Puncture Voltage**
Air	1.0	19.8–22.8
Alsimag A196	5.7	240
Bakelite (paper-base)	3.8–5.5	650–750
Bakelite (mica-filled)	5–6	475–600
Celluloid	4–16	
Cellulose acetate	6–8	300–1000
Fiber	5–7.5	150–180
Formica	4.6–4.9	450
Glass (window)	7.6–8	200–250
Glass (photographic)	7.5	
Glass (Pyrex)	4.2–4.9	335
Lucite	2.5–3	480–500
Mica	2.5–8	
Mica (clear India)	6.4–7.5	600–1500
Mycalex	7.4	250
Paper	2.0–2.6	1250
Polyethylene	2.3–2.4	1000
Polystyrene	2.4–2.9	500–2500
Porcelain	6.2–7.5	40–100
Rubber (hard)	2–3.5	450
Steatite (low-loss)	4.4	150–315
Wood (dry oak)	2.5–6.8	

* In volts per mil (0.001 inch).

Fig. 2-9 — A multiple-plate condenser. Alternate plates are connected together.

greater the capacitance. The capacitance also depends upon the kind of insulating material between the plates; it is smallest with air insulation, but substitution of other insulating materials for air may increase the capacitance of a condenser many times. The ratio of the capacitance of a condenser with some material other than air between the plates, to the capacitance of the same condenser with air insulation, is called the **specific inductive capacity** or **dielectric constant** of that particular insulating material. The material itself is called a **dielectric.** The dielectric constants of a number of materials commonly used as dielectrics in condensers are given in Table 2-III. If a sheet of photographic glass is substituted for air between the plates of a condenser, for example, the capacitance of the condenser will be increased 7.5 times.

Units

The fundamental unit of capacitance is the **farad,** but this unit is much too large for practical work. Capacitance is usually measured in **microfarads** (abbreviated μ**fd.**) or **micromicrofarads** ($\mu\mu$**fd.**). The microfarad is one-millionth of a farad, and the micromicrofarad is one-millionth of a microfarad. Condensers nearly always have more than two plates, the alternate plates being connected together to form two sets as shown in Fig. 2-9. This makes it possible to attain a fairly large capacitance in a small space as compared to a two-plate condenser, since several plates of smaller individual area can be stacked to form the equivalent of a single large plate of the same total area. Also, all plates, except the two on the ends, are exposed to plates of the other group on *both sides*, and so are twice as effective in increasing the capacitance.

The formula for calculating the capacitance of a condenser is:

$$C = 0.224 \frac{KA}{d} (n - 1)$$

where C = Capacitance in $\mu\mu$fd.
K = Dielectric constant of material between plates
A = Area of one side of *one* plate in square inches
d = Separation of plate surfaces in inches
n = Number of plates

If the plates in one group do not have the same area as the plates in the other, use the area of the *smaller* plates.

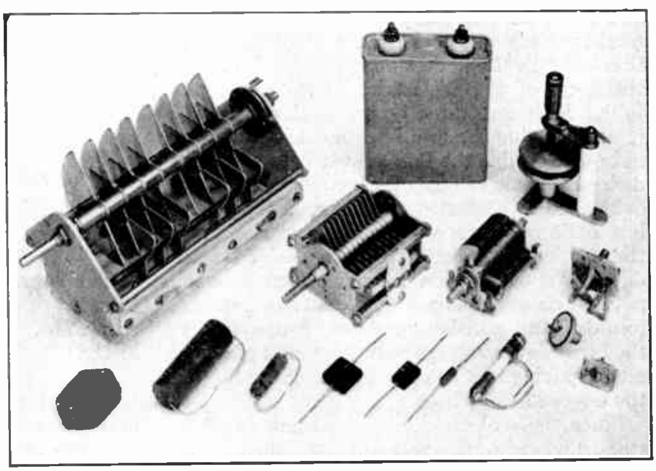
Example: A "variable" condenser has 7 semicircular plates on its rotor, the diameter of the semicircle being 2 inches. The stator has 6 rectangular plates, with a semicircular cut-out to clear the rotor shaft, but otherwise large enough to face the entire area of a rotor plate. The diameter of the cut-out is ½ inch. The distance between the adjacent surfaces of rotor and stator plates is ⅛ inch. The dielectric is air. What is the capacitance of the condenser with the plates fully meshed?

In this case, the "effective" area is the area of the rotor plate minus the area of the cut-out in the stator plate. The area of either semicircle is $\pi r^2/2$, where r is the radius. The area of the rotor plate is $\pi/2$, or 1.57 square inches (the radius is 1 inch). The area of the cut-out is $\pi(\frac{1}{4})^2/2 = \pi/32 = 0.10$ square inch, approximately. The "effective" area is therefore $1.57 - 0.10 = 1.47$ square inches. The capacitance is therefore

$$C = 0.224 \frac{KA}{d} (n-1) = 0.224 \frac{1 \times 1.47}{0.125} (13-1)$$
$$= 0.224 \times 11.76 \times 12 = 31.6 \ \mu\mu\text{fd.}$$

(The answer is only approximate, because of the difficulty of accurate measurement, plus a "fringing" effect at the edges of the plates that makes the actual capacitance a little higher.)

The usefulness of a condenser in electrical circuits lies in the fact that it can be charged



Fixed and variable condensers. The bottom row includes, left to right, a high-voltage mica fixed condenser, a tubular electrolytic, tubular paper, two sizes of "postage-stamp" micas, a small ceramic type (temperature compensating), an adjustable condenser with ceramic insulation (for neutralizing in transmitters), a "button" ceramic condenser, and an adjustable "padding" condenser. Four sizes of variable condensers are shown in the second row. The two-plate condenser with the micrometer adjustment is used in transmitters. The condenser enclosed in the metal case is a high-voltage paper type used in power-supply filters.

with electricity at one time and then discharged at a later time. In other words, it is capable of storing electrical energy that can be released later when it is needed; it is an "electrical reservoir."

Condensers in Radio

The types of condensers used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In "variable" condensers (almost always constructed with air for the dielectric) one set of plates is made movable with respect to the other set so that the capacitance can be varied. "Fixed" condensers — that is, having fixed capacitance — also can be made with metal plates and with air as the dielectric, but usually are constructed from plates of metal foil with a thin solid or liquid dielectric sandwiched in between, so that a relatively large capacitance can be secured in a small unit. The solid dielectrics commonly used are mica and paper. An example of a liquid dielectric is mineral oil, but it is seldom used by itself in present-day condensers. The "electrolytic" condenser uses aluminum-foil plates with a semiliquid conducting chemical compound between them; the actual dielectric is a very thin film of insulating material that "forms" on one set of plates through electrochemical action when a d.c. voltage is applied to the condenser. The capacitance obtained with a given plate area in an electrolytic condenser is very large, compared with condensers having other dielectrics, because the film is so extremely thin — much less than any thickness that is practicable with a solid dielectric.

Voltage Breakdown

When a high voltage is applied to the plates of a condenser, a considerable force is exerted on the electrons and nuclei of the dielectric. Because the dielectric is an insulator the electrons do not become detached from atoms the way they do in conductors. However, if the force is great enough the dielectric will "break down"; usually it will puncture and may char (if it is solid) and permit current to flow. The **breakdown voltage** depends upon the kind and thickness of the dielectric, as shown in the table. It is not directly proportional to the thickness; that is, doubling the thickness does not quite double the breakdown voltage. If the dielectric is air or any other gas, breakdown is evidenced by a spark or arc between the plates, but if the voltage is removed the arc ceases and the condenser is ready for use again. Breakdown will occur at a lower voltage between pointed or sharp-edged surfaces than between rounded and polished surfaces; consequently, the breakdown voltage between metal plates of given spacing in air can be increased by buffing the edges of the plates.

Since the dielectric must be thick to withstand high voltages, and since the thicker the dielectric the smaller the capacitance for a given plate area, a high-voltage condenser must have more plate area than a low-voltage condenser of the same capacitance. High-voltage high-capacitance condensers are physically large. The breakdown voltage of paper-dielectric condensers can be increased by saturating the paper with a special insulating oil and by immersing the condenser in oil. Electrolytic condensers can stand 400 to 500 volts before the dielectric film breaks down.

● CONDENSERS IN SERIES AND PARALLEL

The terms "parallel" and "series" when used with reference to condensers have the same circuit meaning as with resistances. When

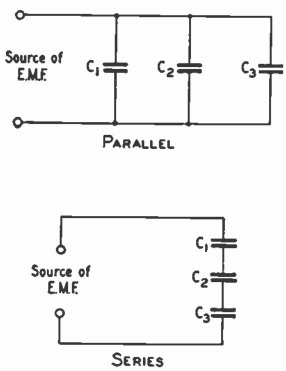


Fig. 2-10 — Condensers in series and parallel.

a number of condensers are connected in parallel, as in Fig. 2-10, the total capacitance of the group is equal to the sum of the individual capacitances, so

$$C\text{ (total)} = C_1 + C_2 + C_3 + C_4 + \dots\dots\dots$$

However, if two or more condensers are connected in series, as in the second drawing, the total capacitance is less than that of the smallest condenser in the group. The rule for finding the capacitance of a number of series-connected condensers is the same as that for finding the resistance of a number of *parallel*-connected resistors. That is,

$$C\text{ (total)} = \cfrac{1}{\cfrac{1}{C_1} + \cfrac{1}{C_2} + \cfrac{1}{C_3} + \cfrac{1}{C_4} + \dots\dots}$$

and, for only two condensers in series,

$$C\text{ (total)} = \frac{C_1 C_2}{C_1 + C_2}$$

The same units must be used throughout; that is, all capacitances must be expressed in either μfd. or $\mu\mu$fd.; you cannot use both units in the same equation.

Condensers are connected in parallel to obtain a larger total capacitance than is available in one unit. The largest voltage that can be applied safely to a group of condensers in parallel is the voltage that can be applied safely to the condenser having the *lowest* voltage rating.

When condensers are connected in series, the applied voltage is divided up among the various condensers; the situation is much the same as when resistors are in series and there is a voltage drop across each. However, the voltage that appears across each condenser of a group connected in series is in *inverse* proportion to its capacitance, as compared with the capacitance of the whole group.

Example: Three condensers having capacitances of 1, 2 and 4 μfd., respectively, are connected in series as shown in Fig. 2-11. The total capacitance is

$$C = \frac{1}{\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}} = \frac{1}{\frac{1}{1}+\frac{1}{2}+\frac{1}{4}} = \frac{1}{\frac{7}{4}} = \frac{4}{7} = 0.571 \text{ μfd.}$$

The voltage across each condenser is proportional to the *total* capacitance divided by the capacitance of the condenser in question, so the voltage across C_1 is

$$E_1 = \frac{0.571}{1} \times 2000 = 1142 \text{ volts}$$

Similarly, the voltages across C_2 and C_3 are

$$E_2 = \frac{0.571}{2} \times 2000 = 571 \text{ volts}$$

$$E_3 = \frac{0.571}{4} \times 2000 = 286 \text{ volts}$$

totaling approximately 2000 volts, the applied voltage.

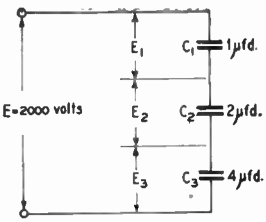


Fig. 2-11 — An example of condensers connected in series. The solution to this arrangement is worked out in the text.

Condensers are frequently connected in series to enable the group to withstand a larger voltage (at the expense of decreased total capacitance) than any individual condenser is rated to stand. One very common application of this arrangement is in the filter circuits of high-voltage power supplies. However, as shown by the previous example, the applied voltage does not divide equally among the condensers (except when all the capacitances are the same) so care must be taken to see that the voltage rating of no condenser in the group is exceeded. It does no good, for example, to connect a condenser in series with another if the capacitance of the second is many times as great as the first; nearly all of the voltage still will appear across the condenser having the smaller capacitance.

Inductance

It is possible to show that the flow of current through a conductor is accompanied by magnetic effects; a compass needle brought near the conductor, for example, will be deflected from its normal north-south position. The stronger the current, the more pronounced is the magnetic effect. The current, in other words, sets up a magnetic field.

If a wire conductor is formed into a coil, the same current will set up a stronger magnetic field than it will if the wire is straight. Also, if the wire is wound around an iron or steel "core" the field will be still stronger. The relationship between the strength of the field and the intensity of the current causing it is expressed by the **inductance** of the conductor or coil. If the same current flows through two coils, for example, and it is found that the magnetic field set up by one coil is twice as strong as that set up by the other, the first coil has twice as much inductance as the second. Inductance is a property of the conductor or coil and is determined by its shape and dimensions. The unit of inductance (corresponding to the ohm for resistance and the farad for capacitance) is the **henry**.

If the current through a conductor or coil is made to vary in intensity, it is found that an e.m.f. will appear across the terminals of the conductor or coil. This e.m.f. is entirely separate from the e.m.f. that is causing the current to flow. The strength of this "induced" e.m.f. becomes greater, the greater the intensity of the magnetic field and the more rapidly the current (and hence the field) is made to vary. Since the intensity of the magnetic field depends upon the inductance, the induced voltage (for a given current intensity and rate of variation) is proportional to the inductance of the conductor or coil.

The fact that an e.m.f. is "induced" accounts for the name "inductance" — or "self-inductance" as it is sometimes called. The induced e.m.f. tends to send a current through the circuit in the *opposite* direction to the current that flows because of the external e.m.f. so long as the latter current is *increasing*. However, if the current caused by the applied e.m.f. *decreases*, the induced e.m.f. tends to send current through the circuit in the *same* direction as the current from the applied e.m.f. The effect of inductance, therefore, is to oppose any *change* in the current flowing in the circuit, regardless of the nature of the change. It accomplishes this by storing energy in its magnetic field when the current in the circuit is being increased, and by releasing the stored energy when the current is being decreased. The effect is the same as the mechanical inertia that prevents an automobile from instantly coming up to speed when the accelerator pedal is pressed, and that prevents it from coming to an instant stop when the brakes are applied.

The values of inductance used in radio equipment vary over a wide range. Inductance of several henrys is required in power-supply circuits (see chapter on Power Supply) and to obtain such values of inductance it is necessary to use coils of many turns wound on iron cores. In radio-frequency circuits, the inductance values used will be measured in **millihenrys** (a millihenry is one one-thousandth of a henry) at low frequencies, and in **microhenrys** (one one-millionth of a henry) at medium frequencies and higher. Although coils for radio frequencies may be wound on special iron cores (ordinary iron is not suitable) most r.f. coils made and used by amateurs are the "air-core" type; that is, wound on an insulating form consisting of nonmagnetic material.

Inductance Formula

The inductance of air-core coils may be calculated from the formula

$$L\,(\mu\text{h.}) = \frac{0.2\,a^2n^2}{3a + 9b + 10c}$$

where L = Inductance in microhenrys
a = Average diameter of coil in inches
b = Length of winding in inches
c = Radial depth of winding in inches
n = Number of turns

The notation is explained in Fig. 2-12. The quantity c may be neglected if the coil only has one layer of wire.

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 (Chapter 24), 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:

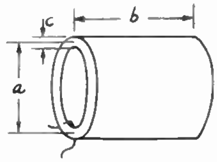


Fig. 2-12 — Coil dimensions used in the inductance formula.

$$N = \sqrt{\frac{3a + 9b}{0.2a^2} \times L}$$

Example: Suppose an inductance of 10 microhenrys is required. The form on which the coil is to be wound has a diameter of one inch and is long enough to accommodate a coil length of 1¼ inches. Then a = 1, b = 1.25, and L = 10. Substituting,

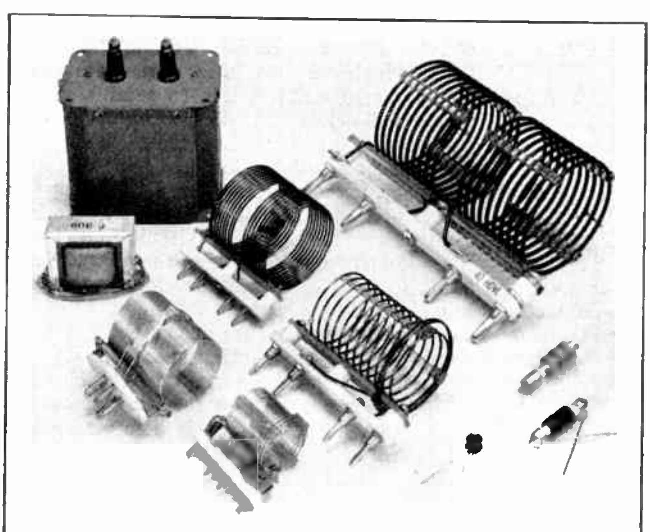
$$N = \sqrt{\frac{(3 \times 1) + (9 \times 1.25)}{0.2 \times 1^2} \times 10}$$

$$= \sqrt{\frac{14.25}{0.2} \times 10} = \sqrt{712.5}$$

$$= 26.6 \text{ turns.}$$

A 27-turn coil would be close enough to the required value of inductance, in practical work. Since the coil will be 1.25 inches long, the number of turns per inch will be 27/1.25 = 21.6. Consulting the wire table, we find that No. 18 enameled wire (or any smaller size) can be used. We obtain the proper inductance by winding the required number of turns on the form and then adjusting the spacing between the turns to make a uniformly-spaced coil 1.25 inches long.

Every conductor has inductance, even though the conductor is not formed into a coil. The inductance of a short length of straight wire is small — but it may not be negligible, because if the current through it changes its intensity rapidly enough the induced voltage may be appreciable. This will be the case in even a few inches of wire when an alternating current having a frequency of the order of 100 Mc. is flowing. However, at much lower frequencies the inductance of the same wire could be left out of any calculations because the induced voltage would be negligibly small.



Inductance coils for power and radio frequencies. The two iron-core coils at the upper left are "chokes" for power-supply filters. The three "pie"-wound coils at the lower right are used as chokes in radio-frequency circuits. The other coils are for r.f. tuned circuits ranging in power from 25 watts to a kilowatt.

IRON-CORE COILS

We mentioned earlier that the inductance of a coil wound on an iron core is much greater than the inductance of the same coil wound on a nonmagnetic core. As a crude analogy, iron has a much lower "resistance" to the magnetic force than nonferrous materials, just as metals have much lower resistance to the flow of electric current than nonmetallic substances.

Permeability

For example, suppose that the coil in Fig. 2-13 is wound on an iron core having a cross-sectional area of 2 square inches. When a certain current is sent through the coil it is found that there are 80,000 lines of force in the core. Since the area is 2 square inches, the flux density is 40,000 lines per square inch. Now suppose that the iron core is removed and the same current is maintained in the coil, and that the flux density without the iron core is found to be 50 lines per square inch. The ratio of the flux density with the given core material to the flux density (with the same coil and same current) with an air core is called the **permeability** of the material. In this case the permeability of the iron is 40,000/50 = 800. The inductance of the coil is increased 800 times by inserting the iron core, therefore.

The permeability of a magnetic material is not constant, unfortunately, but varies with the flux density. At low flux densities (or with an air core) increasing the current through the coil will cause a proportionate increase in flux. For example, if there are 2000 lines per square inch at a given current, doubling the current will increase the flux density to 4000 lines per square inch. But this cannot be carried on indefinitely; at some value of flux density, depending upon the kind of iron, it will be found that doubling the current only increases the flux density by, say, 10 per cent. At very high flux densities, increasing the current may cause no appreciable change in the flux at all. When this is so, the iron is said to be **saturated.** "Saturation" causes a rapid decrease in permeability, because it decreases the ratio of flux lines to those obtainable with the same current and an air core. Obviously, the inductance of an iron-core coil is highly dependent upon the current flowing in the coil. In an air-core coil, the inductance is independent of current because air does not "saturate."

In amateur work, iron-core coils such as the one sketched in Fig. 2-13 are used chiefly in power-supply equipment. They usually have direct current flowing through the winding, and the variation in inductance with current is usually undesirable. It may be overcome by keeping the flux density below the saturation point of the iron. This is done by cutting the core so that there is a small "air gap," as indicated by the dashed lines. The magnetic "resistance" introduced by such a gap is so large — even though the gap is only a small fraction of an inch — compared with that of the iron that the gap, rather than the iron, controls the flux density. This naturally reduces the inductance compared to what it would be without the air gap — but only for *small* currents. It actually results in a *higher* inductance when the current is large; furthermore, the inductance is practically constant regardless of the value of the current. Further information on the construction of such inductance coils will be found in the chapter on Power Supply.

Eddy Currents and Hysteresis

When alternating current flows through a coil wound on an iron core the magnetic flux in the core goes through variations in intensity and direction that correspond to the variations in the alternating current. Variations in a magnetic field cause an e.m.f. to be induced, as previously explained, and since iron is a conductor a current will flow in the core. Such currents (called **eddy currents**) represent a waste of power because they flow through the resistance of the iron and thus cause heating. Eddy-current losses can be reduced by **laminating** the core; that is, by cutting it into thin strips. These strips or **laminations** must be insulated from each other by painting them with some insulating material such as varnish or shellac.

There is also another type of energy loss in an iron core: the iron tends to resist any change in its magnetic state, so a rapidly-changing current such as a.c. is forced continually to supply energy to the iron to overcome this "inertia." Losses of this sort are called **hysteresis** losses.

Eddy-current and hysteresis losses in iron increase rapidly as the frequency of the alternating current is increased. For this reason, we can use ordinary iron cores only at power and audio frequencies — up to, say, 15,000 cycles. Even so, a very good grade of iron or steel is necessary if the core is to perform well at the higher audio frequencies. Iron cores of this type are completely useless at radio frequencies.

For radio-frequency work, the losses in iron cores can be reduced to a satisfactory figure by grinding the iron into a powder and then mixing it with a "binder" of insulating material in such a way that the individual iron particles are insulated from each other. By this means cores can be made that will function satisfac-

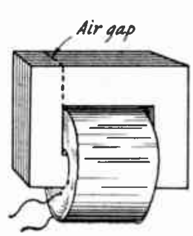


Fig. 2-13 — Typical construction of an iron-core coil. The small air gap prevents magnetic saturation of the iron and increases the inductance at high currents.

torily even through the v.h.f. range — that is, at frequencies up to perhaps 100 Mc. Because a large part of the magnetic path is through a nonmagnetic material, the permeability of the iron is low compared to the values obtained at power-supply frequencies. The core is usually in the form of a "slug" or cylinder which fits inside the insulating form on which the coil is wound. Despite the fact that, with this construction, the major portion of the magnetic path for the flux is in the air surrounding the coil, the slug is quite effective in increasing the coil inductance. By pushing the slug in and out of the coil the inductance can be varied over a considerable range.

INDUCTANCES IN SERIES AND PARALLEL

When two or more inductance coils (or **inductors,** as they are frequently called) are connected in series (Fig. 2-14, left) the total inductance is equal to the sum of the individual inductances, *provided the coils are sufficiently separated so that no coil is in the magnetic field of another.* That is,

$$L_{total} = L_1 + L_2 + L_3 + L_4 + \ldots\ldots$$

If inductances are connected in parallel (Fig. 2-14, right), the total inductance is

$$L_{total} = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \frac{1}{L_4} + \ldots\ldots}$$

and for two inductances in parallel,

$$L = \frac{L_1 L_2}{L_1 + L_2}$$

Thus the rules for combining inductances in series and parallel are the same as for resistances, *if* the coils are far enough apart so that each is unaffected by another's magnetic field. When this is not so the formulas given above cannot be used.

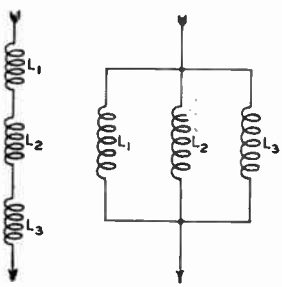


Fig. 2-14 — **Inductances in series and parallel.**

In calculating the total inductance of a combination of iron-core coils to be used in a d.c. circuit, it must be remembered that the inductance of each coil may change with the amount of current that flows through it. With air-core coils there is no such change.

Although there is frequent occasion to combine resistances or capacitances in series or parallel in amateur work, there is relatively little necessity for such combinations of inductances — or rather, the cases that do arise in practice seldom require calculations.

MUTUAL INDUCTANCE

If two coils are arranged with their axes on the same line, as shown in Fig. 2-15, 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, but since it appears in the *second* coil because of current flowing in the *first*, it is a "mutual" effect and

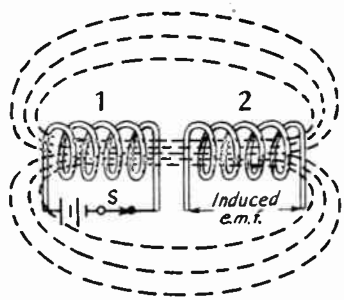


Fig. 2-15 — **Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.**

results from the **mutual inductance** between the two coils.

Mutual inductance may be large or small, depending upon the self-inductances of the coils and the proportion of the flux set up by one coil that 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. If only a small part of the flux set up by one coil cuts the turns of the other the mutual inductance is relatively small. Two coils having mutual inductance are said to be **coupled.**

The ratio of actual mutual inductance to the maximum possible value that could be obtained with two given coils is called the **coefficient of coupling** between the coils. Coils that have nearly the maximum possible mutual inductance are said to be **closely,** or **tightly,** coupled, but 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. 2-15, and are as close together as possible. The coupling is least when the coils are far apart or are placed so their axes are at right angles.

The maximum possible coefficient of coupling is 1. This value is closely approached only when the two coils are wound on a closed iron core. The coefficient with air-core coils may run as high as 0.6 or 0.7 if one coil is wound over the other, but will be much less if the two coils are separated.

If two coils having mutual inductance are connected to the same source of current, the magnetic field of one coil can either aid or oppose the field of the other. In the former case the mutual inductance is said to be "positive"; in the latter case, "negative." Positive mutual inductance means that the total inductance is *greater* than the sum of the two individual inductances. Negative mutual 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.

Time Constant

Both inductance and capacitance possess the property of storing energy — inductance stores magnetic energy and capacitance stores electrical energy. In the case of inductance, electrical energy is converted into magnetic energy when the current through the inductance is increasing, and the magnetic energy is converted back into electrical energy (and thereby restored to the circuit) when the current is decreasing. It is this alternate storing and releasing of energy that makes inductance oppose a change in the current through it. The self-induced e.m.f. is the means by which energy is put into and taken out of the magnetic field.

In the case of capacitance, energy is stored in the condenser (actually in the electric field between the plates) whenever the voltage applied to the condenser is increasing, and restored to the circuit when the applied voltage is decreasing. That is, current flows *into* the condenser in the first case, and *out of* the condenser in the second.

Capacitance and Resistance

In Fig. 2-16A a battery having an e.m.f., *E*, a switch, *S*, a resistor, *R*, and condenser, *C*, are connected in series. Suppose for the moment that *R* has zero resistance — in other words, is short-circuited — and also that there is no other resistance in the circuit. If *S* is now closed, condenser *C* will charge *instantly* to the battery voltage; that is, the electrons that constitute the charge redistribute themselves in a time interval so small that it can be considered to be zero. As soon as the condenser is fully charged the current flow stops completely. But since the condenser became fully charged in zero time, the current during the instantaneous charge must have been very large; mathematically, it would be *infinitely* large if the time actually was zero — this regardless of the actual number of electrons that moved. At the instant of closing the switch, therefore, the condenser can be considered to have a "resistance" of zero, a resistance that becomes an open circuit the instant the charge is complete.

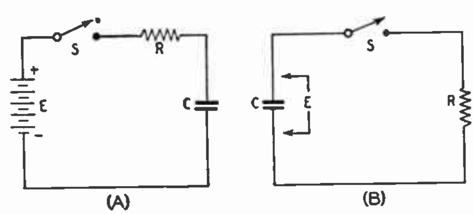


Fig. 2-16 — Schematics illustrating the time constant of an *RC* circuit.

If a finite value of resistance, *R*, is put into the circuit the condenser no longer can be charged instantaneously. If the condenser is initially uncharged, it will have zero "resistance" at the instant *S* is closed, but now the amount of current that can flow is limited by *R*. The infinitely-large current required to charge the condenser in zero time cannot flow through *R*, because even with *C* considered as a short-circuit the current in the circuit as a whole will be determined by Ohm's Law. If the battery e.m.f. is 100 volts, for example, and *R* is 10 ohms, the maximum current that can flow with *C* short-circuited is 10 amperes. Even this much current can flow *only* at the very instant the switch is closed. As soon as *any* current flows, condenser *C* begins to acquire a charge, which means that the voltage across the condenser plates rises. Since the upper plate (in Fig. 2-16A) will be positive and the lower negative, the voltage on the condenser tends to send a current through the circuit in the opposite direction to the current from the battery. The voltage on the condenser, in other words, opposes the battery voltage. Immediately after the switch is closed, therefore, the current drops below its initial Ohm's Law value, and as the condenser continues to acquire charge and its potential rises, the current becomes smaller and smaller.

The length of time required to complete the charging process depends upon the capacitance of the condenser and the resistance in the circuit. More time is taken if either of these quantities is made larger. Theoretically, the charging process is never really finished, but practically the current eventually drops to a value that is smaller than anything that can be measured. The **time constant** of such a circuit is the length of time, in seconds, required for the voltage across the condenser to reach 63 per cent of the applied e.m.f. (this figure is chosen for mathematical reasons). The voltage

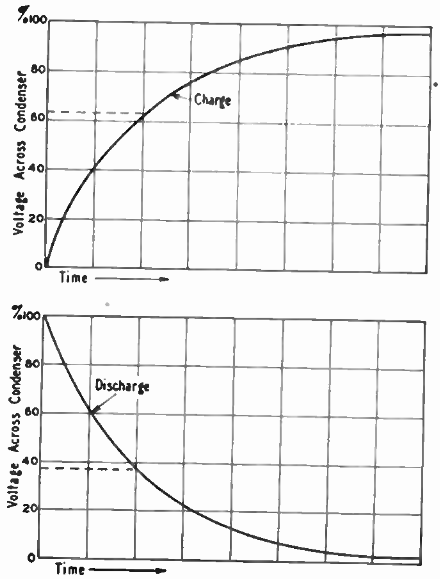


Fig. 2-17 — How the voltage across a condenser rises, with time, when a condenser is charged through a resistor. The lower curve shows the way in which the voltage decreases across the condenser terminals on discharging through the same resistor.

across the condenser rises logarithmically, as shown by Fig. 2-17.

The formula for time constant is

$$T = CR$$

where T = Time constant in seconds
C = Capacitance in farads
R = Resistance in ohms

If C is in microfarads and R in megohms, the time constant also is in seconds. The latter units usually are more convenient.

Example: The time constant of a 2-μfd. condenser and a 250,000-ohm resistor is

$$T = CR = 2 \times 0.25 = 0.5 \text{ second}$$

If the applied e.m.f. is 1000 volts, the voltage across the condenser plates will be 630 volts at the end of ½ second.

If a charged condenser is *discharged* through a resistor, as indicated in Fig. 2-16B, the same time constant applies. If there were no resistance, the condenser would discharge *instantly* when S was closed, and for *instantaneous* discharge the current would have to be infinitely large. However, if R is present the current cannot exceed the value given by Ohm's Law, where E is the voltage to which the condenser is charged and R is the resistance. Since R limits the current flow, the condenser voltage cannot instantly go to zero, but it will decrease just as rapidly as the condenser can rid itself of its charge through R. When the condenser is discharging through a resistance, the time constant (calculated in the same way as above) is the time (in seconds) that it takes for the condenser to *lose* 63 per cent of its voltage; that is, for the voltage to drop to 37 per cent of its initial value.

Example: If the condenser of the example above is charged to 1000 volts, it will discharge to 370 volts in ½ second through the 250,000-ohm resistor.

Inductance and Resistance

A comparable situation exists when resistance and inductance are in series. In Fig. 2-18, first consider L to have no resistance (which would be impossible, since the conductor of which it is composed always has resistance) and also assume that R is zero. Then closing S would tend to send a current through the circuit. However, the instantaneous transition from no current to a finite value, however small, represents a very rapid *change* in current, and a back e.m.f. is developed by the self-inductance of L that is practically equal and opposite to the applied e.m.f. The result is that the initial current is very small. However, the back e.m.f. depends upon the *change* in current and would cease to offer opposition if the current did not *continue* to increase. With no resistance in the circuit (which would lead to an infinitely-large current, by Ohm's Law) the current would increase forever, always increasing just fast enough to keep the e.m.f. of self-induction equal to the applied e.m.f. Since such a circuit never would "settle down," the time constant of an inductive circuit without resistance is infinitely long.

When resistance is in series, Ohm's Law sets a limit to the value that the current can reach. In such a circuit the current is small at first, just as in our hypothetical case without resistance. But as the current increases the voltage drop across R becomes larger. The back e.m.f. generated in L has only to equal the *difference* between E and the drop across R, because that difference is the voltage actually applied to L. This difference becomes smaller as the current approaches the final Ohm's Law value. Theoretically, the back e.m.f. never quite disappears (that is, the current never quite reaches the Ohm's Law value) but practically it becomes unmeasurable after a time. The difference between the actual current and the Ohm's Law value also becomes undetectable. The time required for this to occur is greater the larger the value of L, and is shorter the larger R is made. The time constant of an inductive circuit is the time in seconds required for the current to reach 63 per cent of its final value. The formula is,

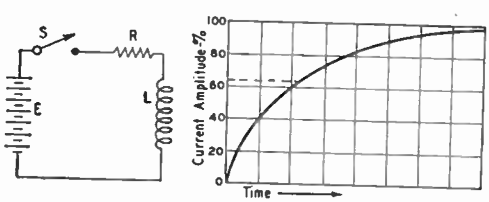


Fig. 2-18 — Time constant of an *LR* circuit.

$$T = \frac{L}{R}$$

where T = Time constant in seconds
L = Inductance in henrys
R = Resistance in ohms

The resistance of the wire in a coil acts as though it were in series with the inductance.

Example: A coil having an inductance of 20 henrys and a resistance of 100 ohms has a time constant of

$$T = \frac{L}{R} = \frac{20}{100} = 0.2 \text{ second}$$

if there is no other resistance in the circuit. If a d.c. e.m.f. of 10 volts is applied to such a coil, the final current, by Ohm's Law, is

$$I = \frac{E}{R} = \frac{10}{100} = 0.1 \text{ amp. or } 100 \text{ ma.}$$

The current would rise from zero to 63 milliamperes in 0.2 second after closing the switch.

An inductor cannot be discharged in the same way as a condenser, because the magnetic field disappears as soon as current flow ceases. Opening S does not leave the inductor "charged." The energy stored in the magnetic field instantly returns to the circuit when S is opened. The rapid disappearance of the field causes a very large voltage to be induced in the coil — ordinarily many times larger than the voltage applied, because the induced voltage is proportional to the *speed* with which the field changes. The common result of opening the switch in a circuit such as the one shown is that a spark or arc forms at the switch contacts at the instant of opening. If the inductance is large and the current in the circuit is high, a great deal of energy is released in a very short period of time. It is not at all unusual for the switch contacts to burn or melt under such circumstances.

"Filter" circuits used in power-supply equipment represent an excellent example of the application of the CR or L/R time constant to practical work, although calculations of the type illustrated above are seldom necessary with such circuits. An understanding of the principles also is necessary in numerous special devices that are coming into widespread use in amateur stations, such as electronic keys, shaping of keying characteristics by vacuum tubes, and timing devices and control circuits. The time constants of circuits are also important in such applications as automatic gain control and noise limiters.

Alternating Currents

PHASE

You cannot really understand alternating currents until you have a clear picture of **phase.** Essentially it means "time," or the *time interval* between the instant when one thing occurs and the instant when a second related thing takes place. As a homely example, when a baseball pitcher throws the ball to the catcher there is a definite interval, represented by the time of flight of the ball, between the act of throwing and the act of catching. The throwing and catching are therefore "out of phase" because they do not occur at exactly the same time.

Time differences are measured in seconds, minutes, hours, and so on. In the baseball example the ball might be in the air two seconds, in which case it could be said that the throwing and catching were out of phase by two seconds. However, simply saying that two events are out of phase does not tell us which one occurred first. To give this information, the later event is said to **lag** the first in phase, while the one that occurs first is said to **lead.** Thus, throwing the ball "leads" the catch by two seconds, or the catch "lags" the throw by two seconds.

In a.c. circuits the current amplitude changes continuously, so the concept of phase or time obviously has utility whenever it becomes necessary to specify the value of the current at a particular instant. Phase can be measured in the ordinary time units, such as the second, but there is a more convenient method: since each a.c. cycle occupies exactly the same amount of time as every other cycle of the same frequency, we can use the cycle itself as the time unit. When this is done it does not matter whether one cycle lasts for a sixtieth of a second or for a millionth of a second so long as *all the cycles are the same.* In other words, using the cycle as the time unit makes the specification or measurement of phase independent of the frequency of the current, so long as only one frequency is under consideration at a time. If there are two or more frequencies, the measurement of phase has to be modified just as the measurements of two lengths must be reconciled if one is given in feet and the other in meters.

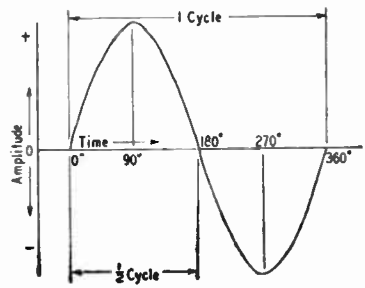


Fig. 2-19 — An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

The time interval or "phase difference" under consideration usually will be less than one cycle. Phase difference could be measured in decimal parts of a cycle, but for many reasons it is more convenient to divide the cycle into 360 parts or **degrees**. A phase degree is therefore 1/360 of a cycle. (The reason for this choice of unit is this: In a sine-wave alternating current, the value of the current at any instant is proportional to the sine of the angle that corresponds to the number of degrees — that is, length of time — from the time the cycle began. There is of course no actual "angle" associated with an alternating current.) Fig. 2-19 should help make this method of measurement clear.

Measuring Phase

In a steady alternating current each cycle is exactly like the preceding one. To compare the phase of two currents of the same frequency, we measure between corresponding parts of cycles of the two currents. This is shown in Fig. 2-20. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees sooner in time. (It is equally correct to say that *B lags* *A* by 45 degrees.) The amplitudes of the individual currents do not affect their relative phases — current *B* is shown as having smaller amplitude than *A*. Regardless of the amplitudes, the lagging current always would begin its cycle (the start of the cycle is considered to be the point at which it is passing through zero and starting to increase in the positive direction) the same number of degrees after the current that leads begins *its* cycle.

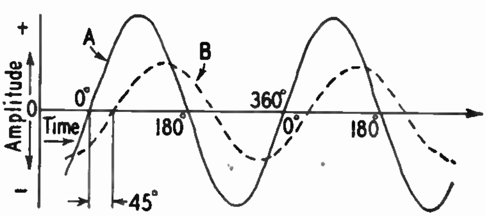
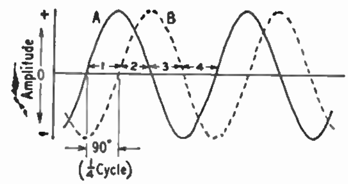


Fig. 2-20 — When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

Two important special cases are shown in Fig. 2-21. In the upper drawing *B* lags 90 degrees behind *A*; that is, its cycle begins just one-quarter cycle later than that of *A*. When one wave is passing through zero, the other is just at its maximum point. Note that (using *A* as a reference) in the first quarter cycle *A* is positive and *B* is negative; in the second quarter cycle both *A* and *B* are positive, but one is decreasing while the other is increasing; in the third quarter cycle *A* is negative while *B* is positive; and in the last quarter cycle both are negative.

In the lower drawing *A* and *B* are 180 degrees out of phase. In this case it does not matter which one we consider to lead or lag. *B* is always positive while *A* is negative, and vice versa. The two waves are thus *completely* out of phase.



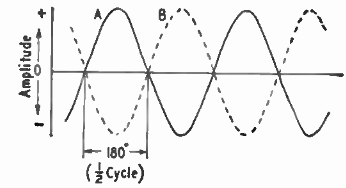


Fig. 2-21 — Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

The waves shown in Figs. 2-20 and 2-21 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage while *B* represented current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the *actual* current (or voltage) would take a *single* value at any instant. This value would equal the sum of the two at that instant. (We must take into account the fact that the sum of positive and negative values is actually equal to the *difference* between them.) The resultant current (or voltage) also is a sine wave, because adding any number of sine waves of the same frequency always results in a sine wave also of the same frequency.

● REACTANCE

The discussion of capacitance and inductance earlier in this chapter was confined to cases where only d.c. voltages were applied. To understand what happens in a condenser or inductance when an *a.c.* voltage is applied, it is necessary to become acquainted with a fundamental *definition* of electric current (as contrasted to the physical *description* of current given earlier). By definition, the amplitude of an electric current is the *rate* at which electric charge is moved past a point in a circuit. If a large quantity of charge moves past the observing point in a given time, the current is large; if the quantity is small in the same amount of time, the current is small.

Alternating Current in Condensers

The quantity of charge that can be placed on a condenser of given capacitance is proportional to the voltage applied to the condenser. As we explained earlier, the condenser becomes charged *instantly* if there is no resistance in the circuit. Suppose a sine-wave a.c. voltage is applied to a condenser in a circuit containing no resistance, as indicated in Fig. 2-22. For convenience, the first half-cycle of the applied voltage is divided into eight equal time intervals. In the period *OA*, the voltage increases from zero to 38 volts; at the end of this period the condenser is charged to that voltage. In the next interval the voltage increases to 71 volts; that is, 33 volts additional. In this second interval a *smaller* quantity of charge has been added than in the first interval, because the voltage rise during the second interval was smaller. Consequently the average current during the second interval is smaller than during the first. In the third interval, *BC*, the voltage rises from 71 to 92 volts, an increase of 21 volts. This is less than the voltage increase during the second interval, so the quantity of electricity added to the charge during the third interval is less than the quantity added during the second. In other words, the average current during the third interval is still smaller. In the fourth interval, *CD*, the voltage increases only 8 volts; the charge added is smaller than in any preceding interval and therefore the current also is smaller. By dividing the first quarter cycle into a very large number of intervals it could be shown that the current charging the condenser has the shape of a sine wave, just as the applied voltage does. But the current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage (the condenser cannot be charged to a higher voltage than the maximum applied, so no further current can flow) so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle of the applied voltage the current is flowing in the normal way through the circuit, since the condenser is being charged. Hence the current is positive during this first quarter cycle, as indicated by the dashed line in Fig. 2-22.

In the second quarter cycle — that is, in the time from *D* to *H*, the voltage applied to the condenser decreases. During this time the condenser *loses* the charge it acquired during the first quarter cycle. Applying the same reasoning, it is plain that the current is small from *D* to *E* and continues to increase during each succeeding interval. However, the current is flowing *against* the applied voltage because the condenser is *discharging into the circuit.* Hence the current is *negative* during this quarter cycle.

The third and fourth quarter cycles repeat the events of the first and second, respectively, with this difference — the polarity of the applied voltage has reversed, and the current changes to correspond. In other words, *an alternating current flows through a condenser when an a.c. voltage is applied to it.* As shown by Fig. 2-22, the current starts its cycle 90 degrees before the voltage, so the current in a condenser *leads the applied voltage by 90 degrees.*

Capacitive Reactance

Remembering the definition of current as given at the beginning of this section, as well as the mechanism of current flow described above, it should be plain that the more rapid the voltage rise the larger the current, because a rapid change in voltage means a rapid transfer of charge into or out of the condenser. The rapidity with which the voltage changes depends upon two things: (1) the amplitude of the voltage (the greater the maximum value, the faster the voltage must rise from zero to reach that maximum in the time of one-quarter cycle if the frequency is fixed); (2) the frequency (the higher the frequency, the more rapidly the voltage goes through its changes in a given time if the maximum amplitude is fixed). Also, the amplitude of the current depends upon the capacitance of the condenser, because the larger the capacitance the greater the amount of charge transferred during a given change in voltage.

The fact that the current flowing through a condenser is directly proportional to the applied a.c. voltage is extremely important. It is exactly what Ohm's Law says about the flow of direct current in a resistive circuit, and so leads us to the conclusion that Ohm's Law may be applied to an alternating-current circuit containing a condenser. Of course, a condenser does not offer "resistance" to the flow of alternating current, because the condenser does not consume power as a resistor does. It merely stores energy in one part of the cycle and returns it to the circuit in the next part. Furthermore, the larger the capacitance the larger the current; this is just the opposite of what we expect with resistance. And finally, the "opposition" offered by a condenser to alternating current depends on the *frequency* of that current. But with a given capacitance and a given frequency, the condenser follows Ohm's Law on a.c.

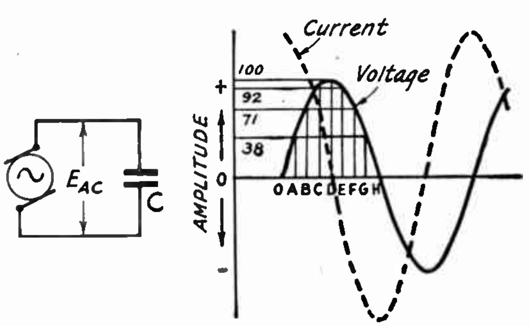


Fig. 2-22 — Voltage and current phase relationships when an alternating voltage is applied to a condenser.

Since the opposition effect of a condenser is not resistance, it is called by another name, **reactance.** But because reactance holds back current flow in a similar fashion to resistance, the unit of reactance also is the ohm. The reactance of a condenser is

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

where X_C = Condenser reactance in ohms
f = Frequency in cycles per second
C = Capacitance in farads
π = 3.14

The fundamental units (cycles per second, farads) are too large for practical use in radio circuits. However, if the capacitance is in microfarads and the frequency is in megacycles, the reactance will come out in ohms in the formula.

Example: The reactance of a condenser of 470 $\mu\mu$fd. (0.00047 μfd.) at a frequency of 7150 kc. (7.15 Mc.) is

$$X = \frac{1}{2\pi fC} = \frac{1}{6.28 \times 7.15 \times 0.00047} = 47.4 \text{ ohms}$$

Inductive Reactance

In the case of an alternating voltage applied to a circuit containing only inductance, with no resistance, it must be remembered that in such a resistanceless circuit the current always changes just rapidly enough to induce a back e.m.f. that equals and opposes the applied voltage. In Fig. 2-23, the cycle is again divided off into equal intervals. Assuming that the current has a maximum value of 1 ampere, the instantaneous current at the end of each interval will be as shown. The value of the *induced* voltage is proportional to the *rate at which the current changes.* It is therefore greatest in the intervals *OA* and *GH* and least in the intervals *CD* and *DE*. The induced voltage actually is a sine wave (if the current is a sine wave) as shown by the dashed curve. The *applied* voltage, because it is always equal to and opposed by the induced voltage, is equal to and 180 degrees out of phase with the induced voltage, as shown by the second dashed curve. The result, therefore, is that the current flowing in an inductance is 90 degrees out of phase with the applied voltage, and lags behind the applied voltage. This is just the opposite of the condenser case.

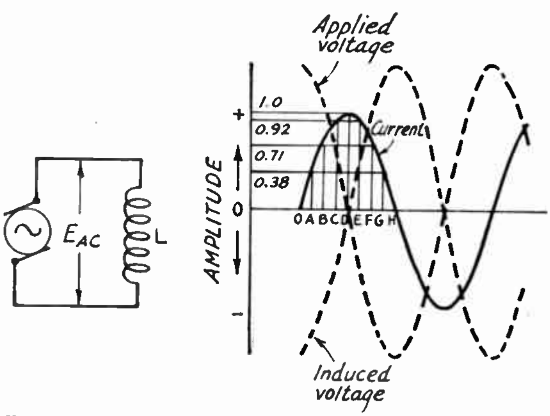


Fig. 2-23 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

Just enough current will flow in an inductance to induce an e.m.f. that just equals the applied e.m.f. Since the value of the induced e.m.f. is proportional to the rate at which the current changes, and this rate of change is in turn proportional to the frequency of the current, it should be clear that a small current changing rapidly (that is, at a high frequency) can generate a large back e.m.f. in a given inductance just as well as a large current changing slowly (low frequency). Consequently, the current that flows through a given inductance will decrease as the frequency is raised, if the applied e.m.f. is held constant. However, with both frequency and inductance fixed, the current will be larger when the applied voltage is increased, because the necessary rate of change in the current to induce the required back e.m.f. can only be obtained by having a greater total current flow under such circumstances. Again, when the applied voltage and frequency are fixed, the value of current required is less, as the inductance is made larger, because the induced e.m.f. also is proportional to inductance.

Just as in the capacitance case, the key point here is that — with the frequency and inductance fixed — an increase in the applied a.c. voltage causes a proportionate increase in the current. This is Ohm's Law again — and, again, the opposition effect is similar to, but not identical to, resistance. It is called **inductive reactance** and, like capacitive reactance, is measured in ohms. There is no energy loss in inductive reactance; the energy is stored in the magnetic field in one quarter cycle and then returned to the circuit in the next.

The formula for inductive reactance is

$$X_L = 2\pi fL$$

where X_L = Inductive reactance in ohms
f = Frequency in cycles per second
L = Inductance in henrys
π = 3.14

Example: The reactance of a coil having an inductance of 8 henrys, at a frequency of 120 cycles, is

$$X_L = 2\pi fL = 6.28 \times 120 \times 8 = 6029 \text{ ohms}$$

In radio-frequency circuits the inductance values usually are small and the frequencies are large. If the inductance is expressed in millihenrys and the frequency in kilocycles, the conversion factors for the two units cancel, and the formula for reactance may be used without first converting to fundamental units. Similarly, no conversion is necessary if the inductance is in microhenrys and the frequency is in megacycles.

Example: The reactance of a 15-microhenry coil at a frequency of 14 Mc. is

$X_L = 2\pi fL = 6.28 \times 14 \times 15 = 1319$ ohms

Ohm's Law for Reactance

Ohm's Law for an a.c. circuit containing *only* reactance is

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

$$E = IX$$

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

where E = E.m.f. in volts
I = Current in amperes
X = Reactance in ohms

The reactance may be either inductive or capacitive.

Example: If a current of 2 amperes is flowing through the condenser of the previous example (reactance = 47.4 ohms) at 7150 kc., the voltage drop across the condenser is

$E = IX = 2 \times 47.4 = 94.8$ volts

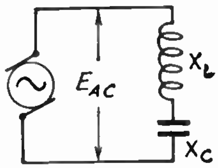
If 400 volts at 120 cycles is applied to the 8-henry inductance of the previous example, the current through the coil will be

$I = \frac{E}{X} = \frac{400}{6029} = 0.0663$ amp. (66.3 ma.)

These examples show that there is nothing complicated about using Ohm's Law for a reactive a.c. circuit. The question naturally arises, though, as to what to do when the circuit consists of an inductance in series with a capacitance. In such a case the same current flows through both reactances. However, the voltage across the coil *leads* the current by 90 degrees, and the voltage across the condenser *lags* behind the current by 90 degrees. The coil and condenser voltages therefore are 180 degrees out of phase.

A simple circuit of this type is shown in Fig. 2-24. The same figure also shows the current (heavy line) and the voltage drops across the inductance (E_L) and capacitance (E_C). It is assumed that X_L is larger than X_C and so has a larger voltage drop. Since the two voltages are completely out of phase the *total* voltage (E_{AC}) is equal to the *difference* between them. This is shown in the drawing as $E_L - E_C$. Notice that, because E_L is larger than E_C, the resultant voltage is exactly in phase with E_L. In other words, the circuit as a whole simply acts *as though it were an inductance* — an inductance of smaller value than the actual inductance present, since the effect of the actual inductive reactance is reduced by the capacitive reactance in series with it. If X_C is larger than X_L, the arrangement will behave like a capacitance — again of smaller reactance than the actual capacitive reactance present in the circuit.

The "equivalent" or total reactance of any circuit containing inductive and capacitive reactances in series is equal to $X_L - X_C$. If



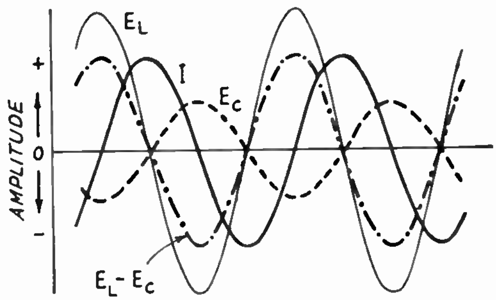


Fig. 2-24 — Current and voltages in a circuit having inductive and capacitive reactances in series.

there are several coils and condensers in series, we simply add up all the inductive reactances, then add up all the capacitive reactances, and then subtract the latter from the former. It is customary to call inductive reactance "positive" and capacitive reactance "negative." If the equivalent or net reactance is positive, the voltage leads the current by 90 degrees; if the net reactance is negative, the voltage lags the current by 90 degrees.

Reactive Power

A curious feature of the drawing in Fig. 2-24 is that the voltage drop across the coil is larger than the voltage applied to the circuit. At first glance this might seem to be an impossible condition. But it is not; the reason is that neither the coil nor condenser *consumes* power. Actually, when energy is being stored in the coil's magnetic field, energy is being returned to the circuit from the condenser's electric field, and vice versa. This stored energy is responsible for the fact that the voltages across reactances in series can be larger than the voltage applied to them.

It will be recalled that in a resistance the flow of current causes heating and a power loss equal to I^2R. The power in a reactance is equal to I^2X, but is not a "loss"; it is simply power that is transferred back and forth between the field and the circuit but not used up in heating anything. In the quarter cycle when the current and voltage in a reactance both have the same polarity, energy is stored in the field; in the quarter cycle when the current and voltage have *opposite* polarity the energy is returned to the circuit. To distinguish this "nondissipated" power from the power which is actually consumed, the unit of reactive power is called the **volt-ampere** instead of the watt. Reactive power is sometimes called "wattless" power.

IMPEDANCE

Although resistance, inductive reactance and capacitive reactance all are measured in ohms, the fact that they all are measured by the same unit does not indicate that they can be combined indiscriminately. Reactance does not absorb energy; resistance does. Voltage and current are in phase in resistance, but differ in phase by a quarter cycle in reactance. Furthermore, in inductive reactance the voltage leads the current, while in capacitive reactance the current leads the voltage. All these things must be taken into account when reactance and resistance are combined together in a circuit.

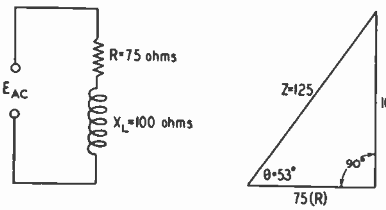


Fig. 2-25 — Resistance and inductive reactance connected in series.

In the simple circuit shown in Fig. 2-25, for example, it is not possible simply to add the resistance and reactance together to obtain a quantity that will indicate the opposition offered by the combination to the flow of current. Inasmuch as both resistance and reactance are present, the total effect can obviously be neither wholly one nor the other. In circuits containing *both* reactance and resistance the opposition effect is called **impedance.** The unit of impedance is also the ohm.

If the inductance in Fig. 2-25 were short-circuited, only the resistance would remain and the circuit would simply have a resistance of 75 ohms. In such a case the current and voltage would be in phase. On the other hand, if the resistance were short-circuited the circuit simply would have a reactance of 100 ohms, and the current would lag behind the voltage by one-quarter cycle or 90 degrees. When both are in the circuit, it would be expected that the impedance would be greater than either the resistance or reactance. It might also be expected that the current would be neither in phase with the voltage nor lagging 90 degrees behind it, but would be somewhere between the complete in-phase and the 90-degree phase conditions. Both things are true. The larger the reactance compared with the resistance, the more nearly the phase angle approaches 90 degrees; the larger the resistance compared to the reactance, the more nearly the current approaches the condition of being in phase with the voltage.

It can be shown that resistance and reactance can be combined in the same way that a right-angled triangle is constructed, if the resistance is laid off to proper scale as the base of the triangle and the reactance is laid off as the altitude to the same scale. This is also indicated in Fig. 2-25. When this is done the hypotenuse of the triangle represents the impedance of the circuit, to the same scale, and the angle between Z and R (usually called θ and so indicated in the drawing) is equal to the phase angle between the applied e.m.f. and the current. It is unnecessary, of course, actually to draw such a triangle when impedance is to be calculated; by geometry,

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

In the case shown in the drawing,

$$Z = \sqrt{(75)^2 + (100)^2} = \sqrt{15{,}625} = 125 \text{ ohms.}$$

The phase angle can be found from simple trigonometry. Its tangent is equal to X/R; in this case $X/R = 100/75 = 1.33$. From trigonometric tables it can be determined that the angle having a tangent equal to 1.33 is approximately 53 degrees. Fortunately, in ordinary amateur work it is seldom necessary to give much consideration to the phase angle because in most practical cases the angle will either be nearly zero (current and voltage in phase) or close to 90 degrees (current and voltage approximately a quarter cycle out of phase).

A circuit containing resistance and capacitance in series (Fig. 2-26) can be treated in the same way. That is, the impedance is

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

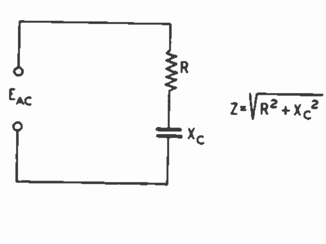
and the phase angle again is the angle whose tangent is equal to X/R. It must be remembered, however, that in this case the current *leads* the applied e.m.f., while in the resistance-inductance case it *lags* behind the voltage.

In neither case is the impedance of the circuit equal to the simple arithmetical sum of the resistance and reactance. With $R = 75$ ohms and $X_L = 100$ ohms, simple addition would give 175 ohms while the actual impedance is 125 ohms. However, if either X or R is very small compared to the other (say, 1/10 or less) the impedance is very nearly equal to the larger of the two quantities. For example, if $R = 1$ ohm and $X = 10$ ohms,

$$Z = \sqrt{R^2 + X^2} = \sqrt{(1)^2 + (10)^2}$$
$$= \sqrt{101} = 10.05 \text{ ohms.}$$

Hence if either X or R is at least 10 times as large as the other, the error in assuming that the impedance is equal to the larger of the two will not exceed ½ of 1 per cent, which is

Fig. 2-26 — Resistance and capacitive reactance in series.



usually negligible. This fact is frequently useful.

In working with impedance, remember that one of its components is reactance and that the reactance of a given coil or condenser changes with the applied frequency. Therefore, impedance also changes with frequency. The change in impedance as the frequency is changed may be very slow if the resistance is considerably larger than the reactance. However, if the impedance is mostly reactance a change in frequency will cause the impedance to change practically as rapidly as the reactance itself changes.

Ohm's Law for Impedance

Since impedance is made up of resistance and reactance, Ohm's Law can be applied to circuits containing impedance just as readily as to circuits having resistance or reactance only. The formulas are

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

$$E = IZ$$

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

where E = E.m.f. in volts
I = Current in amperes
Z = Impedance in ohms

Example: Assume that the e.m.f. applied to the circuit of Fig. 2-25 is 250 volts. Then

$$I = \frac{E}{Z} = \frac{250}{125} = 2 \text{ amperes.}$$

The same current is flowing in both R and X_L, and Ohm's Law as applied to either of these quantities says that the voltage drop across R should equal IR and the voltage drop across X_L should equal IX_L. Substituting,

$$E_R = IR = 2 \times 75 = 150 \text{ volts}$$
$$E_{X_L} = IX_L = 2 \times 100 = 200 \text{ volts}$$

The arithmetical sum of these voltages is greater than the applied voltage. However, the *actual* sum of the two when the phase relationship is taken into account is equal to 250 volts r.m.s., as shown by Fig. 2-27, where the instantaneous values are added throughout the cycle. Whenever resistance and reactance are in series, the individual voltage drops always add up, arithmetically, to more than the applied voltage. There is nothing fictitious about these voltage drops; they can be measured readily by suitable instruments. It is simply an illustration of the importance of phase in a.c. circuits.

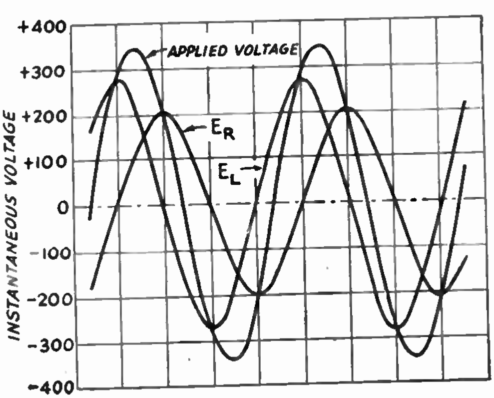


Fig. 2-27 — Voltage drops around the circuit of Fig. 2-25. Because of the phase relationships, the applied voltage is less than the arithmetical sum of the drops across the resistor and inductor.

A more complex series circuit, containing resistance, inductive reactance and capacitive reactance, is shown in Fig. 2-28. In this case it is necessary to take into account the fact that the phase angles between current and voltage differ in all three elements. Since it is a series circuit, the current is the same throughout. Considering first just the inductance and

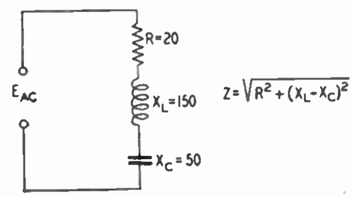


Fig. 2-28 — Resistance, inductive reactance, and capacitive reactance in series.

capacitance and neglecting the resistance, the phase relationships are the same as in Fig. 2-24. The net reactance in Fig. 2-28 is

$$X_L - X_C = 150 - 50 = 100 \text{ ohms (inductive)}$$

Since the series reactances can be lumped into one equivalent reactance, it is easy to find the impedance of the circuit by the rules previously given. The impedance of a circuit containing resistance, inductance and capacitance in series is

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

Example: In the circuit of Fig. 2-28, the impedance is

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$
$$= \sqrt{(20)^2 + (150 - 50)^2} = \sqrt{(20)^2 + (100)^2}$$
$$= \sqrt{10{,}400} = 102 \text{ ohms}$$

The phase angle can be found from X/R, where $X = X_L - X_C$.

Parallel Circuits

Suppose that a resistor, condenser and coil are connected in parallel as shown in Fig. 2-29 and an a.c. voltage is applied to the combination. In any one branch, the current will be unchanged if one or both of the other two branches is disconnected, so long as the applied voltage remains unchanged. For example, I_L, the current through the inductance, will not change if both R and C are removed (although the total current, I, *will* change). Thus the current in each branch can be calculated quite simply by the Ohm's Law

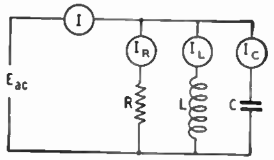


Fig. 2-29 — **Resistance, inductance and capacitance in parallel. Instruments connected as shown will read the total current, *I*, and the individual currents in the three branches of the circuit.**

formulas given in the preceding sections, if the voltage and reactance or resistance are known. The total current, I, is the sum of the currents through all three branches — not the arithmetical sum, but the sum when phase is taken into account.

The currents through the various branches will be as shown in Fig. 2-30, assuming for purposes of illustration that X_L is smaller than X_C and that X_C is smaller than R, thus making I_L larger than I_C, and I_C larger than I_R. The current through C leads the voltage by 90 degrees and the current through L lags the voltage by 90 degrees, so these two currents are 180 degrees out of phase. As shown at E, the total reactive current is the difference between I_C and I_L. This resultant current lags the voltage by 90 degrees, because I_L is larger than I_C. When the reactive current is added to I_R, the total current, I, is as shown at F. It can be seen that I lags the applied voltage by an angle smaller than 90 degrees and that the total current, while less than the simple sum (neglecting phase) of the three branch currents, is larger than the current through R alone.

The impedance looking into the parallel circuit from the source of voltage is equal to the applied voltage divided by the total or "line" current, I. In the case illustrated, I is greater than I_R, so the impedance of the circuit is less than the resistance of R. How much less depends upon the net reactive current flowing through L and C in parallel. If X_L and X_C are very nearly equal the net reactive current will be quite small because it is equal to the *difference* between two nearly equal currents. In such a case the impedance of the circuit will be almost the same as the resistance of R alone. On the other hand, if X_L and X_C are quite different the net reactive current can be relatively large and the total current also will be appreciably larger than I_R. In such a case the circuit impedance will be lower than the resistance of R alone.

The calculation of the impedance of parallel circuits is somewhat complicated. Fortunately, calculations are not necessary in most amateur work except in a special — and simple — case treated in a later section of this chapter.

Power Factor

In the circuit of Fig. 2-25 an applied e.m.f. of 250 volts results in a current of 2 amperes. If the circuit were purely resistive (containing no reactance) this would mean a power dissipation of 250 × 2 = 500 watts. However, the circuit actually consists of resistance and reactance, and only the resistance consumes power. The power in the resistance is

$$P = I^2R = (2)^2 \times 75 = 300 \text{ watts}$$

This is the actual power consumed by the circuit as compared to the apparent power input of 500 watts. The ratio of the power consumed to the apparent power is called the **power factor** of the circuit, and in the case used as an example would be 300/500 = 0.6. Power factor is frequently expressed as a percentage; in this case, the power factor would be 60 per cent.

"Real" or dissipated power is measured in watts; apparent power, to distinguish it from real power, is measured in volt-amperes (just like the "wattless" power in a reactance). It is simply the product of volts and amperes and has no direct relationship to the power actually used up or dissipated unless the power factor of the circuit is known. The power factor of a purely resistive circuit is 100 per cent or 1, while the power factor of a pure reactance is zero. In this illustration, the reactive power is

$$VA \text{ (volt-amperes)} = I^2X = (2)^2 \times 100 = 400 \text{ volt-amperes.}$$

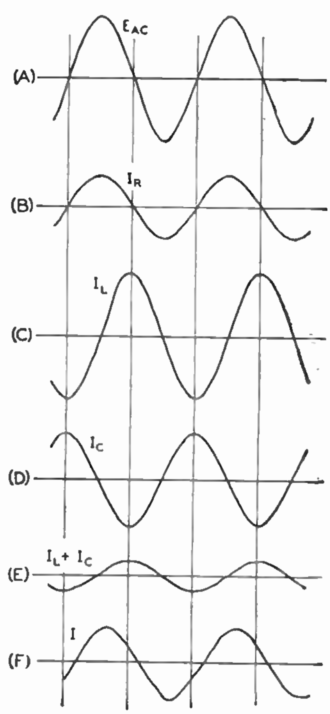


Fig. 2-30 — **Phase relationships between branch currents and applied voltage for the circuit of Fig. 2-29. The total current through *L* and *C* in parallel ($I_L + I_C$) and the total current in the entire circuit (*I*) also are shown.**

Complex Waves

It was pointed out early in this chapter that a complex wave (a "nonsinusoidal" wave) can be resolved into a fundamental frequency and a series of harmonic frequencies. When such a complex voltage wave is applied to a circuit containing reactance, the current through the circuit will not have the same waveshape as the applied voltage. This is because the reactance of a coil and condenser depend upon the applied frequency. For the second-harmonic component of a complex wave, the reactance of the coil is twice and the reactance of the condenser one-half their values at the fundamental frequency; for the third harmonic the coil reactance is three times and the condenser reactance one-third, and so on.

Just what happens to the current waveshape depends upon the values of resistance and reactance involved and how the circuit is arranged. In a simple circuit with resistance and inductive reactance in series, the amplitudes of the harmonics will be reduced because the inductive reactance increases in proportion to frequency. When a condenser and resistance are in series, on the other hand, the harmonics are likely to be accentuated because the condenser reactance becomes lower as the frequency is raised. When both inductive and capacitive reactance are present the shape of the current wave can be altered in a variety of ways, depending upon the circuit and the "constants," or values of L, C and R, selected.

This property of nonuniform behavior with respect to fundamental and harmonics is an extremely useful one. It is the basis of "filtering," or the suppression of undesired frequencies in favor of a single desired frequency or group of such frequencies.

Transformers

It has been shown in the preceding sections that, when an alternating voltage is applied to an inductance, an e.m.f. is induced by 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.

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.

Types of Transformers

The usefulness of the transformer lies in the fact that electrical 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, 115 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.

As shown in Fig. 2-31, the primary and secondary coils of a transformer may be wound on a core of magnetic material. This increases the inductance of the coils so that a relatively small number of turns may be used to induce a given value of voltage with a small current. A closed core (one having a continuous magnetic path) such as that shown in Fig. 2-31 also tends to insure that practically all of the field set up by the current in the primary coil will cut the turns of the secondary coil. However, the core introduces a power loss because of hysteresis and eddy currents so this type of construction is practicable only at power and audio frequencies. The discussion in this section is confined to transformers operating at such frequencies.

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 (which is the case when both are wound on the same closed core) 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 practi-

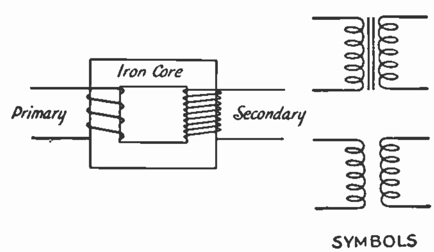


Fig. 2-31 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

cally equal to, and opposes, the applied voltage. Hence, for all practical purposes,

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

where E_s = Secondary voltage
E_p = Primary voltage
n_s = Number of turns on secondary
n_p = Number of turns on primary

The ratio n_s/n_p is called the **turns ratio** of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and 115 volts is applied to the primary. The secondary voltage will be

$$E_s = \frac{n_s}{n_p} E_p = \frac{2800}{400} \times 115 = 7 \times 115 = 805 \text{ volts}$$

Also, if 805 volts is applied to the 2800-turn winding (which then becomes the primary) the output voltage from the 400-turn winding will be 115 volts.

Either winding of a transformer can be used as the primary, *providing* the winding has enough turns to induce a voltage equal to the applied voltage without requiring an excessive current flow.

Effect of Secondary Current

The current that flows in the primary when no current is taken from the secondary is called the **magnetizing current** of the transformer. In any properly-designed transformer the primary inductance will be so large that the magnetizing current will be quite small. The power consumed by the transformer when the secondary is "open" — that is, not delivering power — is only the amount necessary to supply the losses in the iron core and in the resistance of the wire of which the primary is wound.

When current is drawn from the secondary winding, the secondary current sets up a magnetic field of its own in the core. The field from the secondary current always reduces the strength of the original field. But if the induced voltage in the primary is to equal the applied voltage, 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 additional current that sets up a field exactly equal to the field set up by the secondary current, but which opposes the secondary field. The additional primary current is thus 180 degrees out of phase with the secondary current.

In practical 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 the magnetizing current should be very small in comparison with the load current when the latter is near the rated value.

If the magnetic fields set up by the primary and secondary currents are to be equal, the primary current multiplied by the primary turns must equal the secondary current multiplied by the secondary turns. From this it follows that the primary current will be equal to the secondary current multiplied by the turns ratio, secondary to primary, or

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

where I_p = Primary current
I_s = Secondary current
n_p = Number of turns on primary
n_s = Number of turns on secondary

Example: Suppose that the secondary of the transformer in the previous example is delivering a current of 0.2 ampere to a load. Then the primary current will be

$$I_p = \frac{n_s}{n_p} I_s = \frac{2800}{400} \times 0.2 = 7 \times 0.2 = 1.4 \text{ amp.}$$

Although the secondary *voltage* is *higher* than the primary voltage, the secondary *current* is *lower* than the primary current, and by the same ratio.

Power Relationships; Efficiency

A transformer cannot create power; 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. There is always some power loss in the resistance of the coils and in the iron core, so in all practical cases the power taken from the source will exceed that taken from the secondary. Thus,

$$P_o = nP_i$$

where P_o = Power output from secondary
P_i = Power input to primary
n = Efficiency factor

The efficiency, n, always is less than 1. It is usually expressed as a percentage; if n is 0.65, for instance, the efficiency is 65 per cent.

Example: A transformer has an efficiency of 85% at its full-load output of 150 watts. The power input to the primary at full secondary load will be

$$P_i = \frac{P_o}{n} = \frac{150}{0.85} = 176.5 \text{ watts}$$

The efficiency of a transformer is usually — by design — highest at the normal power output for which it is rated. The efficiency decreases with either lower or higher outputs. On the other hand, the *losses* in the transformer are relatively small at low output but increase as more power is taken. The amount of power that the transformer can handle is determined

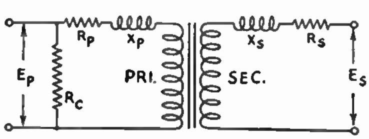


Fig. 2-32 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected in many approximate calculations.

by its own losses, because these heat the wire and core and raise the operating temperature. There is a limit to the temperature rise that can be tolerated, because too-high temperature either will melt the wire or break down the insulation between turns. A transformer always can be operated at reduced output even though the efficiency is low, because the actual loss also will be low under such conditions.

The full-load efficiency of small power transformers such as are used in radio receivers and transmitters usually lies 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 that "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 that is not coupled to another coil; that is, it causes an e.m.f. of self-induction. Consequently, there are small amounts 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 leakage inductance and the frequency. This reactance is called **leakage reactance.**

In the primary, the current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing primary current, hence it increases as more current is drawn from the secondary. The induced voltage consequently decreases, because the applied voltage has been reduced by the voltage drop in the primary leakage reactance. The secondary induced voltage also decreases proportionately.

When current flows in the secondary circuit the secondary leakage reactance causes an additional voltage drop that further reduces the voltage available from the secondary terminals. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the primary and secondary windings of the transformer also cause voltage drops when current is flowing; 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 from open-circuit conditions to full load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies because the leakage reactance increases directly with the frequency.

Impedance Ratio

In an ideal transformer — one without losses or leakage reactance — the following relationship is true:

$$Z_p = Z_s N^2$$

where Z_p = Impedance of primary as viewed from source of power

Z_s = Impedance of load connected to secondary

N = Turns ratio, primary to secondary

That is, a load of any given impedance connected to the *secondary* of the transformer will be changed to a different value "looking into" the *primary* from the source of power. The amount of impedance transformation is proportional to the square of the primary-to-secondary turns ratio.

Example: A transformer has a primary-to-secondary turns ratio of 0.6 (primary has 6/10 as many turns as the secondary) and a load of 3000 ohms is connected to the secondary. The impedance looking into the primary then will be

$$Z_p = Z_s N^2 = 3000 \times (0.6)^2 = 3000 \times 0.36 = 1080 \text{ ohms}$$

By choosing the proper turns ratio, the impedance of a fixed load can be transformed to any desired value, within practical limits. The transformed or "reflected" impedance has the same phase angle as the actual load impedance; if the load is a pure resistance the load presented by the primary to the source of power also will be a pure resistance.

The above relationship is sufficiently accurate in practice to give quite adequate results, even though it is based on an "ideal" transformer. Aside from the normal design requirements of reasonably low internal losses and low leakage reactance, the only other requirement to be met is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary. Despite a common — but mistaken — impression, a transformer operating with

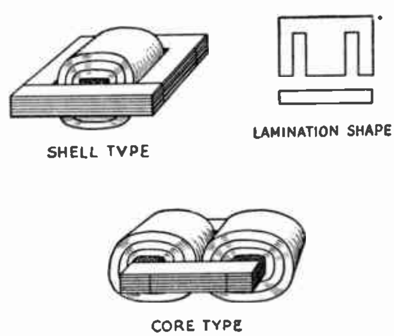


Fig. 2-33 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

a load does not "have" an impedance; the primary impedance — *as it looks to the source of power* — is determined by the load connected to the secondary and by the turns ratio. If the characteristics of the transformer have an appreciable effect on the impedance presented to the power source, the transformer is either poorly designed or is not suited to the voltage applied to it. Most transformers will operate quite well at voltages from slightly above to well below the design figure.

Impedance Matching

Many devices require a specific value of load resistance (or impedance) for optimum operation. The resistance of the actual load that is to dissipate the power may differ widely from this value; so the transformer is frequently called upon to transform the actual load into one of the desired value. This is called **impedance matching.** From the preceding,

$$N = \sqrt{\frac{Z_s}{Z_p}}$$

where N = Required turns ratio, secondary to primary

Z_s = Impedance of load connected to secondary

Z_p = Impedance required

Example: A vacuum-tube a.f. amplifier requires a load of 5000 ohms for optimum performance, and is to be connected to a loudspeaker having an impedance of 10 ohms. The turns ratio, secondary to primary, required in the coupling transformer is

$$N = \sqrt{\frac{Z_s}{Z_p}} = \sqrt{\frac{10}{5000}} = \sqrt{\frac{1}{500}} = \frac{1}{22.4}$$

The primary therefore must have 22.4 times as many turns as the secondary.

Impedance matching means, in general, adjusting the load impedance — by means of a transformer or otherwise — to a desired value. However, there is also another meaning. It is possible to show that any source of power will have its *maximum possible* output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. However, the efficiency is only 50 per cent in such a case; just as much power is used up in the source as is delivered to the load. Because of the poor efficiency, this type of impedance matching is limited to cases where only a small amount of power is available. Getting the most power output may be more important than efficiency in such a case.

Transformer Construction

Transformers usually are designed so that the magnetic path around the core is as short as possible. A short magnetic path means that the transformer will operate with fewer turns, for a given applied voltage, than if the path were long. It also helps to reduce flux leakage and therefore minimizes leakage reactance. The number of turns required also is affected by the cross-sectional area of the core. Transformer design data will be found in Chapter Seven.

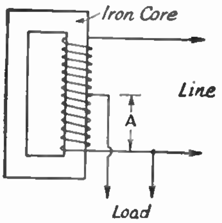


Fig. 2-34 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (A) flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

Two core shapes are in common use, as shown in Fig. 2-33. 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 one of the windings must operate at very high voltage.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of eddy currents. The laminations are overlapped at the ends to make the magnetic path as continuous as possible and thus reduce flux leakage.

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

Autotransformers

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 2-34; the principles just discussed apply equally well. A one-winding transformer is called an **autotransformer.** The section of the winding common to both the line and load circuits carries less current than the remainder of the coil, because the line and load currents are out of phase as explained previously. Hence the common section of the winding may be wound with comparatively small wire.

This advantage of the autotransformer is of practical value only when the primary (line) and secondary (load) voltages are not very different. On the other hand, it is frequently undesirable to have a direct connection between the primary and secondary circuits. For these reasons the autotransformer is used chiefly for boosting or reducing power-line voltage by relatively small amounts.

Radio-Frequency Circuits

RESONANCE

Fig. 2-35 shows a resistor, condenser and coil connected in series with a source of alternating current. Assume that the frequency can be varied over a wide range and that, at any frequency, the voltage of the source always has the same value.

At some *low* frequency the condenser reactance will be much larger than the resistance of R, and the inductive reactance will be small compared with either the reactance of C or the

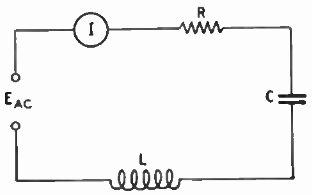


Fig. 2-35 — A series circuit containing L, C and R is "resonant" at the applied frequency when the reactance of C is equal to the reactance of L.

resistance of R. (The resistance, R, is assumed to be the same at all frequencies.) On the other hand, at some very *high* frequency the reactance of C will be very small and the reactance of L will be very large. In the *low*-frequency case the amount of current that can flow will be determined practically entirely by the reactance of C; since X_C is large at the low frequency, the current will be small. In the *high*-frequency case the amount of current that can flow will be determined almost wholly by the reactance of L; X_L is large at the high frequency so the current is again small.

Now condenser reactance *decreases* as the frequency is raised, but inductive reactance *increases* with frequency. At *some* frequency, therefore, the reactances of C and L will be equal. At that frequency the voltage drop across the coil equals the voltage drop across the condenser, and since the two drops are 180 degrees out of phase they cancel each other completely. At that frequency the amount of current flow is determined wholly by the resistance, R. Also, at that frequency the current has its largest possible value (remember that we assumed the source voltage to be constant regardless of frequency). A series circuit in which the inductive and capacitive reactances are equal is said to be **resonant**; or, to be "in resonance" or "in tune" at the frequency for which the reactances are equal.

Resonance is not peculiar to radio-frequency circuits alone. It can occur at any a.c. frequency, including power-line frequencies. However, resonant circuits are used principally at radio frequencies; in fact, at those frequencies the circuits used almost always are resonant.

Resonant Frequency

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 gives

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

where f = Frequency in cycles per second
L = Inductance in henrys
C = Capacitance in farads
π = 3.14

These units are inconveniently large for radio-frequency circuits. A formula using more appropriate units is

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

where f = Frequency in kilocycles (kc.)
L = Inductance in microhenrys (μh.)
C = Capacitance in micromicrofarads (μμfd.)
π = 3.14

Example: The resonant frequency of a series circuit containing a 5-μh. coil and a 35-μμfd. condenser is

$$f = \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}} = \frac{10^6}{6.28 \times 13.2} = \frac{10^6}{83} = 12{,}050 \text{ kc.}$$

The formula for resonant frequency is not affected by the resistance in the circuit.

Resonance Curves

If a plot is drawn of the current flowing in the circuit of Fig. 2-35 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-36. At frequencies very much higher than the resonant frequency the current is limited by the inductive reactance; the condenser and resistor have only a negligible part. At frequencies very much lower than resonance the condenser limits the current, the resistor and inductance playing very little part. Exactly at resonance the current is limited only by the resistance; the smaller the resistance the larger the resonant current. The shape of the **resonance curve** at frequencies *near* resonance is determined by the ratio of reactance to resistance at the particular frequency considered. If the reactance of either the coil or condenser is of the same order of magnitude as the resistance, the current decreases rather slowly as the frequency is moved in either direction away from resonance. Such a curve is said to be **broad**. On the other hand, if the reactance is considerably larger than the resistance the current decreases rapidly as the

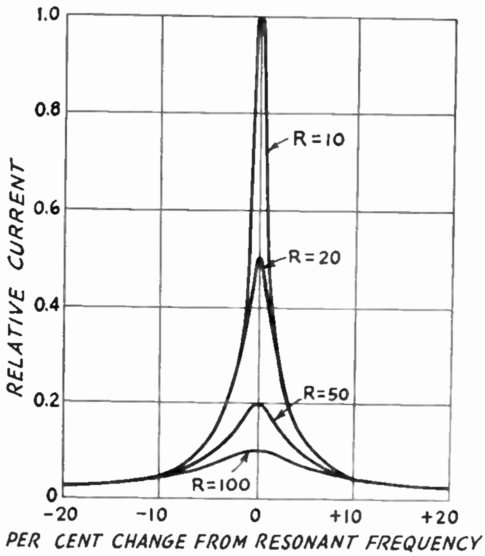


Fig. 2-36 — Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms (minimum $Q = 10$). Note that at frequencies at least plus or minus ten per cent away from the resonant frequency the current is substantially unaffected by the resistance in the circuit.

frequency moves away from resonance and the circuit is said to be **sharp.** Curves of differing sharpness are shown in Fig. 2-36. A sharp circuit will respond a great deal more readily to the resonant frequency than to frequencies quite close to resonance; a broad circuit will respond almost equally well to a group or band of frequencies centering around the resonant frequency.

Both types of resonance curves are useful. A sharp circuit gives good **selectivity** — the ability to select one desired frequency and discriminate against others. A broad circuit is used when the apparatus must give about the same response over a band of frequencies rather than to a single frequency alone.

Q

Most diagrams of resonant circuits show only inductance and capacitance; no resistance is indicated. Nevertheless, resistance is always present. At frequencies up to perhaps 30 Mc. this resistance is mostly in the wire of the coil. Above this frequency energy loss in the condenser (principally in the solid dielectric which must be used to form an insulating support for the condenser plates) becomes appreciable. This energy loss is equivalent to resistance. When maximum sharpness or selectivity is needed the object of design is to reduce the inherent resistance to the lowest possible value.

We mentioned above that the sharpness of the resonance curve is determined by the ratio of reactance to resistance. The value of the reactance of *either* the coil or condenser at the resonant frequency, divided by the resistance in the circuit, is called the Q (quality factor) of the circuit, or

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

where Q = Quality factor
X = Reactance of either coil or condenser, in ohms
R = Resistance in ohms

Example: The coil and condenser in a series circuit each have a reactance of 350 ohms at the resonant frequency. The resistance is 5 ohms. Then the Q is

$$Q = \frac{X}{R} = \frac{350}{5} = 70$$

Since the same current flows in R that flows in X, the Q of the circuit also is the ratio of the reactive power to the "real" power, or power dissipated in the resistance. The term "volt-ampere-to-watt" ratio or, when the power is large, "kva.-to-kw. ratio," therefore is sometimes used instead of "Q." To put it another way, the Q of the circuit is the ratio of the energy stored (in either the magnetic or electric field) to the energy dissipated as heat in the resistance.

The effect of Q on the sharpness of resonance of a circuit is shown by the curves of Fig. 2-37. In these curves the frequency change is shown in percentage above and below the resonant frequency. Qs of 10, 20, 50 and 100 are shown; these values cover much of the range commonly used in radio work.

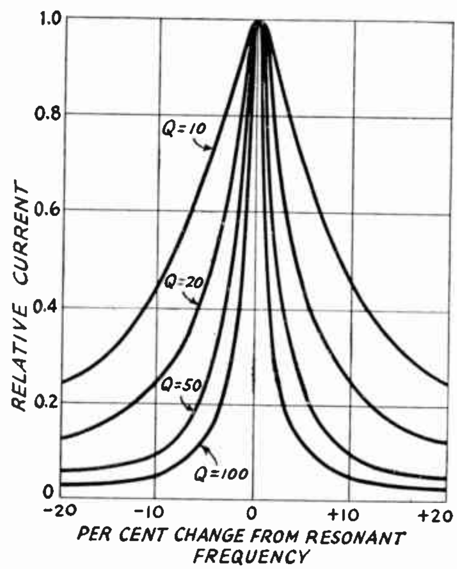


Fig. 2-37 — Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

Voltage Rise

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

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. The two reactances cancel and there will be but 5 ohms of pure resistance to limit the current flow. Thus the current will be 50/5, or 10 amperes. The voltage developed across either the coil or the condenser will be equal to its reactance times the current, or $200 \times 10 = 2000$ volts. An alternate method: The Q of the circuit is $X/R = 200/5 = 40$. The reactive voltage is equal to Q times the applied voltage, or $40 \times 50 = 2000$ volts.

Parallel Resonance

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-38 there is a resonance effect similar to that in a series circuit. However, in this case the current (measured at the point indicated) is *smallest* at the frequency for which the coil and condenser reactances are equal. At that frequency the current through L is exactly canceled by the out-of-phase current through C, as explained in an earlier section, so that only the current taken by R flows in the line. At frequencies *below* resonance the current through L is larger than that through C, because the reactance of L is smaller and that of C higher at low frequencies; there is only partial cancellation of the two reactive currents and the line current therefore is larger than the current taken by R alone. At frequencies *above* resonance the situation is reversed and more current flows through C than through L, so the line current again increases. The current at resonance, being determined wholly by R, will be small if R is large and large if R is small.

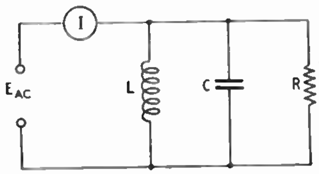


Fig. 2-38 — Circuit illustrating parallel resonance.

The resistance R shown in Fig. 2-38 seldom is an actual physical resistor. In most cases it will be an "equivalent" resistance that corresponds to the effect of an actual energy loss in the circuit. This energy loss can be inherent in the coil or condenser, or may represent energy transferred to a load by means of the resonant circuit. (For example, the resonant circuit may be used for transferring power from a vacuum-tube amplifier to an antenna system.)

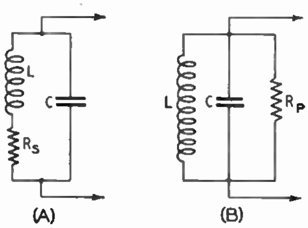


Fig. 2-39 — Series and parallel equivalents when the two circuits are resonant. The series resistor, R_s, in A can be replaced by an equivalent parallel resistor, R_p, in B, and vice versa.

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-39 will behave identically, when an external voltage is applied, if (1) L and C are the same in both cases; and (2) R_p multiplied by R_s equals the square of the reactance (at resonance) of either L or C. When these conditions are met the two circuits will have the same Qs. (These statements are approximate, but are quite accurate if the Q is 10 or more.) Now the circuit at A is a *series* circuit if it is viewed from the "inside" — that is, going around the loop formed by L, C and R — so its Q can be found from the ratio of X to R_s.

What this means is that a circuit like that of Fig. 2-39A has an equivalent **parallel impedance** (at resonance) equal to R_p, the relationship between R_s and R_p being as explained above. Although R_p is not an actual resistor, to the source of voltage the parallel-resonant circuit "looks like" a pure resistance of that value. It is "pure" resistance because the coil and condenser currents are 180 degrees out of phase and are equal; thus there is no reactive current. At the resonant frequency, then, the parallel impedance of a resonant circuit is

$$Z_r = QX$$

where Z_r = Resistive impedance at resonance
Q = Quality factor
X = Reactance (in ohms) of either the coil or condenser

Example: The parallel impedance of a circuit having a Q of 50 and having inductive and capacitive reactances of 300 ohms will be

$$Z_r = QX = 50 \times 300 = 15{,}000 \text{ ohms.}$$

At frequencies off resonance the impedance is no longer purely resistive because the coil and condenser currents are not equal. The off-resonant impedance therefore is complex, and is lower than the resonant impedance for the reasons previously outlined.

The higher the Q of the circuit, the higher the parallel impedance. Curves showing the variation of impedance (with frequency) of a parallel circuit have just the same shape as the curves showing the variation of current with frequency in a series circuit. Fig. 2-40 is a set of such curves.

Q of Loaded Circuits

In many applications of resonant circuits the only power lost is that dissipated in the resistance of the circuit itself. At frequencies below 30 Mc. most of this resistance is in the coil. Within limits, increasing the number of turns on the coil increases the reactance faster than it raises the resistance, so coils for circuits in which the Q must be high are made with relatively large inductance for the frequency under consideration.

However, 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 is shown in Fig. 2-41A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is at least ten times as great as the power lost in the coil and condenser, the parallel impedance of the resonant circuit itself will be so high compared with the resistance of the load that for all practical purposes the impedance of the combined circuit is equal to the load resistance. Under these conditions the Q of a parallel-resonant circuit loaded by a resistive impedance is

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

where Q = Quality factor
Z = Parallel load resistance (ohms)
X = Reactance (ohms) of either the coil or condenser

Example: A resistive load of 3000 ohms is connected across a resonant circuit in which the inductive and capacitive reactances are each 250 ohms. The circuit Q is then

$$Q = \frac{Z}{X} = \frac{3000}{250} = 12$$

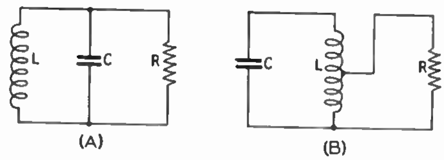


Fig. 2-41 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

The effective Q of a circuit loaded by a parallel resistance becomes higher when the reactances of the coil and condenser are decreased. A circuit loaded with a relatively low resistance (a few thousand ohms) must have low-reactance elements (large capacitance and small inductance) to have reasonably high Q.

The effect of a given load resistance on the Q of a circuit can be changed by connecting the load across only part of the circuit. A common method is to tap the load across part of the coil, as shown in Fig. 2-41B. 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. In high-frequency resonant circuits the impedance ratio does not vary exactly as the square of the turns 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.

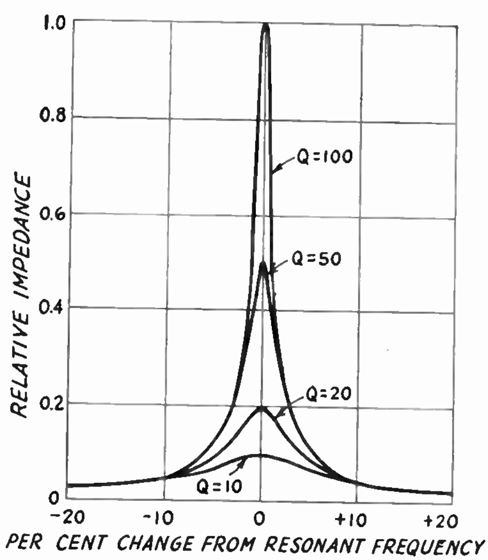


Fig. 2-40 — Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-37 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

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

As pointed out in the preceding paragraph, the product of inductance and capacity is constant for any given frequency. It is frequently convenient to use the numerical value of the *LC* **constant** when a number of calculations have to be made involving different L/C ratios for the same frequency. The constant for any frequency is given by the following equation:

$$LC = \frac{25{,}330}{f^2}$$

where L = Inductance in microhenrys (μh.)
C = Capacitance in micromicrofarads ($\mu\mu$fd.)
f = Frequency in megacycles.

Example: Find the inductance required to resonate at 3650 kc. (3.65 Mc.) with capacitances of 25, 50, 100, and 500 $\mu\mu$fd. The LC constant is

$$LC = \frac{25{,}330}{(3.65)^2} = \frac{25{,}330}{13.35} = 1900$$

With 25 $\mu\mu$fd. $L = 1900/C = 1900/25 = 76$ μh.
50 $\mu\mu$fd. $L = 1900/C = 1900/50 = 38$ μh.
100 $\mu\mu$fd. $L = 1900/C = 1900/100 = 19$ μh.
500 $\mu\mu$fd. $L = 1900/C = 1900/500 = 3.8$ μh.

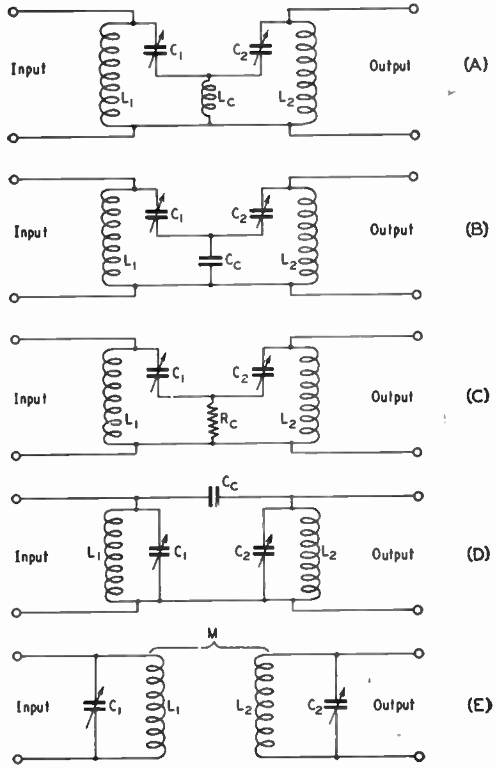


Fig. 2-42 — Basic methods of circuit coupling.

COUPLED CIRCUITS

Energy Transfer and Loading

Two circuits are **coupled** when energy can be transferred from one to the other. The circuit delivering power is called the **primary** circuit; the one receiving power is called the **secondary** circuit. The power may be practically all dissipated in the secondary circuit itself (this is usually the case in receiver circuits) or the secondary may simply act as a medium through which the power 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. The matching can be accomplished by adjusting the loading on the secondary and by varying the amount of coupling between the primary and secondary.

A general understanding of coupling methods is essential in amateur work, but there is seldom, if ever, need for *calculation* of the performance of coupled circuits. Very few radio amateurs have the equipment necessary for measuring the quantities that enter into such calculations. In actual practice, the adjustment of a coupled circuit is a cut-and-try process. Satisfactory results readily can be obtained if the principles are understood.

Coupling by a Common Circuit Element

One method of coupling between two resonant circuits is through a circuit element common to both. The three variations of this type of coupling shown at A, B and C of Fig. 2-42, utilize a common inductance, capacitance 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.

If both circuits are resonant to the same frequency, as is usually the case, the value of impedance — reactance or resistance — required for maximum energy transfer is generally quite small compared to the other reactances in the circuits. The common-circuit-element method of coupling is used only occasionally in amateur apparatus.

Capacitive Coupling

In the circuit at D the coupling increases as the capacitance of C_c, the "coupling condenser," is made greater (reactance of C_c is decreased). When two resonant circuits are coupled by this means, the capacitance 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, a reactance of 10,000 ohms or so in the condenser will give ample coupling. The corresponding capacitance required is only a few micromicrofarads at high frequencies.

Inductive Coupling

Fig. 2-42E shows inductive coupling, or coupling by means of the magnetic field. A circuit of this type resembles the iron-core transformer, but because only a small percentage of the magnetic 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.

Three common types of inductively-coupled circuits are shown in Fig. 2-43. In the first two, only one circuit actually is resonant. 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 that has both reactance and resistance.

In circuits A and B the coupling between the primary and secondary coils usually is "tight" — that is, the coefficient of coupling between the coils is large. With tight coupling either circuit operates much as though the device to which the untuned coil is connected were simply tapped across a corresponding number of turns on the tuned-circuit coil. 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 "coupled" resistance increases the effective series resistance of the tuned circuit, thereby lowering its Q and selectivity. 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. The coupled reactance makes it necessary to readjust the tuning whenever the coupling is changed, because coupled reactance tunes the circuit just as the actual coil and condenser reactance does.

These circuits may be used for impedance matching by adjusting the mutual inductance between the coils. This can be done by varying the coupling, changing the number of turns in the untuned coil, or both. The parallel impedance of the tuned circuit is affected by the coupled-in resistance in the same way as it would be by a corresponding increase in the actual series resistance. The larger the value of coupled-in resistance the lower the parallel impedance. By proper choice of the number of turns on the untuned coil, and by adjustment of the coupling, the parallel impedance of the tuned circuit may be adjusted to the value required for the proper operation of the device to which it is connected.

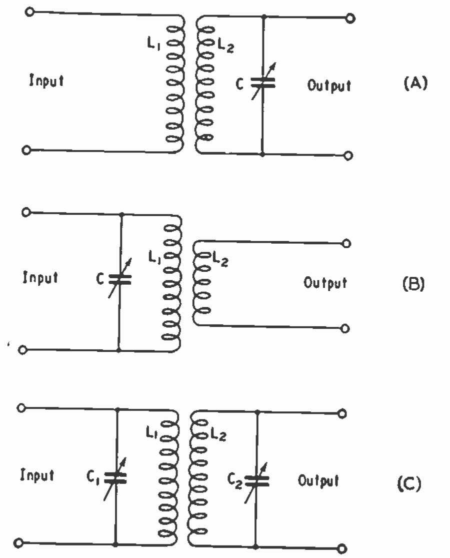


Fig. 2-43 — Types of inductively-coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.

Coupled Resonant Circuits

When the primary and secondary circuits are both tuned, as in Fig. 2-43C, the resonance effects in both circuits make the operation somewhat more complicated than in the simpler circuits just considered. Imagine first that the two circuits are not coupled and that each is independently tuned to the resonant frequency. The impedance of each will be purely resistive. If the two are then coupled, the secondary will couple resistance into the primary, causing its parallel impedance to decrease. As the coupling is made greater (without changing the tuning of either circuit) the coupled resistance becomes larger and the parallel impedance of the primary continues to decrease. Also, as the coupling is made tighter the amount of power transferred from the primary to the secondary will increase — but only up to a certain point. The power transfer becomes maximum at a "critical" value of coupling, but then decreases if the coupling is tightened beyond the critical point. At **critical coupling**, the resistance coupled into the primary circuit is equal to the resistance of the primary itself. This represents the matched-impedance condition and gives maximum power transfer.

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

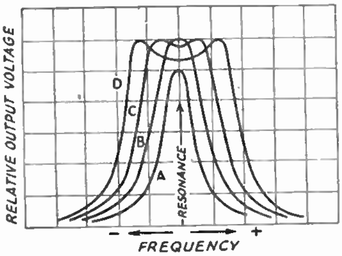


Fig. 2-44 — Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output voltage is measured across the secondary.

as physically close 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. 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.

Selectivity

In A and B, Fig. 2-43, only one circuit is tuned and the selectivity curve will be that of a single resonant circuit having the appropriate Q. As stated, the effective Q depends upon the resistance connected to the untuned coil.

In Fig. 2-43C, the selectivity is the same as that of a single tuned circuit having a Q equal to the *product* of the Qs of the individual circuits — *if* the coupling is below critical and both circuits are tuned to resonance. The Qs of the individual circuits are affected by the degree of coupling, because each couples resistance into the other; the tighter the coupling, the lower the individual Qs and therefore the lower the over-all selectivity.

If both circuits are independently tuned to resonance, the over-all selectivity will vary about as shown in Fig. 2-44 as the coupling is varied. At loose coupling, A, the output voltage (across the secondary circuit) is small and the selectivity is high. As the coupling is increased the secondary voltage also increases until critical coupling, B, is reached. At this point the output voltage at the resonant frequency is maximum but the selectivity is lower than with looser coupling. At still tighter coupling, C, the output voltage at the resonant frequency decreases, but as the frequency is varied either side of resonance it is found that there are two "humps" to the curve, one on either side of resonance. With very tight coupling, D, there is a further decrease in the output voltage at resonance and the "humps" are farther away from the resonant frequency. Resonance curves such as those at C and D are called **flat-topped** because the output voltage does not change much over an appreciable band of frequencies.

Note that the off-resonance humps have the same maximum value as the resonant output voltage at critical coupling. These humps are caused by the fact that at frequencies off resonance the secondary circuit is reactive and couples reactance as well as resistance into the primary. The coupled resistance decreases off resonance and the humps represent a new condition of impedance matching — at a frequency to which the primary is detuned by the coupled-in reactance from the secondary.

When the two circuits are tuned to slightly *different* frequencies a double-humped resonance curve results even though the coupling is below critical. This is to be expected, because each circuit will respond best to the frequency to which it is tuned. Tuning of this type is called **stagger tuning,** and often is used when substantially uniform response over a wide band of frequencies is desired.

Link Coupling

A modification of inductive coupling, called **link coupling,** is shown in Fig. 2-45. This gives the effect of inductive coupling between two coils that have no mutual inductance; the link is simply a means for providing the mutual inductance. The total mutual inductance between two coils coupled by a link cannot be made as great as if the coils themselves were coupled. This is because the coefficient of coupling between air-core coils is considerably less than 1, and since there are two coupling points the over-all coupling coefficient is less than for any *pair* of coils. In practice this need not be disadvantageous because the power transfer can be made great enough by making the tuned circuits sufficiently high-Q. Link coupling is convenient when ordinary inductive coupling would be impracticable for constructional reasons. It finds wide use in transmitters, for example.

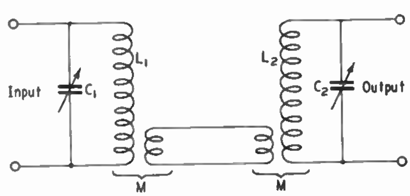


Fig. 2-45 — Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

The link coils usually have a small number of turns compared with the resonant-circuit coils. The number of turns is not greatly important, because the coefficient of coupling is relatively independent of the number of turns on either coil; it is more important that both link coils should have about the *same* number of turns. The length of the link between the coils is not critical if it is very small compared with the wavelength; if the length becomes an appreciable fraction of a wavelength the link operates more as a transmission line than as a means for providing mutual inductance. In such case it should be treated by the methods described in Chapter Ten.

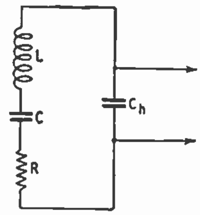


Fig. 2-46 — Equivalent circuit of a crystal resonator. *L*, *C* and *R* are the electrical equivalents of mechanical properties of the crystal; C_h is the capacitance of the electrodes with the crystal plate between them.

Piezoelectric Crystals

A number of crystalline substances found in nature have the ability to transform mechanical strain into an electrical charge, and vice versa. This property is known as piezoelectricity. A small plate or bar cut in the proper way from a quartz crystal, for example, and placed between two conducting electrodes, will be mechanically strained when the electrodes are connected to a source of voltage. Conversely, if the crystal is squeezed between two electrodes a voltage will develop between the electrodes.

Piezoelectric crystals can be used to transform mechanical energy into electrical energy, and vice versa. They are used, for example, in microphones and phonograph pick-ups, where mechanical vibrations are transformed into alternating voltages of corresponding frequency. They are also used in headsets and loudspeakers, transforming electrical energy into mechanical vibration. Crystal plates for these purposes are cut from large crystals of Rochelle salts.

Crystalline plates also are mechanical vibrators that have natural frequencies of vibration ranging from a few thousand cycles to several megacycles per second. The vibration frequency depends on the kind of crystal, the way the plate is cut from the natural crystal, and on the dimensions of the plate. Such a crystal is, in fact, the mechanical counterpart of an electrical tuned circuit; its resonant frequency is the natural frequency of the mechanical vibration. Because of the piezoelectric effect, the crystal plate can be coupled to an electrical circuit and made to substitute for a coil-and-condenser resonant circuit. The thing that makes crystals valuable as "resonators" is the fact that they have extremely high *Q*, ranging from 5 to 10 times the *Q*s obtainable with *LC* resonant circuits.

Analogies can be drawn between various mechanical properties of the crystal and the electrical characteristics of a tuned circuit. This leads to an "equivalent circuit" for the crystal. The electrical coupling to the crystal is through the electrodes between which it is sandwiched; these electrodes form, with the crystal as the dielectric, a small condenser like any other condenser constructed of two plates with a dielectric between. The crystal itself is an equivalent to a series-resonant circuit, and together with the capacitance of the electrodes forms the equivalent circuit shown in Fig. 2-46. The equivalent inductance of the crystal is extremely large and the series capacitance, *C*, is correspondingly low; this is the reason for the high *Q* of a crystal. The electrode capacitance, C_h, is so very large compared with the series capacitance of the crystal that it has only a very small effect on the resonant frequency. It will be realized, also, that because C_h is so large compared with *C* the electrical coupling to the crystal is quite loose.

Crystal plates for use as resonators in radio-frequency circuits are almost always cut from quartz crystals, because quartz is by far the most suitable material for this purpose. Quartz crystals are used as resonators in receivers, to give highly-selective reception, and as frequency-controlling elements in transmitters.

Practical Circuit Details

● COMBINED A.C. AND D.C.

Most radio circuits are built around vacuum tubes, and it is the nature of these tubes to require direct current (usually at a fairly high voltage) for their operation. They *convert* the direct current into an alternating current (and sometimes the reverse) at frequencies varying from ones well down in the audio range to well up in the superhigh range. The conversion process almost invariably requires that the direct and alternating currents meet somewhere in the circuit.

In this meeting, the a.c. and d.c. are actually combined into a single current that "pulsates" (at the a.c. frequency) about an *average* value equal to the direct current. This is shown in Fig. 2-47. It is easier, though, to think of them separately and to consider that the alternating current is superimposed on the direct current. Thus we look upon the actual current as having two components, one d.c. and the other a.c.

If the alternating current is a sine wave, its positive and negative alternations have the same maximum amplitude. When the wave is superimposed on a direct current the latter is alternately increased and decreased *by the same amount*. There is thus no *average* change in the direct current. If a d.c. instrument is being used to read the current, the reading will be exactly the same whether or not the sine-wave a.c. is superimposed.

However, there is actually more *power* in such a combination current than there is in the direct current alone. This is because power

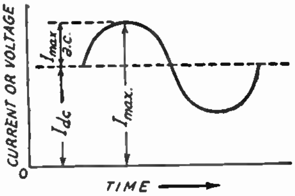


Fig. 2-47 — Pulsating current, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

varies as the *square* of the instantaneous value of the current, so more power is added to the circuit on the half-cycle of the a.c. wave that *increases* the instantaneous current than is subtracted on the half-cycle that *decreases* it. If the peak value of the alternating current is just equal to the direct current, the average power in the circuit is 1.5 times the power in the direct current alone.

In many circuits, also, we may have two alternating currents of different frequencies; for example, an audio frequency and a radio frequency may be combined in the same circuit. The two in turn may be combined with a direct current. In some cases, too, two r.f. currents of widely-different frequencies may be combined in the same circuit.

Series and Parallel Feed

Fig. 2-48 shows in simplified form how d.c. and a.c. may be combined in a vacuum-tube circuit. (The tube is shown only in bare outline; so far as the d.c. is concerned, it can be looked upon as a resistance of rather high value. On the other hand, the tube may be looked upon as the *generator* of the a.c. The mechanism of tube operation is described in the next chapter.) In this case, we have assumed that the a.c. is at radio frequency, as suggested by the coil-and-condenser tuned circuit. We also assume that r.f. current can easily flow through the d.c. supply; that is, the impedance of the supply at radio frequencies is so small as to be negligible.

In the circuit at the left, the tube, tuned circuit, and d.c. supply all are connected in series. The direct current flows through the r.f. tuned circuit to get to the tube; the r.f. current generated by the tube flows through the d.c. supply to get to the tuned circuit. This is **series feed.** It works because the impedance of the d.c. supply at radio frequencies is so low that it does not affect the flow of *r.f.* current, and because the d.c. resistance of the coil is so low that it does not affect the flow of *direct* current.

In the circuit at the right the direct current does not flow through the r.f. tuned circuit, but instead goes to the tube through a second coil, *RFC* **(radio-frequency choke).** Direct current cannot flow through *L* because a **blocking condenser,** *C*, is placed in the circuit to prevent it. (Without *C*, the d.c. supply would be short-circuited by the low resistance of *L*.) On the other hand, the r.f. current generated by the tube can easily flow through *C* to the tuned circuit because the capacitance of *C* is intentionally chosen to have low reactance (compared with the impedance of the tuned circuit) at the radio frequency. The r.f. current cannot flow through the d.c. supply because the inductance of *RFC* is intentionally made so large that it has a very high reactance at the radio frequency. The resistance of *RFC*, however, is too low to have an appreciable effect on the flow of direct current. The two currents are thus in *parallel*, hence the name **parallel feed.**

Both types of feed are in use. They may be used for both a.f. and r.f. circuits. In parallel feed there is no d.c. voltage on the a.c. circuit (the blocking condenser prevents that); this is a desirable feature from the viewpoint of safety to the operator, because the voltages applied to tubes — particularly transmitting tubes — are dangerous to human beings. On the other hand, it is somewhat difficult to make an r.f. choke work well over a wide range of frequencies. Series feed is usually preferred, therefore, because it is relatively easy to keep the impedance between the a.c. circuit and the tube low.

By-Passing

In the series-feed circuit just discussed, it was assumed that the d.c. supply had very low impedance at radio frequencies. This is not likely to be true in a practical power supply — if for no other reason than that the normal physical separation between the supply and the r.f. circuit would make it necessary to use rather long connecting wires or leads. At radio frequencies, even a few feet of wire can have fairly large reactance — too large to be considered a really "low-impedance" connection.

To get around this, an actual circuit would be provided with a **by-pass condenser,** as shown in Fig. 2-49. Condenser *C* is chosen to have low reactance at the operating frequency, and is installed right in the circuit where it can be wired to the other parts with quite short connecting wires. (The condenser will be an open circuit for the d.c. voltage across which it is connected, of course.) Since condenser *C* offers a low-impedance path, the r.f. current will tend to flow through it rather than through the d.c. supply; thus the current is confined to a known path rather than one of dubious impedance through the power supply.

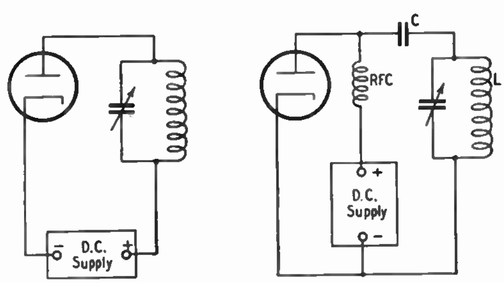


Fig. 2-48 — Illustrating series and parallel feed.

To be effective, a by-pass should have very low impedance compared to the impedance of the circuit element around which it is supposed to shunt the current. The reactance of the condenser should not be more than one-tenth of the impedance of the by-passed part of the circuit. Very often the latter impedance is not known, in which case it is desirable to use the largest capacitance in the by-pass that circumstances permit. To make doubly sure that r.f. current will not flow through a non-r.f. circuit such as a power supply, an r.f. choke may be connected in the lead to the latter, as shown in Fig. 2-49. The choke, having high reactance, will prevent the r.f. from going where it is not wanted and thereby ensure that it goes where it *is* wanted — i.e., through the by-pass condenser.

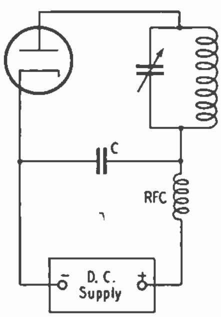


Fig. 2-49 — Typical use of a by-pass condenser in a series-feed circuit.

The use of a by-pass condenser is not confined only to circuits where r.f. is to be kept out of a d.c. source. The same type of by-passing is used when audio frequencies are present in addition to r.f. Because the reactance of a condenser changes with frequency, it is readily possible to choose a capacitance that will represent a very low reactance at radio frequencies but that will have such high reactance at audio frequencies that it is practically an open circuit. A capacitance of 0.001 μfd. is practically a short-circuit for r.f., for example, but is almost an open circuit at audio frequencies. (The actual value of capacitance that is usable will be modified by the impedances concerned.)

By-pass condensers also are used in audio-frequency circuits, to carry the audio frequencies around a d.c. supply. In this case a capacitance of several microfarads is needed if the reactance is to be low enough at the lower audio frequencies.

Distributed Capacitance and Inductance

In the discussions earlier in this chapter it was assumed that a condenser has only capacitance and that a coil has only inductance. Unfortunately, this is not strictly true. There is always a certain amount of inductance in a conductor of any length, and since a condenser is made up of conductors it is bound to have a little inductance in addition to its intended capacitance. Also, there is always capacitance between two conductors or between parts of the same conductor, and so we find that there is appreciable capacitance between the turns of an inductance coil.

This distributed inductance in a condenser and the distributed capacitance in a coil have important practical effects. Actually, every condenser is a tuned circuit, resonant at the frequency where its capacitance and distributed inductance have the same reactance. The same thing is true of a coil and its distributed capacitance. At frequencies well below these "natural" resonances, the condenser will act like a normal capacitance and the coil will act like a normal inductance. Near the natural resonant points, the coil and condenser act like self-tuned circuits. Above resonance, the condenser acts like an inductance and the coil acts like a condenser. If we want our circuit components to behave properly, they must always be used at frequencies well on the low side of their natural resonances.

Because of these effects, there is a limit to the amount of capacitance that can be used at a given frequency. There is a similar limit to the inductance that can be used. At audio frequencies, capacitances measured in microfarads and inductances measured in henrys are practicable. At low and medium radio frequencies, inductances of a few millihenrys and capacitances of a few thousand micromicrofarads are the largest practicable. At high radio frequencies, usable inductance values drop to a few microhenrys and capacitances to a few hundred micromicrofarads.

Distributed capacitance and inductance are important not only in r.f. tuned circuits, but in by-passing and choking as well. It will be appreciated that a by-pass condenser that actually acts like an inductance, or an r.f. choke that acts like a condenser, cannot work as it is intended they should. That is why you will find, in the circuits described later in this *Handbook*, by-pass condenser capacitances and r.f.-choke inductances that may look rather small — considering that, theoretically, a larger condenser or larger coil should be even more effective at its job.

Grounds

Throughout this book you will find frequent references to **ground** and **ground potential.** When a connection is said to be "grounded" it does not mean that it actually goes to earth (although in many cases such earth connections *are* used). What it means, more often, is that an actual earth connection *could* be made to that point in the circuit without disturbing the operation of the circuit in any way. The term also is used to indicate a "common" point in the circuit where power supplies and metallic supports (such as a metal chassis) are electrically tied together. It is customary, for example, to "ground" the negative terminal of a d.c. power supply, and to "ground" the filament or heater power supplies for vacuum tubes. Since the cathode of a vacuum tube is a junction point for grid and plate voltage supplies, it is a natural point to "ground." Also, since the various circuits connected to the tube elements have at least one point connected to cathode, these points also are "returned to ground."

"Ground" is therefore a common reference point in the circuit. In circuit diagrams, it is customary (for the sake of making the diagrams easier to read) to show such common connections by the ground symbol rather than by showing a large number of wires all connected together.

"Ground potential" means that there is no "difference of potential" — that is, no voltage — between the circuit point and the earth. A direct earth connection at such a point would cause no disturbance to the operation of the circuit.

Single-Ended and Balanced Circuits

With reference to ground, a circuit may be either single-ended (unbalanced) or balanced. In a single-ended circuit, one *side* of the circuit is connected to ground. In a balanced circuit, the *electrical midpoint* of the circuit is connected to ground, so that the circuit has two ends each at the same voltage "above" ground. A balanced circuit also is called a "symmetrical" circuit.

Typical single-ended and balanced circuits are shown in Fig. 2-50. R.f. circuits are shown in the upper line, while iron-core transformers (such as are used in power-supply and audio circuits) are shown in the lower line. The r.f. circuits may be balanced either by connecting the center of the coil to ground or by using a "balanced" or "split-stator" condenser — that is, one having two identical sets of stator and rotor plates with the rotor plates on the same shaft — and connecting the condenser rotor to ground. In the iron-core transformer, one or both windings may be tapped at the center of the winding to provide the ground connection.

In the single-ended circuit, only one side of the circuit is "hot" — that is, has a voltage that differs from ground potential. In the balanced circuit, both ends are "hot" and the grounded center point is "cold" — that is, at ground potential. The applications of both types of circuits are discussed in later chapters.

Nonlinear Circuits; Beats

The circuits that have been discussed in this chapter are, essentially, ones obeying Ohm's Law. That is, an increase or decrease of the applied voltage causes an exactly proportional increase or decrease in current. (This neglects relatively minor effects such as the temperature rise and consequent change in resistance of conductors with increasing current, etc.) However, many devices (such as vacuum tubes under some conditions of operation) do not obey any such straightforward rules. There may be no current flow at all with an applied voltage of one polarity, but the current may be large if the polarity of the voltage is reversed. Also, the current may increase with increasing voltage up to a certain point and then stay at a fixed value no matter how much more the voltage is raised. Such devices, and the circuits in which they are used, are called **nonlinear.**

One important result of nonlinearity is the behavior of the circuit when two or more alternating currents of different frequencies are flowing in it. In a normal circuit, the two frequencies will have no particular effect on each other. However, if two (or more) alternating currents of different frequencies are present in a nonlinear circuit, additional currents having frequencies equal to the sum, and difference, of the original frequencies will be set up. 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 nonlinear circuit 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.

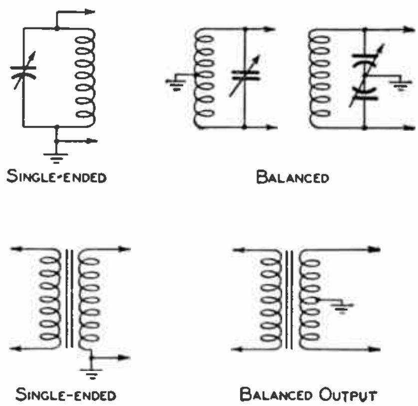


Fig. 2-50 — Single-ended and balanced circuits.

Beat frequencies are generated, and used to advantage, in very many radio circuits. For example, all of our modern reception methods are based on the use of beat frequencies.

Shielding

Two circuits that are physically near each other usually will be coupled to each other in some degree even though no coupling is intended. The metallic parts of the two circuits form a small capacitance through which energy can be transferred by means of the electric field. Also, the magnetic field about the coil or wiring of one circuit can couple that circuit to a second through the latter's coil and wiring. In many cases these unwanted couplings must be prevented if the circuits are to work properly.

Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called **shields.** The electric field from the circuit components does not penetrate the shield, because the lines of force are short-circuited by the metal. A metallic plate, called a **baffle shield**, inserted between two components also may suffice to prevent electrostatic coupling between them. Very little of the field tends to bend around such a shield if it is large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. In this case the magnetic field induces a current in the shield; this current in turn sets up its own magnetic field opposing the original field. The amount of current induced is proportional to the frequency and also to the conductivity of the shield; therefore 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.

Shielding a coil reduces its inductance, because part of its field is canceled. Also, there is always a small amount of resistance in the shield, and there is therefore an energy loss. This loss raises the effective resistance of the coil. The decrease in inductance and increase in resistance lower the Q of the coil. The reduction in inductance and Q will be small if the shield is sufficiently far away from the coil; the spacing between the sides of the coil and the shield should be at least half the coil diameter, and the spacing at the ends of the coil should at least equal the coil diameter. The higher the conductivity of the shield material, the less the effect on the inductance and Q. Copper is the best material, but aluminum is quite satisfactory.

At low (audio) frequencies this type of magnetic shielding does not work, because the current induced in the shield is too small. For good shielding at audio frequencies it is necessary to enclose the coil in a container of high-permeability iron or steel. This provides a much better path for the magnetic flux than air — so much so that most of the stray flux stays in the iron in preference to spreading out in the space around the coil. In this case the shield can be quite close to the coil without harming its performance.

CHAPTER 3

Vacuum-Tube Principles

Present-day methods of radio communication rely heavily on the vacuum tube. The tube is used to generate radio-frequency power, to amplify it in transmitters, to amplify and detect weak radio signals picked up from distant stations, to magnify the human voice, to change alternating current into direct current for power supplies — in fact, to do innumerable things that, without it, could not be done. An understanding of vacuum-tube principles is just as necessary to the radio amateur as an understanding of the circuit principles discussed in Chapter Two.

In this chapter we shall confine ourselves to the *fundamentals* of vacuum-tube operation. The special circuits and special types of tubes that find application in amateur radio will be taken up in later chapters.

The operation of vacuum tubes can be predicted mathematically, just as the operation of circuits can be predicted from mathematical formulas. It happens, though, that the amateur rarely has need to perform any calculations in connection with vacuum tubes, other than simple ones having to do with the power supplies for the tube elements. These are straightforward applications of Ohm's Law. Tube manufacturers invariably supply sets of data that give optimum operating conditions for their tubes, and thus save any need for calculation. What you need, to get the most out of your tubes, is mostly a picture of how they work.

Diodes and Rectification

● CURRENT IN A VACUUM

The outstanding difference between the vacuum tube and most other electrical devices is that the electric current does not flow through a conductor but through empty space — a vacuum. This is only possible when "free" electrons — that is, electrons that are not attached to atoms — are somehow introduced into the vacuum. It will be recalled from Chapter Two that electrons are particles of negative electricity. Free electrons in an evacuated space therefore can be attracted to a positively-charged object within the same space, or can be repelled by a negatively-charged object. The movement of the electrons under the attraction or repulsion of such charged objects constitutes the current in the vacuum.

The most practical way to introduce a sufficiently-large number of electrons into the evacuated space is by **thermionic emission.**

Thermionic Emission

If a thin wire or filament is heated to incandescence in a vacuum, electrons near the surface are given enough energy of motion to fly off into the surrounding space. The higher the temperature, the greater the number of electrons **emitted.** A more general name for the filament is **cathode.**

If the cathode is the only thing in the vacuum, most of the emitted electrons stay in its immediate vicinity, forming a "cloud" about the cathode. The reason for this is that the electrons in the space, being negative electricity, form a negative charge **(space charge)** in the region of the cathode. The negatively-charged space repels those electrons nearest the cathode, tending to make them fall back on it.

Now suppose a second conductor is introduced into the vacuum, but not connected to anything else inside the tube. If this second conductor is given a positive charge with respect to the cathode, electrons in the space will be attracted to the positively-charged conductor. The conductor can be given the requisite charge by connecting a source of e.m.f. between it and the cathode, as indicated in Fig. 3-1. The electrons emitted by the cathode and attracted to the positively-

charged conductor then constitute an electric current, with the circuit completed through the source of e.m.f. In Fig. 3-1 this e.m.f. is supplied by a battery (**"B" battery**); a second battery (**"A" battery**) is also indicated for heating the cathode or filament to the proper operating temperature.

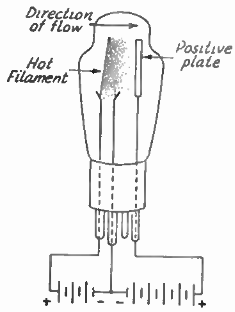


Fig. 3-1 — Conduction by thermionic emission in a vacuum tube. One battery is used to heat the filament to a temperature that will cause it to emit electrons. The other battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons captured by the plate flow back through the battery to the filament.

The positively-charged conductor is usually a metal plate or cylinder (surrounding the cathode) and is called an **anode** or **plate.** Like the other working parts of a tube, it is a **tube element** or **electrode.** The tube shown in Fig. 3-1 is a **two-element** or **two-electrode** tube, one element being the cathode or filament and the other the anode or plate.

Since electrons are *negative* electricity, they will be attracted to the plate *only* when the plate is positive with respect to the cathode. If the plate is given a negative charge, the electrons will be repelled back to the cathode and no current will flow in the vacuum. The vacuum tube therefore can conduct *only in one direction.*

Cathodes

Before electron emission can occur, the cathode must be heated to a high temperature. The only satisfactory way to heat it is by electricity. However, it is not essential that the heating current flow through the actual metal that does the emitting. The filament or heater can be electrically separate from the emitting cathode, and very many tubes are built that way. Such a cathode is called **indirectly heated,** while an emitting filament is called **directly heated.** Fig. 3-2 shows both types in the forms in which they are commonly used.

Obviously, the cathode should emit as many electrons as possible with the least possible heating power. A plain metal cathode is quite inefficient in this respect. Much greater electron emission can be obtained, at relatively low temperatures, by using special cathode materials. One of these is **thoriated tungsten,** or tungsten in which thorium is dissolved. Still greater efficiency is achieved in the **oxide-coated** cathode, a cathode in which rare-earth oxides form a coating over a metal base.

Although the oxide-coated cathode has much the highest efficiency, it can be used successfully only in tubes that operate at rather low plate voltages. Its use is therefore confined to receiving-type tubes and to the smaller varieties of transmitting tubes. The thoriated filament, on the other hand, will operate well in high-voltage tubes and is therefore found in most of the transmitting types used by amateurs.

Plate Current

The number of electrons attracted to the plate depends upon the strength of the positive charge on the plate — that is, on the amount of voltage between the cathode and plate. The electron current — called the **plate current** — increases as the plate voltage is increased (although the relationship is not the simple proportionality of Ohm's Law). Actually, this statement is true only up to a certain point; if the plate voltage is made high enough, *all* the electrons emitted by the cathode would be attracted to the plate. Obviously, when this occurs, a further increase in plate voltage cannot cause an increase in plate current.

Fig. 3-3 shows a typical plot of plate current with increasing plate voltage for a two-element tube or **diode.** A curve of this type can be obtained with the circuit shown, if the plate voltage can be increased in small steps and a current reading taken (by means of the current-indicating instrument — a "milliammeter") at each voltage. The plate current is zero with no plate voltage and the curve rises almost in a straight line until a "saturation point" is reached. This is where the positive

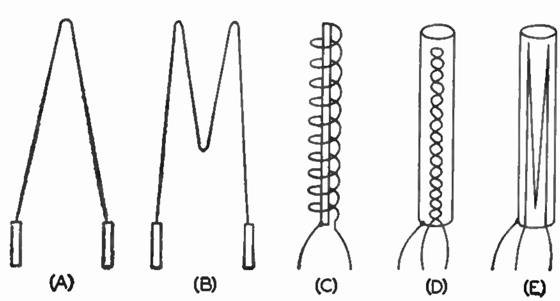


Fig. 3-2 — Types of cathode construction. Directly-heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and 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.

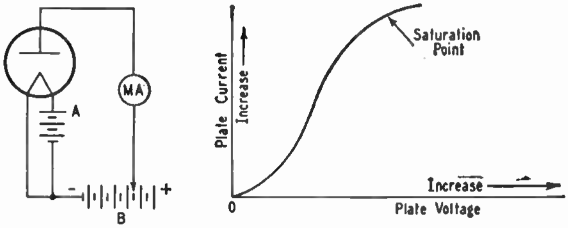
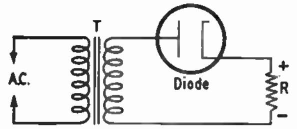


Fig. 3-3 — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

charge on the plate has completely overcome the space charge and practically all the electrons are going to the plate. At any higher voltages the plate current stays at the same value.

The curve of Fig. 3-3 does not show actual values of plate voltage and plate current, since these will vary with the type of tube. The *shape* of the curve, however, is typical of all diodes.

The plate voltage multiplied by the plate current is the *power input* to the tube. In a circuit like that of Fig. 3-3 this power is all used in heating the plate. If the power input is large, the plate temperature may rise to a very high value (the plate may become red or even white hot). The heat developed in the plate is radiated to the bulb of the tube, and in turn radiated by the bulb to the surrounding air.



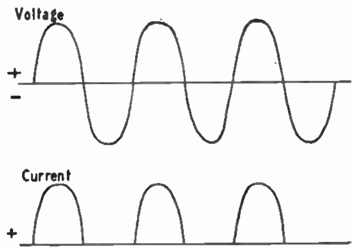


Fig. 3-4 — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

RECTIFICATION

Since current can flow through a tube in only one direction, a diode can be used to change alternating current into direct current. It does this by permitting current to flow when the plate is positive with respect to the cathode, but by shutting off current flow when the plate is negative.

Fig. 3-4 shows a representative circuit. Alternating voltage from the secondary of the transformer, *T*, is applied to the diode tube in series with a **load resistor,** *R*. The voltage varies as is usual with a.c., but current flows through the tube and *R only* when the plate is positive with respect to the cathode — that is, during the half-cycle when the upper end of the transformer winding is positive. During the negative half-cycle there is simply a gap in the current flow. This **rectified** alternating current therefore is an *intermittent* direct current. (The "humps" in the output current may be smoothed out by a "filter." A filter uses inductance and capacitance to store up energy during the time that current flows through the diode, energy that is then released to the circuit during the period when the diode is nonconducting. Filters of this type are discussed in later chapters.)

The load resistor, *R*, represents the actual circuit in which the rectified alternating current does work. All tubes work into a load of one type or another; in this respect a tube is much like a generator or transformer. A circuit that did not provide a load for the tube would be like a short-circuit across a transformer; no useful purpose would be accomplished and the only result would be the generation of heat in the transformer. So it is with vacuum tubes; they must *deliver* power to a load in order to serve a useful purpose. Also, to be *efficient* most of the power must do useful work in the load and not be used in heating the plate of the tube. This means that most of the voltage should appear as a drop across the load rather than as a drop between the plate and cathode of the diode. That is, the "resistance" of the tube should be small compared to the resistance of the load.

Notice that, with the diode connected as shown in Fig. 3-4, the polarity of the voltage drop across the load is such that the end of the load nearest the cathode is positive. If the connections to the diode elements are reversed, the direction of rectified current flow also will be reversed through the load.

Vacuum-Tube Amplifiers

TRIODES

Grid Control

It was shown in Fig. 3-3 that, within the normal operating range of a tube, the plate current will increase when the plate voltage is increased. The reason why all the electrons are not drawn to the plate when a *small* positive voltage is placed on it is that the space charge (which is negative) counteracts the effect of the positive charge on the plate. The higher the positive plate voltage, the more

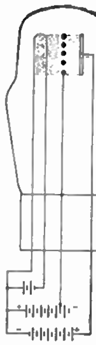


Fig. 3-5 — **Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.**

effectively the space charge is overcome.

If a third element — called the **control grid**, or simply **grid** — is inserted between the cathode and plate as in Fig. 3-5, it can be used to control the effect of the space charge. If the grid is given a positive voltage with respect to the cathode, the positive charge will tend to neutralize the negative space charge. The result is that, at any selected plate voltage, more electrons will flow to the plate than if the grid were not present. On the other hand, if the grid is made negative with respect to the cathode the negative charge on the grid will *add* to the space charge. This will *reduce* the number of electrons that can reach the plate at any selected plate voltage.

The grid is inserted in the tube to control the space charge and not to attract electrons to itself, so it is made in the form of a wire mesh or spiral. Electrons then can go through the open spaces in the grid and to the plate.

Characteristic Curves

For any particular tube, the effect of the grid voltage on the plate current can be shown by a set of **characteristic curves.** A typical set of curves is shown in Fig. 3-6, together with the circuit that is used for getting them. With several fixed values of plate voltage (in these curves, the plate voltage is increased in 50-volt steps, starting at 100 volts) the grid voltage is varied in small steps and a plate-current reading taken at each value of grid voltage. The curves show the result. In Fig. 3-6, the grid voltage is varied between zero and 25 volts negative with respect to the cathode. It can be seen that, for each value of plate voltage, there is a value of negative grid voltage that will reduce the plate current to zero; that is, there is a value of negative grid voltage that will **cut off** the plate current.

The curves could be extended by making the grid voltage positive as well as negative. The practical effect would be to lengthen each of the curves upward along the same line. However, in some types of operation the grid is always kept negative with respect to the cathode, and the particular tube used as an illustration happens to be one that normally would be used that way. Whenever the grid is negative, it repels electrons and therefore none of them reaches it; in other words, no current flows in the grid circuit. When the grid is positive, it attracts electrons and a current **(grid current)** flows, just as current flows to the positive plate. Whenever there is grid current there is an accompanying power loss in the grid circuit, but so long as the grid is negative there is no current and therefore no power is used.

It is obvious that the grid can act as a valve to control the flow of plate current. Actually, the grid has a much greater effect on plate current flow than does the plate voltage. 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 not obtained 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 load must be connected in the plate or "output" circuit, just as in the diode case. The load may be either a resistance or an impedance. The term "impedance" is frequently used even when the load is purely resistive.

Tube Characteristics

The physical construction of a triode determines the relative effectiveness of the grid and plate in controlling the plate current. If a very small change in the grid voltage has just as much effect on the plate current as a very large change in plate voltage, the tube is said

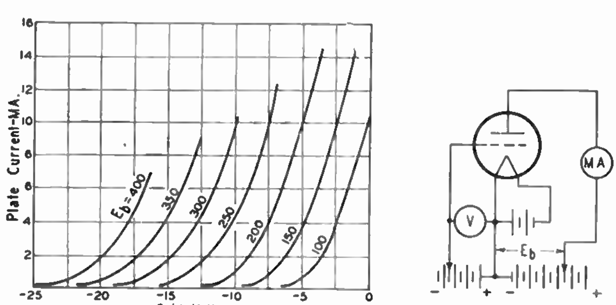


Fig 3-6 — **Grid-voltage-*vs.*-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.**

to have a high **amplification factor**. Amplification factor is commonly designated by the Greek letter μ. An amplification factor of 20, for example, means this: if the grid voltage is changed by 1 volt, the effect on the plate current will be the same as when the plate voltage is changed by 20 volts. The amplification factors of triode tubes range from 3 to something of the order of 100. A **high-μ** tube is one with an amplification factor of perhaps 30 or more; **medium-μ** tubes have amplification factors in the approximate range 8 to 30, and low-μ tubes in the range below 7 or 8.

It would be natural to think that a tube that has a large μ would be the best amplifier, but such is not necessarily the case. If the μ is high it is difficult for the plate to attract large numbers of electrons. Quite a large change in the plate voltage must be made to effect a given change in plate current. This means that the resistance of the plate-cathode path — that is, the **plate resistance** — of the tube is high. Since this resistance acts in series with the load, the amount of current that can be made to flow through the load is relatively small. On the other hand, the plate resistance of a low-μ tube is relatively low. Whether or not a high-μ tube is better than one with a low μ depends on the operation we want the tube to perform.

The best all-around indication of the effectiveness of the tube as an amplifier is its **transconductance** — also called **mutual conductance**. This characteristic takes account of both amplification factor and plate resistance, and therefore is a sort of figure of merit for the tube. Actually, transconductance is the change in plate *current* divided by the change in grid *voltage* that causes the plate-current change (the plate voltage being fixed at a desired value). Since current divided by voltage is equal to conductance, transconductance is measured in the unit of conductance, the mho. Practical values of transconductance are very small, so the micromho (one-millionth of a mho) is the commonly-used unit. Different types of tubes have transconductances ranging from a few hundred to several thousand. The higher the transconductance the greater the possible amplification.

AMPLIFICATION

To understand amplification, it is first necessary to become acquainted with a type of graph called the **dynamic characteristic**. Such a graph, together with the circuit used for obtaining it, is shown in Fig. 3-7. The curves are taken with the plate-supply voltage fixed at the desired operating value. The difference between this circuit and the one shown in Fig. 3-6 is that there is a load resistance connected in series with the plate of the tube in Fig. 3-7, while there is none in Fig. 3-6. Fig. 3-7 thus shows how the plate current will vary, with different grid voltages, *when the plate current is made to flow through a load and thus do useful work.*

The several curves in Fig. 3-7 are for various values of load resistance. The effect of the amount of load resistance is worth noting. When the resistance is small (as in the case of the 5000-ohm load) the plate current changes rather rapidly with a given change in grid voltage. If the load resistance is high (as in the 100,000-ohm curve), the change in plate current for the same grid-voltage change is relatively small, so the curve tends to be straighter.

Going now to Fig. 3-8, we have the same type of curve, but with the circuit arranged so that a source of alternating voltage **(signal)** is inserted between the grid and the grid battery **("C" battery)**. The voltage of the grid battery is fixed at −5 volts, and from the curve

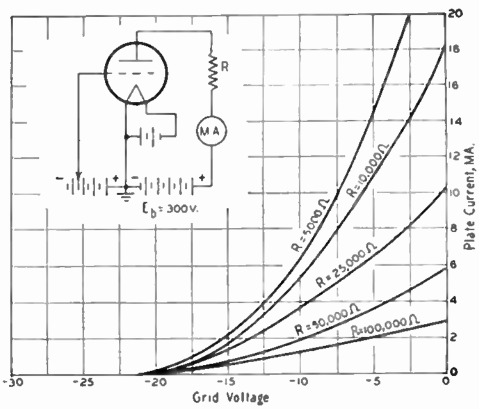


Fig. 3-7 **— Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.**

it is seen that the plate current at this grid voltage is 2 milliamperes. This current flows when the load resistance is 50,000 ohms, as indicated in the circuit diagram. If there is no a.c. signal in the grid circuit, the voltage drop in the load resistor is $50{,}000 \times 0.002 = 100$ volts, leaving 200 volts between the plate and cathode.

Now when 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 will occur at the instant the grid voltage is −3 volts. As shown by the graph, it 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

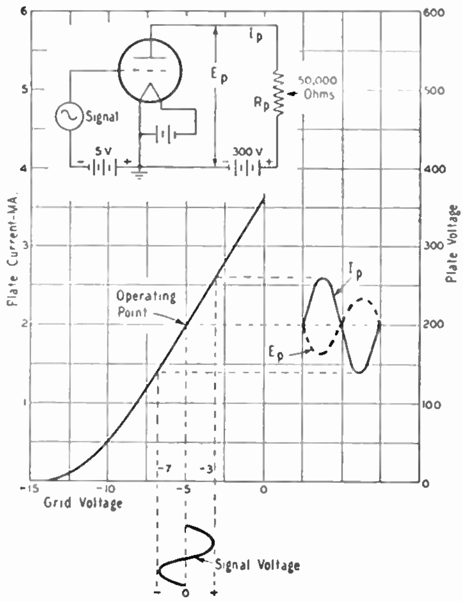


Fig. 3-8 — **Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.**

graph. When the plate current is maximum, the instantaneous voltage drop in R_p is 50,000 $\times$ 0.00265 = 132.5 volts; when the plate current is minimum the instantaneous voltage drop in R_p is 50,000 $\times$ 0.00135 = 67.5 volts. The actual voltage between plate and cathode is the difference between the plate-supply potential, 300 volts, and the voltage drop in the load resistance. The plate-to-cathode voltage is therefore 167.5 volts at maximum plate current and 232.5 volts at minimum plate current.

This varying plate voltage is an **a.c.** voltage superimposed on the steady plate-cathode potential of 200 volts (as 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; that is, 32.5 volts in either case. 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 voltage is obtained from the plate circuit as is applied to the grid circuit.

One feature of the alternating component of plate voltage is worth special note. As shown by the drawings in Fig. 3-8, the positive swing in the grid signal voltage is accompanied by a *downward* swing in the voltage (E_p) between the plate and cathode of the tube. Also, when the alternating grid voltage swings in the *negative* direction, the plate-to-cathode voltage swings to a *higher* value. In other words, the alternating component of the plate voltage swings in the *negative* direction (with reference to the no-signal value of plate-cathode voltage) when the grid swings in the *positive* direction, and vice versa. This means that the alternating component of plate voltage (that is, the amplified signal) is 180 degrees out of phase with the signal voltage on the grid.

Bias

The fixed negative grid voltage (called **grid bias**) in Fig. 3-8 serves a very useful purpose. In the first place, one of the things we want to do in the type of amplification shown in this drawing is to obtain, from the plate circuit, an alternating voltage that has the *same waveshape* as the signal voltage applied to the grid. To do so, we must choose an operating point on the *straight* part of the curve; not only that, the curve must be straight in both directions from the operating point at least far enough to accommodate the maximum value of the signal applied to the grid. If the grid signal swings the plate current back and forth, over a part of the curve that is not straight, as in Fig. 3-9, the shape of the a.c. wave in the plate circuit will not be the same as the shape of the grid-signal wave. In such a case the output waveshape will be **distorted.**

The second reason for using negative grid bias is this: The grid will not attract electrons — that is, there will be no grid current — if the grid is always negative with respect to the cathode. When the grid has a negative bias, any signal whose peak *positive* voltage does not exceed the fixed *negative* voltage on the grid cannot cause grid current to flow. With no current flow there is no power consumption, so the tube will amplify *without taking any power from the signal source.* However, if the positive

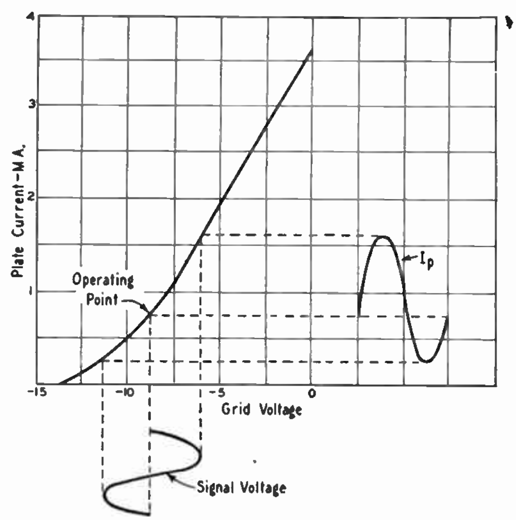


Fig. 3-9 — **Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.**

peak of the signal does exceed the negative bias, current will flow in the grid circuit during the time the grid is positive. While it is perfectly possible to operate the tube in the "positive-grid region," in many cases we do not want the grid to consume power.

Distortion of the output waveshape that results from working over a part of the curve that is not straight (that is, a **nonlinear** part of the curve) has the effect of transforming a sine-wave grid signal into a more complex waveform. As explained in Chapter Two, a complex wave can be resolved into a fundamental and a series of harmonics. In other words, distortion from nonlinearity causes the generation of harmonic frequencies — frequencies that are not present in the signal applied to the grid. Harmonic distortion is undesirable in most amplifiers, although there are occasions when harmonics are deliberately generated and used. This is particularly so in certain types of r.f. transmitting circuits.

Amplifier Output Circuits

The thing that is wanted from the output circuit of a vacuum-tube amplifier is the *alternating* component of plate current or plate voltage. The d.c. voltage on the plate of the tube is essential, of course, for the tube's operation. However, it almost invariably would cause difficulties if it were applied, along with the a.c. output voltage, to the load. The output circuits of vacuum tubes are therefore arranged so that the a.c. is transferred to the load but the d.c. is not.

Three types of coupling are in common use at audio frequencies. These are **resistance coupling, impedance coupling,** and **transformer coupling.** They are shown in Fig. 3-10. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits can be used to couple to other devices than tubes.

In the resistance-coupled circuit, the a.c. voltage developed across the plate resistor R_p (that is, between the plate and cathode of the tube) is applied to a second resistor, R_g, through a coupling condenser, C_c. The condenser "blocks off" the voltage on the plate of the first tube and prevents it from being applied to the grid of tube *B*. The latter tube should have negative grid bias, of course, and this is supplied by the battery shown. No current flows in the grid circuit of tube *B* and there is therefore no d.c. voltage drop in R_g; in other words, the full voltage of the bias battery is applied to the grid of tube *B*.

The grid resistor, R_g, usually has a rather high value (0.5 to 2 megohms). The reactance of the coupling condenser, C_c, must be low enough compared to the resistance of R_g so that the a.c. voltage drop in C_c is negligible at the lowest frequency to be amplified. If R_g is at least 0.5 megohm, a 0.1-μfd. condenser will be amply large for the usual range of audio frequencies.

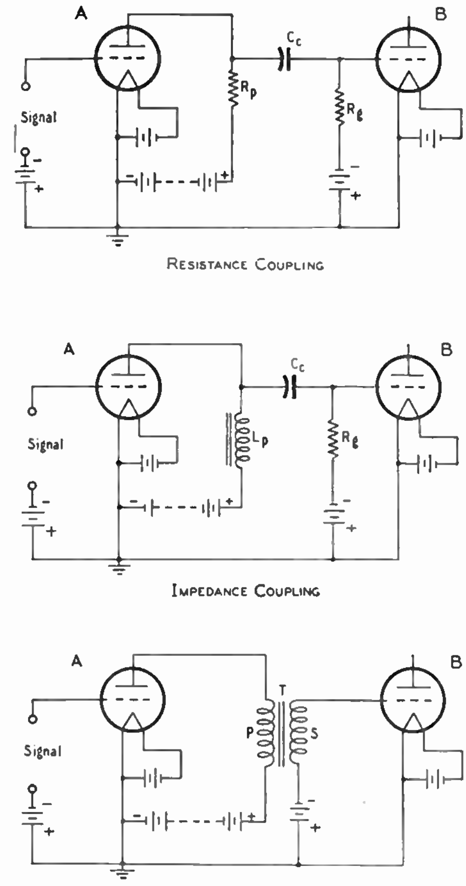


Fig. 3-10 — Three basic forms of coupling between vacuum-tube amplifiers.

So far as the alternating component of plate voltage is concerned, it will be realized that if the voltage drop in C_c is negligible then R_p and R_g are effectively in parallel (although they are quite separate so far as d.c. is concerned). The resultant parallel resistance of the two is therefore the actual load resistance for the tube. That is why R_g is made as high in resistance as possible; then it will have the least effect on the load represented by R_p.

The impedance-coupled circuit differs from that using resistance coupling only in the substitution of a high-inductance coil (usually several hundred henrys) for the plate resistor. The advantage of using an inductance rather than a resistor is that its impedance is high for alternating currents, but its resistance is relatively low for d.c. (A resistor, of course, has the same resistance for d.c. that it does for a.c.). It thus permits us to obtain a high value of load impedance for a.c., but without an excessive d.c. voltage drop that would use up a good deal of the voltage from the plate supply.

The transformer-coupled amplifier uses a transformer with its primary connected in the plate circuit of the tube and its secondary connected to the load (in the circuit shown, a following amplifier). There is no direct connection between the two windings, so the plate voltage on tube A is isolated from the grid of tube B. The transformer-coupled amplifier has the same advantage as the impedance-coupled circuit with respect to loss of voltage from the plate supply. There is an additional advantage as well: if the secondary has more turns than the primary, the output voltage will be "stepped up" in proportion to the turns ratio.

All three circuits have good points. Resistance coupling is simple, inexpensive, and will give the same amount of amplification — or **voltage gain** — over a wide range of frequencies; it will give substantially the same amplification at any frequency in the audio range, for example Impedance coupling will give somewhat more gain, with the same tube and same plate-supply voltage, than resistance coupling. However, it is not quite so good over a wide frequency range; it tends to "peak," or give maximum gain, over a comparatively narrow band of frequencies. With a good transformer the gain of a transformer-coupled amplifier can be kept fairly constant over the audio-frequency range. On the other hand, transformer coupling is best suited to triodes

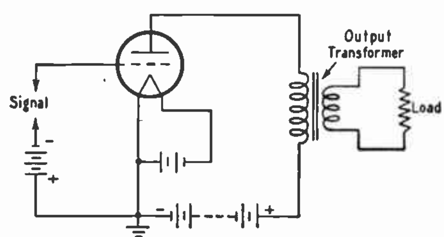


Fig. 3-11 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

having amplification factors of about 10 or less, for the reason that the primary inductance of a practicable transformer cannot be made large enough to work well with a tube having high plate resistance.

An amplifier in which voltage gain is the primary consideration is called a **voltage amplifier.** Maximum voltage gain is secured when the load resistance or impedance is made as high as possible in comparison with the plate resistance of the tube. In such a case, the major portion of the voltage generated will appear across the load and only a relatively small part will be "lost" in the plate resistance.

Voltage amplifiers belong to a group called **Class A amplifiers.** A Class A amplifier is one operated so that the waveshape of the output voltage is the same as that of the signal voltage applied to the grid. If a Class A amplifier is biased so that the grid is always negative, even with the largest signal to be handled by the grid, it is called a **Class A_1 amplifier.** Voltage amplifiers are always Class A_1 amplifiers, and their primary use is in driving a following Class A_1 amplifier.

Power Amplifiers

The end result of any amplification is that the amplified signal does some *work*. As a familiar example, an audio-frequency amplifier usually drives a loudspeaker that in turn produces sound waves. The greater the amount of a.f. *power* supplied to the 'speaker, the louder the sound it will produce.

In some amplifiers, therefore, *power* output rather than voltage is the primary consideration. It was mentioned in Chapter Two that any source of power will deliver the largest possible output when the resistance of the load is equal to the internal resistance of the source. In the case of a vacuum tube, the "source" resistance is the plate resistance of the tube. Therefore if we want the utmost power from the tube the load resistance should be equal to the plate resistance of the tube. Actually, however, this is not the best operating condition because the use of such a relatively low value of load resistance generally results in more distortion than we want. For this reason the load resistance for a power amplifier usually is two or three times the plate resistance; this represents a good compromise between distortion and power output.

Fig. 3-11 shows an elementary power-amplifier circuit. It is simply a transformer-coupled amplifier with the load connected to the secondary. Although the load is shown as a resistor, it actually would be some device, such as a loudspeaker, that employs the power usefully. The resistance of the actual load is rarely the right value for "matching" the load resistance that the tube wants for optimum power output. Therefore the transformer turns ratio is chosen to reflect the proper value of resistance into the primary. The turns ratio may be either step-up or step-down, depending on whether the actual load resistance is higher or lower than the load the tube wants.

The **power-amplification ratio** of an amplifier is the ratio of the power output obtained from the plate circuit to the power required from the a.c. signal in the grid circuit. There is no power lost in the grid circuit of a Class A_1 amplifier, so such an amplifier has an infinitely large power-amplification ratio. However, it is quite possible to operate a Class A amplifier in such a way that current flows in its grid circuit during at least part of the cycle. In such a case power is used up in the grid circuit and the power amplification ratio is not infinite. A tube operated in this fashion is known as a **Class A_2 amplifier.** It is necessary to use a power amplifier to drive a Class A_2 amplifier, because a voltage amplifier cannot deliver power without serious distortion of the waveshape.

Another term used in connection with power

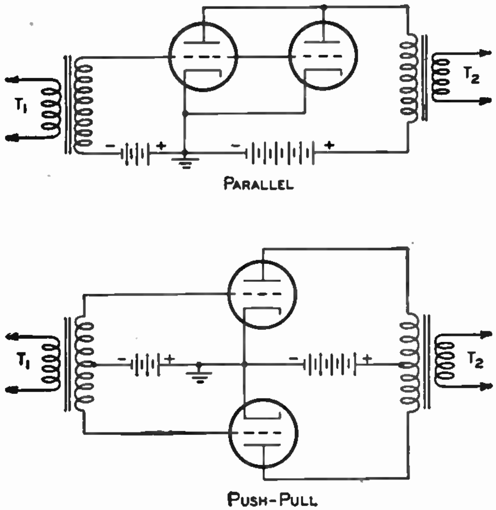


Fig. 3-12 — **Parallel and push-pull a.f. amplifier circuits.**

amplifiers is **power sensitivity.** In the case of a Class A_1 amplifier, it means the ratio of power output to the grid signal voltage that causes it. If grid current flows, the term usually means the ratio of plate power output to grid power input.

The a.c. power that is delivered to a load by an amplifier tube has to be paid for in power taken from the source of plate voltage and current. In fact, there is always more power going into the plate circuit of the tube than is coming out as useful output. The difference between the input and output power is used up in heating the plate of the tube, as explained previously. If we want a great deal of power output, therefore, it is advantageous to make this difference as small as possible. The ratio of useful power output to d.c. plate input is called the **plate efficiency.** The higher the plate efficiency, the greater the amount of power that can be taken from a tube having a fixed plate-dissipation rating.

Parallel and Push-Pull

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. 3-12 for a transformer-coupled amplifier. The power output is in proportion to the number of tubes used; the grid signal or "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.

An increase in power output also can be secured by connecting two tubes in **push-pull.** In this case the grids and plates of the two tubes are connected to opposite ends of a balanced circuit as shown in Fig. 3-12. At any instant the ends of the secondary winding of the input transformer, T_1, will be at opposite polarity with respect to the cathode connection, so 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 amplifier the voltages and currents of one tube are out of phase with those of the other tube.

In push-pull operation the even-harmonic (second, fourth, etc.) distortion is balanced out in the plate circuit. This means 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 amplifier is twice that taken by either tube alone.

Cascade Amplifiers

It is of course thoroughly possible to take the output of one amplifier and apply it as a signal on the grid of a second amplifier, then take the second amplifier's output and apply it to a third, and so on. Each amplifier is called a **stage,** and a number of amplifier stages used to increase successively the amplitude of the signal are said to be in **cascade.**

The number of amplifiers that can be connected in cascade is not unlimited. If the over-all amplification becomes too great, there is danger that some of the output voltage will get back into one of the early stages. This "feed-back," discussed in a later section, may make the amplifier unstable and prevent it from functioning as it should.

Class B Amplifiers

Fig. 3-13 shows two tubes connected in a push-pull circuit. If the grid bias is set at the point where (when no signal is applied) the plate current is just cut off, then a signal can cause plate current to flow in either tube *only* when the signal voltage applied to that particular tube is positive. In the balanced grid circuit, the signal voltages on the grids of the two tubes always have opposite polarities; that is, when the signal swings the instantaneous voltage in the positive direction on the grid of tube *A*, it is at the same time swinging the grid of tube *B* more negative. On the next half-cycle the polarities reverse and the grid of tube *B* is more positive and that of tube *A* more negative. Since the fixed bias is just at the cut-off point, this means that plate current flows only in one tube at a time.

The graphs show the operation of such an amplifier. The plate current of tube *B* is drawn inverted to show that it flows in the opposite direction, through the primary of the output transformer, to the plate current of tube *A*. Thus each half of the output-transformer primary works alternately to induce a half-cycle of voltage in the secondary. In the secondary of T_2, the original waveform is restored. This type of operation is called **Class B amplification.**

The Class B amplifier is considerably more efficient than the Class A amplifier. Furthermore, the d.c. plate current of a Class B amplifier is proportional to the signal voltage on the grids, so the power input is small with small signals. The d.c. plate power input to a Class A amplifier is the same whether the signal is large, small, or absent altogether; therefore the maximum input that can be applied to a Class A amplifier is the rated plate dissipation of the tube or tubes. Two tubes in a Class B amplifier can deliver approximately twelve times as much audio power as the same two tubes in a Class A amplifier.

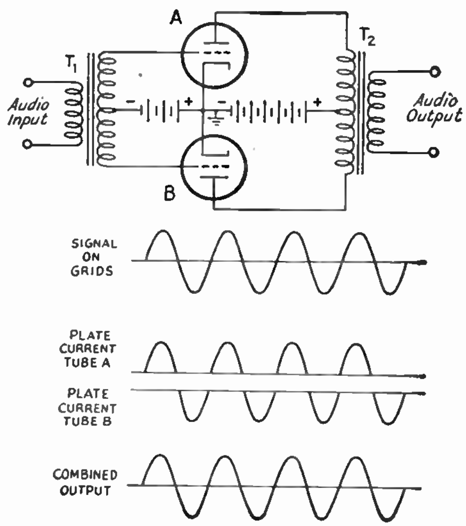


Fig. 3-13 — Class B amplifier operation.

A Class B amplifier usually is operated in such a way as to secure the maximum possible power output. This requires that the grids be driven positive with respect to the cathode during at least part of the cycle, so grid current flows and the grid circuit consumes power. While the power requirements are fairly low (as compared with the power output), the fact that the grids are positive during only *part* of the cycle means that the load on the "driver" stage varies in magnitude during the cycle; the effective load resistance is high when the grids are not drawing current and relatively low when they do take current. This must be allowed for when designing the driver.

Certain types of tubes have been designed specifically for Class B service and can be operated without fixed or other form of grid bias ("zero-bias" tubes). The amplification factor is so high that the plate current is small without signal. Because there is no fixed bias, the grids start drawing current immediately whenever a signal is applied, so the grid-current flow is continuous throughout the cycle. This makes the load on the driver much more constant than is the case with tubes of lower μ biased to plate-current cut-off.

Class AB Amplifiers

A **Class AB amplifier** is one operated midway between Class A and Class B conditions. A Class AB amplifier is a push-pull amplifier with higher bias than would be normal for pure Class A operation, but less than the cut-off bias required for Class B. At low signal levels the tubes operate practically as Class A amplifiers, and the plate current is the same with or without signal. At higher signal levels, the plate current of one tube is cut off during part of the *negative* cycle of the signal applied to its grid, and the plate current of the other tube rises with the signal. The plate current for the whole amplifier also rises above the no-signal level when a large signal is applied.

In a properly-designed Class AB amplifier the distortion is as low as with a Class A stage, but the efficiency and power output are considerably higher than with pure Class A operation. A Class AB amplifier can be operated either with or without driving the grids into the positive region. A **Class AB_1 amplifier** is one in which the grids are never positive with respect to the cathode; therefore, no driving power is required — only voltage. A **Class AB_2 amplifier** is one that has grid-current flow during part of the cycle, when the applied signal is large; it takes a small amount of driving power. The Class AB_2 amplifier will deliver somewhat more power (using the same tubes) but the Class AB_1 amplifier avoids the problem of designing a driver for it that will deliver power, without distortion, into a load of highly-variable resistance.

Class C Amplifiers

Inspection of Fig. 3-13 shows that either of the two tubes actually is working for only half the a.c. cycle and idling during the other half. It is convenient to describe the amount of time during which plate current flows in terms of electrical degrees. In Fig. 3-13 each tube has "180-degree" excitation, a half-cycle being equal to 180 degrees. The number of degrees during which plate current flows is called the **operating angle** of the amplifier. From the descriptions given above, it should be clear that a Class A amplifier has 360-degree excitation, because plate current flows during the whole cycle. In a Class AB amplifier the operating angle is between 180 and 360 degrees (in each tube) depending on the particular operating conditions chosen. The greater the amount of negative grid bias, the smaller the operating angle becomes.

An operating angle of less than 180 degrees obviously would lead to a considerable amount of distortion, because there is no way for the tube to reproduce even a half-cycle of the signal on its grid. Using two tubes in push-pull, as in Fig. 3-13, would not overcome this distortion; it would merely put together two distorted half-cycles. An operating angle of less than 180 degrees therefore cannot be used if distortionless output is wanted.

However, in certain types of amplifiers distortion does not matter particularly. One example is an amplifier used to generate r.f. power. The power output of such an amplifier is delivered to a tuned circuit, and it is characteristic of a tuned circuit that it will have a high impedance at the frequency to which it is resonant, but low impedance to all other frequencies. The tuned circuit can be made to have a high impedance at the frequency applied to the grid of the amplifier, thus providing a load of the optimum value for the tube. At harmonics of this fundamental frequency the impedance of the tuned circuit will be low, and thus will be a poor load for the tube for those frequencies set up by distortion; the distortion is "filtered out." The result is that the output voltage and current are practically pure sine waves.

Using an operating angle less than 180 degrees increases the plate efficiency, because it is characteristic of tube operation that the smaller the time during which plate current flows the smaller the amount of power lost in the plate. Also, when the proper angle and other operating conditions are chosen the power output of the amplifier is proportional to the square of the voltage applied to its plate. That is, the amplifier has the linear characteristics of a resistor insofar as its behavior when the plate voltage is varied is concerned. This is an important consideration when the amplifier is to be "modulated," as described in Chapter Nine. Such an amplifier is called a **Class C amplifier.** In Class C operation the operating angle usually is in the range 120–150 degrees, and the plate efficiency is 70 to 80 per cent.

● FEED-BACK

As we have shown, there is more energy in the plate circuit of an amplifier than there is in the grid circuit. It is easily possible to take a part of the plate-circuit energy and insert it into the grid circuit. When this is done the amplifier is said to have **feed-back.**

There are two types of feed-back. If the voltage that is inserted in the grid circuit is 180 degrees out of phase with the signal voltage acting on the grid, the feed-back is called **negative,** or **degenerative.** On the other hand, if the voltage is fed back *in* phase with the grid signal, the feed-back is called positive, or **regenerative.** With negative feed-back the voltage that is fed back *opposes* the signal voltage; this decreases the amplitude of the voltage acting between the grid and cathode. With a smaller signal voltage, of course, the output also is smaller. The effect of negative feed-back, then, is to *reduce* the amount of amplification.

Negative Feed-Back

The circuit shown at A in Fig. 3-14 gives degenerative feed-back. Resistor R_c is in series with the regular plate resistor, R_p, and thus is a part of the load for the tube. Therefore, part of the output voltage will appear across R_c. However, R_c also is connected in series with the *grid* circuit, and so the output voltage that appears across R_c is in series with the signal voltage. In this circuit, the output voltage across R_c opposes the signal voltage and the actual a.c. voltage between the grid and cathode therefore is equal to the *difference* between the two voltages.

While it would be natural to assume that there could be no point in reducing the amplification by negative feed-back, it does have uses. The greater the amount of negative feed-back (when properly applied) the more independent the amplification becomes of tube characteristics and circuit conditions. This means that the frequency-response characteristic of the amplifier becomes **flat** — that is, amplification tends to be the same at all frequencies within the range for which the amplifier is designed. Also, any distortion generated in the plate circuit of the tube tends to "buck itself out" when some of the output voltage is fed back to the grid. Amplifiers with negative feed-back are therefore comparatively free of harmonic distortion. These advantages, secured at the expense of voltage amplification, are worth while if the amplifier otherwise has enough gain for its intended use.

The circuit shown at B in Fig. 3-14 can be used to give either negative or positive feed-back. In this case the secondary of a transformer is connected back into the grid circuit to insert a desired amount of feed-back voltage. Reversing the terminals of either the

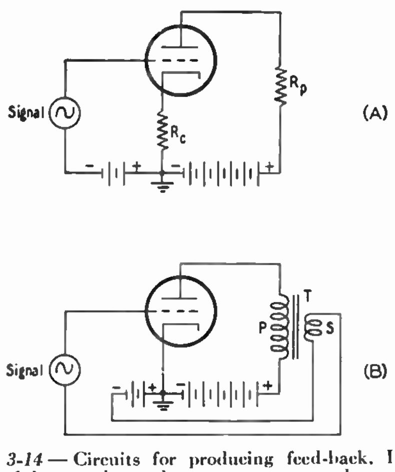


Fig. 3-14 — Circuits for producing feed-back. In A, part of the a.c. plate voltage appears across the cathode resistor, R_c, and is therefore also applied between grid and cathode. The feed-back is negative in this case. In B, the voltage that is generated in the secondary of the transformer is inserted in series in the grid circuit. Feed-back may be either positive or negative, depending upon the transformer connections.

primary or secondary of the transformer (but not both windings simultaneously) will reverse the phase of the voltage fed back. Thus either type of feed-back is available.

Positive Feed-Back

Positive feed-back *increases* the amplification because the fed-back voltage adds to the original signal voltage and the resulting larger voltage on the grid causes a larger output voltage. It has the opposite characteristics to negative feed-back; the amplification tends to be greatest at one frequency (depending upon the particular circuit arrangement) and harmonic distortion is increased. If the energy fed back becomes large enough, a self-sustaining oscillation will be set up at one frequency; in this case all the signal voltage on the grid is supplied from the plate circuit; no external signal is needed. It is not even necessary to have an external signal to start the oscillation; any small irregularity in the plate current — and there are always some such irregularities — will be amplified and thus give the oscillation an opportunity to build up. Oscillations obviously would be undesirable in an audio-frequency amplifier, and for that reason (as well as the others mentioned above) positive feed-back is never used in a.f. amplifiers. Positive feed-back finds its use in "oscillators" at both audio and radio frequencies, as described in a subsequent section.

The two circuits shown in Fig. 3-14 are only two of many that can be used to provide feed-back. Despite differences in appearance, such circuits are alike in this fundamental — energy is fed back from the output circuit to the grid circuit in the proper phase to give the type of feed-back that is wanted.

INTERELECTRODE CAPACITANCES

Each pair of elements in a tube actually forms a small "condenser," with each element acting as a condenser "plate." There are three such capacitances in a triode — that between the grid and cathode, that between the grid and plate, and that between the plate and cathode. The capacitances are very small — only a few micromicrofarads at most — but they frequently have a very pronounced effect on the operation of an amplifier circuit.

Input Capacitance

It was explained previously that the a.c. grid voltage and a.c. plate voltage of an amplifier are 180 degrees out of phase, using the cathode of the tube as a reference point. However, these two voltages are *in* phase if we go around the circuit from plate to grid as shown in Fig. 3-15. This means that their sum is acting between the grid and plate; that is, across the grid-plate capacitance of the tube. When an a.c. voltage is applied to a condenser, a current flows through the condenser. As viewed from the source of the signal on the grid, this current is flowing because of the signal voltage.

The larger the current, the lower the effective reactance in the grid circuit. The larger the grid-plate capacitance the larger the current; also, the greater the voltage amplification the larger the current, because this puts more voltage across the grid-plate condenser. The result is that the source of signal "sees" a capacitive reactance that is much smaller than the actual reactance of the capacitance between the grid and cathode.

Since a small reactance is equivalent to a large capacitance, the input capacitance of an amplifier may be many times its actual grid-cathode capacitance. In practice, the input capacitance of a triode may be as much as a few hundred micromicrofarads, particularly if the triode has a large amplification factor. Such a capacitance is not negligible, even at audio frequencies, when it is placed in parallel with a resistor of 50,000 ohms or more.

Tube Capacitance at R.F.

At radio frequencies the reactances of the interelectrode capacitances drop to such low values that they must always be taken into account in circuit design. A resistance-coupled amplifier cannot be used at r.f., for example, because the reactances of the interelectrode "condensers" are so low that they, and not the resistors, would be the actual load. Furthermore, they are so low that they practically short-circuit the input and output circuits and thus the tube is unable to amplify. We get around this at radio frequencies by using *tuned* circuits for the grid and plate, and making the tube capacitances part of the tuning capacitances. In this way the circuits can have the high impedances necessary for satisfactory amplification.

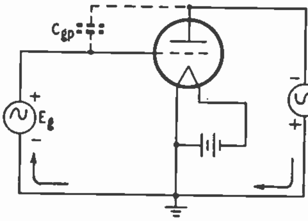


Fig. 3-15 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

The grid-plate capacitance is important at radio frequencies because it is, in effect, a coupling condenser between the grid and plate circuits. Since its reactance is relatively low at r.f., it offers a path over which energy can be fed back from the plate to the grid. In practically every case the feed-back is in the right phase and of sufficient amplitude to cause oscillation, so the amplifier becomes useless. Special circuits can be used to prevent feed-back but they are, in general, not too satisfactory when used in radio receivers. (They are, however, widely used in transmitters.) A better solution to this problem is found in the use of the screen-grid tube.

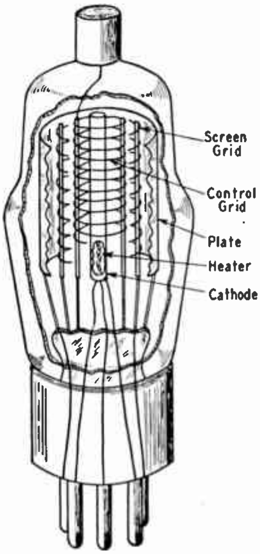


Fig. 3-16 — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen grid cut away. In this drawing the control-grid connection is made through a cap on the top of the tube, thus eliminating the capacitance that would exist between the plate- and grid-lead wires if both passed through the base. Some modern tubes that have both leads going through the base use special shielding and construction to eliminate interlead capacitance.

SCREEN-GRID TUBES

The grid-plate capacitance 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. 3-16. The second grid, called the **screen grid,** acts as a shield between the control grid and plate. It is made in the form of a grid or coarse screen so that electrons can pass through it; a solid shield would entirely prevent the flow of plate current. The screen grid is usually grounded through a by-pass condenser that has low reactance at the radio frequency being amplified.

Because of the shielding action of the screen grid, 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 voltage (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 such velocity that most of them shoot between the screen wires and go on to the plate. A certain proportion do strike the screen, however, with the result that some current also flows to the screen-grid circuit of the tube.

A tube having a cathode, control grid, screen grid and plate (four elements) is called a **tetrode.**

Pentodes

When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons which "splash" from the plate into the interelement space. This is called **secondary emission.** In a triode the negative grid repels the secondary electrons back into the plate and they 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.

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 usually 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. A five-element tube of this type is called a **pentode.**

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, the control grid still can control the plate current in essentially the same way that it does in a triode. 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. 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. A voltage gain in the vicinity of 50 to 200 is typical of a pentode stage.

Pentode R.F. Amplifier

Fig. 3-17 shows a simplified form of r.f. amplifier circuit, using a pentode tube. Radio-frequency energy in the small coil coupled to L_1 is built up in voltage in the tuned circuit, L_1C_1, when L_1C_1 is tuned to resonance with the frequency of the incoming signal. The voltage that appears across L_1C_1 is applied to the grid and cathode of the tube and is amplified by the tube. A second resonant circuit, L_2C_2, is the load for the plate of the tube, its parallel impedance being high because it is tuned to resonance with the frequency applied to the

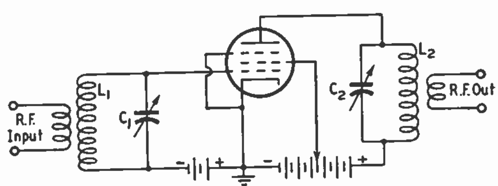


Fig. 3-17 — Simplified pentode r.f.-amplifier circuit. L_1C_1 and L_2C_2 are tuned to the same frequency.

grid. R.f. output can be taken from the coil coupled to L_2. The screen-grid voltage is obtained from a tap on the plate battery; most tubes are designed for operation with the screen voltage considerably lower than the plate voltage. In this circuit the batteries are assumed to have low impedance for the r.f. current; in a practical circuit, by-pass condensers would be used to make sure that the impedances of the return paths actually are low enough to be negligible.

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 compared to triodes of the same power output. Harmonic distortion is somewhat greater with pentodes and tetrodes than with triodes, however.

Variable-μ Tubes

The mutual conductance of a vacuum tube decreases with increasing negative grid bias, assuming that the other electrode voltages are held constant. Since the mutual conductance controls the amount of amplification, it is possible to adjust the gain of the amplifier by adjusting the grid bias. This method of gain control is universally used in radio-frequency amplifiers designed for receivers. Some means of controlling the r.f. gain is essential in a receiver having a number of amplifiers, because of the wide range in the strengths of the incoming signals.

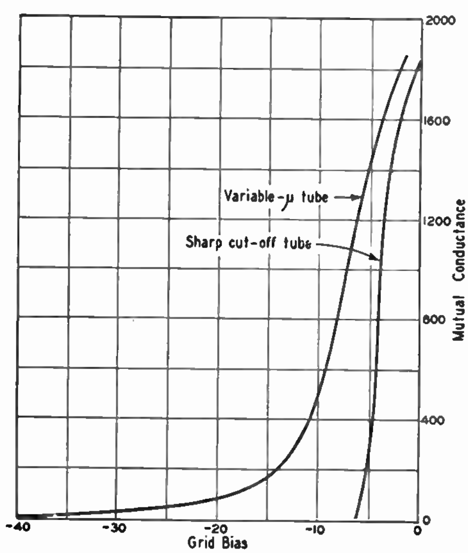


Fig. 3-18 — Curves showing the relationship between mutual conductance and negative grid bias for two small receiving pentodes, one a sharp cut-off type and the other a variable-μ type.

The ordinary type of tube has what is known as a **sharp cut-off** characteristic. The mutual conductance decreases at a uniform rate as the negative bias is increased, as shown in Fig. 3-18. The amount of signal voltage that such a tube can handle without causing distortion is quite limited, and not sufficient to take care of very strong signals. To overcome this, some tubes are made with a **variable-μ** characteristic (that is, the amplification factor changes with the grid bias), resulting in the type of curve shown in Fig. 3-18. It is evident that the variable-μ tube can handle a much larger signal than the sharp cut-off type before the signal swings either beyond the zero grid-bias point or the plate-current cut-off point.

OTHER TYPES OF AMPLIFIERS

In the amplifier circuits so far discussed, the signal has been applied between the grid and cathode and the amplified output has been taken from the plate-to-cathode circuit. That is, the *cathode* has been the common point, or meeting point, for the input and output circuits. However, since there are three elements (the screen and suppressor in a pentode ordinarily do not enter *directly* into the amplifying action) it is possible to use any one of the three as the common point. This leads to two different kinds of amplifiers, commonly called the **grounded-grid amplifier** (or **grid-separation** circuit) and the **cathode follower.**

These two circuits are shown in simplified form in Fig. 3-19. In both circuits the resistor R represents the load into which the amplifier works; the actual load may be resistance-capacitance-coupled, transformer-coupled, may be a tuned circuit if the amplifier operates at radio frequencies, and so on. Also, in both circuits the batteries that supply grid bias and plate power are assumed to have such negligible impedance that they do not enter into the operation of the circuits.

Grounded-Grid Amplifier

In the grounded-grid amplifier the input signal is applied between the cathode and grid, and the output is taken between the plate and grid. The grid is thus the common element. The plate current (including the a.c. component) has to flow through the signal source to reach the cathode. Since this source always has appreciable impedance, the alternating plate current causes a voltage drop that acts between the grid and cathode. Because of the phase relationship between the signal and output voltages, the circuit is degenerative. Also, since the source of signal is in series with the load through the plate-to-cathode resistance of the tube, some of the power in the load is supplied by the signal source. The result is that the signal source is called upon to furnish a considerable amount of power.

The grounded-grid amplifier finds its chief application at v.h.f. and u.h.f., where the more conventional amplifier circuit fails to work properly. With a triode tube designed for

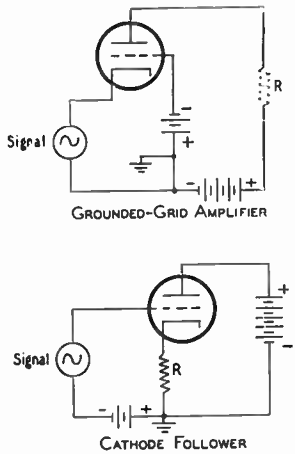


Fig. 3-19 — In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, *R*, and may be coupled to a following amplifier by the usual methods.

this type of operation, an r.f. amplifier can be built that is free from the type of feed-back that causes oscillation. This requires that the grid act as a shield between the cathode and plate, reducing the plate-cathode capacitance to a very low value.

Cathode Follower

The cathode follower uses the plate of the tube as the common element. The input signal is applied between the grid and plate (assuming negligible impedance in the batteries) and the output is taken from between cathode and plate. This circuit, like the grounded-grid amplifier, is degenerative. In fact, *all* of the output voltage is fed back into the input circuit to buck the applied signal. The input signal therefore has to be larger than the output voltage; that is, the cathode follower not only gives no voltage gain but actually results in a *loss* in voltage. (It can still give just as much *power* gain as ever, though.)

The cathode follower has two advantages: It has a very high input impedance (impedance between grid and ground — in the customary cathode-follower circuit the plate is at ground for signal voltage); and its output impedance is very low. (The large amount of negative feed-back has the effect of greatly reducing the plate resistance of the tube.) These two characteristics are valuable in an amplifier that must work over a very wide range of frequencies. Also, the high input impedance and low output impedance can be used to obtain an impedance step-down over wide ranges of frequencies that could not possibly be covered by a transformer. The cathode follower is useful both at audio and radio frequencies.

CATHODE CIRCUITS AND GRID BIAS

Most of the equipment used by amateurs is powered by the a.c. line. This includes the filaments or heaters of vacuum tubes. Although supplies for the plate (and sometimes the grid) are usually rectified and filtered to give "pure" d.c. — that is, direct current that is constant and without a superimposed a.c. component — the relatively large currents required by filaments and heaters make a d.c. supply impracticable.

Filament Hum

Alternating current is just as good as direct current from the heating standpoint, but some of the a.c. voltage is likely to get on the grid and cause a low-pitched "a.c. hum" to be superimposed on the output. The voltage can get on the grid either by a direct circuit connection, through the electric field about the heater, or through the magnetic field set up by the current.

Hum troubles are worst with directly-heated cathodes or filaments, because with such cathodes there has to be a direct connection between the source of heating power and the rest of the circuit. The hum can be minimized by either of the connections shown in Fig. 3-20. In both cases the grid- and plate-return circuits are connected to the electrical midpoint (center-tap) of the filament supply. Thus, so far as the grid and plate are concerned, the voltage and current on one side of the filament are balanced by an equal and opposite voltage and current on the other side. This balances out the hum. The balance is never quite perfect, however, so filament-type tubes are never completely hum-free. For this reason directly-heated filaments are employed for the most part in transmitting power tubes, where the amount of hum introduced is extremely small in comparison to the power-output level.

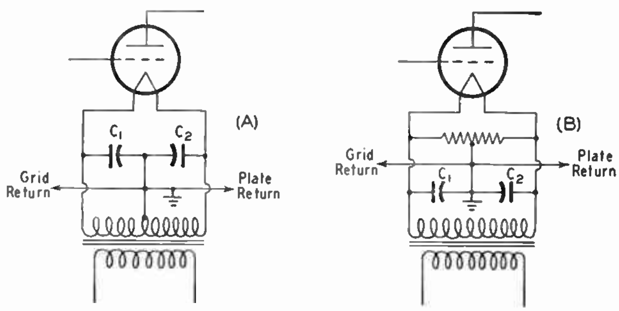


Fig. 3-20 — Filament center-tapping methods for use with directly-heated tubes.

With indirectly-heated cathodes the source of heating power does not introduce hum by a direct connection. The chief problem with such tubes is the magnetic field set up by the heater. Occasionally, also, there is leakage between the heater and cathode; leakage that allows a small a.c. voltage to get to the grid. Both these things are principally a matter of tube design. However, it is found in practice that, if hum appears, grounding one side of the heater supply will help to reduce it. Sometimes better results are obtained if the heater supply is center-tapped and the center-tap grounded, as in Fig. 3-20.

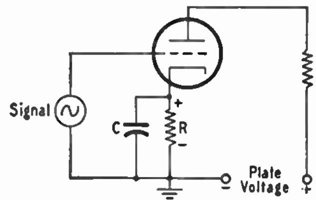


Fig. 3-21 — Cathode biasing. *R* is the cathode resistor and *C* is the cathode by-pass condenser.

Cathode Bias

In the simplified amplifier circuits discussed in this chapter, grid bias has been supplied by a battery. However, it is seldom obtained that way in an actual piece of equipment that operates from the power line. **Cathode bias** is the type commonly used.

The cathode-bias method uses a resistor connected in series with the cathode, as shown at *R* in Fig. 3-21. The direction of plate-current flow is such that the end of the resistor nearest the cathode is positive. The voltage drop across *R* therefore places a *negative* voltage on the grid. This negative bias is obtained from the steady d.c. plate current.

If the alternating component of plate current flows through *R* when the tube is amplifying, the voltage drop caused by the a.c. will be degenerative (note the similarity between this circuit and that of Fig. 3-14A). To prevent this the resistor is by-passed by a condenser, *C*, that has very low reactance compared to the resistance of *R*. The capacitance required at *C* depends upon the value of *R* and the frequency being amplified. Depending on the type of tube and the particular kind of operation, *R* may be between about 250 and 3000 ohms. For good by-passing at the low audio frequencies, *C* should be 10 to 50 microfarads (electrolytic condensers are used for this purpose). At radio frequencies, capacitances of about 100 $\mu\mu$fd. to 0.1 μfd. are used; the small values are sufficient at very high frequencies and the largest at low and medium frequencies. In the range 3 to 30 megacycles a capacitance of 0.01 μfd. is satisfactory.

The value of cathode resistor can easily be calculated from the known operating conditions of the tube. The proper grid bias and plate current always are specified by the manufacturer. Knowing these, the required resistance can be found by applying Ohm's Law.

Example: It is found from tube tables that the tube to be used should have a negative grid bias of 8 volts and that at this bias the plate current will be 12 milliamperes (0.012 amp.). The required cathode resistance is then

$$R = \frac{E}{I} = \frac{8}{0.012} = 667 \text{ ohms.}$$

The nearest standard value, 680 ohms, would be close enough. The power used in the resistor is

$$P = EI = 8 \times 0.012 = 0.096 \text{ watt.}$$

A ¼-watt or ½-watt resistor would have ample rating.

The current that flows through *R* is the *total* cathode current. In an ordinary triode amplifier this is the same as the plate current, but in a screen-grid tube the cathode current is the sum of the plate and screen currents. Hence these two currents must be added when calculating the value of cathode resistor required for a screen-grid tube.

Example: A receiving pentode requires 3 volts negative bias. At this bias and the recommended plate and screen voltages, its plate current is 9 ma. and its screen current is 2 ma. The cathode current is therefore 11 ma. (0.011 amp.). The required resistance is

$$R = \frac{E}{I} = \frac{3}{0.011} = 272 \text{ ohms.}$$

A 270-ohm resistor would be satisfactory. The power in the resistor is

$$P = EI = 3 \times 0.011 = 0.033 \text{ watt.}$$

The cathode-resistor method of biasing is convenient because it avoids the use of batteries or other source of fixed voltage. However, that is not its only advantage: it is also self-regulating, because if the tube characteristics vary slightly from the published values (as they do in practice) the bias will increase if the plate current is slightly high, or decrease if it is slightly low. This tends to hold the plate current at the proper value. For the same reason, the value of the cathode resistance is not highly critical. Cathode bias also avoids any tendency toward unwanted feed-back that might occur when a single fixed-bias source is used to furnish bias for several amplifiers. Even a very small a.c. voltage drop in the impedance of a bias source can cause oscillation (if the feed-back is positive) or loss of gain (if the feed-back is negative) when the voltage is applied to the first stage of amplification in an amplifier having several stages, simply because the gain in a multistage amplifier is likely to be very large.

The calculation of the bias resistor in a resistance-coupled amplifier is not as easy as the examples above. This is because the actual voltages that should be used on the plate and grid are not ordinarily known. The difficulty is that the voltage drop in the plate resistor causes the actual voltage at the plate of the tube to be considerably less than the plate-supply voltage, and the lower plate voltage requires a different value of bias than that given in the published operating conditions for the tube. The proper voltages can be found by a cut-and-try process from the tube characteristic curves. However, representative data for the tubes commonly used as resistance-coupled amplifiers are given in Chapter Nine, including cathode-resistor values.

Screen Supply

In practical circuits using tetrodes and pentodes the voltage for the screen frequently is taken from the plate supply through a resistor. A typical circuit for an r.f. amplifier is shown in Fig. 3-22. Resistor R is the screen **dropping resistor**, and C is the **screen by-pass condenser**. In flowing through R, the screen current causes a voltage drop in R that reduces the plate-supply voltage to the proper value for the screen. When the plate-supply voltage and the screen current are known, the value of R can be calculated from Ohm's Law.

Example: An r.f. receiving pentode has a rated screen current of 2 milliamperes (0.002 amp.) at normal operating conditions. The rated screen voltage is 100 volts, and the plate supply gives 250 volts. To put 100 volts on the screen, the drop across R must be equal to the difference between the plate-supply voltage and the screen voltage; that is, 250 — 100 = 150 volts. Then

$$R = \frac{E}{I} = \frac{150}{0.002} = 75{,}000 \text{ ohms.}$$

The power to be dissipated in the resistor is

$$P = EI = 150 \times 0.002 = 0.3 \text{ watt.}$$

A ½- or 1-watt resistor would be satisfactory.

The reactance of the screen by-pass condenser, C, should be low compared with the screen-to-cathode impedance. For radio-frequency applications a capacitance of 0.01 μfd. is amply large.

In some circuits the screen voltage is obtained from a voltage divider connected across the plate supply. The design of voltage dividers is discussed in Chapter Seven.

SPECIAL TUBE TYPES

Beam Tubes

"Beam tetrodes" are tetrode tubes constructed in such a way that the power sensitivity is very high. Beam tubes are useful as both radio-frequency and audio-frequency power amplifiers, and are available in output ratings from a few watts up to several hundred watts. The grids in a beam tube are so constructed and aligned as to form the electrons traveling to the plate into concentrated beams. This makes it possible to draw large plate currents at relatively low plate voltages, and also reduces the number of electrons that are captured by the screen. Additional design features overcome the effects of secondary emission, so that a suppressor grid is not needed.

Multipurpose Tubes

A number of "combination" tubes is available to perform more than one function, particularly in receiver circuits. For the most part these are simply multiunit tubes made up of individual tube-element structures, combined in a single bulb for compactness and economy.

Among the simplest multipurpose types are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb. More-complex types include duplex-diode triodes (two diodes and a triode in one structure), duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on.

Mercury-Vapor Rectifiers

For a given value of plate current, the power lost in a diode rectifier will be reduced if it is possible to decrease the voltage drop from plate to cathode. A small amount of mercury in the tube will vaporize when the cathode is heated and, further, will ionize 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.

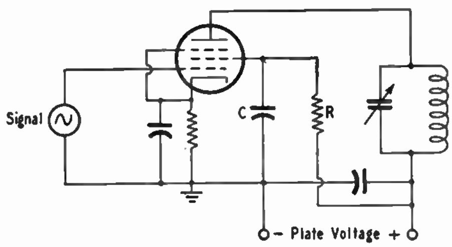


Fig. 3-22 — Screen-voltage supply for a pentode tube through a dropping resistor, R. The screen by-pass condenser, C, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

Since this voltage drop is smaller than can be attained with purely thermionic conduction, there is less power loss in a mercury-vapor rectifier than in a vacuum rectifier. Also, the voltage drop in the tube is constant despite variations in load current. Mercury-vapor tubes are widely used in rectifiers built to deliver large power outputs.

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. However, this is true *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, because the space charge disappears when ionization occurs. The grid can assume control again only after the plate voltage is reduced below the ionizing voltage.

The same phenomenon also occurs in triodes filled with other gases that ionize at low pressure. Grid-control rectifiers or **thyratrons** find considerable application in "electronic switching."

Oscillators

It was mentioned earlier in this chapter that if there is enough positive feed-back in an amplifier circuit, self-sustaining oscillations will be set up. When an amplifier is arranged so that this condition exists it is called an **oscillator.**

Oscillations normally take place at only one frequency, and a desired frequency of oscillation can be obtained by using a resonant circuit tuned to that frequency. The proper phase for positive feed-back can be obtained quite easily from a single tuned circuit. For example, in Fig. 3-23A the circuit LC is tuned to the desired frequency of oscillation. The coil L is tapped and the cathode of the tube is connected to the tap. The grid and plate are connected to opposite ends of the tuned circuit. There will be a voltage drop across the tuned circuit, a voltage drop that increases progressively along the turns of the coil when viewed from one end. At an instant when the upper end of L is positive, for instance, the lower end is negative. However, the tap on the coil is at an intermediate voltage and so is negative with respect to the upper end of L, and positive with respect to the lower end. Or, viewed from the tap, the upper end of L is positive and the lower end is negative. Therefore the grid and plate ends of the coil are opposite in polarity, or opposite in phase. This is the right phase relationship for positive feed-back.

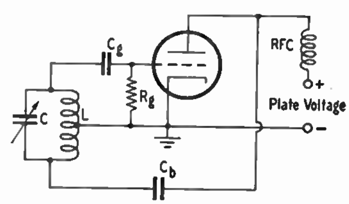
The *amount* of feed-back depends on the *position* of the tap. If the tap is too close to either end of the coil the circuit will not oscillate. If the tap is too near the grid end the voltage drop is too small to give enough feed-back, and if it is too near the plate end the impedance between the cathode and plate is too small to permit good amplification. Maximum feed-back usually is obtained when the tap is somewhere near the center of the coil.

It will be observed that the circuit of Fig. 3-23A is parallel-fed, C_b being the blocking condenser. The value of C_b is not critical so long as its reactance is low at the operating frequency.

Condenser C_g is the **grid condenser.** It and R_g (the **grid leak**) are used for the purpose of obtaining grid bias for the tube. In this (and practically all) oscillator circuits the tube generates its own bias. When the grid end of the tuned circuit is positive with respect to the cathode, the grid attracts electrons from the cathode. These electrons cannot flow through L back to the cathode because C_g "blocks" direct current. They therefore have to flow or "leak" through R_g to cathode, and in doing so cause a voltage drop in R_g that places a negative bias on the grid. The amount of bias so developed is equal to the grid current multiplied by the resistance of R_g (Ohm's Law). The value of grid-leak resistance required depends upon the kind of tube used and the purpose for which the oscillator is intended. Values range all the way from a few thousand to several hundred thousand ohms. The capacitance of C_g should be large enough to have low reactance at the operating frequency.

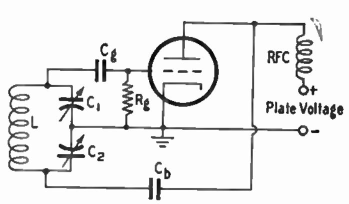
The circuit shown at B in Fig. 3-23 uses the voltage drops across two condensers in series in the tuned circuit to supply the feed-back. Other than this, the operation is the same as just described. The feed-back can be varied by varying the ratio of the reactances of C_1 and C_2 (that is, by varying the ratio of their capacitances). To maintain the same oscillation frequency the total *capacitance* across L must be constant; this means that every time C_1, for example, is adjusted to change the feed-back, C_2 must be adjusted in the *opposite* sense to return the total capacitance and thereby the frequency to the original value.

Another type of oscillator, called the **tuned-plate tuned-grid** circuit, is shown in Fig. 3-24. Resonant circuits tuned approximately to the same frequency are connected between grid and cathode and between plate and cathode. The two coils, L_1 and L_2, are not magnetically-coupled. The feed-back is through the grid-plate capacitance of the tube, and will be in the right phase to be positive when the plate circuit, C_2L_2, is tuned to a slightly higher frequency than the grid circuit, L_1C_1. The amount of feed-back can be adjusted by varying the tuning of either circuit. The frequency of oscillation is determined by the tuned circuit that has the higher Q. The grid leak and grid condenser have the same functions as in the other circuits. In this case it is convenient



(A)

Hartley Circuit



(B)

Colpitts Circuit

Fig. 3-23 — Basic oscillator circuits. Feed-back voltage is obtained by tapping the grid and cathode across a portion of the tuned circuit. In the Hartley circuit the tap is on the coil, but in the Colpitts circuit the voltage is obtained from the drop across a condenser.

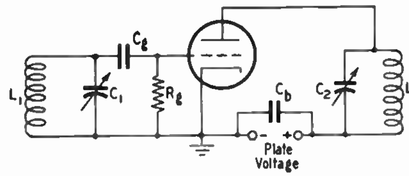


Fig. 3-24 — The tuned-plate tuned-grid oscillator.

to use series feed for the plate circuit, so C_b is a by-pass condenser to guide the r.f. current around the plate supply.

Practically all feed-back oscillator circuits (and there is an endless variety of them) are variations of these general types. They differ in details and appearance, and some use two or more tubes to accomplish the purpose. However, the basic feature of all of them is that there is positive feed-back in the proper amplitude to sustain oscillation.

Oscillator Operating Characteristics

As a general rule, oscillators are *power-generating* devices. There are exceptions: in some cases the oscillator is used primarily to generate a *voltage* that is then applied to an amplifier that does not require power in its grid circuit. This type of oscillator is used principally in certain types of measuring equipment; the oscillators used in transmitters and receivers usually are called upon to deliver some power.

When an oscillator is delivering power to a load, the adjustment for proper feed-back will depend on how heavily the oscillator is loaded. If the feed-back is not large enough — that is, if the **grid excitation** is too small — a slight change in load may tend to throw the circuit into and out of oscillation. On the other hand, too much feed-back will make the grid current excessively high, with the result that the power loss in the grid circuit is larger than necessary. The oscillator itself supplies this grid power, so excessive feed-back lowers the over-all efficiency because whatever power is used in the grid circuit is not available as useful output.

One of the most important considerations in oscillator design is frequency stability. Almost invariably we want the generated frequency to be as constant as possible. The principal factors that 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 capacitances. 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 capacitance 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 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 and that the circuit should be lightly loaded. Dynamic stability also can be improved by using a high value of grid leak; this increases the grid bias and raises the effective resistance of the tube as seen by the tank circuit. Using relatively high plate voltage and low plate current also helps.

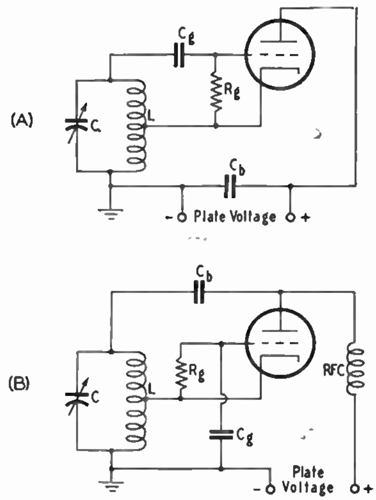


Fig. 3-25 — Showing how the r.f. ground on a typical oscillator circuit (Hartley) may be placed on either the plate (A) or grid (B) instead of the more conventional method of grounding the cathode. Provided the proper provisions are made for supplying cathode and plate voltages, the circuit operation is unchanged by shifting the r.f. ground to any desired point.

Mechanical variations, usually caused by vibration, cause changes in inductance and/or capacitance that in turn cause the frequency to "wobble" in step with the vibration.

Methods of minimizing frequency variations in oscillators are taken up in detail in later chapters.

Ground Point

In the oscillator circuits shown in Figs. 3-23 and 3-24 the cathode is connected to ground. It is not actually essential that the radio-frequency circuit should be grounded at the cathode; in fact, there are many times when an *r.f.* ground on some other point in the circuit is desirable. The r.f. ground can be placed at any point so long as proper provisions are made for feeding the supply voltages to the tube elements.

Fig. 3-25 shows the Hartley circuit with (A) the plate end of the circuit grounded, and (B) the grid end. In A, no r.f. choke is needed in the plate circuit because the plate already is at ground potential and there is no r.f. to choke off. All that is necessary is a by-pass condenser, C_b, across the plate supply. Direct current flows to the cathode through the lower part of the tuned-circuit coil, *L*.

The grounded-grid circuit at B is essentially the same as the circuit in Fig. 3-23A except that the ground point and negative plate-voltage connection have been placed at the grid end of the tuned circuit.

One advantage of either type of circuit (the one in Fig. 3-25A is widely used) is that the frame of the tuning condenser can be grounded. With a grounded-cathode oscillator, both ends of the tuned-circuit are "hot"; that is, there is an r.f. voltage to ground from both ends of the circuit. When the ordinary type of tuning condenser is used in such a circuit there is a slight change in capacitance when the hand is brought near the tuning shaft for adjustment of capacitance. This "hand capacitance" or "body capacitance" is annoying because the oscillator frequency changes when the hand is brought near the tuning control. It is overcome by grounding (for r.f.) the condenser shaft and by using a condenser that has a frame with metal end plates.

Tubes having indirectly-heated cathodes are more easily adaptable to circuits grounded at other points than the cathode than are tubes having directly-heated filaments. With the latter tubes special precautions have to be taken to prevent the filament from being by-passed to ground by the capacitance of the filament-heating transformer.

NEGATIVE-RESISTANCE OSCILLATORS

If a tuned circuit could be built without resistance, a small amount of energy introduced into the circuit would start an oscillation that would continue indefinitely. It would do so because, in a circuit having no power losses, the power never diminishes and therefore is always available to keep the oscillation going. Of course, such a circuit cannot be built.

However, it was explained in Chapter Two that a resonant circuit has a definite value of parallel impedance at resonance, and that that impedance is a pure resistance. If we could connect across the circuit a value of "negative" resistance equal to the parallel resistance of the circuit, the negative resistance would cancel the "positive" (real) resistance of the circuit and we would have a circuit that is, in effect, without resistance.

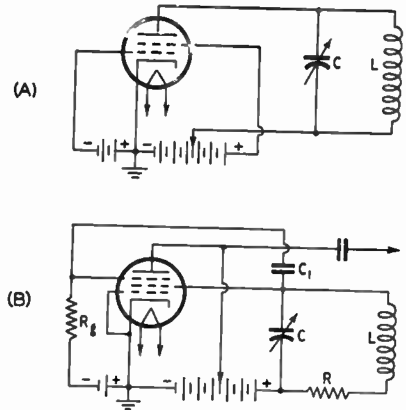


Fig. 3-26 — Negative-resistance oscillator circuits. A, dynatron; B, transitron.

A negative resistance is one having the opposite characteristics to real or positive resistance. In a negative resistance the current *increases* when the voltage is decreased, and vice versa. Also, a negative resistance does not consume power; it generates it. Under certain conditions a vacuum tube can be made to operate like a negative resistance, and thus can be connected to a tuned circuit to set up oscillations. Two circuits for doing this are shown in Fig. 3-26.

The circuit at A is called the **dynatron** oscillator. It functions because of the secondary emission from the plate that occurs in certain types of screen-grid tetrodes. It makes use of the fact that, at certain values of screen voltage, the plate current of a screen-grid tetrode decreases when the plate voltage is increased. This gives a negative plate-resistance characteristic.

In Fig. 3-26B, negative resistance is produced by virtue of the fact that, if the suppressor grid of a pentode is given negative bias, electrons that normally would pass through the suppressor to the plate are turned back to the screen, thus increasing the screen current and reversing normal tube action. The negative resistance produced between the screen and suppressor grids is sufficiently low so that ordinary tuned circuits will oscillate readily up to 15 Mc. or so. This circuit is known as the **transitron**.

For most amateur applications, negative-resistance oscillators do not have enough advantages to bring them into wide use. Feedback oscillators are generally more adaptable to wide frequency ranges, can generate more power, and are more readily adjusted to meet varying conditions. The transitron oscillator is used occasionally in measuring equipment.

CHAPTER 4

High-Frequency Communication

Much of the appeal of amateur communication on the high frequencies lies in the fact that the results are not always predictable. Transmission conditions on the same frequency vary with the year and even with the time of day. Although these variations usually follow certain established cycles, many peculiar effects can be observed from time to time. Every radio amateur should have some understanding of the known facts about radio-wave propagation so that he will stand some chance of interpreting the unusual conditions when they occur. The observant amateur is in an excellent position to make worth-while contributions to the science, provided he has sufficient background to understand his results. The serious amateur can, for example, coöperate with the National Bureau of Standards in its NBS-ARRL 28-Mc. Observing Project, about which more is said at the end of this chapter. Or he may develop a new theory of propagation for the very-high frequencies or the microwave region, as did the late Ross Hull. By making extensive observations of 56-Mc. conditions over a long-distance path and correlating the results with various weather conditions, Mr. Hull was able to establish the now-accepted theory of "tropospheric bending."

What To Expect on the Various Amateur Bands

The 3.5-Mc., or "80-meter," band is a more useful band during the night than during the daylight hours. In the daytime, one can seldom hear signals from a distance of greater than 100 miles or so, but during the darkness hours distances up to several thousand miles are not unusual, and transoceanic contacts are regularly made during the winter months. During the summer, the static level is high in some parts of the world, and the sharp crashes of static often make reception difficult. The 3.5-Mc. band supports the majority of the traffic nets throughout the country, and it is also a great gathering place for "rag-chewers." Low power and simple antennas can be used here with good results.

The 7-Mc., or "40-meter," band has many of the same characteristics as 3.5, except that the distances that can be covered during the day and night hours are increased. During daylight, distances up to a thousand miles can be covered under good conditions, and during the dawn and dusk periods in winter it is possible to work stations as far as the other side of the world, the signals following the darkness path. The winter months are somewhat better than the summer ones. Rag-chewing, traffic handling and DX (working foreign countries) are popular activities on the band, in the order named. Here again antennas are not too important, although results will be improved in proportion to the effectiveness of the antenna system. In general, summer static is much less of a problem than on 80 meters, although it can be serious in the semitropical zones.

The 14-Mc., or "20-meter," band is probably the best one for long-distance work. During portions of the sunspot cycle it is open to some part of the world during practically all of the 24 hours, while at other times it is generally useful only during daylight hours and the dawn and dusk periods. DX activity is paramount, with rag-chewing next. Being less consistent, day by day, traffic handling is not too general, although many long-distance schedules are kept on the band. Effective antennas are more necessary than on the lower frequencies, but many amateurs enjoy excellent results with simple antennas and low power. Automobile ignition and other types of man-made interference begin to be a problem on this band.

The 28-Mc. band is generally considered to be a DX band during the daylight hours and a local rag-chewer's band during the hours of darkness. However, during parts of the sunspot cycle, the band is "open" into the late evening hours for DX communication. The band is even less consistent than 14 Mc., but this very fact is what makes it so fascinating for its many followers. It is not unusual for a foreign station to appear suddenly with a loud signal when only U. S. stations, or none at all, are being heard. High-performance antennas are almost a necessity for best results, but its small dimensions make the rotary beam a popular choice for the band. These antennas can be turned to direct the radiation in the desired direction, and they are used to provide useful gain on reception as well. A good antenna is far more important on this band than high power.

Characteristics of Radio Waves

Radio waves differ from other forms of electromagnetic radiation (such as light and heat) in the manner in which they are generated and detected and in their wavelength. The wavelength spectrum of radio waves is greater than either heat or light, and ranges from approximately 30,000 meters to a small fraction of a centimeter. This corresponds to a frequency range of about 10 kc. to 1,000,000 Mc. They travel at the same velocity as light waves (about 186,000 miles per second in free space) and can be reflected, refracted and diffracted the way light and heat waves can.

The passage of radio energy through space is explained by a concept of traveling electrostatic and electromagnetic waves. The energy is evenly divided between the two types of fields, and the lines of force of these fields are at right angles to each other, in a plane perpendicular to the direction of travel. A simple representation of this is shown in Fig. 4-1.

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.

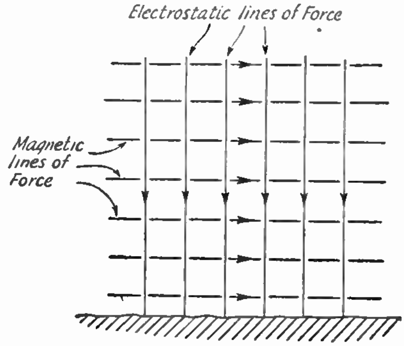


Fig. 4-1 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

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 (or any insulator having a dielectric constant differing from that of the medium) offers such a discontinuity if its dimensions are at least comparable to the wavelength. The surface of the earth and the boundaries between ionospheric layers are examples of such discontinuities. Objects as small as an airplane, a tree or even a man's body will readily reflect the shorter waves.

Refraction

As in the case of light, a radio wave is bent when it moves obliquely into any medium having a refractive index different from that of the medium it leaves. Since the velocity of propagation differs in the two mediums, that part of the wave front that enters first travels faster if the new medium has a higher velocity of propagation. This tends to swing the wave front around, or "refract" it, in such a manner that the wave is directed in a new direction. If the wave front is one that is traveling obliquely away from the earth, and it encounters a medium with a higher velocity of propagation, the wave will be directed back toward the earth. If the new medium has a lower velocity of propagation, the opposite effect takes place, and the wave is directed away from the earth. Refraction may take place either in the ionosphere (ionized upper atmosphere) or the troposphere (lower atmosphere), or both.

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 that is directed toward the ionosphere. Depending upon variable conditions in that region, as well as upon transmitting wavelength, 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 that 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 that 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. 4-2.

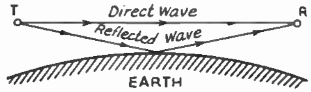


Fig. 4-2 — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

Ionospheric Propagation

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 that 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 that 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 (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 that 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 that 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 a higher frequency that 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 by special equipment, 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 that actually takes place, as illustrated in Fig. 4-3. 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.

In the daytime there is a still lower ionized region, the D layer. The D-layer intensity is proportional to the height of the sun and is greatest at noon. Low-frequency waves (80 meters) are almost completely absorbed by this layer while it exists, and only the high-angle radiation is reflected by the E layer. (Lower-angle radiation travels farther through the D layer and is absorbed.)

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 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, seasonal changes result in higher critical frequencies in the E layer in summer, averaging about 4 Mc. as against a winter average of 3 Mc. The F layer shows little variation, the critical frequency being of the order of 4 to 5 Mc. in the evening. 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.

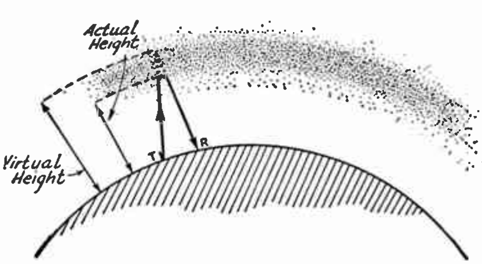


Fig. 4-3 — **Bending in the ionosphere, and the echo or reflection method of determining virtual height.**

Seasonal transition periods occur in spring and fall, when ionospheric conditions are found highly variable.

There are at least two other regular cycles in ionization. One such cyclic period covers 28 days, which corresponds with the period of the sun's rotation. For a short time in each 28-day cycle, 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, corresponding 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 long-distance 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 the winter of 1947–48.

Magnetic Storms and Other Disturbances

Unusual disturbances in the earth's magnetic field (magnetic storms) usually are accompanied by disturbances in the ionosphere, when the layers apparently break up and expand. There is usually also an increase in absorption during such a period. Radio transmission is poor and there is a drop in critical frequencies so that lower frequencies must be used for communication. A magnetic storm may last for several days.

Unusually high ionization in the region of the atmosphere below the normal ionosphere may increase absorption to such an extent that sky-wave transmission becomes impossible on high frequencies. The length of such a disturbance may be several hours, with a gradual falling off of transmission conditions at the beginning and an equally gradual building up at the end of the period. **Fade-outs**, similar to the above in effect, are caused by sudden disturbances on the sun. They are characterized by very rapid ionization, with sky-wave transmission disappearing almost instantly, occur only in daylight, and do not last as long as the first type of absorption.

Magnetic storms frequently are accompanied by unusual auroral displays, creating an ionized "curtain" in the polar regions which can act as a reflector of radio waves. **Auroral reflection** is occasionally observed at frequencies as high as 54 Mc. It is characterized on 28 Mc. by a flutter on all signals which makes voice work difficult but not impossible. Directive antennas must be pointed toward the north and not in the direction of the station being worked.

Sporadic-E Layer Ionization

Occasionally scattered patches of 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 1250 miles may be covered at 50 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.

Scatter

Scatter signals are heard on any band, but are more easily recognizable on the higher frequencies because of the extended skip zone. They are signals reflected from large discontinuities at a distance, such as sharp concentrations of ionization in any of the normal layers, sporadic-*E* clouds or (rarely) large land objects. They result in one's hearing signals within the normal skip zone. Scatter signals are never very loud, and have a slight flutter characteristic. A further indication of scatter reflection is that, when beam antennas are used to indicate the direction of arrival of the wave, the ray path is not necessarily the direct route but can even be at right angles or in the opposite direction.

Meteor Trails

Another phenomenon generally encountered in the 28-Mc. band, but also observed in the 14- and 50-Mc. bands, is one characterized by sudden bursts of intensity of a signal. These bursts last less than a second, generally, and are caused by reinforced reflection from the ionized trail of a meteor. The meteor, entering the earth's atmosphere at high velocity, heats by friction against the atmosphere and leaves a trail of ionized atmosphere. It takes a finite time for the ionized molecules to recombine, and during this time a small ionized cloud exists. If it is in the ray path of a signal, it may serve to reinforce the signal and cause the burst in intensity. When the meteor is moving in a direction somewhat parallel to the ray path, it can induce a rising or falling "whistle" on the signal, for a second or so. The effects of bursts and whistles can be observed at any time during the day or night, if there is any marked meteor activity, and during rare "meteor showers" the ionized clouds can serve in almost the same manner that sporadic-*E* does to make long-distance work possible on 50 Mc.

Wave Angle

The smaller the angle at which a wave leaves the earth, the less will 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. 4-4. 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 is illustrated in Fig. 4-4, where waves at angles of *A* or less give useful signals while waves sent at

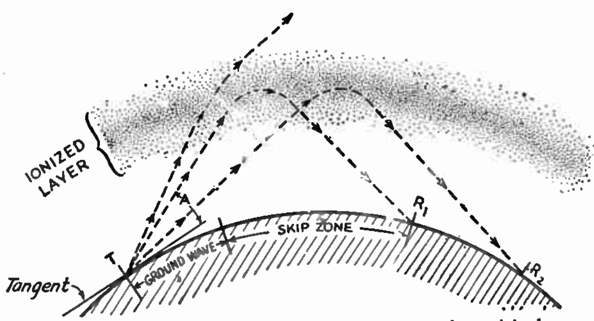


Fig. 4-4 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than *A*) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances.

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.

Multihop 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. Multihop 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 that 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 multihop 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 happens that transmission conditions are different for waves of slightly different frequencies, so that in the case of voice-modulated transmission, involving sidebands differing slightly from the carrier in frequency, the carrier and various sideband 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.

Tropospheric Propagation

Changes in refractive index of air masses in the lower atmosphere often permit work over greater-than-normal distances on 28 Mc. and higher frequencies. The effect can be observed on 28 Mc., but it is generally more marked on 50 and 144 Mc. The subject is treated in detail in Chapter Eleven.

● PREDICTION CHARTS

The National Bureau of Standards offers prediction charts three months in advance, for use in predicting and studying long-distance communication on the usable frequencies above 3.5 Mc. By means of these charts, it is possible to predict with considerable accuracy the maximum usable frequency that will hold over any path on the earth during a monthly period. The charts are based on ionosphere soundings made at a number of stations throughout the world, coupled with considerable statistical data. The charts are conservative enough to enable the amateur to anticipate and plan his best operating times, particularly on the 14- and 28-Mc. bands. Amateurs who work on 50 Mc. and are interested in the occasional F_2 "openings" in this band watch the charts with great interest. They can be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. for 10 cents a copy or $1.00 per year on subscription. They are called "CRPL-D Basic Radio Propagation Predictions."

● N.B.S.-A.R.R.L. 28-MC. OBSERVING PROJECT

Any amateur with 28-Mc. receiving and transmitting equipment can apply to the National Bureau of Standards and ask to take part in the 28-Mc. Observing Project. This service provides valuable information for the NBS in their propagation work, and only requires that the amateur make regular monthly reports on a minimum number of schedules or receiving observations. The program is not confined to amateurs in the United States, and stations throughout the world are taking an active part. Address your application to Central Radio Propagation Laboratory, National Bureau of Standards, Washington 25, D. C.

CHAPTER 5

High-Frequency Receivers

A good receiver in the amateur station makes the difference between mediocre contacts and solid QSOs, and its importance cannot be emphasized too much. In the v.h.f. bands that are not too crowded, **sensitivity** (the ability to bring in weak signals) is the most important factor in a receiver. In the more crowded amateur bands, good sensitivity must be combined with **selectivity** (the ability to distinguish between signals separated by only a small frequency difference) for best results and general ease of reception. Using only a simple receiver, old and experienced operators can copy signals that would be missed entirely by newer amateurs, but their success is because of their experience and not the receiving equipment. On the other hand, a less-experienced operator can use modern techniques to obtain the same degree of success, provided he understands the operation of his more advanced type of receiver and how to get the most out of it.

A number of signals may be picked up by the receiving antenna, and the receiver must be able to separate them and allow the operator to copy the one he wants. This ability is called "selectivity." To receive weak signals, the receiver must furnish enough **amplification** to amplify the minute signal power delivered by the antenna up to a useful amount of power that will operate a loudspeaker or set of headphones. Before the amplified signal can operate the 'speaker or 'phones, however, it must be converted to audio-frequency power by the process of **detection.** The sequence of amplification is not too important — some of the amplification can take place (and usually does) before detection, and some can be used after detection.

There are two major differences between receivers for 'phone reception and for c.w. reception. A 'phone signal has sidebands that make the signal take up about 6 or 8 kc. in the band, and the audio quality of the received signal is impaired if the passband of the receiver is less than half of this. On the other hand, a c.w. signal occupies only a few hundred cycles at the most, and consequently the passband of a c.w. receiver can be small. In either case, if the passband of the receiver is more than is necessary, signals adjacent to the desired one can be heard, and the selectivity of the receiver is said to be poor. The other difference is that the detection process delivers directly the audio frequencies present as modulation on a 'phone signal, but there is no modulation on a c.w. signal and additional technique is required to make the signal audible. 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. 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. Modern superheterodyne receivers (described later) generally use a separate oscillator to generate the beat-note. Summing up the two differences, 'phone receivers can't use as much selectivity as c.w. receivers, and c.w. receivers require some kind of beating oscillator to give an audible signal. Broadcast receivers can receive only 'phone signals because no beat oscillator is included. On the other hand, communications receivers include beat oscillators and often some means for varying the selectivity.

Receiver Characteristics

Sensitivity

Confusion exists among some radio men when talking about the "sensitivity" of a receiver. In commercial circles it is defined as the strength of the signal (in microvolts) at the input of the receiver that is required to produce a specified audio power output at the 'speaker or headphones. This is a perfectly-satisfactory definition for broadcast and communications receivers operating below about 20 Mc., where general atmospheric and man-made electrical noises normally mask any noise generated by the receiver itself.

Another commercial definition of sensitivity measures the merit of a receiver by defining the sensitivity as the signal at the input of the receiver required to give an audio output some stated amount (generally 10 db.) above the noise output of the receiver. This is a much more useful sensitivity measure for the amateur, since it indicates how well a weak signal will be reproduced and is not merely a measure of the over-all gain, or amplification, of the receiver. However, it is still not an absolute method for comparing two receivers, because the passband width of the receiver plays a large part in the result.

The random motion of the molecules in the antenna and receiver circuits generates small voltages called **thermal-agitation noise** voltages. The frequency of this noise is random and the noise exists across the entire radio spectrum. Its amplitude increases with the temperature of the circuits. Only the noise in the antenna and first stage of a receiver is normally significant, since the noise developed in later stages is masked by the amplified noise from the first stage. Since the only noise that is amplified is that which falls within the passband of the receiver, the noise appearing in the output of a receiver is less when the passband is reduced (the effect of the "tone control" of a broadcast receiver). Similar noise is generated by the current flow within the first tube itself; this effect can be combined with the thermal noise and called **receiver noise.** Since the passband of two receivers plays an important part in the sensitivity measured on a signal-to-noise basis as described in the preceding paragraph, such a sensitivity measurement puts more emphasis on passband width than on the all-important "front-end" design of the receiver.

The limit of a receiver's ability to detect weak signals is the thermal noise generated in the input circuit. Even if a perfect noise-free tube were developed and used throughout the receiver, the limit to reception would be the thermal noise. (Atmospheric-and-man-made noise is a *practical* limit below 20 Mc., but we are looking for a measure of comparison of receivers.) The degree to which a receiver approaches this ideal is called the **noise figure** of the receiver, and it is expressed as the ratio of noise power at the input of the receiver required to increase the noise output of the receiver 3 db. Since the noise power passed by the receiver is dependent on the passband (which is the same for the receiver noise and the noise introduced to the receiver), the figure is one that shows how far the receiver departs from the ideal. The ratio is generally expressed in db., and runs around 6 to 12 db. for a good receiver, although figures of 2 to 4 db. have been obtained with special techniques. Comparisons of noise figures can be made by the amateur with simple equipment, as described in Chapter Sixteen.

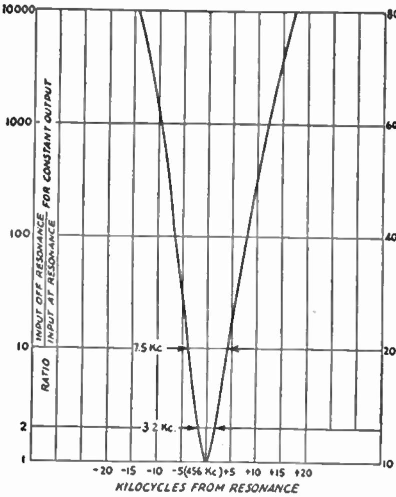


Fig. 5-1 — Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.

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 that 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. 5-1. The **bandwidth** is the width of the resonance curve (in cycles or kilocycles) of a receiver at a specified ratio; in Fig. 5-1, the bandwidths are indicated for ratios of response of 2 and 10 ("2 times down" and "10 times down").

A receiver is more selective if the bandwidth (or passband) is less, but the bandwidth must be sufficient to pass the signal and its sidebands if faithful reproduction of the signal is desired. In the crowded amateur bands, it is generally advisable to sacrifice fidelity for selectivity, since the added selectivity reduces adjacent-channel interference and also the noise passed by the receiver. If the selectivity curve has steep sides, it is said to have good **skirt selectivity,** and this feature is very useful in listening to a weak signal that is adjacent to a strong one. Good skirt selectivity can only be obtained by using a large number of tuned circuits.

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, and also its ability to remain tuned to a signal under varying conditions of gain-control setting, temperature, supply-voltage changes and mechanical shock and distortion. In other words, it means the ability "to stay put" on a given signal. The term "unstable" is also applied to a receiver that breaks into oscillation or a regenerative condition with some settings of its controls that are not specifically intended to control such a condition. This type of instability is sometimes encountered in high-gain amplifiers.

Fidelity

Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, voice, etc.) carried by the incoming signal. For exact reproduction the bandwidth must be great enough to accommodate the carrier and all of the sidebands before detection, and all of the frequency components of the modulation after detection. For perfect fidelity, the relative amplitudes of the various components must not be changed by passing through the receiver. However, fidelity plays a very minor rôle in amateur communication, where the important requirement is to transmit intelligence and not "high-fidelity" signals.

Detection and Detectors

Detection is the process of recovering the modulation from a signal. Any device that is "nonlinear" (i.e., whose output is not *exactly* proportional to its input) will act as a detector. It can be used as a detector if an impedance for the desired modulation frequency is connected in the output circuit, so that the detector output can develop across this impedance.

Detector sensitivity is the ratio of desired detector output to the input. Detector linearity is a measure of the ability of the detector to reproduce the exact form of the modulation on the incoming signal. The resistance or impedance of the detector is the resistance or impedance it presents to the circuits it is connected to. The input resistance is important in receiver design, since if it is relatively low it means that the detector will consume power, and this power must be furnished by the preceding stage. The signal-handling capability means the ability of the detector to accept signals of a specified amplitude without overloading or distortion.

Diode Detectors

The simplest detector is the diode rectifier. A galena, silicon or germanium crystal is an imperfect form of diode (a small current can pass in the reverse direction), and the principle of detection in a crystal is similar to that in a vacuum-tube diode.

Circuits for both half-wave and full-wave diodes are given in Fig. 5-2. The simplified half-wave circuit at 5-2A 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 causes a d.c. voltage to develop across the terminals of R_1, 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

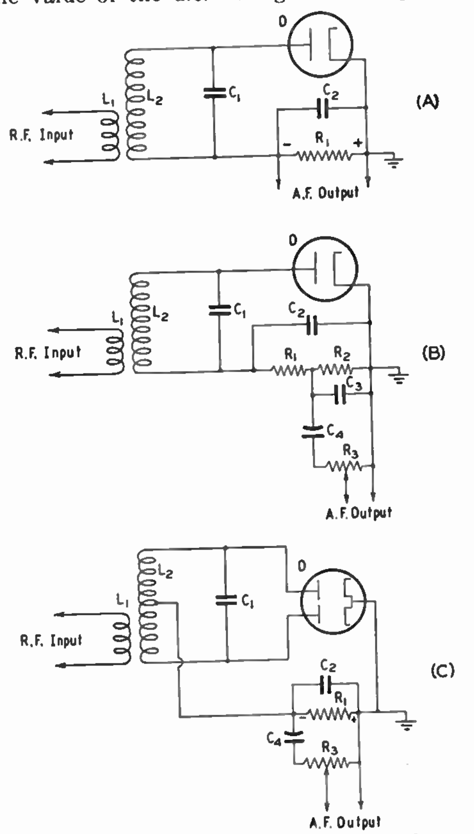


Fig. 5-2 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values for C_2 and R_1 in A and C are 250 μμfd. and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 μμfd. 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

load resistor, R_1, usually 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. 5-3. A typical modulated signal as it exists in the tuned circuit is shown at A. When this signal is applied to the rectifier tube, current will flow 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 that 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. 5-2B), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 5-2B, 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. 5-2B, to a load resistor, R_3, which usually is a "potentiometer" so that the volume can be adjusted to a desired level.

Coupling to the potentiometer (gain control) through a condenser also avoids any flow of d.c. through the gain control. The flow of d.c. through a high-resistance gain control often tends to make the control noisy (scratchy) after a short while.

The full-wave diode circuit at 5-2C 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., and r.f. filtering is easier than in the half-wave circuit.

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. 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 sensitiv-

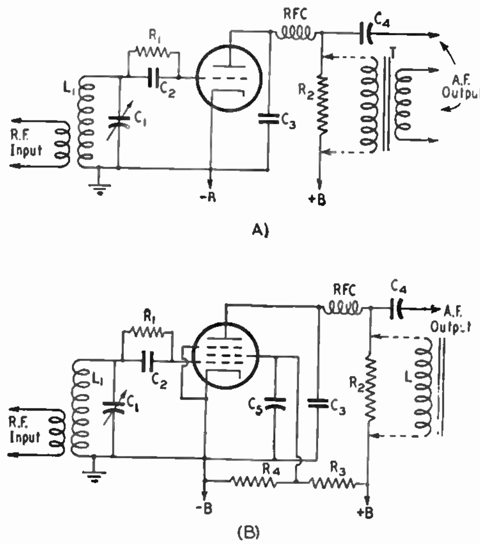


Fig. 5-4 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit 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.
L		500-henry choke.
RFC	2.5 mh.	2.5 mh.
T	Audio transformer.	

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.

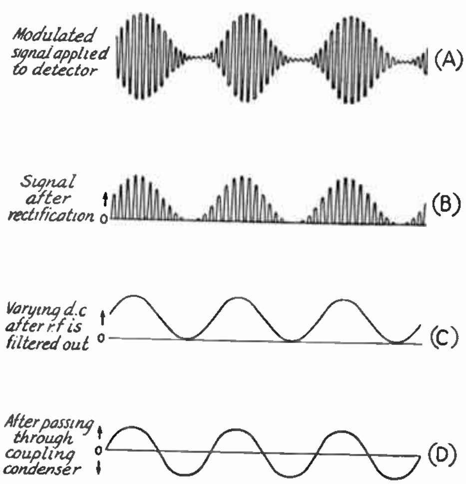


Fig. 5-3 — Diagrams showing the detection process.

ity of the diode is low. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity. 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 circuits of Fig. 5-4, 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, C_3 is a by-pass condenser and *RFC* an r.f. choke to eliminate 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 5-4B. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, C_5, should have low reactance for both radio and audio frequencies. R_3 and R_4 constitute a voltage divider on 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. The reactance of *RFC* will be high for r.f. and low for audio frequencies.

Because of the high plate resistance of the screen-grid tube, transformer coupling from the plate circuit of a screen-grid detector is not satisfactory. An impedance (L in Fig. 5-4B) 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, 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. The signal-handling capability can be improved by reducing R_1 to 0.1 megohm, but the sensitivity will be decreased.

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. 5-5. C_3 is the plate by-pass condenser, and, with *RFC*, prevents r.f. from appearing in the output. R_1 is the cathode resistor which provides the operating grid bias, and C_2 is a by-pass for both radio and audio frequencies across R_1. R_2 is the plate load resistance 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. Impedance coupling may be used in place of the resistance

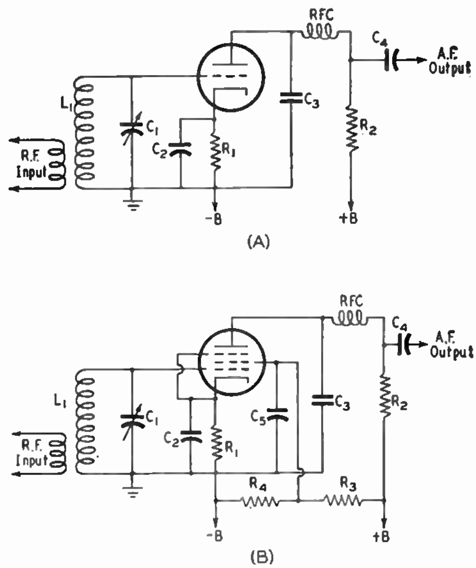


Fig. 5-5 — Circuits for plate detection. A, triode; B, pentode. The input circuit, L_1C_1, is tuned to the signal frequency. Typical values for the other components are:

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.
RFC	2.5 mh.	2.5 mh.

Plate voltages from 100 to 250 volts may be used. Effective screen voltage in B should be about 30 volts.

coupling shown in Fig. 5-5. The same order of inductance is required as with the pentode grid-leak detector described previously.

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.

Infinite-Impedance Detector

The circuit of Fig. 5-6 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 it connects to. 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_2) but not for audio, while the plate circuit is by-passed to ground for both audio and radio frequencies. R_2 forms, with C_3, an RC filter to isolate the plate from the "B" supply at a.f. An r.f. filter, consisting of a series r.f. choke and a shunt condenser, can be connected between the cathode and C_4 to eliminate any r.f. that might otherwise appear in the output.

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.

REGENERATIVE DETECTORS

By providing controllable r.f. feed-back or regeneration 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 increases the selectivity because 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, and the critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. In the oscillating condition, a regenerative detector can be detuned slightly from an incoming c.w. signal to give *autodyne* reception.

Fig. 5-7 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 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 5-7B is for a pentode 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. The feed-back is adjusted by varying the number of turns on L_3 or the coupling between L_2 and L_3, until the tube just goes into oscillation at a screen potential of approximately 30 volts.

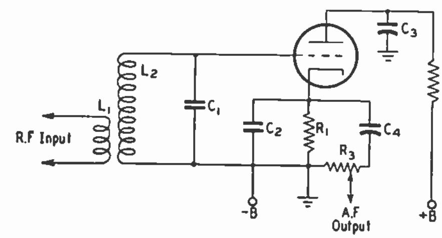


Fig. 5-6 — The infinite-impedance detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:

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.

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type. 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.

Smooth Regeneration Control

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 usually 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 that 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 feedback 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 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 that 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 that is present only when the antenna is connected is caused by resonance effects in the antenna, which tend to 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. The hum may result from the use of a.c. on the tube heater, but effects of this type normally are troublesome

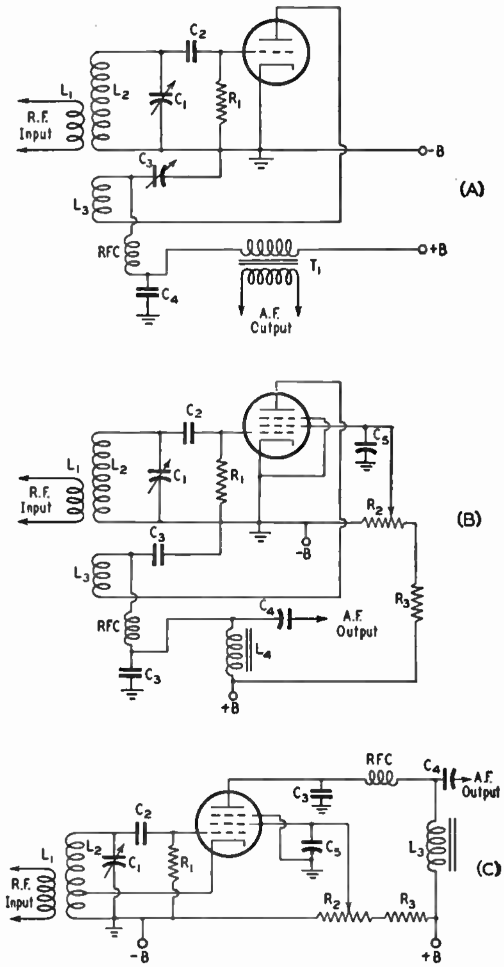


Fig. 5-7 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 $\mu\mu$fd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily 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 $\mu\mu$fd. or more; by-pass condensers C_3 in B and C are likewise 100 $\mu\mu$fd. 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 potential may be 100 to 250 volts.

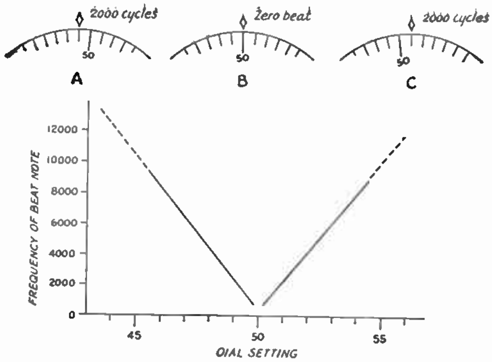


Fig. 5-8 — As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.

only when the circuit of Fig. 5-7C 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 in the power-supply system).

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, when the receiver is put "above ground" by the antenna, as described in a preceding paragraph. The effect 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. 5-8. 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 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 satisfactorily eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the detector 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 detector 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.

Audio-Frequency Amplifiers

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 are sufficient for headphones, but a loudspeaker requires a watt or more for good room volume.

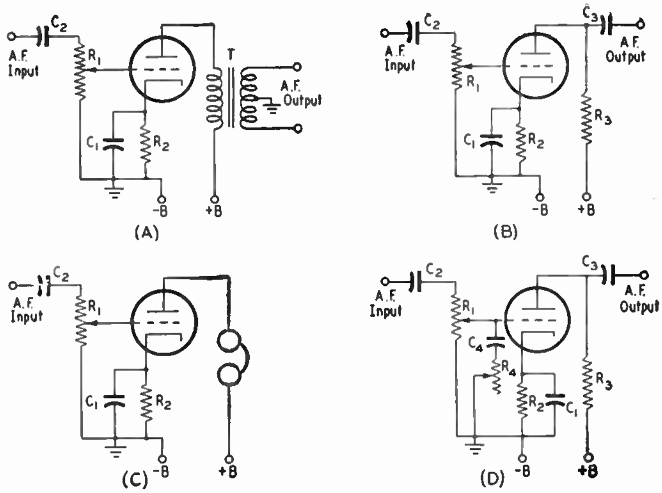


Fig. 5-9 — Audio-amplifier circuits used for voltage amplification and to provide power for headphone output. The tubes are operated as Class A amplifiers. A simple tone-control circuit (R_4C_4) is shown in D. For R_1 = 1 megohm, R_4 can be 0.1 megohm and C_4 0.01 μfd.

◆

Headset and Voltage Amplifiers

The circuits shown in Fig. 5-9 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones. 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. 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. 5-4 and 5-5, if the amplifier is coupled to a detector. In 5-9D, C_4 and R_4 are a simple "tone-control" circuit. As R_4 is made smaller, C_4 by-passes more of the high audio frequencies. R_4 should be large compared to the reactance of C_4 at the highest audio frequency.

In all receivers using tubes with indirectly-heated cathodes, the negative grid bias of audio amplifiers usually is secured from the voltage drop in a cathode resistor. 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). In battery-operated receivers, which use filament-type tubes, a separate grid-bias battery generally is used.

Power Amplifiers

A popular type of power amplifier is the single tetrode, operated Class A or AB; the circuit diagram is given in Fig. 5-10A. The grid resistor, R_1, may be a potentiometer for volume control, as shown at R_1 in Fig. 5-9. The output transformer, T, should have a turns ratio 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 tetrodes or pentodes may be connected in push-pull, as shown in Fig. 5-10B. 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, in Class A operation since the a.f. current does not flow through the resistor as it does in single-tube circuits.

Headphones and Loudspeakers

Two types of headphones are in general use, the **magnetic** and **crystal** types. They are shown in cross-section in Fig. 5-11. 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. 5-11, are called "double-

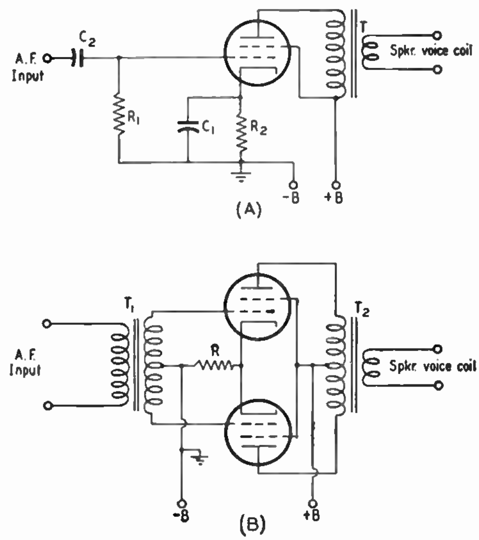


Fig. 5-10 — Power-output audio-amplifier circuits. Either Class A or AB amplification may be used.

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

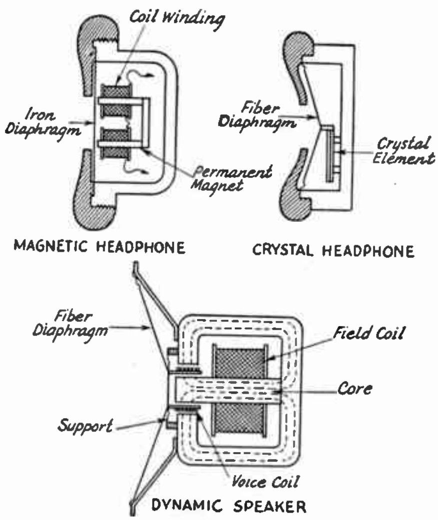


Fig. 5-11 — Headphone and loudspeaker construction.

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 made of Rochelle salts are cemented together in such a way that the pair tend 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 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, provided 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, that 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. 5-11. 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. 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, 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. 5-11 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.

Tuning and Band-Changing Methods

Band-Changing

The resonant circuits that 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

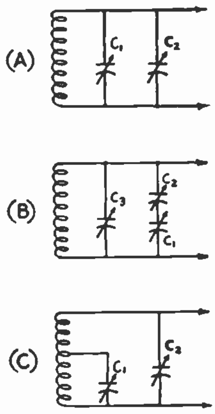


Fig. 5-12 — Essentials of the three basic band-spread tuning systems.

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.

One method of changing inductances is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. Another 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. 5-12.

In A, a small **bandspread condenser,** C_1 (15- to 25-$\mu\mu$fd. maximum capacity), is used in parallel with a condenser, C_2, which is usually large enough (100 to 140 $\mu\mu$fd.) 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 nonharmonic 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 $\mu\mu$fd. 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 $\mu\mu$fd. 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 preadjusted 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 $\mu\mu$fd. 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 preset, 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 inductances and 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. 5-13, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer necessarily

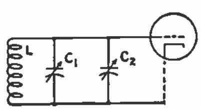


Fig. 5-13 — Showing the use of a trimmer condenser, to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget condensers having maximum capacities of 15 to 30 $\mu\mu$fd. are commonly used.

The same methods are applied to bandspread circuits that must be tracked. The circuits are identical with those of Fig. 5-12. 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. 5-12B, and C_2 in Fig. 5-12C, serve as trimmers.

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

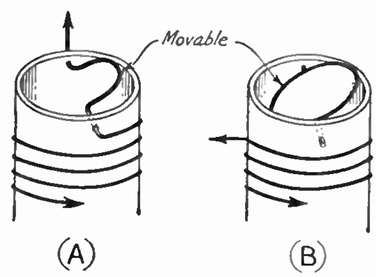


Fig. 5-14 — Methods of adjusting the inductance for ganging. The half-turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the axis of the loop is perpendicular to the axis of the coil, and will give maximum reduction of the coil inductance when rotated 90°. The loop can be a solid disk of metal and give exactly the same effect.

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. 5-14.

Still another method for trimming the inductance is to use an adjustable brass (or copper) or powdered-iron core. The brass core acts like a single shorted turn, and the inductance of the coil is decreased as the brass core, or **"slug,"** is moved into the coil. The powdered-iron core has the opposite effect, and *increases* the inductance as it is moved into the coil. The Q of the coil is not affected materially by the use of the brass slug, provided the brass slug has a clean surface or is silverplated. The use of the powdered-iron core will actually raise the Q of a coil, provided the iron core is of a type suitable for the frequency in use. Good powdered-iron cores can be obtained for use up to about 50 Mc.

The Superheterodyne

For many years (up to about 1932) practically the only type of receiver to be found in amateur stations consisted of a regenerative detector and one or more stages of audio amplification. Receivers of this type can be made quite sensitive but they are lacking in stability and selectivity, particularly on the higher frequencies. Strong signals block them easily and, in our present crowded bands, they are seldom used except in emergencies. They have been replaced by **superheterodyne** receivers, generally called "superhets."

The Superheterodyne Principle

In a superheterodyne receiver, the frequency of the incoming signal is changed to a new radio frequency, the **intermediate frequency** (abbreviated "i.f."), then amplified, and finally detected. The frequency is changed by means of the heterodyne process, the output of a tunable oscillator (the **high-frequency, or local, 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. The audio-frequency signal is obtained at the **second detector.** C.w. signals are made audible by autodyne or heterodyne reception at the second detector.

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 high-frequency oscillator frequency may be set to 7455 kc., in order that the beat frequency (7455 minus 7000) will be 455 kc. The high-frequency oscillator could also be set to 6545 kc. and give the same difference frequency. To produce an audible c.w. signal at the second detector of, say, 1000 cycles, the autodyning or heterodyning oscillator would be set to either 454 or 456 kc.

The frequency-conversion process permits r.f. amplification at a relatively low frequency, the i.f. High selectivity and gain can be obtained at this frequency, and this selectivity and gain are constant. The separate oscillators can be designed for 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 tune to a 7000-kc. signal, for example, the receiver can respond also to a signal on 7910 kc., which likewise gives a 455-kc. beat. The resultant undesired signal of the two frequencies 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 or eliminates the response to the image signal. 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 further away from the resonance peak of the signal-frequency input circuits. Most receiver designs represent a compromise between economy (few r.f. stages) and image rejection (large number of r.f. stages).

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 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.) intermediate frequency, 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.**

FREQUENCY CONVERTERS

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, to offer a high impedance to the i.f. voltage that is developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are by-passed 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.

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. 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 intermediate frequencies. Another type of pulling is caused by regulation in the power supply. Strong signals cause the supply voltage to change, and this in turn shifts the oscillator frequency.

Circuits

If the first detector and high-frequency oscillator are separate tubes, the first detector is called a "mixer." If the two are combined in one envelope (as is often done for reasons of economy or efficiency), the first detector is called a "converter." In either case the function is the same, however.

Typical mixer circuits are shown in Fig. 5-15. The variations are chiefly in the way in which the oscillator voltage is introduced. In 5-15A, a pentode functions as a plate detector; the oscillator voltage is capacity-coupled to the grid of the tube through C_2. 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. If the signal frequency is only 5 or 10 times the i.f., it may be difficult to develop enough oscillator voltage at the grid (because of the selectivity of the tuned input circuit). However, the circuit is a sensitive one and makes a good mixer, particularly with high-G_m tubes like the 6AC7 and 6AK5. A good triode also works well in the circuit, and

TABLE 5-I
Circuit and Operating Values for Converter Tubes

Tube	*Plate Volts*	*Screen Volts*	*Cathode Resistor*	*Screen Resistor*	*Grid Leak*	*Grid I*
6BE6	250	100	100[1,2]	22,000	22,000	0.5 ma.
6K8	250	100	240[1]	27,000	47,000	0.15-0.2
6SA7	250	100	0[1] 160[2]	18,000	22,000	0.5
6SB7Y 6BA7	250	100	0[1] 75[2]	12,000 15,000	22,000	0.35

[1]Self-excited. [2]Separate excitation.

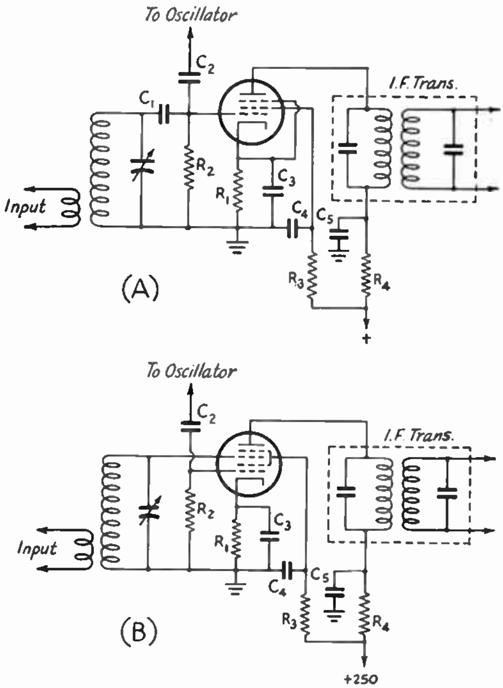


Fig. 5-15 — Typical circuits for separately-excited mixers. Grid injection of a pentode mixer is shown at A, and separate excitation of a pentagrid converter is given in B. Typical values for B will be found in Table 5-I — the values below are for the pentode mixer of A.

C_1 — 10 to 50 $\mu\mu$fd.	R_2 — 1.0 megohm.
C_2 — 5 to 10 $\mu\mu$fd.	R_3 — 0.47 megohm.
C_3, C_4, C_5 — 0.001 μfd.	R_4 — 1500 ohms.
R_1 — 6800 ohms.	

Positive supply voltage can be 250 volts with a 6AC7, 150 with a 6AK5.

tubes like the 7F8 (one section), the 6J6 (one section) and the 6J4 work well. When a triode is used, care should be taken to see that the signal frequency is short-circuited in the plate circuit, and this is done by mounting the tuning capacitor of the i.f. transformer directly from plate to cathode.

It is difficult to avoid "pulling" in a triode or pentode mixer, however, and a pentagrid converter tube used as a mixer provides much better isolation. A typical circuit is shown in Fig. 5-15B, and tubes like the 6SA7, 7Q7 or 6BE6 are commonly used. The oscillator voltage is introduced into the electron stream of the tube through an "injection" grid. Measurement of the rectified current flowing in R_2 is used as a check for proper oscillator-voltage amplitude. Tuning of the signal-grid circuit can have little effect on the oscillator frequency because the injection grid is isolated from the signal grid by a screen grid that is at r.f. ground potential. The pentagrid mixer is not quite as sensitive as a triode or pentode mixer, but its splendid isolating characteristics make it a very useful circuit.

Many receivers use pentagrid converters, and two typical circuits are shown in Fig. 5-16. The circuit shown in Fig. 5-16A, which is suitable for the 6K8, 7D7, 7J7 or 7S7, is for a "triode-hexode" converter. A triode oscillator tube is mounted in the same envelope with a hexode, and the control grid of the oscillator portion is connected internally to an injection grid in the hexode. The isolation between oscillator and converter tube is reasonably good, and very little pulling results, except on signal frequencies that are quite large compared with the i.f.

The pentagrid-converter circuit shown in Fig. 5-16B can be used with a tube like the 6SA7, 7Q7 or 6BE6. Generally the only care necessary is to adjust the feed-back of the oscillator circuit to give the proper oscillator r.f. voltage. This condition is checked by measuring the d.c. current flowing in grid resistor R_2.

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 very little difference to be realized from the cost standpoint.

Typical circuit constants for converter tubes are given in Table 5-I. The grid leak referred to is the oscillator grid leak or injection-grid return, R_2 of Figs. 5-15 and 5-16.

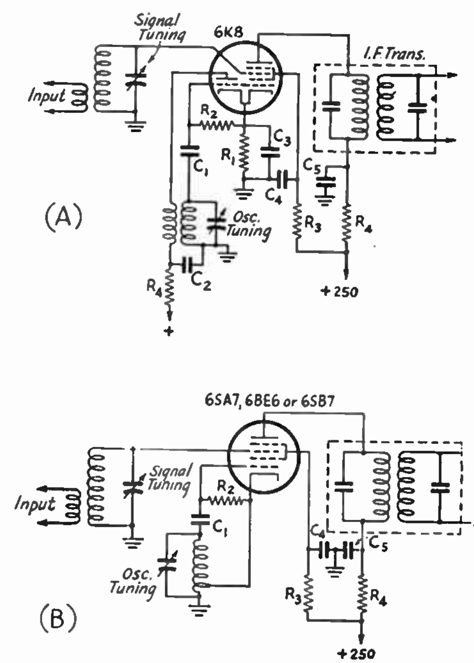


Fig. 5-16 — Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for R_1, R_2 and R_3 can be found in Table 5-I; others are given below.

C_1 — 47 $\mu\mu$fd.	C_3 — 0.01 μfd.
C_2, C_4, C_5 — 0.001 μfd.	R_4 — 1000 ohms.

THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver 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 mechanical shock and changes in voltage and loading. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. They can be reduced by using ceramic instead of bakelite insulation in the r.f. circuits, a large cabinet compared with the chassis (to provide for good radiation of developed heat), minimizing the number of high-wattage resistors in the receiver itself and putting them in the power supply, and not mounting the oscillator coils and tuning condenser too close to a tube.

Sensitivity to vibration and shock can be a bother, and should be minimized by using good mechanical support for coils and tuning condensers, a heavy chassis, and by not hanging any of the oscillator-circuit components in the air on long leads. Tie-points should be used wherever necessary to avoid long leads on components in the oscillator circuits. Stiff long wires used for wiring components are no good if they can vibrate, and stiff *short* leads are excellent because they can't be made to vibrate.

Smooth tuning is a great convenience to the operator, and can be obtained by taking pains with the mounting of the dial and tuning condensers. They should have good alignment and no back-lash. If the condensers are mounted off the chassis on posts instead of brackets, it is almost impossible to avoid some back-lash unless the posts have extra-wide bases. The condensers should be selected with good wiping contacts to the rotor, since with age the rotor contacts can be a source of erratic tuning. All joints in the oscillator tuning circuit should be carefully soldered, since a loose connection or "rosin joint" can develop trouble that is sometimes hard to locate. The chassis and panel materials 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 of a receiver is repaid many times over by increased frequency stability.

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 the possibility of spurious response.

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will reduce tube heating and thereby lower the 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.

If the h.f.-oscillator frequency is affected by

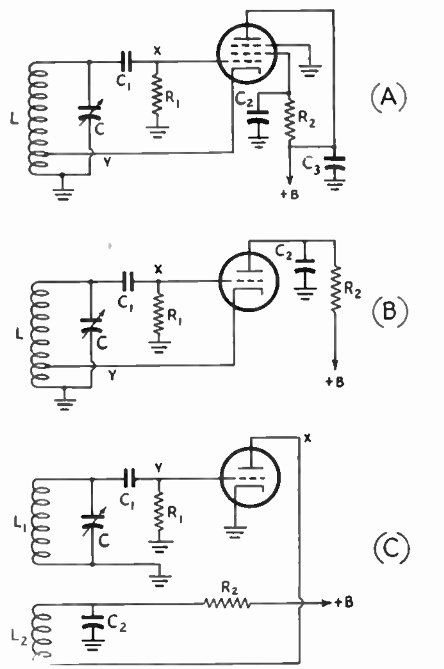


Fig. 5-17 — **High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from *Y* will reduce pulling effects, but gives less voltage 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 —	47,000 ohms.	47,000 ohms.	47,000 ohms.
R_2 —	47,000 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.

changes in plate voltage, it is good practice to use a voltage-regulated plate supply employing a VR tube except, of course, in receivers operated from batteries. Changes in plate-supply voltage are caused not only by variations in the line voltage but by poor regulation in the power supply. When a.v.c. is used, the controlled tubes draw less current from the power supply as the signal increases, and this change in power-supply load causes the power-supply voltage to vary if it isn't regulated. The use of Class AB audio amplification may also cause severe changes in the power-supply voltage.

Circuits

Several oscillator circuits are shown in Fig. 5-17. 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 indirectly-heated-cathode tubes with a.c. on the heaters are used. The circuit of Fig. 5-17C reduces hum because the cathode is grounded. It is a simple circuit to adjust, and it is also the best circuit to use with filament-type tubes. With filament-type tubes, the other two circuits would require r.f. chokes to keep the filament above r.f. ground.

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; 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. Using the oscillator shown at C, feed-back is obtained by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

The Intermediate-Frequency Amplifier

One major advantage of the superhet is that high gain and selectivity can be obtained by using a good i.f. amplifier. This can be a one-stage affair in simple receivers, or two or three stages in the more complex sets.

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. A low i.f. also increases pulling of the oscillator frequency. 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 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 50 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 frequencies of 28 Mc. and higher, the best solution is to use a double superheterodyne, 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.

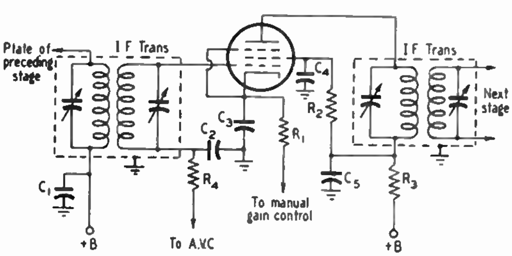


Fig. 5-18 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

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, R_2 — See Table 5-II. R_3 — 1800 ohms.
R_4 — 0.27 megohm.

Fidelity; Sideband Cutting

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 that 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, 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 bandwidths can be used for communication work where intelligibility rather than fidelity is the primary objective. 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 communications effectiveness.

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. At 455 kc. two stages generally give all the gain usable, and also give suitable selectivity for good-quality 'phone reception.

A typical circuit arrangement is shown in Fig. 5-18. A second stage would simply duplicate the circuit of the first. The i.f. amplifier practically always uses a remote cut-off pentode-type tube operated as a Class A amplifier. For maximum selectivity, double-tuned transformers are used for interstage coupling, although single-tuned circuits or transformers with untuned primaries can be used for coupling, with a consequent loss in selectivity. All other things being equal, the selectivity of an i.f. amplifier is proportional to the number of tuned circuits in it. The use of too many high-Q tuned circuits in an amplifier is not generally feasible, however, because of stability problems.

In Fig. 5-18, the gain of the stage is reduced by introducing a negative voltage to the lead marked "to a.v.c." or a positive voltage to R_1 at the point marked "to manual gain control." In either case, the voltage increases the bias on the tube and reduces the mutual conductance and hence the gain. When two or more stages are used, these voltages are generally obtained from common sources. The decoupling resistor, R_3, helps to isolate the amplifier from the power supply and thus prevents stray feed-back. C_2 and R_4 are part of the automatic volume-control circuit (described later); if no a.v.c. is used, the lower end of the i.f.-transformer secondary is simply connected to ground.

TABLE 5-II
Cathode and Screen-Dropping Resistors for R.F. or I.F. Amplifiers

Tube	*Plate Volts*	*Screen Volts*	*Cathode Resistor*	*Screen Resistor*
6AB7	300		200 ohms	33,000 ohms
6AC7	300		160	62,000
6AK5	180	120	200	27,000
6AU6	250	150	68	33,000
6BA6	250	100	68	33,000
6J7	250	100	1200	270,000
6K7	250	125	240	47,000
6SG7	250	125	68	27,000
6SG7	250	150	200	47,000
6SH7	250	150	68	39,000
6SJ7	250	100	820	180,000
6SK7	250	100	270	56,000
7G7/1232	250	100	270	68,000
7H7	250	150	180	27,000

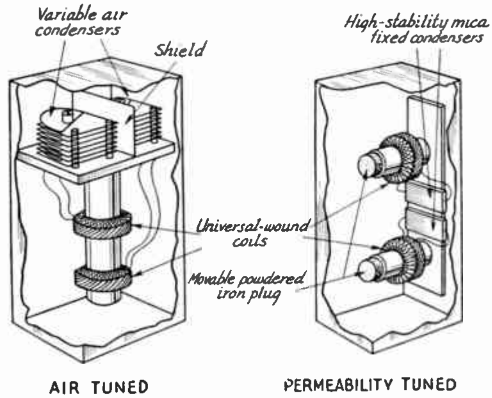


Fig. 5-19 — Representative i.f.-transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores 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.

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 across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

Typical values of cathode and screen resistors for common tubes are given in Table 5-II. The 6K7, 6SK7, 6SG7, 6BA6 and 7H7 are recommended for i.f. work.

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, are 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 with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multilayer winding, a fairly large capacity can exist between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects.

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected

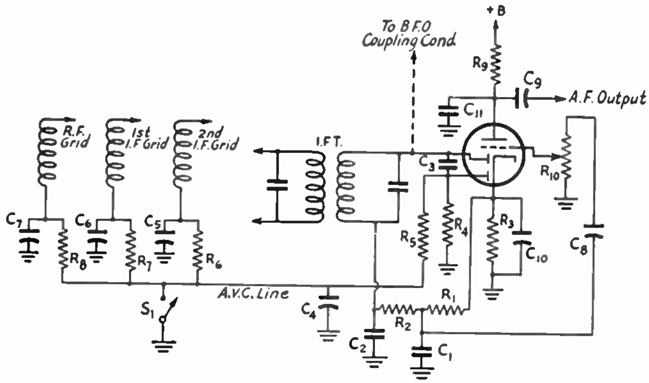


Fig. 5-20 — Automatic volume-control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first a.f. amplifier.
R_1 — 0.27 megohm.
R_2 — 50,000 to 250,000 ohms.
R_3 — 1800 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.5 to 1 megohm.
R_6, R_7, R_8, R_9 — 0.25 megohm.
R_{10} — 0.5-megohm variable.
C_1, C_2, C_3 — 100 $\mu\mu$fd.
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} — 270 $\mu\mu$fd.

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. 5-19.

Besides the type of i.f. transformer shown in Fig. 5-19, special units to give desired selectivity characteristics are available. For higher-than-ordinary adjacent-channel selectivity **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 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.

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 bandwidths to be expected with good-quality transformers in amplifiers so constructed as to keep regeneration at a minimum:

Intermediate Frequency	*Bandwidth in Kilocycles* 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 (remote cut-off) pentodes are almost invariably used in i.f. amplifier stages, since grid-bias gain control 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).

When single-ended tubes 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, crosswise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be connected to ground.

● THE SECOND DETECTOR AND BEAT OSCILLATOR

Detector Circuits

The second detector of a superheterodyne receiver with an i.f. amplifier performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great amplification ahead of it. 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. The basic circuits have been described, although in many cases the diode elements are incorporated in a multipurpose tube that contains an amplifier section in addition to the diode.

The Beat Oscillator

Any standard oscillator circuit may be used for the beat oscillator required for heterodyne reception. 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. 5-17A and B, with the output taken from Y. A variable condenser of about 25-μμfd. 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 μμfd. 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 with desired 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.

When an oscillating second detector is used to give the audio beat note, the detector must be detuned from the i.f. by an amount equal to the frequency of the beat note. The selectivity and signal strength will be reduced, while blocking will be pronounced because of the high signal level at the second detector.

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, to vary the bias on the r.f. and i.f. amplifier tubes. 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. 5-20. 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, 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 amplifiers 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. 5-20. 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. 5-20 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 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. will be unable to follow rapid fading. The capacitance and resistance values indicated in Fig. 5-20 will give a time constant that is satisfactory for high-frequency reception.

C.W.

A.v.c. can be used for c.w. reception but the circuit is more complicated. The a.v.c. voltage must be derived from a rectifier that is isolated from the beat-frequency oscillator (otherwise the rectified b.f.o. voltage will reduce the receiver gain even with no signal coming through). This is generally done by using a separate a.v.c. channel connected to an i.f. amplifier stage ahead of the second detector (and b.f.o.). If the i.f. selectivity ahead of the a.v.c. rectifier isn't good, strong adjacent signals will develop a.v.c. voltages that will reduce the receiver gain while listening to weak signals. When clear channels are available, however, c.w. a.v.c. will hold the receiver output constant over a wide range of signal input. A.v.c. systems designed to work on c.w. signals must have fairly long time constants to work with slow-speed sending, and often a selection of time constants is made available.

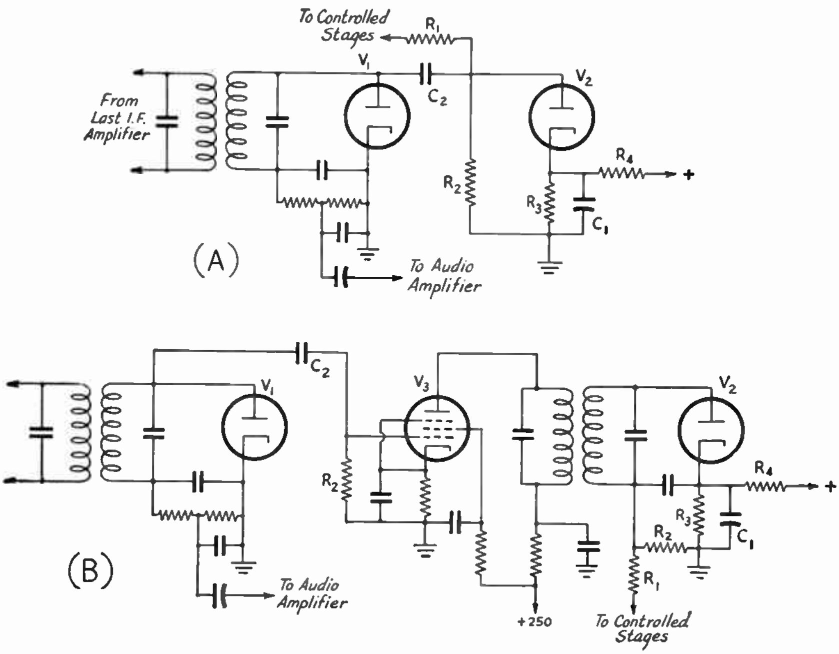


Fig. 5-21 — Delayed a.v.c. is shown at A, and amplified and delayed a.v.c. is shown in B. The circuit at B gives excellent a.v.c. action over a wide range, with no impairment of sensitivity for weak signals. For either circuit, typical values are:

C_1 — 0.001 μfd. R_1, R_2 — 1.0 megohm.
C_2 — 100 μμfd. R_3, R_4 — Voltage divider.

Resistors R_3 and R_4 are carefully proportioned to give the desired delay voltage at the cathode of diode V_2. Bleeder current of 1 or 2 ma. is ample, and hence the bleeder can be figured on 1000 or 500 ohms per volt. The delay voltage should be in the vicinity of 3 or 4 for a simple receiver and 20 or 30 in the case of a multitube high-gain affair.

Amplified A.V.C.

The a.v.c. system shown in Fig. 5-20 will not hold the audio output of the receiver exactly constant, although the variation becomes less as more stages are controlled by the a.v.c. voltage. The variation also becomes less as the delay voltage is increased, although there will, of course, be variation in output if the signal intensity is below the delay-voltage level at the a.v.c. rectifier. In the circuit of Fig. 5-20, the delay voltage is set by the proper operating bias for the triode portion of the tube. However, a separate diode may be used, as shown in Fig. 5-21A. Since such a system requires a large voltage at the diode, a separate i.f. stage is sometimes used to feed the delayed a.v.c. diode, as in Fig. 5-21B. A system like this, sometimes called an amplified a.v.c. system, gives excellent control once the delay voltage is reached, and yet maintains full receiver sensitivity up to that point.

Noise Reduction

Types of Noise

In addition to tube and circuit noise, 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" 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 with 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 with 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 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 without fading. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous circuits.

SECOND-DETECTOR NOISE-LIMITER CIRCUITS

The circuit of Fig. 5-22 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes nonconducting 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 nonconducting 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 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. 5-22A, using an infinite-impedance detector, gives a positive voltage on rectification. When the rectified voltage is negative, as it is from the usual diode detector,

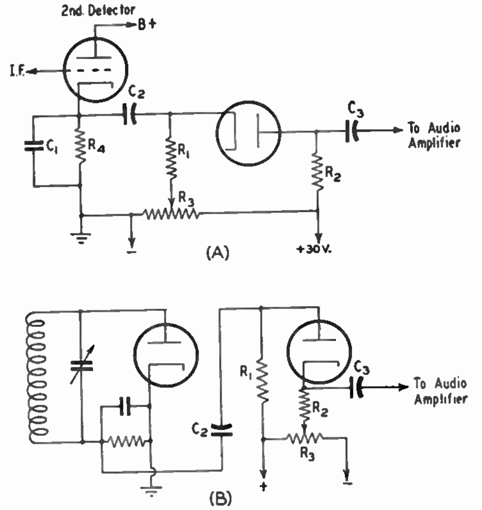


Fig. 5-22 — Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

R_1 — 0.27 megohm.	R_4 — 20,000 to 50,000 ohms.
R_2 — 47,000 ohms.	C_1 — 270 $\mu\mu$fd.
R_3 — 10,000 ohms.	C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

the circuit arrangement shown in Fig. 5-22B 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.

Second-detector noise-limiting circuits that automatically adjust themselves to the receiver carrier level are shown in Fig. 5-23. In either circuit, V_1 is the usual diode second detector, R_1R_2 is the diode load resistor, and C_1 is an r.f. by-pass. A negative voltage proportional to the carrier level is developed across C_2, and this voltage cannot change rapidly because R_3 and C_2 are both large. In the circuit at A, diode V_2 acts as a conductor for the audio signal up to the point where its anode is negative with respect to the cathode. Noise peaks that exceed the maximum carrier-modulation level will drive the anode negative instantaneously, and during this time the diode does not conduct. The large time constant of C_2R_3 prevents any rapid change of

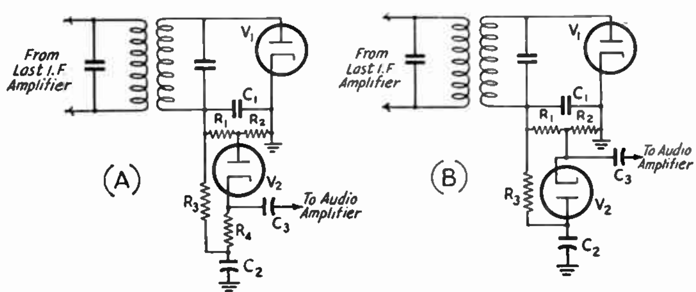


Fig. 5-23 — Self-adjusting series (A) and shunt (B) noise limiters. The functions of V_1 and V_2 can be combined in one tube like the 6H6 or 6AL5, or Type 1N34 crystals can be used.

C_1 — 100 $\mu\mu$fd.
C_2, C_3 — 0.05 μfd.
R_1 — 0.27 meg. in A; 47,000 ohms in B.
R_2 — 0.27 meg. in A; 0.15 meg. in B.
R_3 — 1.0 megohm.
R_4 — 0.82 megohm.

this reference voltage. In the circuit at B, the diode V_2 is inactive until its cathode voltage exceeds its anode voltage. This condition will obtain under noise peaks and, when it does, the diode V_2 short-circuits the signal and no voltage is passed on to the audio amplifier. Practical values for the circuit at B will be found in the eight-tube super-heterodyne described later in this chapter. Diode rectifiers such as the 6H6 and 6AL5, or the 1N34 germanium crystal diode, can be used for these types of noise limiters. Neither circuit is useful for c.w. reception, but they are both quite effective for 'phone work.

I.F. Noise Silencer

In the circuit shown in Fig. 5-24, 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. In a typical instance, 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.

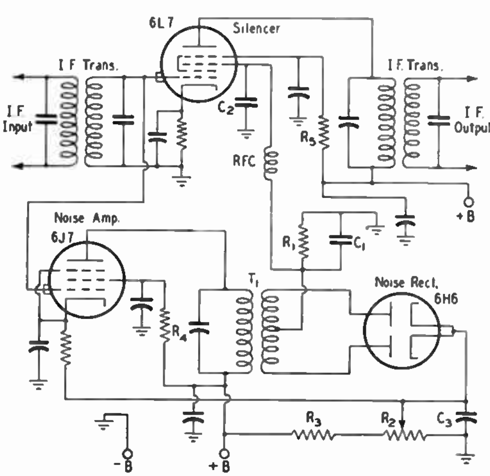


Fig. 5-24 — **I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:**

C_1 — 50–250 $\mu\mu$fd. (use smallest value possible without r.f. feed-back).
C_2 — 47 $\mu\mu$fd.
C_3 — 0.1 μfd.
R_1 — 0.1 megohm.
R_2 — 5000-ohm variable.
R_3 — 22,000 ohms.
R_4, R_5 — 0.1 megohm.
T_1 — Special i.f. transformer for noise rectifier.

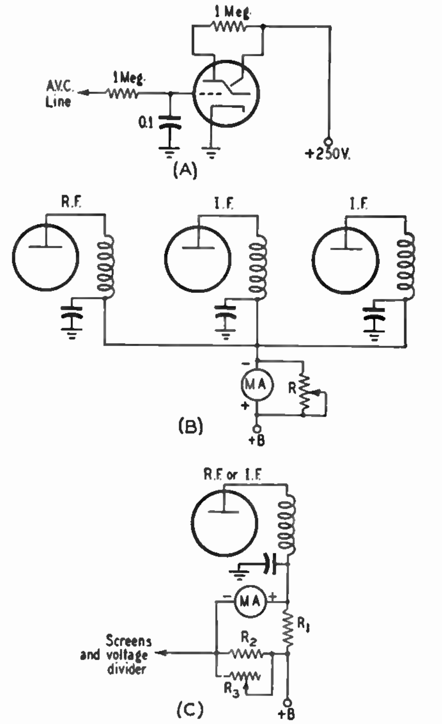


Fig. 5-25 — **Tuning-indicator or S-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor *R* should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: R_1, 270 ohms; R_2, 330 ohms; R_3, 1000-ohm variable.**

SIGNAL-STRENGTH AND TUNING INDICATORS

A useful accessory to the receiver is an indicator that 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. 5-25. That at A uses an electron-ray tube, 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, 6N5 or 6AD6G) 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 downward 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 milliammeter in a bridge circuit, arranged so that the 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. 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.

Improving Receiver Selectivity

● INTERMEDIATE-FREQUENCY AMPLIFIERS

As mentioned earlier in this chapter, one of the big advantages of the superheterodyne receiver is the improved selectivity that is possible. This selectivity is obtained in the i.f. amplifier, where the lower frequency allows more selectivity per stage than at the higher signal frequency. For 'phone reception, the limit to useful selectivity in the i.f. amplifier is the point where so many of the sidebands are cut that intelligibility is lost, although it is possible to remove completely one full set of sidebands without impairing the quality at all. Maximum receiver selectivity in 'phone reception requires excellent stability in both transmitter and receiver, so that they will both remain "in tune" during the transmission. The limit to useful selectivity in code work is around 50 or 100 *cycles* for hand-key speeds, but it is difficult to use this much selectivity because it requires remarkable stability in both transmitter and receiver, and to tune in a signal becomes a major problem.

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., or if the receiver is tuned to heterodyne the incoming signal to 457 kc., it will also be heterodyned by the beat oscillator to produce a 1000-cycle beat. Hence every signal can be tuned in at two places that will give a 1000-cycle beat (or any other low audio frequency). This **audio-frequency image** effect can be reduced if the i.f. selectivity is such that the incoming signal, when heterodyned to 457 kc., is attenuated to a very low level.

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 nonregenerative amplifiers using ordinary tuned circuits unless a very low i.f. 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 bandwidth 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 receiver 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. 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. 5-26 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 bandwidth, reduces the response of the receiver to noise both from sources external to the receiver and in the radio-frequency stages of the receiver itself.

Crystal-Filter Circuits; Phasing

Several crystal-filter circuits are shown in Fig. 5-27. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit, 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 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. 5-26, 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.

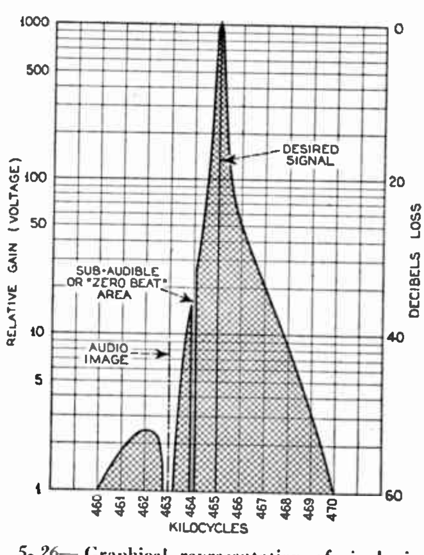


Fig. 5-26— Graphical representation of single-signal selectivity. The shaded area indicates the over-all bandwidth, or region in which response is obtainable.

Variable Selectivity

In circuits such as A and B, Fig 5-27, 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. 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 over-all selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

● RADIO-FREQUENCY AMPLIFIERS

While selectivity to reduce audio-frequency images can be built into the i.f. amplifier, discrimination against radio-frequency images can only be obtained in circuits ahead of the first detector. These tuned circuits and their associated vacuum tubes are called **radio-frequency amplifiers.** For top performance of a communications receiver on frequencies above 7 Mc., it is mandatory that it have one or two stages of r.f. amplification, for image rejection and improved sensitivity. (The improvement in sensitivity that can be obtained will be discussed later.)

Receivers with an i.f. of 455 kc. can be expected to have some r.f. image response at a signal frequency of 14 Mc. and higher if only one stage of r.f. amplification is used. (Regeneration in the r.f. amplifier will reduce image

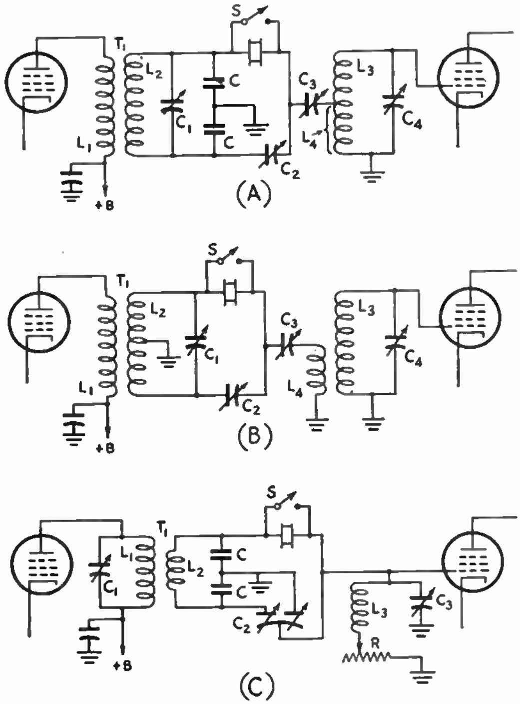


Fig. 5-27 — Crystal-filter circuits of three types. All give variable bandwidth, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely 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_2L_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).

response, but regeneration is often a tricky thing to control.) With two stages of r.f. amplification and an i.f. of 455 kc., no images should be apparent at 14 Mc., but they will show up on 28 Mc. and higher. Three stages or more of r.f. amplification, with an i.f. of 455 kc., will reduce the images at 28 Mc., but it really takes four or more stages to do a good job. The better solution at 28 Mc. is to use a "triple-detection" superheterodyne, with one stage of r.f. amplification and a first i.f. of 1600 kc. or higher. A regular receiver with an i.f. of 455 kc. can be converted to a triple-detection superhet by connecting a "converter" (to be described later) ahead of the receiver.

For best selectivity, r.f. amplifiers should use high-Q circuits and tubes with high input and output resistance. Variable-μ pentodes are practically always used, although triodes (neutralized or otherwise connected so that they won't oscillate) are often used on the higher frequencies because they introduce less noise. Pentodes are better where maximum image rejection is desired, because they have less loading effect on the circuits.

Feed-Back

Feed-back giving rise to regeneration and oscillation can occur in a single stage or it may appear as an over-all feed-back through several stages that are on the same frequency. To avoid feed-back in a single stage, the output must be isolated from the input in every way possible, with the vacuum tube furnishing the only coupling between the two circuits. For example, an oscillation can be obtained in an r.f. or i.f. stage if there is any undue capacitive or inductive coupling between output and input circuits, if there is too high an impedance between cathode and ground or screen and ground, or if there is any appreciable impedance through which the grid and plate currents can flow in common. This simply means good shielding of coils and condensers in r.f. and i.f. circuits, the use of good by-pass condensers (mica at 14 Mc. and higher, and with short leads), and returning all by-pass condensers (grid, cathode, plate and screen) with short leads to one spot on the chassis. If single-ended tubes are used, the screen or cathode by-pass condenser should be mounted across the socket, to serve as a shield between grid and plate pins. Less care is required as the frequency is lowered, but in high-impedance circuits, it is sometimes necessary to shield grid and plate leads and to be careful not to run them close together.

To avoid over-all feed-back in a multistage amplifier, strict attention must be paid to avoid running any part of the output circuit back near the input circuit without first filtering it carefully. Since the signal-carrying parts of the circuit (the "hot" grid and plate leads) can't be filtered, the best design for any multistage amplifier is a straight line, to keep the output as far away from the input as possible. For example, an r.f. amplifier might run along a chassis in a straight line, run into a mixer where the frequency is changed, and then the i.f. amplifier could be run back parallel to the r.f. amplifier, provided there was a very large frequency difference between the r.f. and the i.f. amplifiers. However, to avoid any possible coupling, it would be better to run the i.f. amplifier off at right angles to the r.f.-amplifier line, just to be on the safe side. Good shielding is important in preventing over-all oscillation in high-gain-per-stage amplifiers, but it becomes less important when the stage gain drops to a low value. In a high-gain amplifier, the power leads (including the heater circuit) are common to all stages, and they can provide the over-all coupling if they aren't properly filtered. Good by-passing and the use of series isolating resistors will generally eliminate any possibility of coupling through the power leads. R.f. chokes, instead of resistors, are used in the heater leads where necessary.

● CROSS-MODULATION

Since a one- or two-stage r.f. amplifier will have a passband measured in hundreds of kc. at 14 Mc. or higher, strong signals will be amplified through the r.f. amplifier even though it is not tuned exactly to them. If these signals are strong enough, their amplified magnitude may be measurable in volts after passing through several r.f. stages. If an undesired signal is strong enough after amplification in the r.f. stages to shift the operating point of a tube (by driving the grid into the positive region), the undesired signal will modulate the desired signal. This effect is called **cross-modulation,** and is often encountered in receivers with several r.f. stages that are working at high gain. It is readily detectable as a superimposed modulation on the signal being listened to, and often the effect is that a signal can be tuned in at several points. It can be reduced or eliminated by greater selectivity in the antenna and r.f. stages (difficult to obtain), the use of variable-μ tubes in the r.f. amplifier, reduced gain in the r.f. amplifier, or reduced antenna input to the receiver.

Gain Control

To avoid cross-modulation and other overload effects in the first detector and r.f. stages, the gain of the r.f. stages is usually made adjustable. This is accomplished by using variable-μ tubes and varying the d.c. grid bias, either in the grid or cathode circuit. If the gain control is automatic, as in the case of a.v.c., the bias is controlled in the grid circuit. Manual control of r.f. gain is generally done in the cathode circuit. A typical r.f. amplifier stage with the two types of gain control is shown in Fig. 5-28.

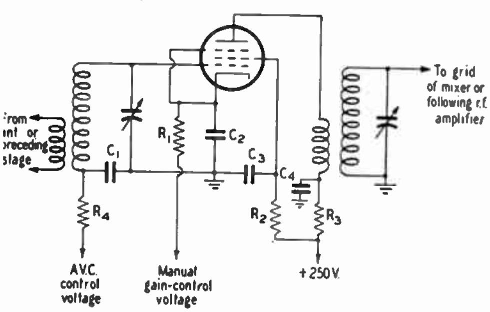


Fig. 5-28 — Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1, C_2, C_3, C_4 — 0.01 μfd. below 15 Mc., 0.001 μfd. at 30 Mc.
R_1, R_2 — See Table 5-II.
R_3 — 1800 ohms.
R_4 — 0.22 megohm.

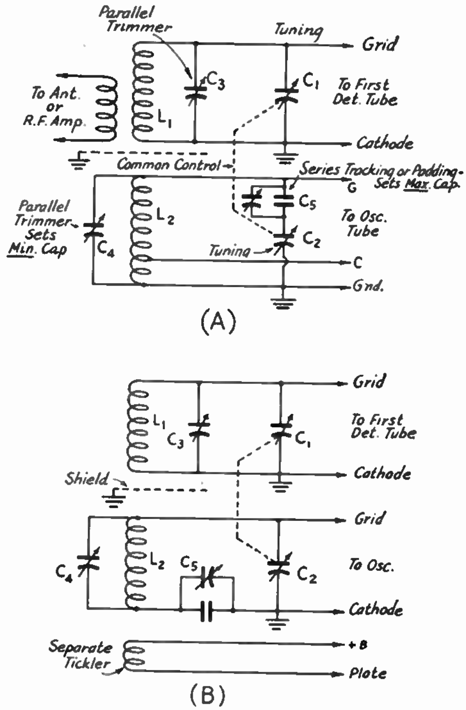


Fig. 5-29 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s. with tuning ranges of approximately 2.15-to-1 and C_2 having 140-$\mu\mu$fd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 35 $\mu\mu$fd.

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

Tracking

In a simple receiver with no r.f. stage, it is no inconvenience to adjust the high-frequency oscillator and the mixer circuit independently, because the mixer tuning is broad and requires little attention over an amateur band. However, when r.f. stages are added ahead of the mixer, the selectivity of the r.f. stages and mixer makes it awkward to use a two-control receiver over an entire amateur band, even though the mixer and r.f. stages are ganged and require only one control. Hence most receivers with one or more r.f. stages gang all of the tuning controls to give a single-tuning-control receiver. Obviously there must exist a constant difference in frequency (the i.f.) between the oscillator and the mixer/r.f. circuits, and when this condition is achieved the circuits are said to **track.**

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 5-29. 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. 5-29. The coils can be calculated quite closely by using the *ARRL Lightning Calculator*, but they will have to be trimmed in the circuit for best tracking.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit that 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. 5-12C, 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 5-12A if the value of C_1 is small compared with C_2.

Improving Receiver Sensitivity

Early in this chapter it was pointed out that the sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies above 20 Mc. is dependent upon the bandwidth of the receiver and the noise contributed by the "front end" of the receiver. Neglecting the fact that the image rejection is poor, a receiver with no r.f. stage is generally satisfactory, from a sensitivity point, in the 3.5- and 7-Mc. bands. However, as the frequency is increased and the atmospheric noise becomes less, the advantage of a good "front end" becomes apparent. Hence at 14 Mc. and higher it is worth while to use at least one stage of r.f. amplification ahead of the first detector for best sensitivity as well as image rejection. The multigrid converter tubes have very poor noise figures, and even the best pentodes and triodes are three or four times noisier when used as mixers as they are when used as amplifiers.

If the purpose of an r.f. amplifier is to improve the receiver noise figure at 14 Mc. and higher, a high-G_m pentode or triode should be used. Among the pentodes, the best tubes are the 6AC7, 6AK5 and the 6SG7, in the order named. The 6AK5 takes the lead above 30 Mc. The 6J4, 6J6 and 7F8 are the best of the triodes. For best noise figure, the antenna circuit should be coupled a little heavier than optimum. This condition leads to poor selectivity in the antenna circuit, so it is rather futile to try to combine best sensitivity with maximum selectivity in this circuit.

When a receiver is satisfactory in every respect (stability and selectivity) except sensitivity on 14 and/or 28 Mc., the best solution for the amateur is to add a **preamplifier,** a stage or two of r.f. amplification designed expressly to improve the sensitivity. If image rejection is lacking in the receiver, some selectivity should be built into the preamplifier (it is then called a preselector). If, however, the receiver operation is poor on the higher frequencies but is satisfactory on the lower ones, a "converter" is the best solution.

Some commercial receivers that appear to lack sensitivity on the higher frequencies can be improved simply by tighter coupling to the antenna. Since the receiver manufacturer has no way to predict the type of antenna that will be used, he generally designs the input for some compromise value, usually around 300 or 400 ohms in the high-frequency ranges. If your antenna matches to something far different from this, the receiver effectiveness can be improved by proper matching. This can be accomplished by changing the antenna to the right value (as determined from the receiver instruction book) or by using a simple matching device as described later in this chapter. Overcoupling the input circuit will often improve sensitivity but it will, of course, always reduce the image-rejection contribution of the antenna circuit.

Commercial receivers can also be "hopped up" by substituting a high-G_m tube in the first r.f. stage if one isn't already there. The amateur must be prepared to take the consequences, however, since the stage may oscillate, or not track without some modification. A simpler solution is to add the "hot" r.f. stage ahead of the receiver.

Regeneration

Regeneration in the r.f. stage of a receiver (where only one stage exists) will often improve the sensitivity because the greater gain it provides serves to mask more completely the first-detector noise, and it also provides a measure of automatic matching to the antenna through tighter coupling. However, accurate ganging becomes a problem, because of the increased selectivity of the regenerative r.f. stage, and the receiver almost invariably becomes a two-handed-tuning device. Regeneration should not be overlooked as an expedient, however, and many amateurs have used it with considerable success. High-G_m tubes are the best as regenerative amplifiers, and the feed-back should not be controlled by changing the operating voltages (which should be the same as for the tube used in a high-gain amplifier) but by changing the loading or the feed-back coupling. This is tricky and another reason why regeneration is not too widely used.

Gain Control

In a receiver front end designed for best signal-to-noise ratio, it is advantageous in the reception of weak c.w. signals to eliminate the gain control from the first r.f. stage and allow it to run "wide open" all of the time. If the first stage is controlled along with the i.f. (and other r.f. stages, if any), the signal-to-noise ratio of the receiver will suffer. As the gain is reduced, the G_m of the first tube is reduced, and its noise figure becomes higher. An elaborate receiver might well have separate gain controls for the first r.f. stage and for all i.f. stages.

Extending the Tuning Range

As mentioned earlier, when a receiver doesn't cover a particular frequency range, either in fact or in satisfactory performance, a simple solution is to use a **converter**. A converter is another "front end" for the receiver, and it is made to tune the proper range or to give the necessary performance. It works into the receiver at some frequency between 1.6 and 10 Mc. and thus forms with the receiver a "triple-detection" superhet.

There are several different types of converters in vogue at the present time. The commonest type, since it is the oldest, uses a regular tunable oscillator, mixer, and r.f. stages as desired, and works into the receiver at a fixed frequency. A second type uses broad-banded r.f. stages in the r.f. and mixer stages of the converter, and only the oscillator is tuned. Since the frequency the converter works into is high (7 Mc. or more), little or no trouble with images is experienced, despite the broad-band r.f. stages. A third type of converter uses broad-banded r.f. and output stages and a fixed-frequency oscillator (self- or crystal-controlled). The tuning is done with the receiver the converter is connected to. This is an excellent system if the receiver itself is well shielded and has no external pick-up of its own. Many war-surplus receivers fall in this category. A fourth type of converter uses a fixed oscillator with ganged mixer and r.f. stages, and requires two-handed tuning, for the r.f. stages and for the receiver. The r.f. tuning is not critical, however, unless there are many stages.

The broad-banded r.f. stages have the advantage that they can be built with short leads, since no tuning capacitors are required and the unit can be tuned initially by trimming the inductances. They are a little more prone to cross-modulation than the gang-tuned r.f. stages, however, because of the lack of selectivity. The fourth type of converter, although the most difficult to build, is probably the most satisfactory, particularly if a crystal-controlled high-frequency oscillator is used. It not only has the advantage of the best selectivity and protection against images and cross-modulation, but the crystal gives it a stability unobtainable with self-controlled oscillators. Amateurs who specialize in operation on 28 and 50 Mc. often develop good converters for use ahead of conventional communications receivers, and the extra trouble often pays off in outstanding performance for the station.

While converters can extend the operating range of an existing receiver, their greatest advantage probably lies in the opportunity they give for getting the best performance on any one band. By selecting the best tubes and techniques for any particular band, the amateur is assured of top receiver performance. With separate converters for each of several bands, changes can be made in any one without disabling or impairing the receiver performance on another band. The use of converters ahead of the low-frequency receiver is rapidly becoming standard practice on the bands above 14 Mc.

Tuning a Receiver

C.W. Reception

For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency. 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. is not generally satisfactory in c.w. reception, except in receivers expressly designed for the purpose, 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 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, except when there are few loud signals on the band and a low noise 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 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 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." 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 often 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 that prevents "blocking" by the stronger signal.

A crystal filter will do much toward reducing interference in 'phone reception. Although the high selectivity cuts sidebands 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 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 rece.ver 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 that necessary with legitimate signals. In poorly-shielded receivers it is often possible to find b.f.o. harmonics below 2 Mc., but they should be very weak at higher frequencies.

Narrow-Band Frequency- and Phase-Modulation Reception

FM Reception

In the reception of NFM signals by a normal communications receiver, the a.v.c. is switched off and the incoming signal is not tuned "on the nose," as indicated by maximum reading of the S-meter, but slightly off to one side or the other. This puts the carrier of the incoming signal on one side or the other of the i.f. selectivity characteristic (see Fig. 5-1). As the frequency of the signal changes back and forth over a small range with modulation, these variations in frequency are translated to variations in amplitude, and the consequent AM is detected in the normal manner. The signal is tuned in (on one side or the other of maximum carrier strength) until the audio quality appears to be best. The audio output from the signal depends on the *slope* of the i.f. characteristic and the amount of *swing* (deviation) of the signal. If the audio is too weak, the transmitting operator should be advised to increase his swing slightly, and if the audio quality is bad ("splashy" and with serious distortion on volume peaks) he should be advised to reduce his swing. Coöperation between transmitting and receiving operators is a necessity for best audio quality. The transmitting station should always be advised immediately if at any time his bandwidth exceeds that of an AM signal, since this is a violation of FCC regulations, except in those portions of the bands where wide-band FM is permitted.

If the receiver has a discriminator or other detector designed expressly for FM reception, the signal is *peaked* on the receiver (as indicated by maximum S-meter reading or minimum background noise). There is also a spot on either side of this tuning condition where audio is recovered through slope detection, but the signal will not be as loud and the background noise will be higher.

PM Reception

Phase-modulated signals can be received in the same way that NFM (narrow-band FM) signals are, except that in this case the audio output will appear to be lacking in "lows," because of the differences in the deviation-*vs.*-audio characteristics of the two systems. This can be remedied to a considerable degree by advancing the tone control of the receiver to the point where more nearly normal speech output is obtained.

NPM signals can also be received on communications receivers by making use of the crystal filter, in which case there is no need for audio compensation. The crystal filter should be set to the sharpest position and the carrier should be tuned in on the crystal peak, *not* set off to one side. The phasing condenser should be set not for exact neutralization but to give a rejection notch at some convenient side frequency such as 1000 cycles off resonance. There is considerable attenuation of the side bands with such tuning, but it can readily be overcome by using additional audio gain. NFM signals received through the crystal filter in this fashion will have a "boomy" characteristic because the lower frequencies are accentuated.

Reception of Single-Sideband Signals

Single-sideband signals are generally transmitted with little or no carrier, and it is necessary to furnish the carrier at the receiver before proper reception can be obtained. Because little or no carrier is transmitted, the a.v.c. in the receiver is not useful, and manual variation of the r.f. gain control is required.

A single-sideband signal can be identified by the absence of a strong carrier and by the severe variation of the S-meter at a syllabic rate. When such a signal is encountered, it should first be peaked with the main tuning dial. (This centers the signal in the i.f. passband.) After this operation, do not touch the main tuning dial. Then set the r.f. gain control at a very low level and switch off the a.v.c. Increase the audio volume control to maximum, and bring up the r.f. gain control until the signal can be heard weakly. Switch on the beat oscillator, and carefully adjust the frequency of the beat oscillator until proper speech is heard. If there is a slight amount of carrier present, it is only necessary to *zero-beat* the oscillator with this weak carrier. It will be noticed that with an incorrect setting of the beat oscillator, the speech will sound high- or low-pitched or even inverted (very garbled), but no trouble will be had in getting the correct setting, once a little experience has been obtained. The use of minimum r.f. gain and maximum audio gain will insure that no distortion (overload) occurs in the receiver.

Another method of receiving single-sideband signals is to reinsert the carrier at the signal frequency. If, for example, you wish to copy a single-sideband signal that is on 3990 kc., you can supply the carrier at that frequency (with a small auxiliary oscillator or frequency meter) and leave your receiver in the normal condition for AM reception (a.v.c. on, b.f.o. off). This method of reception is advantageous in "round-table" contacts that include a single-sideband station, because it calls only for careful tuning of the auxiliary oscillator and not of the receiver. Further, only the auxiliary oscillator must be stable.

Servicing Superhet Receivers

I.F. Alignment

A calibrated signal generator or test oscillator is a very useful device for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver is required. If the receiver has a tuning meter, its indications will serve the purpose. Lacking an S-meter, a high-resistance voltmeter or preferably a vacuum-tube voltmeter can be connected across the second-detector load resistor, if the second detector is a diode. Alternatively, if the signal generator is a modulated type, an a.c. voltmeter can be connected across the primary of the transformer feeding the 'speaker, or from the plate of the last audio amplifier through a 0.1-μfd. blocking condenser to the receiver chassis. Lacking an a.c. voltmeter, the audio output can be judged by ear, although this method is not as accurate as the others. If the tuning meter is used as an indication, the a.v.c. of the receiver should be turned on, but any other indication requires that it be turned off. Lacking a test oscillator, a steady carrier tuned through the input of the receiver (if the job is one of just touching up the i.f. amplifier) will be suitable. However, with no oscillator and tuning an amplifier for the first time, one's only recourse is to try to peak the i.f. transformers on "noise," a difficult task if the transformers are badly off resonance, as they are apt to be. It would be much better to spend a little time and haywire together a simple oscillator for test purposes.

Initial alignment of a new i.f. amplifier is as follows: The test oscillator is set to the correct frequency, and its output is connected to the grid of the last i.f. amplifier tube and to the chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum output, as shown by the indicating device being used. The oscillator output lead is then clipped on to the grid of the next-to-the-last i.f. amplifier tube, and the second-from-the-last transformer trimmer adjustments are peaked 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 test oscillator as more of the i.f. amplifier is brought into use, because the increased gain is likely to cause overloading and consequent inaccurate adjustments. It is desirable in all cases to use the minimum oscillator signal that will give useful output readings. The i.f. transformer in the plate circuit of the mixer is aligned with the signal introduced to the grid of the mixer. Since the tuned circuit feeding the mixer grid may have a very low impedance at the i.f., it may be necessary to boost the test generator output or to disconnect the circuit temporarily from the mixer grid.

If the i.f. amplifier has a crystal filter, the filter should first be switched out and the alignment carried out as above, setting the test oscillator as closely as possible to the crystal frequency. When this is completed, 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 the exact frequency, as indicated by a sharp rise in output. Leaving the test oscillator set on the crystal peak, the i.f. trimmers should be realigned for maximum output. The necessary readjustment should be small. The oscillator 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 is used as the tuning indication. Lacking the a.v.c. tuning meter, the transformers may be conveniently 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 i.f. transformers for maximum audio output.

An amplifier that is only slightly out of alignment, as a result of normal drift or aging, can be realigned by using any steady signal, such as a local broadcast station, instead of the test oscillator. One's 100-kc. standard makes an excellent signal source for "touching up" an i.f. amplifier. Allow the receiver to warm up thoroughly, tune in the signal, and trim the i.f. for maximum output.

If you bought your receiver instead of making it, be sure to read the instruction book carefully before attempting to realign the receiver. Most instruction books include alignment details, and any little special tricks that are peculiar to that particular type of receiver will also be described.

R.F. Alignment

The objective in aligning the r.f. circuits of 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, with harmonics from your 100-kc. standard or other known oscillator, 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 inductance 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, it indicates that 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.

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. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. 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 that appears when the gain is advanced with the c.w. beat oscillator on. It can result from defects in i.f.-amplifier circuits similar to those above. Inadequate cathode by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1 to 0.25 μfd. often 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. This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feed-back, 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, 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 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. The better receivers use VR-type tubes to stabilize the oscillator voltage — a defective tube will cause trouble with oscillator instability.

A One-Tube Regenerative Receiver

The receiver shown in Figs. 5-30, 5-31, 5-32 and 5-33 represents close to the minimum requirements of a useful short-wave receiver. Under suitable conditions, it is capable of receiving signals from many foreign countries. It is an excellent receiver for the beginner, because it is easy to build and the components are not expensive.

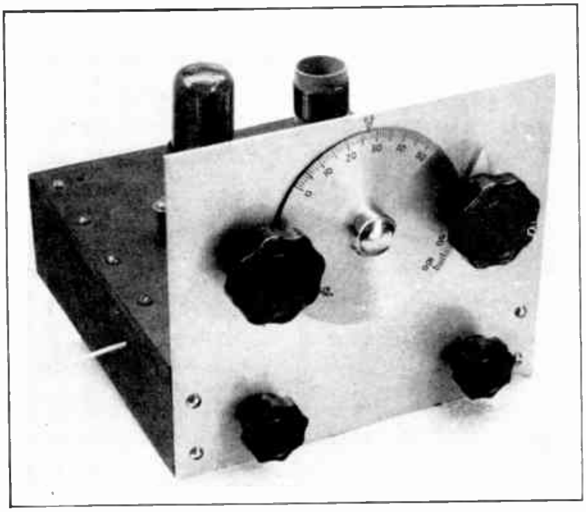


Fig. 5-30 — The simple one-tube regenerative receiver is built on a wood-and-Presdwood chassis, with an aluminum panel. The large left-hand knob drives the calibrated scale on the bandspread condenser. The large right-hand knob is for the band-set condenser.

From the circuit in Fig. 5-32, it can be seen that the only tube in the receiver is a 6SN7 twin triode. One section is used as a regenerative detector, the other triode section serving as an audio amplifier to the headphones. A variable antenna-coupling condenser, C_1, minimizes "dead spots" in the tuning range that might be caused by antenna-resonance effects. Two tuning condensers are used. The band-set condenser, C_4, tunes to the desired frequency band, and the bandspread condenser, C_2 C_3, allows the operator to tune slowly through the band. The bandspread condenser is a dual condenser made from a single midget variable, and on all of the amateur bands except 3.5 Mc. only the C_3 portion is connected in the circuit. The 3.5-Mc. coil includes a jumper that connects C_2 on that band. Regeneration is controlled by varying the plate voltage on the detector with R_4.

The mechanical design is made as simple as possible. Work on the chassis and the front panel can be done with only a No. 8 drill, a ¼-inch drill, and a round file. There is no complicated metal work or bending. To reduce the panel size, the knob on the band-set condenser overlaps the friction-driven tuning dial.

The front panel is a 7 × 7-inch sheet of 1/16-inch aluminum. It carries the tuning controls, the regeneration adjustment and the antenna-coupling condenser shaft. The sides of the chassis are soft wood strips, 7 × 2 × 5/8 inches. The deck of the chassis is a 7 × 7-inch sheet of ¼-inch Presdwood (or Masonite). The 6SN7 socket is supported on 5/8-inch-long mounting pillars, and the 5-

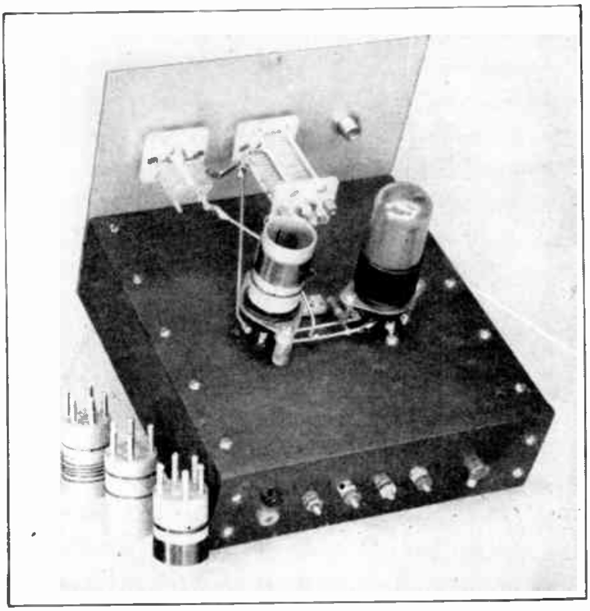


Fig. 5-31 — Another view of the one-tube regenerative receiver shows how the tube and coil sockets are mounted. The headphone tips plug into the two small tip jacks on the rear panel — the set of four machine screws and nuts is for connecting to the power supply.

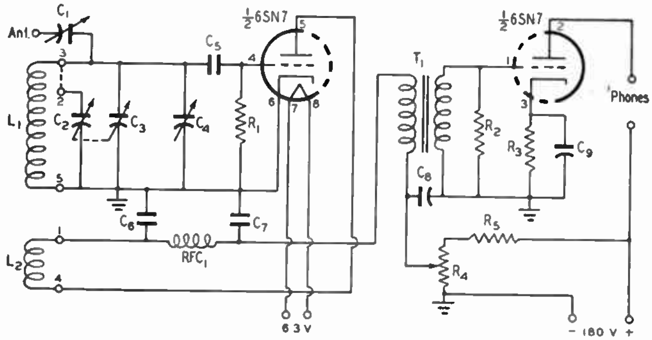


Fig. 5-32 — Wiring diagram of the one-tube regenerative receiver.

- C_1 — Homemade adjustable condenser. See text.
- C_2, C_3 — Reworked midget variable (Millen 21935). See text.
- C_4 — 100-μμfd. midget variable (Millen 20100).
- C_5 — 100-μμfd. mica.
- C_6, C_7 — 470-μμfd. mica.
- C_8 — 12-μfd. 150-volt electrolytic.
- C_9 — 10-μfd. 25-volt electrolytic.
- R_1 — 1.5 megohms, ½ watt.
- R_2 — 0.15 megohm, ½ watt.
- R_3 — 1500 ohms, ½ watt.
- R_4 — 50,000-ohm wire-wound potentiometer.
- R_5 — 33,000 ohms, 1 watt.
- RFC_1 — 2.5-mh. r.f. choke (National 100U).
- T_1 — Interstage audio transformer (Stancor A-4723).

prong coil socket is on ⅞-inch pillars. The grid leak, R_1, and grid condenser, C_5, are located above the deck. The back panel is made of ¼-inch Presdwood and carries the binding posts. The binding posts are ¾-inch 6–32 machine screws with suitable nuts and washers. The chassis is assembled with ¾-inch No. 6 round-head wood screws. Upon completion, the assembly is given a coat of flat black paint. The front panel is secured to the chassis side members with No. 6 round-head wood screws.

The bandspread condenser, C_2/C_3, is made by modifying a Millen 21935 variable condenser. Using a hack-saw blade, the stator bars are carefully cut between the eighth and ninth plates (counting back from the front panel). The ninth plate is removed by twisting it loose with long-nosed pliers.

Coil sizes and data are given in the coil table. All coils are wound on 1-inch diameter 5-pin coil forms. The coil for the 80-meter range is close-wound and requires no treatment, but the spaced-turns coils should be secured by running a thin line of Duco cement across the wire at several points. Before cementing the turns in place, each coil should be tried in the receiver. To obtain smooth regeneration, it may be necessary to make minor coupling adjustments (changes in spacing) between L_1 and L_2.

The antenna condenser, C_1, is made from two 1-inch squares of sheet copper. One plate is secured to the underside of the deck on a tie-point. The other plate is carried by a ¼-inch diameter polystyrene rod. Rotating the shaft swings the moving plate away from the fixed plate and provides a capacity of from 5 to less than 1 μμfd. The polystyrene rod passes through the front panel and out the back panel. It is secured at the back by a ¼-inch shaft collar. The panel end carries a tuning knob, and a rubber grommet under slight compression, placed between the knob and the panel, acts as a friction lock. The moving plate is secured to the polystyrene rod by a copper-wire hairpin soldered to the plate and fixed into a pair of holes drilled in the rod. A flexible

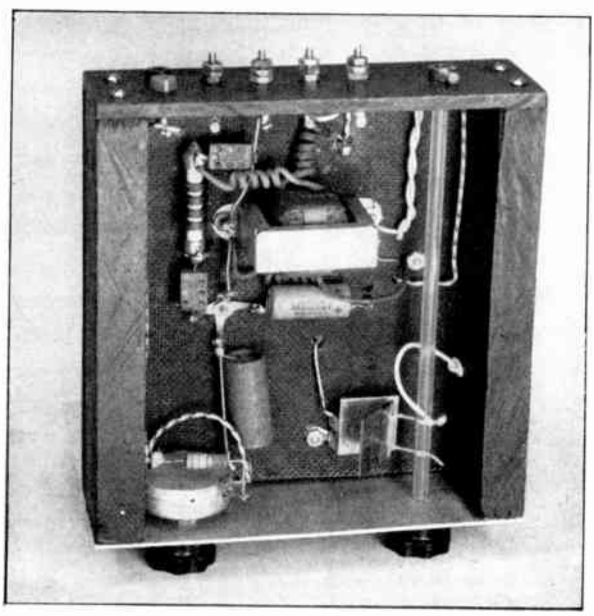


Fig. 5-33 — This view underneath the one-tube regenerative receiver shows the arrangement of parts and the construction of the variable antenna-coupling condenser.

COIL TABLE FOR THE ONE-TUBE REGENERATIVE RECEIVER

All coils wound on Millen 45005 1-inch diameter coil forms. Both L_1 and L_2 should be wound in the same direction, with L_2 closer to the pins of the form. The grid end of L_1 and the plate end of L_2 should be on the outside ends of the coils.

Range	L_1	L_2	*Sep.* L_1-L_2
2.8 — 6 Mc. (80 meters)	25 t. No. 26 enam., close-wound	4 t. No. 26 enam., close-wound	3/8 inch
5.9 — 13.5 Mc. (40 meters)	13½ t. No. 22 enam., spaced to occupy 5/8 inch	1¼ t. No. 26 enam., close-wound	¼ inch
13.6 — 30 Mc. (20 and 14 meters)	5¼ t. No. 22 enam., spaced to occupy 5/8 inch	1¾ t. No. 26 enam., close-wound	3/8 inch
24.5 — 40 Mc. (10 and 11 meters)	1½ t. No. 22 enam., close-wound	1¾ t. No. 26 enam., close-wound	5/16 inch

lead is soldered to the protruding wire, and the lead passes out through a hole in the side of the chassis to make connection to the antenna. Knots in this wire, on either side of the chassis wall, secure the wire firmly in place. The fixed plate is covered with a single layer of cellophane Scotch Tape, to prevent a short-circuit when the condenser is positioned at maximum capacity.

All wiring is No. 14 tinned copper. Direct leads from the condensers to the coil socket add to the strength and rigidity of the receiver. The r.f. choke RFC_1, by-pass condensers, and the audio transformer all are fastened to the underside of the deck.

The power supply for the receiver, shown in Figs. 5-34 and 5-35, is simple to assemble because it is built on a wooden chassis. Two strips of 1½ × ¾-inch wood, 12 inches long, are nailed to two short end pieces. The separation between strips is just enough (1¼ inches) to clear the tube socket and electrolytic condensers, and the leads from the transformer and choke also pass through this opening. Binding posts are made in the same manner as on the receiver, with No. 6 machine screws and suitable nuts and washers.

Although it is satisfactory to mount the power supply on the same table with the receiver, it should be at least one or two feet away, to avoid the possibility of a.c. hum pick-up. For the same reason, the antenna lead should not pass too close to any a.c. wiring from or to the power supply.

Using the parts listed in Fig. 5-35 should result in a power supply that gives about 180 volts when connected to the receiver. However, if the 6SN7 in the receiver appears to run too hot (as tested by touching the tube after the receiver has been running for 5 or 10 minutes), the output voltage can be reduced by increasing the resistance at R_1 (Fig. 5-35). Adding

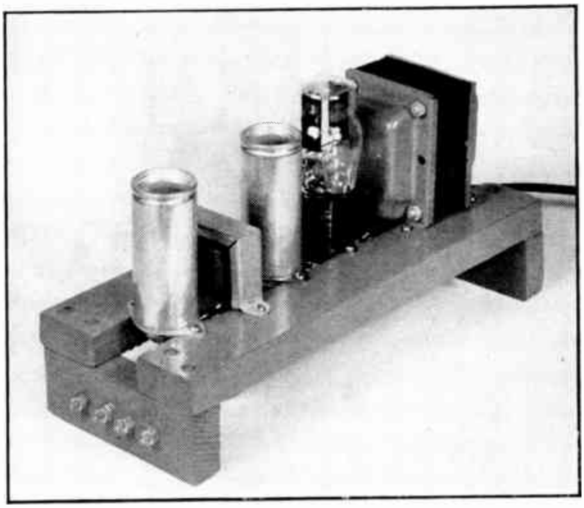


Fig. 5-34 — The power supply for the regenerative receiver is built on a simple wooden chassis.

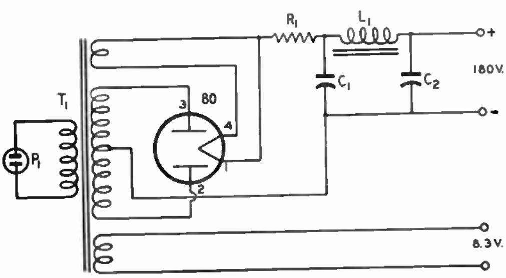


Fig. 5-35 — Circuit diagram of the power supply for the regenerative receiver.
C_1, C_2 — 16-μfd. 450-volt electrolytic (Mallory RS-217).
R_1 — 20,000-ohm 10-watt wire-wound.
L_1 — 15-henry 50-ma. filter choke (Stancor C-1080)
P_1 — 115-volt line plug.
T_1 — 275-0-275 volts at 50 ma., 6.3 v. at 2.5 amp., 5 v. at 2 amp. (Thordarson T22R30).

5000 or 10,000 ohms in series with R_1 should do the trick. Or it may be possible to borrow a voltmeter for measuring the output voltage.

The tuning procedure for a regenerative receiver is given earlier in this chapter. Even a short piece of wire hung inside the operating room will serve as an antenna, but for best results an antenna from 30 to 75 feet long, strung as high as possible, should be used.

In buying headphones for use with this receiver, one should avoid the "low-impedance" headphones offered in many of the surplus outlets. While these headsets are excellent when used in the proper circuits, this simple receiver requires the use of "high-impedance" headphones for maximum signal output. Good, inexpensive headphones of this type can be found in any radio store.

An Amateur-Band Eight-Tube Superheterodyne

An advanced type of amateur receiver incorporating one r.f. amplifier stage, variable i.f. selectivity and audio noise limiting is shown in Figs. 5-36, 5-38 and 5-39. As can be seen from the circuit in Fig. 5-37, a 6SG7 pentode is used for the tuned r.f. stage ahead of the 6K8 converter. An antenna compensator, C_4, controlled from the panel, allows one to trim up the r.f. stage when using different antennas that might modify the tracking. The cathode bias resistor of the r.f. stage is made as low as possible consistent with the tube ratings, to keep the gain and hence the signal-to-noise ratio of the stage high. The oscillator portion of the 6K8 mixer is tuned to the high-frequency side of the signal except on the 28-Mc. band, the usual custom nowadays in communications receivers. The oscillator tuning condenser, C_{17}, is of higher capacity than the r.f. and mixer tuning condensers, in the interest of better oscillator stability.

The i.f. amplifier is tuned to 455 kc., and the first stage is made regenerative by soldering a short length of wire to the plate terminal of the socket and running it near the grid terminal, as indicated by C_{C1} in the diagram. Regeneration is controlled by reducing the gain of the tube, and R_{12}, a variable cathode-bias control, serves this function. The second i.f. stage uses a 6K7, selected because high gain is not necessary at this point.

Manual gain-control voltage is applied to the r.f. and second i.f. stages. It is not applied to the mixer because it might pull the oscillator frequency, and it is not tied in with the first i.f. amplifier because it would interlock with the regeneration control used for controlling the selectivity. However, the a.v.c. voltage is applied to the r.f. and both i.f. stages, with the result that the selectivity of the regenerative stage decreases with loud signals and gives a measure of automatic selectivity control. Using a negative-voltage power supply for the manual gain control is more expensive than the familiar cathode control, but it allows a wide range of control with less dissipation in the components. The a.v.c. is of the delayed type, the a.v.c. diode being biased about 1½ volts by the cathode resistor of the diode-triode detector-audio stage.

The second-detector-and-first-audio is the usual diode-triode combination and uses a 6SQ7. A 1N34 crystal diode is used as a noise limiter, and is left in the circuit all of the time. As is common with this type of circuit, it has little or no effect when the b.f.o. is on, but it is of considerable help to 'phone reception on the bands where automobile ignition is a factor. The constructor can satisfy himself on its operation when first building the receiver and working on it out of the case. By leaving one end of the 1N34 floating and touching it to the proper point in the circuit, a marked drop in ignition noise will be noted.

The b.f.o. is capacity-coupled to the detector by soldering one end of an insulated wire to the a.v.c. diode plate and wrapping several turns of the wire around the b.f.o. grid lead. This capacity is designated C_{C2} in the diagram. The wire was connected to the a.v.c. diode plate lead only for wiring convenience — the a.v.c. coupling condenser, C_{32}, passing the b.f.o. voltage without introducing appreciable attenuation.

Headphone output is obtained from the plate circuit of the 6SQ7 at J_1, and loudspeaker output is available from the 6F6 audio-amplifier stage. High-impedance or crystal headphones are recommended for maximum headphone output.

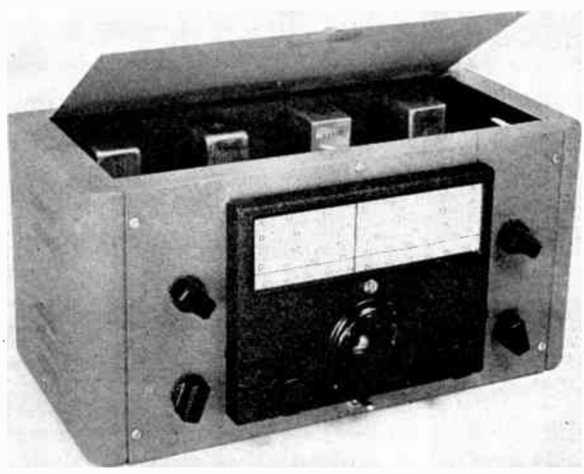


Fig. 5-36 — **An amateur-band eight-tube receiver. The knobs on the left control audio volume (upper) and b.f.o. pitch, and the two on the right handle r.f. and i.f. gain (upper) and i.f. regeneration. The knob to the left of the large tuning knob is fastened to the *MAN.-A.V.C.-B.F.O* switch, and the one on the right is for the antenna trimmer. The toggle switch under the dial throws high negative bias on the r.f. stage during transmission periods.**

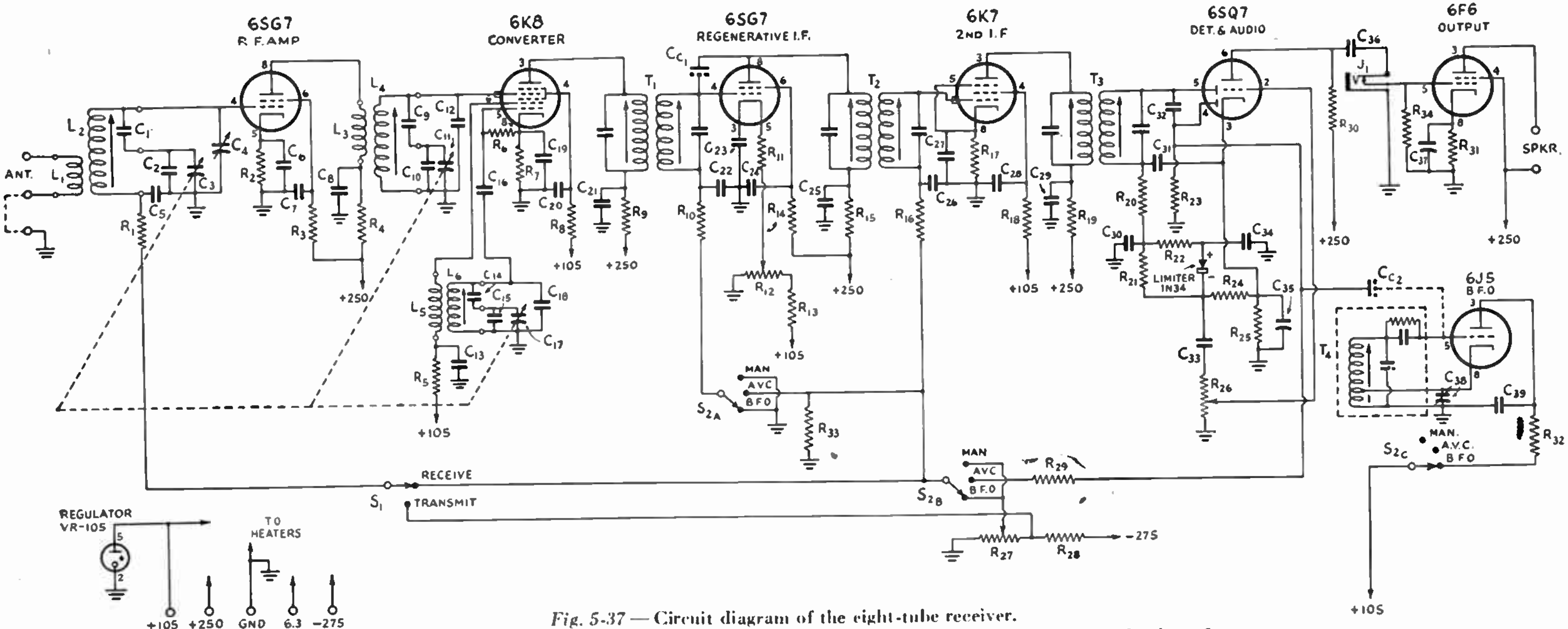


Fig. 5-37 — Circuit diagram of the eight-tube receiver.

C_1, C_9, C_{14} — See coil table.
C_2, C_{10}, C_{12}, C_{18} — 10-μμfd. ceramic.
C_3, C_{11} — 15-μμfd. midget variable (National UM-15).
C_4 — 15-μμfd. midget variable (Hammarlund HF-15).
C_5, C_6, C_7, C_8, C_{13}, C_{19}, C_{20}, C_{21}, C_{22}, C_{23}, C_{24}, C_{25}, C_{26}, C_{27}, C_{28}, C_{29}, C_{39} — 0.01-μfd. mica.
C_{15} — 37-μμfd. ceramic (10 and 27 in parallel).
C_{16}, C_{30}, C_{32} — 100-μμfd. mica.
C_{17} — 35-μμfd. midget variable (National UM-35).
C_{31} — 220-μμfd. mica.
C_{33} — 0.05-μfd. paper, 200 volts.
C_{34} — 0.1-μfd. paper, 200 volts.
C_{35}, C_{37} — 10-μfd. 25-volt electrolytic.
C_{36} — 0.1-μfd. paper, 400 volts.
C_{38} — 35-μμfd. midget variable (Hammarlund HF-35).

C_{C1}, C_{C2} — See text.
R_1, R_{10}, R_{16}, R_{30} — 0.1 megohm.
R_2 — 68 ohms.
R_3, R_{14} — 33,000 ohms.
R_4, R_5, R_8, R_9, R_{15}, R_{18}, R_{19} — 470 ohms.
R_6, R_{13}, R_{20}, R_{21} — 47,000 ohms.
R_7 — 220 ohms.
R_{11} — 180 ohms.
R_{12} — 2000-ohm wire-wound potentiometer.
R_{17} — 330 ohms.
R_{22}, R_{23}, R_{29}, R_{33} — 1.0 megohm.
R_{24}, R_{28} — 0.15 megohm.
R_{25} — 2700 ohms.
R_{26} — 1.0-megohm carbon potentiometer.
R_{27} — 25,000-ohm carbon potentiometer.

R_{31} — 470 ohms, 1 watt.
R_{32} — 27,000 ohms.
R_{34} — 0.22 megohm.
All resistors ½-watt unless otherwise noted.
L_1 through L_6 — See coil table.
J_1 — Closed-circuit jack.
S_1 — S.p.d.t. toggle switch.
$S_{2A\text{-}B\text{-}C}$ — Three-pole 3-position wafer switch (Centralab 2507).
T_1, T_2 — 456-kc. interstage i.f. transformer, permeability-tuned (Millen 64456).
T_3 — 456-kc. diode transformer, permeability-tuned (Millen 64454).
T_4 — 456-kc. b.f.o. assembly, permeability-tuned (Millen 65456).

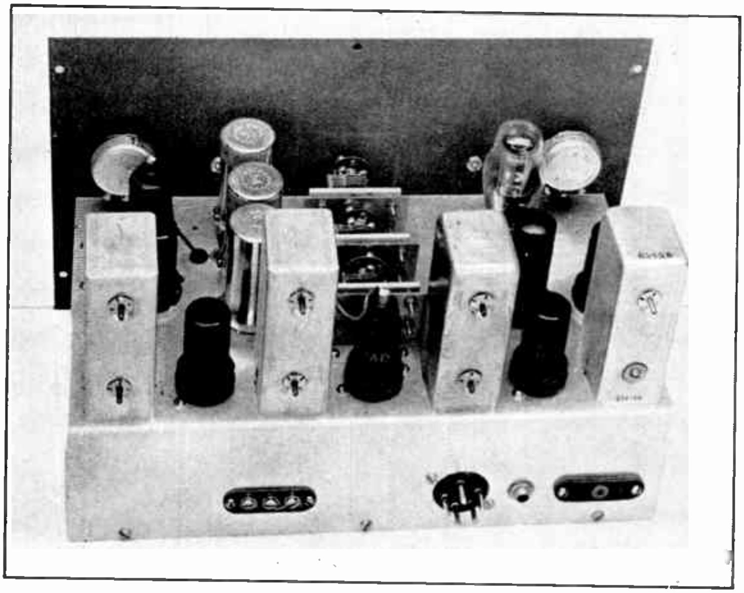


Fig. 5-38 — This view of the eight-tube receiver chassis shows the mounting of the tuning condensers and the placement of most of the large components. The three shielded plug-in coil assemblies can be seen to the left of the tuning gang. The 6K8 converter is the tube on the left nearest the panel.

The antenna terminal strip, power-supply plug, headphone jack and 'speaker terminals are mounted on the rear (foreground in this view) of the chassis.

Construction

The receiver is built on an aluminum chassis mounted in a Par-Metal CA-202 cabinet, and a Millen 10035 dial is used for tuning. The chassis is made of 1/16-inch-thick stock, bent into a "U"-channel, and measures 13 inches wide and 7 1/4 inches deep on the top. It is 3 3/8 inches deep at the rear and 1/8 inch less at the front. The rear edge is reinforced with a piece of 3/8-inch square dural rod that is tapped for screws through the bottom of the cabinet, further to add to the strength of the structure when finally assembled. The various components that are common to the front lip of the chassis and the panel are used to tie the two together.

The shield panel used to mount the antenna-compensator condenser is also made of 1/16-inch aluminum with a 5/8-inch lip on the side for mounting. Part of the lip must be cut away to clear wires and mounting plates on some sockets, so it is advisable to put in the panel after most of the assembly and wiring have been completed. Flexible couplings and bakelite rod couple the condenser to the panel bushing.

The three tuning condensers are mounted on individual brackets of 1/16-inch aluminum. The brackets measure 2 1/2 inches wide and 1 9/16 high, with 1/2-inch lips. A cover of thin aluminum — not shown in the photographs — slides over the condenser assembly to dress up the top view a bit. The dust cover is not necessary for satisfactory operation of the receiver.

Ceramic sockets are used for the plug-in coils and for the r.f. amplifier, converter and b.f.o. tubes. Mica condensers were used throughout the receiver for by-passing wherever feasible, because they lend themselves well to compact construction. Paper condensers could be used in the i.f. amplifier but they would crowd things a bit more.

In wiring the receiver, small tie-points were used wherever necessary to support the odd ends of resistors and condensers, and rubber grommets were used wherever wires run through the chassis, with the exception of the tuning-condenser leads. The latter leads, being of No. 14 wire, are self-supporting through the 5/16-inch clearance holes and do not require grommets. The same heavy wire was used for the grid and plate leads of the r.f. stage and the plate lead of the oscillator, to reduce the inductance in these leads. The tuning condensers are grounded back at the coil sockets and not above the chassis as might be the tendency. Screen, cathode and plate by-pass condensers are grounded at a single point for any tube wherever possible, although C_2 is grounded at the r.f.-coil socket, C_8 is grounded at the converter-coil socket, and C_{13} is returned at the oscillator-coil socket. The plate and B+ leads from T_1 are brought back to the converter socket through shield braid, and C_{21} is returned to ground at the converter socket.

The b.f.o. pitch condenser, C_{38}, is insulated from the chassis and panel by fiber washers, and the rotor is connected back to the tube socket by braid that shields the stator lead. This is done to reduce radiation from the b.f.o. which might get in at the front end of the i.f. amplifier.

The coils are wound on Millen 74001 permeability-tuned coil forms, according to the coil table. Series condensers are mounted inside the forms on all bands except the 80-meter range, where no condenser is required and the tuning condenser is jumped directly to the grid end of the coils. In building the coils, the washers are first drilled for the leads and then cemented to the form with Duco or other cement. The bottom washer is cemented close to the terminal pins, leaving just enough room to get the soldering iron in to fasten the coil ends and to leave room for the series condenser. The large coils, L_2, L_4 and L_6, were wound first in every case, and then a layer of polystyrene Scotch Tape wrapped over the coil, after which the smaller winding was put on and the ends of the windings soldered in place. Since for maximum range of adjustment it is desirable to allow the powdered-iron slug to be fully withdrawn from the coil, keeping the coils at the base end of the form allows the iron slug to travel out at the other end, under which condition the adjusting screw on the slug projects the least. To secure the wires after winding, drops of cement should be placed on them where they feed through the polystyrene washers.

Alignment

If a signal generator is available, it can be used to align the i.f. amplifier on 455 kc. in the usual manner. If one is not available, the coupling at C_{C1} can be increased to the point where the i.f. stage oscillates readily and the b.f.o. transformer is then tuned until a beat note is heard. The other transformers can then be aligned until the signal is loudest, after which C_{C1} should be decreased until the i.f. oscillates with the regeneration control, R_{12}, about 5 degrees from maximum. The trimmers on T_1 then should be tuned to require maximum advancing of the regeneration control for oscillation, with a set value of C_{C1}. When properly tuned, the oscillation frequency of the i.f. stage and the frequency for maximum gain in the regenerative condition will be the same.

With a set of coils in the front end, set the tuning dial near the high-frequency end and tune in a strong signal or marker with the adjustment screw on the oscillator coil. The converter and r.f. coils can then be peaked, with the antenna compensator set at about half capacitance. Then tune to the other end of the band and see if you have enough bandspread. If the bandspread is inadequate, it means that C_{14} is too large, and it should be reduced by using a smaller size of condenser or a combination that gives slightly less capacitance. The tracking of the converter and r.f. coils can be checked by repeaking the position of the slugs in the coils at the low-frequency end. If the converter- or r.f.-coil tuning slugs have to be advanced farther into the coil (to increase the inductance) it indicates that C_9 or C_1 should be larger. Tracking by the method described is at best a compromise, although to all intents and purposes the loss from some slight misalignment is completely unimportant. Another method would be to tap the tuning condensers on the coil in the familiar bandspreading manner, but this requires considerable time and patience. However, with the series condensers as used in this receiver, the tuning curve is more crowded at the high-frequency end of a range than at the low, and this would be reduced somewhat by the tapped-coil bandspread.

COIL DATA FOR THE EIGHT-TUBE SUPERHETERODYNE

Coil	*3.5 Mc.*	*7 Mc.*	*14 Mc.*	*28 Mc.*
L_1	15 t.	9 t.	6 t.	4 t.
L_2, L_4	76 t.	33 t.	19 t.	8 t.
C_1, C_9	short	27 μμfd.	15 μμfd.	20 μμfd.
L_3	25 t.	11 t.	7 t.	4 t.
L_5	10 t.	8 t.	4 t.	2 t.
L_6	47 t.	32 t.	14 t.	6 t.
C_{14}	short	42 μμfd.	27 μμfd.	51 μμfd.

All coils wound on Millen 74001 forms, close-wound. 3.5-Mc. coils wound with No. 30 enam.; 7-Mc. coils wound with No. 30 d.s.c.; 14- and 28-Mc. coils wound with No. 30 d.s.c. on primaries and ticklers and No. 24 enam. on secondaries. C_{14} for 7-Mc. range made by connecting 27- and 15-μμfd. condensers in parallel. C_1, C_9 and C_{14}, Erie Ceramicons, mounted in coil form.

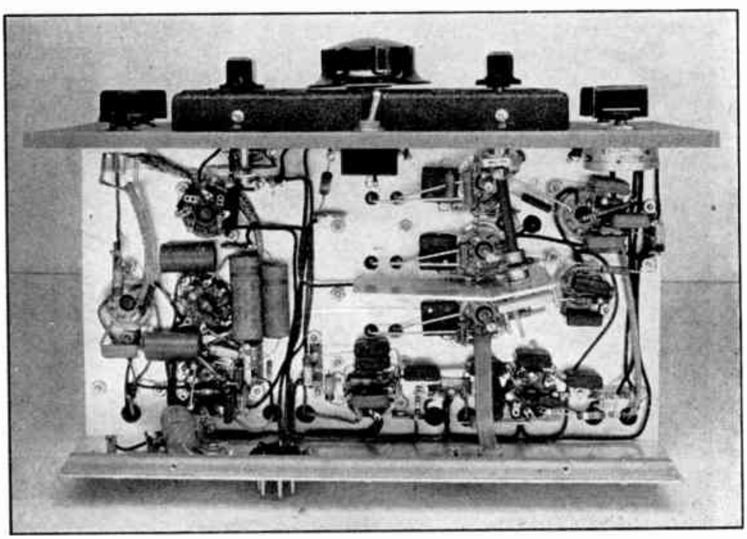


Fig. 5-39 — The mica by-pass condensers used throughout the r.f. and i.f. stages are grouped around the sockets of their respective tubes. Tie-points are used wherever necessary to support small resistors and condensers. The antenna trimmer condenser is mounted on a bracket which also serves as shielding between the mixer- and r.f.-coil sockets, and it is offset to allow access to the trimmer screws on the coil forms. The plate and B+ leads from the first i.f. transformer, T_1, are run in shielded braid, as are the leads from the b.f.o. pitch-control condenser and the volume control.

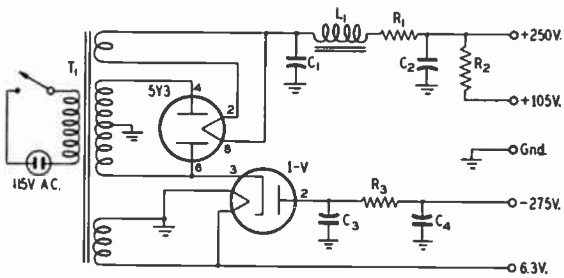


Fig. 5-40 — Wiring diagram of power supply for the eight-tube receiver.

C_1, C_2 — 16-μfd. 450-volt electrolytic.
C_3, C_4 — 8-μfd. 450-volt electrolytic.
R_1 — 500 ohms, 10 watts, wire-wound.
R_2 — 5000 ohms, 10 watts, wire-wound.
R_3 — 0.1 megohm, 1 watt, composition.
L_1 — 30-henry 110-ma. filter choke (Stancor C-1001).
T_1 — 350-0-350 volts, 90 ma.; 5 volts at 3 amp., 6.3 volts at 3.5 amp.

♦

The adjustment of L_5 can be made, if deemed necessary, by lifting the cathode end of R_6 and inserting a 0–1 milliammeter. If the tickler coil has the right number of turns, the current will be from 0.15 to 0.2 ma., and it won't change appreciably over the band. Although such a grid-current check is a fine point and not really necessary, it is a simple way to determine that the oscillator portion is working, since the cold ends of L_5 and L_6 are at the same end of the form — the plug end — and this necessitates winding the two coils in opposite directions.

Some trouble may be experienced with oscillation in the r.f. stage at 28 Mc. However, a grounding strap of spring brass, mounted under one of the screws holding the mixer-coil socket to ground the shield when the coil is plugged in, will normally clear up the trouble. Inadequate coupling to the antenna will also let the r.f. stage oscillate under some tuning conditions, and close coupling is highly recommended for stability in this stage and also for best signal response. A 10-ohm resistor from L_2 to the grid of the 6SG7 will also do the trick.

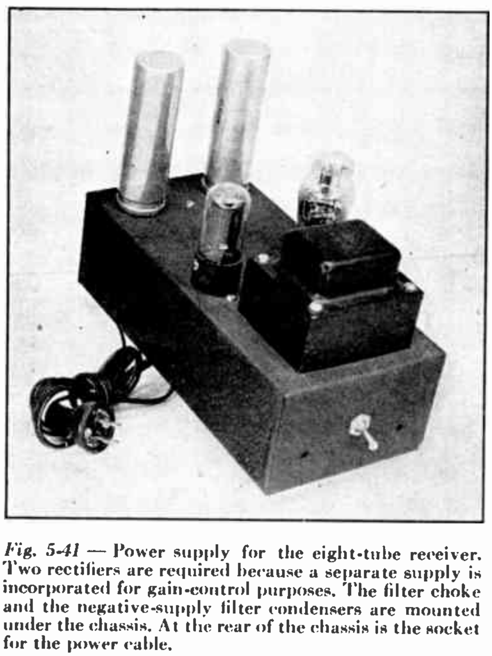


Fig. 5-41 — Power supply for the eight-tube receiver. Two rectifiers are required because a separate supply is incorporated for gain-control purposes. The filter choke and the negative-supply filter condensers are mounted under the chassis. At the rear of the chassis is the socket for the power cable.

It will be found that the over-all gain of the receiver is quite high on the lower-frequency bands, requiring that the r.f. gain be cut down to prevent overloading on strong signals. For c.w. reception, the regeneration control is advanced to the point just below oscillation and the b.f.o. is detuned slightly to give the familiar single-signal effect. For 'phone reception, S_2 is switched to "A.V.C." and volume-control adjustments made with the audio control, R_{26}. If desired, the regeneration control can be advanced until the i.f. is oscillating weakly, and then a heterodyne will be obtained on weak carriers, making them easy to spot. Strong carriers will pull the i.f. out of oscillation because the developed a.v.c. voltage reduces the gain, and hence a simple form of automatic selectivity control is obtained. If it is considered desirable to reduce the i.f. gain when switched to the "A.V.C." position, the regeneration control can be used for this purpose. The "MAN." position permits manual gain-control operation with the b.f.o. off.

The switch S_1 is used for receive-transmit and throws about 40 volts negative on the grid of the first r.f. stage, saving the first tube a little if the transmitter is pouring some power into the receiver.

Power Supply

A power supply suitable for the eight-tube receiver is shown in Figs. 5-40 and 5-41. An idea of the parts arrangement can be obtained from Fig. 5-41, although there is nothing critical about this portion of the receiver. If one wants a neat-looking station with no loose power supplies in sight, the power supply can be built into one corner of the loudspeaker cabinet.

The filtering of the power supply is quite adequate and no trace of hum should be found in the completed receiver when used with this power supply. If any a.c. hum is noticed, it is being introduced in the audio section if it is still present with the r.f. gain control set at minimum. Probable sources of hum in the audio system are leads to C_{33}, R_{26}, C_{36} or J_1 running too close to a "hot" (ungrounded) heater lead, and the correction is to remove these leads from the field of the heater wiring. If signals are modulated with a.c. hum, particularly at the higher frequencies, it is possible that the grid circuit of the 6K8 converter is picking up hum from a nearby heater lead.

A Simple Audio Noise Limiter

The limiter shown in Fig. 5-42 is plugged into the receiver headphone jack and the headphones are plugged into the limiter, with no work required on the receiver. The limiter will cut down serious noise on 'phone signals, and it will keep the strength of c.w. signals at a constant level. It will do much to relieve the operating fatigue caused by long hours of listening to static crashes, key clicks encountered on the air and with break-in operation, and the like.

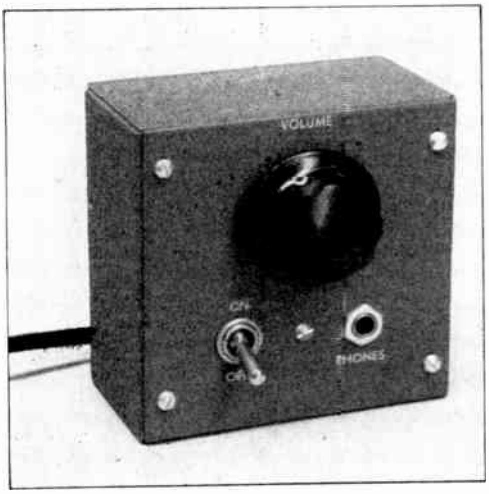


Fig. 5-42 — A simple audio noise limiter for reducing operator fatigue caused by ignition noises, key clicks and static crashes.

The wiring diagram, Fig. 5-43, shows how two 1N34 crystal diodes are individually biased by 1½-volt flashlight cells. The crystals short circuit any audio signal that has an amplitude of more than 3 volts peak-to-peak. A 10,000-ohm potentiometer, R_2, allows the operator to control the output from the limiter to his headphones and is useful in establishing the optimum relationship between the receiver volume-control setting and the headphone signal strength. A 6AL5 twin diode can be substituted for the two crystals, but a heater supply will be required, and it is generally more convenient to build the limiter as shown. No current is drawn from the two bias cells, and their useful life will be their shelf life.

The limiter can be built in a 4 × 4 × 2-inch cabinet, as shown in Fig. 5-44. The front panel carries the "on-off" switch, the headphone jack and the potentiometer. The 1N34 crystals are mounted on their own leads. Care must be taken while soldering to hold the leads of the crystal diodes with long-nose pliers placed between the point being soldered and the body of the crystal. The pliers conduct away the heat that might otherwise damage the crystal.

The back panel carries the batteries. A wooden stirrup has contacts of folded copper braid that make contact to one end of the batteries, and a strip of Presdwood with similar contacts is used at the opposite end. The batteries are secured to the panel and the two strips under tension with rubber bands tied to hooks made from soldering lugs.

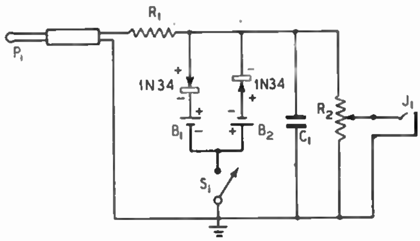


Fig. 5-43 — Wiring diagram of the audio noise limiter.
C_1 — 0.0022-μfd. mica.
R_1 — 15,000 ohms, 1 watt.
R_2 — 10,000-ohm potentiometer, wire-wound.
B_1, B_2 — 1½-volt flashlight cell.
J_1 — Open-circuit Jack.
P_1 — Headphone plug.
S_1 — S.p.s.t. toggle switch.

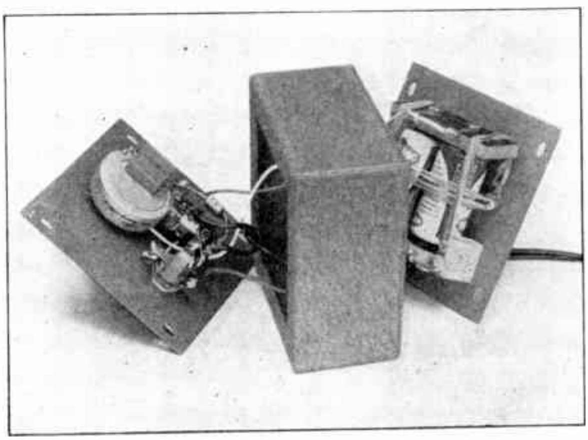


Fig. 5-44 — The audio noise limiter is built on the two removable panels of a small cabinet. The dry cells are held in place by rubber bands.

A Signal-Strength Indicator (S-Meter)

If your receiver has no built-in S-meter and you would like one for comparing signal strengths (and for help in aligning your receiver), the unit shown in Figs. 5-45 and 5-46 can be used. The wiring diagram, Fig. 5-47, is an adaptation of Fig. 5-25C, and uses a 0–1 milliammeter as the indicator. A variable shunt, R_1, allows the meter sensitivity to be regulated to suit the particular receiver, and R_4 is for setting the meter to zero with no signal. The meter can be connected in the plate circuit of any amplifier controlled by the a.v.c. If possible and desirable, the meter and circuit can be built into the receiver.

It is customary to calibrate in terms of S-units up to about midscale, and then in "decibels above S9" over the upper half of the scale. Although there are no standards, current practice is to use about 6-db. steps in the S-scale, and a 100-microvolt signal for "S9."

Such a calibration requires an accurate r.f. signal generator, and relatively few amateurs have access to laboratory equipment of this type. Also, the scale will be accurate only on the radio frequency at which the calibration is made. On different bands — or even in different parts of the same band — the r.f. gain of the receiver will change and the calibration will not hold.

An S-meter is principally useful for making comparisons between signals on or near the same frequency. For this purpose it is entirely satisfactory to choose arbitrarily a signal that seems to you to be about the right strength to represent "S9," adjust the meter sensitivity to give a suitable reading on that signal, and then divide off the scale into equal intervals from zero to 9.

Alternatively, points can be taken by comparing with another receiver that does have a calibrated S-meter. The two receivers may be connected to the same antenna so that simultaneous measurements can be made on incoming signals, provided their antenna input impedances are not widely different. Local signals should be used to avoid fading effects.

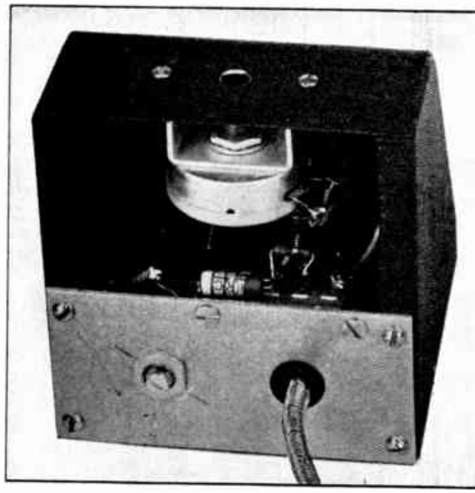


Fig. 5-46 — This rear view of the S-meter shows the meter shunt, R_1, and a tie-point strip mounted on a metal strip attached to the rear side of the meter cabinet. Resistors R_2, R_3 and R_5 are mounted on the tie-point strip. A three-wire cable, running out of the case through a rubber grommet, connects the meter to the receiver.

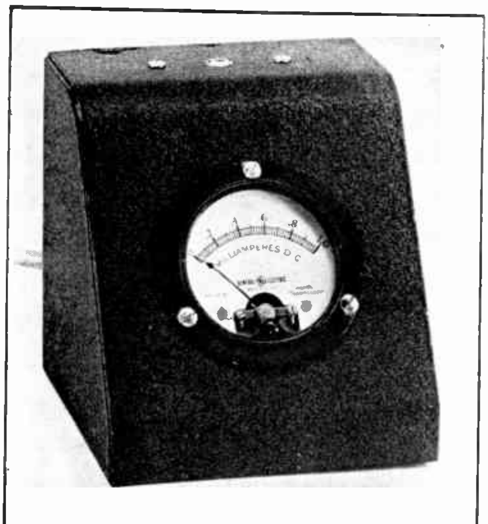


Fig. 5-45 — Front view of the signal-strength indicator. The 0–1 milliammeter is mounted in a metal meter case. The zero-adjustment potentiometer, R_4, is mounted below the top of the cabinet by means of a "U"-shaped bracket; the potentiometer shaft is slotted so that it can be adjusted with a screwdriver. A new face, calibrated in S-units, can be pasted to the 0–1 ma. scale, or a calibration chart can be attached to the cabinet.

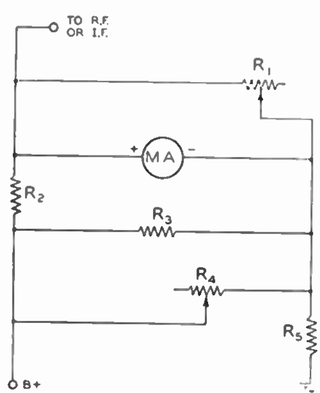


Fig. 5-47 — Wiring diagram of the signal-strength indicator.

R_1 — 100-ohm wire-wound potentiometer.
R_2 — 220 ohms, ½ watt.
R_3 — 680 ohms, ½ watt.
R_4 — 1000-ohm wire-wound potentiometer.
R_5 — 47,000 ohms, 1 watt.
MA — 0–1 ma. d.c. meter.

A Peaked Audio Amplifier

The peaked audio amplifier shown in Figs. 5-48 and 5-50 uses only resistors and condensers to obtain a high degree of selectivity. The circuit, Fig. 5-49, consists of an ordinary audio amplifier and a simple twin-"T" resistance-capacitance bridge. The bridge has a null at the desired audio frequency, and the bridge is connected in a negative-feed-back loop in the amplifier. As a result, the amplifier is highly degenerative at all frequencies except that at which the bridge circuit shows a null. By controlling the amount of negative feed-back, varying degrees of selectivity can be obtained.

The unit, minus its power supply, is housed in a 3 × 4 × 5-inch standard steel box. To simplify construction, most of the components are mounted on a piece of 4 × 5 × 1/16-inch aluminum that replaces one of the removable panels of the box.

When completed and connected to a source of plate and heater power — the plate demand is about 20 ma. at 250 volts — plug P_1 into the receiver output jack and the headphones into J_1. Set the selectivity control, R_8, at maximum, i.e., with the arm farthest away from the grounded end. Tune in a stable c.w. signal and adjust C_8 until the amplifier "rings" or indicates a tendency toward oscillation. Back off on R_8 until you can tune through a peak on C_8 with oscillation, and the audio amplifier is adjusted. In operation, the control for R_8 can be advanced or backed off to give the desired amount of selectivity.

If the amplifier is used with a single-signal superheterodyne in which the crystal filter already contributes considerable selectivity, it is essential that the b.f.o. be adjusted to give peak audio response from the receiver at the frequency for which the audio amplifier shows maximum gain.

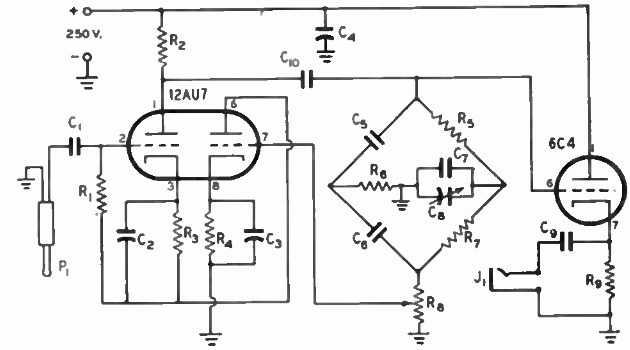


Fig. 5-49 — Wiring diagram of the peaked audio amplifier.

C_1, C_{10} — 0.01-μfd. paper.
C_2, C_3 — 25-μfd. 25-volt electrolytic.
C_4 — 8-μfd. 450-volt electrolytic.
C_5, C_6 — 680-μμfd. mica.
C_7 — 0.001-μfd. mica.
C_8 — 280–1050-μμfd. mica compression trimmer (El-Menco 306).
C_9 — 0.1-μfd. 200-volt paper.
R_1 — 1.0 megohm.
R_2 — 56,000 ohms, 1 watt.
R_3, R_4 — 1200 ohms.
R_5, R_7 — 0.22 megohm.
R_6 — 0.1 megohm.
R_8 — 2.0-megohm volume control.
R_9 — 10,000 ohms, 1 watt.
Resistors are ½-watt composition unless specified otherwise.
J_1 — Open-circuit jack.

Fig. 5-48 — A peaked audio amplifier for increased c.w. selectivity. It is connected to the receiver at the headphone jack, and the headphooes plug into the unit. The knob controls the degree of selectivity.

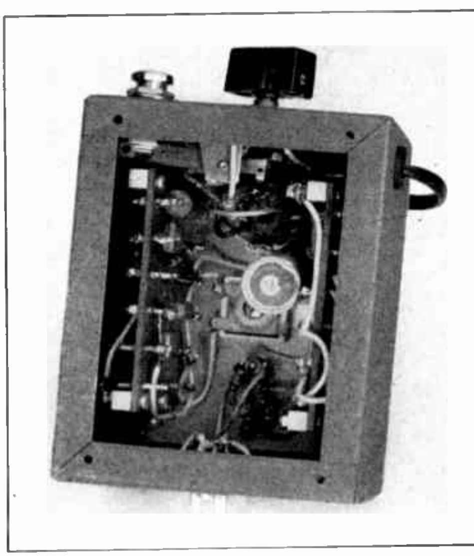


Fig. 5-50 — Construction of the peaked audio amplifier is facilitated by mounting the parts on an aluminum panel that replaces the normal panel of the cabinet. Two resistor boards, supported by square posts mounted on the panel, are used to support most of the small components. A Jones P-304-AB base-mounting plug on the cabinet is used for connecting to the power supply.

A Selective I.F. Amplifier

Most commercial communications receivers do not have sufficient selectivity for amateur use, and their performance can be greatly improved by adding additional selectivity. One popular method is to couple a BC-453 aircraft receiver (war surplus, tuning range 190 to 550 kc.) to the tail end of the 465-kc. i.f. amplifier in the communications receiver and use the resultant output of the BC-453. The aircraft receiver uses an 85-kc. i.f. amplifier that is quite sharp — 6.5 kc. wide at −60 db. — and it helps tremendously in separating 'phone signals and in backing up crystal filters for improved c.w. reception. (See *QST*, January, 1948, page 40.) All that is required to connect a BC-453 into an ordinary communications receiver is a small power supply for the unit (24-volt heater supply and 250-volt d.c. plate supply) and a small control panel. A wire wrapped around a tube pin in the receiver i.f. amplifier is then run to the antenna post of the BC-453, and the auxiliary unit is then tuned to the i.f. of the communications receiver. This addition to a receiver, and any other that increases the i.f. selectivity, has acquired the nickname of "Q5-er," because it adds to the readability report of any signal.

If a BC-453 is not available, it is still a simple matter to enjoy the benefits of improved selectivity. Some of the standard 175-kc. i.f. transformers are fairly sharp, and by cascading enough of them a good selectivity characteristic can be obtained. It is only necessary to heterodyne to 175 kc. the 465-kc. signal existing in the receiver i.f. amplifier and then rectify it after passing it through the 175-kc. amplifier. Little additional gain is required, since the signal level is already quite high in the receiver.

The circuit in Fig. 5-52 shows how such an "outrigger" i.f. amplifier can be connected. The actual layout of parts is not very important, although it is advisable to arrange the parts along a straight line, so that the input to the amplifier will be removed from the output. One example of construction is shown in Fig. 5-51, where the amplifier is shown mounted alongside the chassis of a BC-312 receiver. Power for the amplifier is "stolen" from the receiver (if the power supply is not already overloaded) and the audio output of the "Q5-er" can be brought out directly or fed back through the audio system of the receiver. It is advisable to shunt C_3 with a small variable condenser that is adjustable from the front panel of the unit. This control permits a slight variation in the relative alignment of the 465- and 175-kc. channels, and is useful in getting the maximum benefit from an amplifier of this type. If the unit is to be used for c.w. reception as well as for 'phone (and such units have been found to be very useful in backing up the selectivity of a crystal filter), a beat oscillator should be added to the circuit and its tuning control brought out to the front panel. Although the b.f.o. in the regular receiver can be used, it is generally advisable to introduce the oscillator just before converting to the audio signal.

Generally, no gain control is necessary in the "Q5-er" unit, because the gain is made just enough to give a signal at the 6SR7 of the same order as that obtained at the second detector of the normal receiver. The gain can be reduced, if necessary, by increasing the value of R_6 or by decreasing C_7 and C_{11}.

When the amplifier is completed, the transformers T_1, T_2, T_3 and T_4 should be aligned accurately at their nominal frequency. The signal from the receiver's i.f. amplifier is obtained by connecting the short length of shielded wire from C_1 to the "hot" side of the last i.f. transformer. This generally runs to a diode plate in the receiver, and it is conven-

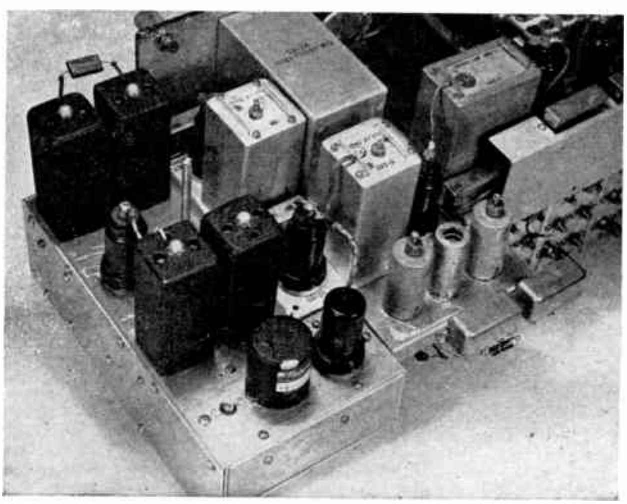


Fig. 5-51 — An outrigger selective i.f. amplifier ("Q5-er") can be located at any convenient point in the receiver. This shows one mounted alongside the chassis of a BC-312. In this version, the second detector and audio amplifier of the BC-312 are connected in the circuit, instead of using a separate tube as shown in Fig. 5-52. (Rand, "The Q5-er," *QST*, Dec., 1947).

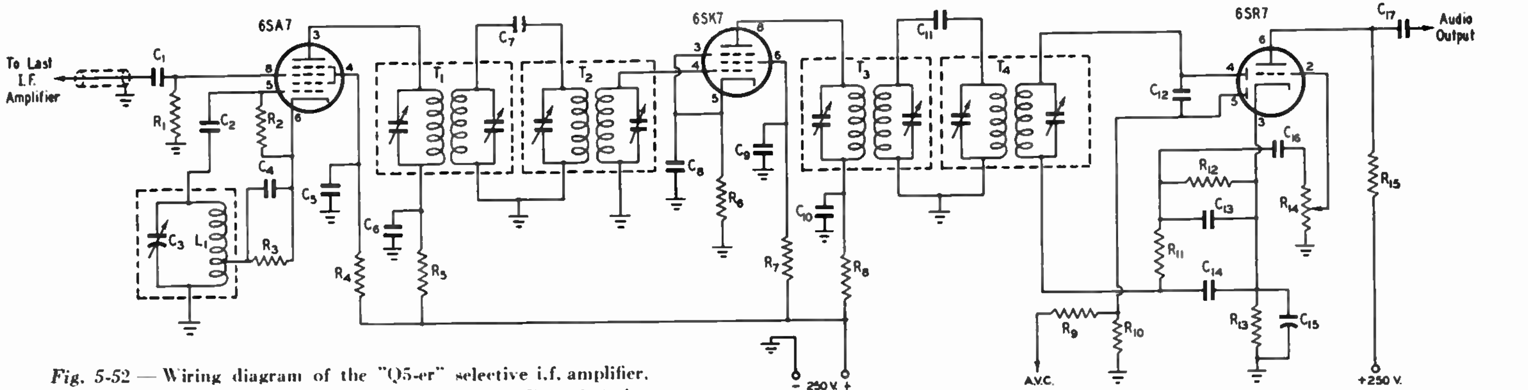


Fig. 5-52 — Wiring diagram of the "Q5-er" selective i.f. amplifier.

C_1 — 1 to 10 μμfd. Use smallest workable value.
C_2, C_{12} — 100 μμfd.
C_3 — See L_1.
C_4, C_5, C_6, C_8, C_9, C_{10}, C_{16}, C_{17} — 0.05 μfd. paper.
C_7, C_{11} — 10 μμfd.
C_{13}, C_{14} — 220 μμfd.
C_{15} — 10-μfd. 25-volt electrolytic.
R_1, R_{10} — 1.0 megohm.
R_2 — 22,000 ohms.
R_3 — 160 ohms.
R_4 — 18,000 ohms.
R_5, R_8 — 470 ohms.
R_6 — 680 ohms.
R_7 — 0.12 megohm.
R_9 — 2.0 megohms.
R_{11} — 47,000 ohms.
R_{12} — 0.27 megohm.
R_{13} — 1800 ohms.
R_{14} — 1.0-megohm volume control.
R_{15} — 47,000 ohms.
All resistors are ½-watt.
L_1 — Tunes with C_3 to receiver i.f. plus 175 kc. For 465-kc. i.f., tunes to 640 kc.
T_1, T_4 — 175-kc. i.f. transformer (Meissner 16-5702).
T_2 — 175-kc. i.f. transformer (Meissner 16-5728).
T_3 — 175-kc. i.f. transformer (Meissner 16-5730).

ient to slip the wire around the tube pin for this connection. Tune in a signal on the receiver, and then adjust C_3 of the "Q5-er" until the signal is a maximum in the output of the 6SR7. That is all of the adjustment necessary, with the exception of reducing the coupling to the receiver (by making C_1 small) to the minimum usable value. This is important because it reduces the chances for overload in the 175-kc. amplifier. The lead from R_9, marked "A.V.C.," should run to the regular a.v.c. line in the receiver. Although it is possible to use the regular a.v.c. circuit in the receiver without modification, it is highly advisable to take the a.v.c. voltage from the more selective "Q5-er," so that strong signals close in frequency to the desired signal do not control the gain of the receiver. If insufficient a.v.c. voltage is developed at R_9, it will manifest itself by overload (distortion) in the regular receiver, and it may be necessary to increase the gain through the "Q5-er" or to use an "amplified-a.v.c." system.

Special transformers are available for the "Q5-er" application. The transformers used in the BC-453 can sometimes be obtained through outlets for surplus goods, and these can be used to build a selective amplifier operating at 85 kc. The Hammarlund Company and the J. W. Miller Company both offer transformers tuning to 50 kc., for the same application.

Still more elaborate systems for utilizing high i.f. selectivity will be found in past issues of *QST*. These include:

McLaughlin, "Exit Heterodyne QRM," *QST*, Oct., 1947.

McLaughlin, "Selectable Single-Sideband Reception Simplified," *QST*, April, 1948.

A Bandswitching Preselector for 14 to 30 Mc.

The performance of many receivers begins to drop off at 14 and 30 Mc. The signal-to-noise ratio is reduced, and trouble with r.f.-image signals becomes apparent. The preselector shown in Figs. 5-53 and 5-55 can be added ahead of any receiver without making any changes within the receiver, and a self-contained power supply eliminates the problem of furnishing heater and plate power.

As can be seen from the wiring diagram, Fig. 5-54, a 6AK5 r.f. pentode is used in the preselector. Both the grid and plate circuits are tuned, but the tuning condensers are ganged and only one control is required. The gain through the amplifier is controlled by changing the cathode voltage, through R_3. A selenium rectifier is used to supply plate power, and the heater power comes from a step-down transformer. The chassis is at r.f. ground but the d.c. circuit is isolated, to prevent short-circuiting the a.c. line through external connections to the preselector.

A two-section ceramic switch selects either the 14- to 21-Mc. or the 28-Mc. coil, or the antenna can be fed through directly to the receiver input. When operating in an amateur band between 14 and 30 Mc., switching to the band not in use will attenuate one's own signal sufficiently to permit direct monitoring, in most cases.

As shown in Fig. 5-53, the ganged condensers are controlled from the front panel by a National MCN dial, and a small knob to the right of this dial is connected to the antenna trimmer, C_4, for peaking the tuning with various antennas. The a.c. line is controlled by S_2, a toggle switch mounted on the panel.

The preselector is built on a 3 × 5 × 10-inch chassis, and a 6 × 6-inch plate of thin metal is used for a panel. A 1¾ × 3-inch aluminum bracket mounted about 3½ inches behind the front panel supports the tuning condenser, C_5, and the antenna trimmer, C_4. Millen 39005 flexible couplings are required to handle the offset shaft of C_4. Both C_5 and C_8 are mounted on the chassis with 6-32 screws, but the chassis should be scraped free of paint before installation, to insure good contact.

The shield partition between the two switch sections (Fig. 5-55) straddles the tube socket and shields the grid from the plate circuit. The switched ends of all coils are supported by their respective switch points, and the other ends are soldered to tie points mounted on the chassis. The mica trimmers, C_9 and C_{10}, are supported on short lengths of stiff wire, and a hole in the side of the chassis is required to reach C_{10} with an aligning tool.

The power-supply components are mounted as near the rear of the chassis as possible. The selenium rectifier must be insulated from the chassis.

COIL TABLE FOR THE PRESELECTOR

L_1	5 t. No. 24, ¾-inch diameter (B & W 3012)
L_2	5 t. No. 24, 1-inch diameter (B & W 3016)
L_3	6 t. No. 24, ¾-inch diameter (B & W 3012)
L_4	7 t. No. 20, 1-inch diameter (B & W 3014)
L_5	7½ t. No. 20, ¾-inch diameter (B & W 3014)
L_6	3 t. No. 24, 1-inch diameter (B & W 3015)
L_7	11 t. No. 24 d.c.c., close-wound, ½-inch diameter
L_8	4 t. No. 28 d.c.c., close-wound, ½-inch diameter

L_7 and L_8 are wound adjacent on a ½-inch diameter polystyrene form (National PRD-2

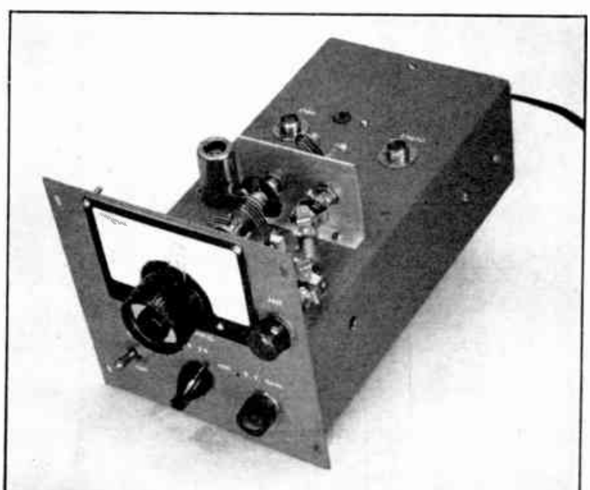


Fig. 5-53 — A bandswitching preselector for 14 and 28 Mc. A single 6AK5 amplifier is used, and the power supply is included in the unit. The antenna-trimming condenser is mounted on the small aluminum partition.

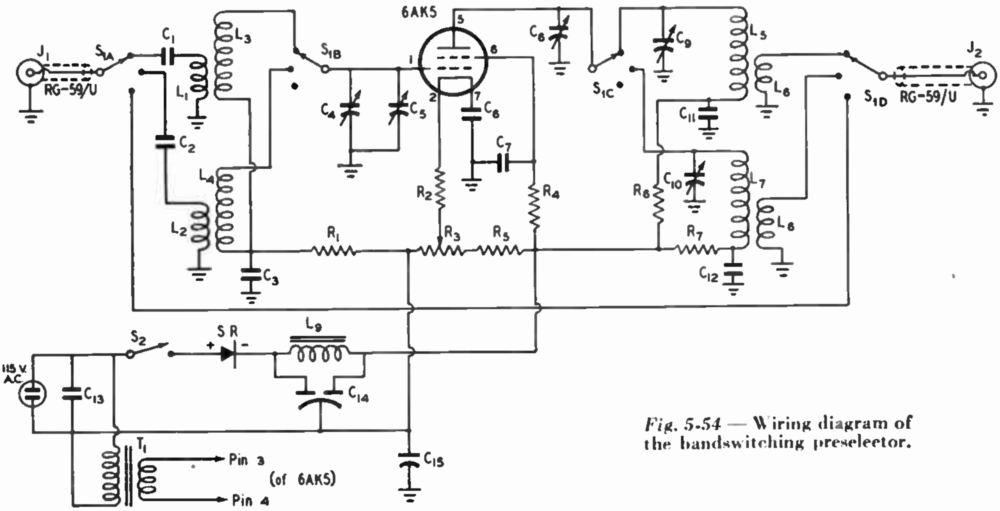


Fig. 5-54 — **Wiring diagram of the bandswitching preselector.**

C_1, C_2 — 10-μμfd. mica.
C_3, C_6, C_7, C_{11}, C_{12} — 680-μμfd. mica.
C_4 — 15-μμfd. midget variable (Millen 20015).
C_5, C_8 — 50-μμfd. midget variable (Millen 19050).
C_9, C_{10} — 3- to 30-μμfd. mica trimmer.
C_{13}, C_{15} — 0.01-μfd. paper, 400 volts.
C_{14} — Dual 40-μfd. 150-volt electrolytic.
R_1 — 27,000 ohms.
R_2 — 330 ohms.
R_3 — 5000-ohm wire-wound potentiometer.
R_4 — 4700 ohms.
R_5 — 18,000 ohms, 2 watts.
R_6, R_7 — 470 ohms.
L_1–L_8 — See coil table.
L_9 — 20-henry 30-ma. filter choke.
J_1, J_2 — Coaxial-cable jack (Jones S-101).
S_1 — 2-gang 2-circuit 5-position ceramic (Mallory 177C).
S_2 — S.p.s.t. toggle.
SR — 50 ma. selenium rectifier.
T_1 — 6.3-volt transformer.

The coils are made from B & W "Miniductors," as shown in the coil table, with the exception of one plate and coupling coil which are wound on a polystyrene form. The ground returns for the cathode and plate by-pass condensers are made to a common terminal, a soldering lug under one of the mounting screws for C_8.

When the wiring has been completed and checked, the antenna is connected to J_1 and a cable from J_2 is run to the receiver input. Tune the receiver to the 14-Mc. band and set S_1 to the proper point. Then turn the main tuning dial until the noise or signal increases to a maximum. This should occur with C_5 and C_8 set at close to maximum capacity. Then peak the noise by adjusting C_{10} and C_4.

The 28-Mc. range is adjusted in the same way, with the exception that C_9 is touched up. It may be found necessary to touch up C_4 when different antennas are used. The preselector may oscillate with no antenna connected, but with any type of wire or feedline the operation of the amplifier should ordinarily be perfectly stable.

As shown, the preselector is intended for use with coaxial-line feed to the antenna and to the receiver. If a balanced two-wire line is used from the antenna, it is recommended that a suitable two-wire connector be substituted for J_1. The grounded sides of L_1 and L_2 should be disconnected from ground and returned to one side of the connector. The output connector can be left as shown, since at the lower frequencies the proper antenna connection isn't so important.

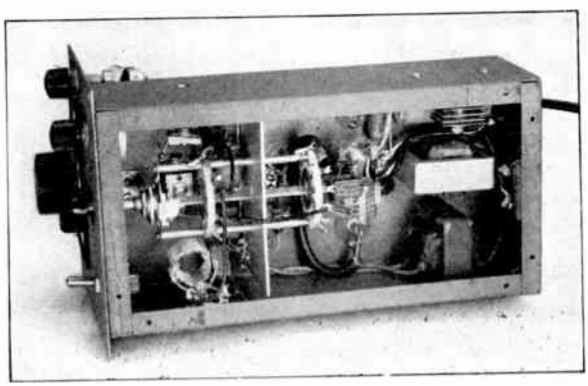


Fig. 5-55 — **A view underneath the chassis of the bandswitching preselector, showing the shield partition between switch sections and the selenium rectifier and associated filter.**

An Antenna-Coupling Unit for Receiving

It will often be found advantageous on the 14- and 28-Mc. bands to tune (or match) the receiving-antenna feedline to the receiver, in order to get the most out of the antenna. One way to do this is to use, in reverse, any of the line-coupling devices advocated for use with a transmitter. Naturally the components can be small, because the power involved is negligible, and small receiving condensers and coils are quite satisfactory. Some provision for adjustable coupling is recommended, as in the transmitting case, because the signal-to-noise ratio at 14 and 28 Mc. is dependent, to a large extent, on the degree of coupling to the antenna system. The tuning unit can be built on a small chassis located near the receiver, or it can be mounted on the wall and a piece of RG-59/U run from the unit to the receiver input, in the manner of a link line in transmitting practice. For ease in changing bands, the coils can be switched or plugged into a suitable socket. Adjustable coupling not only offers an opportunity to adjust for best signal-to-noise ratio, but the coupling can be decreased when a strong local signal is on the air, to eliminate "blocking" and cross-modulation effects in the receiver.

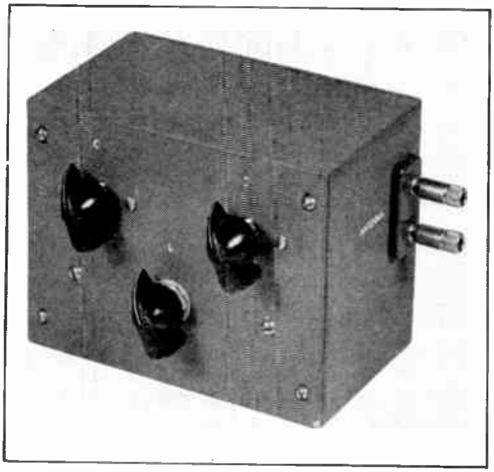


Fig. 5-56 — A compact coupling network for matching a balanced line to the receiver on 14 and 28 Mc.

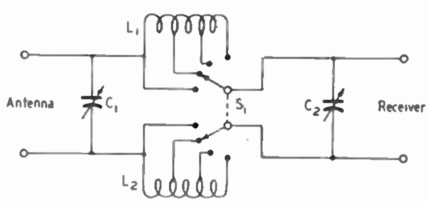


Fig. 5-57 — Circuit diagram of the coupling unit.
C_1, C_2 — 100-μμfd. midget variable (Millen 22100).
L_1, L_2 — 30 turns No. 18 d.c.c. close-wound on ½-inch diameter polystyrene form, tapped at 2½, 6½ and 14½ turns.
S_1 — 2-circuit 5-position single-section ceramic wafer switch (Mallory 173C).

Another convenient type of antenna-coupling unit for receivers uses the familiar pi-section filter circuit, and can be used to match a wide range of antenna impedances. The diagram of a compact unit of this type is shown in Fig. 5-57. Through proper selection of condensers and inductances, a match can be obtained over a wide range of values. The device can be placed close to the receiver and left connected all of the time, since it will have little or no effect on the lower frequencies. A short length of 300-ohm Twin-Lead is convenient for connecting the antenna coupler to the receiver.

The antenna coupler is built in a 3 × 4 × 5-inch metal cabinet. All of the components except the two pairs of terminals are mounted on one panel. The condensers are mounted off the panel by the spacers furnished with the condensers, and a clearance hole for the shaft prevents any short-circuit to the panel. The coils, wound on National PRD-2 polystyrene forms, are fastened to the panel with brass screws, and the coils should be wound on the forms as far away as possible from the mounting end. If this still leaves the coil ends within ½ inch of the panel, the forms should be spaced away from the panel by National XP-6 buttons. The switch should be wired so that the switching sequence puts in, in each coil, 0 turns, 2½ turns, 6½ turns, 14½ turns and 30 turns. All of the wiring, with the exception of the leads to the input and output terminals, can be done with the panel removed from the box.

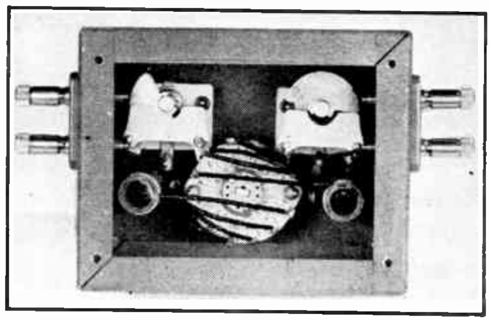


Fig. 5-58 — Rear view of the antenna-coupling unit. The two coils can be seen directly below the two tuning condensers.

The unit is adjusted for maximum signal by switching to different coil positions and adjusting C_1 and C_2. It will not be necessary to retrim the condensers except when going from one end of a band to the other, and when the unit is not in use, as on 7 and 3.5 Mc., the coils should be switched out of the circuit and the condensers set at minimum. The small capacity remaining has a negligible effect.

A One-Tube Converter for 10 and 11 Meters

The 10- and 11-meter converter shown in Figs. 5-59 and 5-61 is a simple unit that can be built in a few hours, for a cost of less than ten dollars. The converter uses a fixed-frequency oscillator and tunable input and output circuits. The fixed oscillator frequency is selected to take advantage of the calibration and bandspread offered by the communications receiver into which the converter works. Because of the light current consumption — 10 to 12 ma. — it is usually possible to operate the converter from the receiver power supply.

The circuit diagram, Fig. 5-60, shows that a Type 6BE6 miniature pentagrid-converter tube is used. The tuning range of the oscillator allows the oscillator to be set 4 to 6 Mc. below the frequency of the signal (input) circuit, and the receiver into which the converter works must be able to cover the range 4–6 Mc.

A Hartley circuit is used in the oscillator portion of the 6BE6. Coil L_3 is connected in parallel with condensers C_2 and C_4, and the frequency of the oscillator is determined by the values of these three components. The frequency of the oscillator must remain fixed after the converter has once been adjusted, and, as a result, stability is an important requirement. This condition is obtained by using a high-C tank circuit, with the 100-$\mu\mu$fd. condenser, C_4, providing the major portion of the capacity. The variable condenser, C_2, is used as a vernier control for selection of a spot-frequency within the oscillator-frequency range. Feed-back control for the oscillator is obtained by moving the 6BE6 cathode tap on

Fig. 5-59 — **A front view of the ten-meter converter. The components and controls on the front wall of the case, from left to right, are as follows: top row, r.f. tuning control, oscillator tuning knob, and i.f.-circuit control; bottom row, dial-light assembly, antenna change-over switch, and filament switch.**

L_3. Bias voltage for the oscillator is developed across resistor R_1, and C_6 is the grid-blocking condenser. Condenser C_6 keeps the screen grid at ground r.f. potential, and the dropping resistor, R_2, reduces the receiver supply voltage to 100 volts — the value recommended for the 6BE6 screen grid. The exact value for this resistor cannot be suggested at this time because the receiver supply voltage must be known before the resistance can be calculated. However, the resistor will carry about 7 ma., and it will probably have a resistance somewhere between 10,000 and 22,000 ohms.

The input circuit consists of coils L_1 and L_2 and condenser C_1. The antenna coil, L_1, is center-tapped to allow changing from the doublet to a single-wire type of antenna without the necessity for grounding one of the input terminals.

The output circuit uses a parallel tank circuit, C_3L_4, an output link, L_5, and a decoupling network formed by condenser C_7 and resistor R_3.

Antenna change-over and stand-by switching is done with the selector switch, $S_{1A\text{-}B\text{-}C\text{-}D}$. When set at one of the two positions, sections A and B will connect the antenna to the converter input coil while section C will connect the output link, L_5, to the output jack, J_1. At the same time, section D will complete the high-voltage connection between the input jack, J_2, and the plate and screen circuits. When the selector switch is thrown to the second position the antenna will be connected to the receiver and plate and screen voltage will be removed from the 6BE6. This action of disconnecting the antenna and high voltage during transmission periods prevents converter-tube overload and damage to the input coils that might be caused by the strong transmitter signal. A toggle switch, S_2, is used as the heater on-off control.

Construction

A utility box, measuring 3 × 4 × 5 inches, serves as the chassis and cabinet for the converter. The variable condensers, switches, pilot-light assembly and jacks should be mounted on the front and rear walls as shown in Fig. 5-60. The condensers are mounted in line on the front wall, with the shafts centered exactly 1 inch down from the top of the box. The pilot-light assembly and switches are mounted below the condensers and, in each case, are centered 11/16 inch above the bottom edge of the case.

The tube socket is mounted on the top cover of the utility box and is located 1⅝ inches from the front edge. Holes to pass the coil-form mounting screws are drilled on either side of the tube socket; these holes are ⅞ inch in from the ends of the cover. A tie-point strip is mounted to the rear and right of the tube socket.

Wiring of the unit will be greatly simplified if the wiring is divided into two jobs. The first half includes the wiring associated with the parts mounted on the case walls. This includes the jumper connections on the selector switch and the connections between this switch and the input and output jacks and terminals. Amphenol 300-ohm Twin-Lead is used between the antenna terminals and switch sections *A* and *B* but ordinary hook-up wire, twisted to form a low-impedance line, can be used. The lead from the switch to the output jack should be placed up against the rolled-over edge of the box, to obtain as much shielding as possible. The pilot light and toggle switch can be wired at this time, and a 6-inch lead should be left hanging from the switch side of the pilot light so that the tube filament circuit can be completed when the unit is assembled. The plate by-pass condenser, C_7, can be connected between the rotor terminals of the i.f. and oscillator condensers, and the decoupling resistor, R_3, can be mounted between C_3 and section *D* of the selector switch.

The input and output coils should now be wound on the forms suggested in the parts list. Holes, separated by the recommended distance, are drilled straight through the forms, and the ends of the windings are pulled through these holes and cemented in place. The antenna coil is wound directly below the grounded end of the grid coil, L_2, and the output link is wound over the cold end of L_4. It will not be possible to pass the top end of the output link, L_5, through a hole because L_4 is directly below this winding and, as a result, the free end of the link should be held in place with Scotch Tape or cement until the coil is mounted and wired. The oscillator coil, L_3, can be wound on a dowel or tube of ⅝-inch diameter; the coil will expand to a ¾-inch diameter when it has been slipped off the form.

The tube socket, tie-point strip and coils are now mounted in place on the box cover. Soldering lugs are placed under each of the tube-socket mounting nuts. The oscillator coil is soldered between one of the lugs and one of the tie-point terminals. Condenser C_4 is connected across the ends of L_3, and the grid resistor, grid-blocking condenser and screen by-pass are wired into the circuit. If the receiver supply voltage is known at this time it is possible to calculate the correct value for the screen-dropping resistor, and the resistor can be mounted on the tie-point strip. The resistor value is obtained from the equation

$$R\text{ (ohms)} = \frac{\text{supply voltage} - 100}{0.0073}$$

Example: Supply voltage 250; the resistor value is $\frac{250 - 100}{0.0073} = 20{,}500$ ohms. Anything within 10% of this figure would be satisfactory.

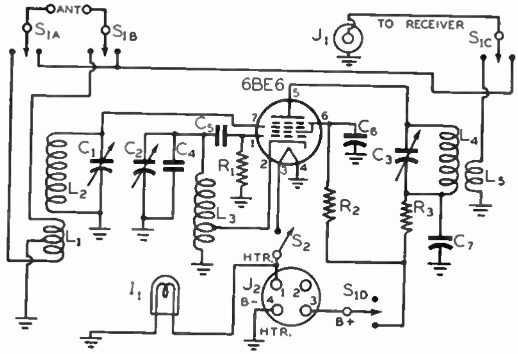


Fig. 5-60—Circuit diagram of the low-cost ten-meter converter.

C_1, C_2 — 15-μμfd. variable (Millen 20015).
C_3 — 75-μμfd. variable (Millen 20075).
C_4 — 100-μμfd. silver mica.
C_5 — 47-μμfd. mica.
C_6, C_7 — 0.01-μfd. paper.
R_1 — 22,000 ohms, ½ watt.
R_2 — Screen resistor; see text.
R_3 — 1,000 ohms, 1 watt.
L_1 — 5 turns No. 22 d.c.c., 9/16-inch diam., close-wound and center-tapped.
L_2 — 13 turns No. 22 d.c.c., 9/16-inch diam., ⅞ inch long.
L_3 — 6 turns No. 14 tinned, ¾-inch diam., ¾ inch long. Cathode tap 1¾ turns from cold end.
L_4 — 78 turns No. 32 d.c.c., 9/16-inch diam., 1¼ inches long.
L_5 — 10 turns No. 32 d.c.c., close-wound.
Coils L_1, L_2, L_4 and L_5 wound on National Type PRE-3 forms.
I_1 — 6.3-volt pilot-lamp-and-socket assembly.
J_1 — Panel-mounting female socket (Jones S-101).
J_2 — Panel-mounting male socket (Amphenol 86-CP4).
$S_{1A\text{-}B\text{-}C\text{-}D}$ — 4-pole double-throw selector switch (Mallory 3242J).
S_2 — S.p.s.t. toggle switch.

An 8-inch lead should be connected to the high-voltage end of the screen-resistor mounting terminal; the free end of this lead will be connected to the selector switch during the final stage of the wiring. The grounded ends of L_2 and L_5, and the center-tap of L_1, are connected to the grounded soldering lugs, and 2-inch tinned wire leads are connected to the following points: one to each soldering lug and one each to Pins 5 and 7 of the tube socket. A connection is now made between the cathode prong of the tube socket and the tap on coil L_3, and a connection is made between the screen dropping-resistor, R_2, and the screen-grid pin (No. 6) of the socket.

The top cover is now attached to the case and the wiring completed. Few connections remain to be made and, in each case, wires are already provided and soldered in place at one end. After the wiring has been completed it should be given a final check before the testing is started, paying special attention to the heater and plate circuits. Extreme care must be taken while soldering leads that terminate at the ends of L_1, L_2, L_4 and L_5. These coils are wound on polystyrene forms which melt and lose shape if subjected to intense heat for any length of time.

Testing

Adjustment of the converter is convenient if a test oscillator is available, but it is not necessary. Power for the unit can be obtained from the receiver with which the converter is to be used, or from a separate power supply. The converter requires 6.3 volts at 0.45 ampere for the heater and pilot lamp, and 200 to 250 volts d.c. at 10 to 12 ma. to supply the plate and screen power.

After the power supply has been connected, it is advisable to check the screen and plate voltages with a voltmeter. It may be necessary to change the screen-dropping resistor, R_2, if the voltage at Pin 6 isn't in the recommended range of 90 to 110.

A coaxial or shielded cable should be connected from the converter output jack to the receiver input terminals. The cable must be shielded to avoid the pick-up of unwanted signals. If your transmitter uses VFO, set it to 28 Mc. and your receiver to 4 Mc. If you don't have VFO but use crystal control, set the receiver to your crystal frequency minus 24 Mc. If, for example, your crystal gives a harmonic at 28,650 kc., set the receiver to 4650 kc. The converter oscillator condenser, C_2, should now be adjusted until the VFO or crystal harmonic can be heard. If the harmonic can't be heard, run a wire from the antenna posts of the converter close to the transmitter oscillator. If the signal from the transmitter oscillator is too loud, reduce the length of the wire or remove it entirely. When the signal is reasonably weak in the converter, the input and output tuning capacitors, C_1 and C_3, can be tuned to make sure that the coils don't need trimming to bring the tuning ranges within the limits of the bands.

Once the converter has been carefully set up on a known frequency within the 10- or 11-meter bands, C_2 is left fixed and the tuning is done with the receiver. The frequency of the incoming signal can be read directly from the receiver, by adding 24 to the receiver frequency in Mc. For example, a 28-Mc. signal will tune at 4 Mc., and a 29.250-Mc. signal will fall at 5.250 Mc. When tuning the 11-meter band, the setting of C_2 is changed so that a signal frequency of 27 Mc. corresponds to 4.0 Mc. on the receiver.

The converter, when properly aligned and working into an average receiver, gives a signal-to-noise ratio of 10 to 1 with an input signal of about 10 microvolts. In operation, C_1 and C_3 need not be touched over a tuning range of about 150 or 200 kc. on the receiver. Therefore, these controls should be touched up at intervals if the entire 10-meter band is being combed, but they require little or no adjusting in the 11-meter band.

It is important that the link between the converter be well shielded, to avoid picking up any signals in the tuning range of the receiver. A length of RG-58/U or RG-59/U should be used between the converter and the receiver and, if necessary, a small shield should be mounted over the antenna binding post on the receiver. If it is found to be impossible to keep out some particularly strong local signal that is being picked up on the coupling lead, it may be necessary to shift the tuning range of the receiver (by resetting C_2) to avoid this signal. Such a condition is very unusual, however, if care is taken with the coupling lead.

If no communications receiver is available, a war-surplus BC-454 aircraft receiver (tuning range of 3 to 6 Mc.) makes an inexpensive receiver for use with this converter.

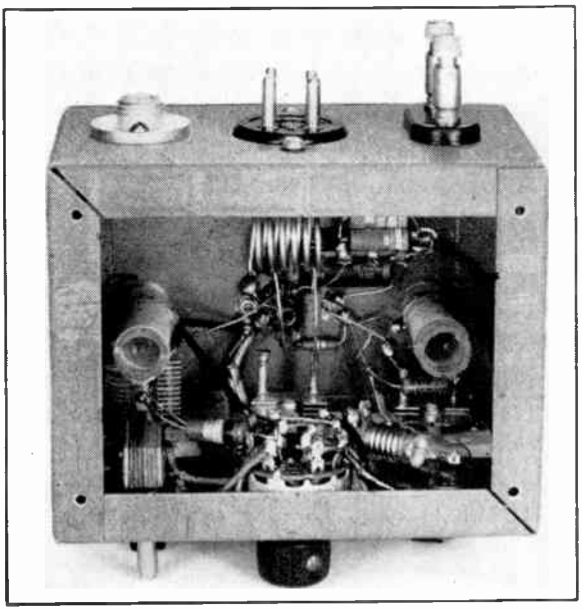


Fig. 5-61 — An inside view of the ten-meter converter. The r.f. and i.f. coils are at the right and left ends of the box, respectively. The oscillator coil may be seen to the rear of the tube socket. This view also shows the arrangement of the components mounted on the front wall of the case and the location of the input and output connectors which are mounted on the rear wall. The plate by-pass condenser is in a vertical position between the oscillator and the i.f. tuning condensers.

Crystal-Controlled Converters for 14, 21 and 28 Mc.

The principle of using a fixed high-frequency oscillator in a converter and tuning the receiver the converter works into can be elaborated upon by using a stage of r.f. amplification ahead of the mixer and by using a crystal-controlled oscillator for maximum stability. Since such a converter is generally used on a high frequency where fundamental crystals are not available, it is necessary to use a harmonic of a lower-frequency crystal. A crystal-controlled converter of this type is shown in Figs. 5-62 and 5-64. A separate converter is required for the 14-, 21- and 27-/28-Mc. bands, since by using separate converters it is possible to simplify their construction and to maximize their performance.

The converter uses the harmonic of a crystal oscillator to provide an exceedingly stable high-frequency oscillator signal. For example, in the 10-meter converter a 12.25-Mc. crystal doubles to 24.5 Mc., and this signal is fed to the mixer. By tuning the amplifier (your present receiver) following the mixer over the range 3.5 to 5.2 Mc., you are, in effect, tuning across the 28-Mc. band. The r.f. circuits in the converter are tuned to 28 Mc., and only have to be touched up when going from one end of the band to the other.

The wiring diagram is shown in Fig. 5-63. A neutralized triode-connected 6AK5 is used for the r.f. amplifier. There is some question as to its necessity on 14 and 21 Mc., where the atmospheric noise is generally high enough to limit the maximum usable sensitivity. A pentode-connected 6AK5 could probably be used with no detectable difference in performance on 14 and 21, but the triode is easy to handle and you don't lose anything by using it. Using high-impedance circuits with the pentode might give trouble from regeneration, unless the stage were neutralized. Adjustable antenna coupling and a Faraday screen are included to accommodate various antenna systems and to eliminate capacity coupling to the antenna line. The r.f. stage runs at 105 volts on the plate, since this gives the best noise figure. The separate plate lead also offers an opportunity to kill the converter by opening this circuit. The 6AK5 pentode mixer is easy to handle and quiet enough so that its noise doesn't impair the over-all performance. A triode mixer might be used, but the pentode runs with low current and is quiet.

The plate circuit of the mixer is tuned to the center of the receiver tuning range by setting L_4 to resonate with the various shunt circuit capacities. The circuit has a low Q and there is little variation in gain over the range. A 6C4 cathode follower is used as a low-impedance coupling to the receiver input.

One section of a 6J6 twin triode is used for the crystal oscillator, and the other half serves as a frequency multiplier. To minimize the other harmonics existing in the plate circuit of the multiplier, the plate is tapped down on L_6.

To get the best possible r.f. circuits, within the space limitations, B & W "Miniductors" are used for L_1, L_2 and L_3. Their Q is well above that obtainable with smaller-diameter coils, and they are easy to handle. To insure good shielding and low-resistance ground paths, an aluminum chassis is used in preference to the more common steel units.

The converter is built on a 5 × 9½ × 3-inch aluminum chassis, with several shield partitions to reduce unwanted interstage coupling. The most important shield is the one that straddles the r.f. amplifier socket and separates the grid and plate circuits of this stage. The grid tuning condenser, C_2, is mounted on bakelite insulating washers, and its ground lead returns to the common ground at the tube socket, to eliminate stray coupling through chassis cur-

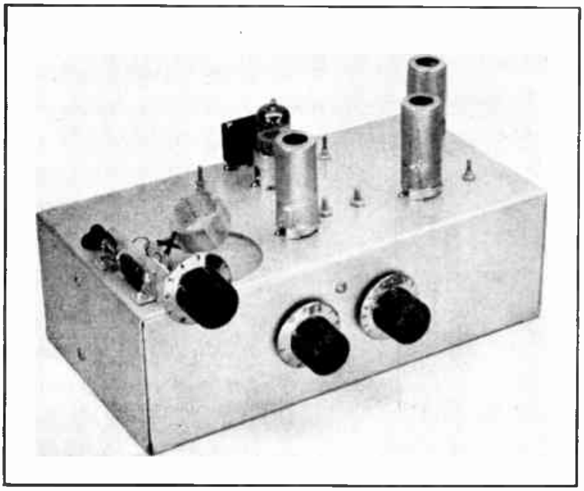


Fig. 5-62 — A 28-Mc. crystal-controlled converter. The adjustable antenna coupling can be seen at the left front. The tube shields, from left to right, cover the triode-connected 6AK5 r.f. amplifier, the 6AK5 mixer and the 6C4 cathode follower. The unshielded tube is the 6J6 oscillator-multiplier.

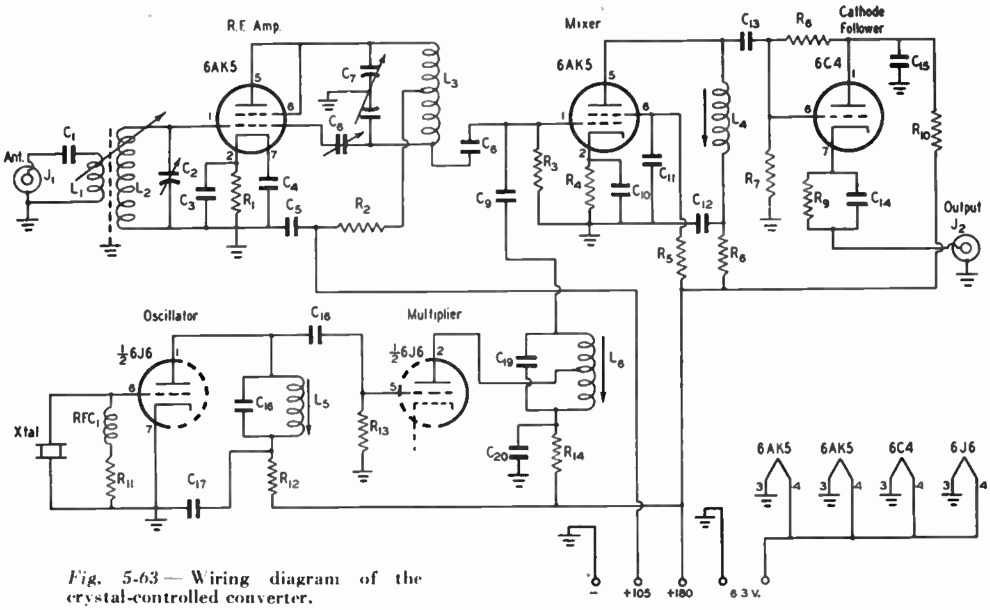


Fig. 5-63 — Wiring diagram of the crystal-controlled converter.

C_1 — 10-μμfd. mica.
C_2 — 20-μμfd. midget variable (Johnson 160-110).
C_3, C_4, C_5, C_{10}, C_{11}, C_{12}, C_{14}, C_{15}, C_{17}, C_{20} — 680-μμfd. mica.
C_6 — 5-μμfd. midget variable (Johnson 160-102).
C_7 — 11-μμfd. midget butterfly (Johnson 160-211).
C_8, C_{13} — 470-μμfd. mica.
C_9 — Twisted wire. See text.
C_{16}, C_{19} — See coil table.
C_{18} — 47-μμfd. mica.
R_1, R_9 — 220 ohms.
R_2 — 2200 ohms, 1 watt.
R_3 — 56,000 ohms.
R_4 — 6800 ohms.
R_5 — 0.1 megohm.
R_6, R_{10}, R_{12}, R_{14} — 470 ohms.
R_7, R_{11} — 4700 ohms.
R_8 — 0.18 megohm.
R_{13} — 82,000 ohms.
All resistors ½-watt unless otherwise specified.
L_1, L_2, L_3, L_4, L_5, L_6 — See coil table.
J_1, J_2 — Cable-connector sockets (Jones S-101).
RFC_1 — 750-μh. r.f. choke (National R-50).
XTAL — See coil table.

rents. If this isn't done, you may have trouble neutralizing the amplifier.

A 2¼-inch diameter hole is punched in the chassis, so that the externally-mounted antenna coil, L_1, can be coupled to the grid coil, L_2. The Faraday screen is then mounted across this hole on the underside of the chassis. To construct the Faraday shield, first cut a piece of ⅛-inch-thick polstyrene (Millen Quartz-Q) to measure 2½ by 3¼ inches, and drill a pair of holes at one end to clear No. 6 screws, for mounting the finished shield. (These are the same screws that hold the mounting strip for the antenna condenser, C_1, visible in Fig. 5-62.) At the opposite end of the poly sheet, drill a small hole in each corner, for securing the wire used in making the shield. Then wind No. 20 tinned wire tightly around the poly sheet in the long direction, spacing it with string or more No. 20 wire. When the winding is finished and secured at both ends, unwind the spacing string (or wire) and remove it. If you have done the job carefully, you will have neat parallel lines of wire across the polystyrene, all equally spaced and all lying fairly flat. Then apply two or three heavy coats of Duco cement to *one side only*, allowing sufficient time between coats for the cement to harden thoroughly. When this has been done, it will be found an easy job to cut each wire on the uncemented side. Straighten out the wires so that you now have a flat sheet of parallel wires, and trim off the wires at the mounting holes end of the sheet along a line inside the mounting holes. Figs. 5-64 and 5-65 show what this looks like. When trimming these wires, be careful to see that no wire is left touching an adjacent one. Trim the wire ends at the other end to about ½ inch from the polystyrene. Clamp the shield in a vise, between two pieces of wood, and wrap each wire end around a piece of No. 12 tinned copper, as shown in Fig. 5-65. With a good hot iron, run a bead of solder along the bus, and your shield is finished. Work fast, and no heat will reach the poly. The shield is mounted with the smooth side exposed through the hole, and one end of the No. 12 bus is grounded at the r.f. tube socket.

The grid coil, L_2, is supported by its leads and a couple of drops of Duco cement that hold its grounded end to the Faraday shield. The antenna coil, L_1, is mounted by its leads on a piece of ¼-inch diameter polystyrene rod. The rod is supported by a shaft bushing. A small wire pin through the rod at the back of the bushing and a rubber grommet between the bushing and the control knob give a soft friction lock that holds the coupling in any position. Flexible leads run from the coil to C_1 and the shield of the RG-59/U coaxial line.

The r.f. plate coil, L_3, is cemented to a small piece of polystyrene sheet that is supported by two small brackets. The neutralizing condenser, C_6, is supported by one terminal of C_7 and a stiff wire lead back to the grid pin on the tube socket. The coupling condenser, C_9, is simply an insulated wire wrapped once around the lead from C_8 to the grid of the mixer. It is brought out of the oscillator compartment through a polystyrene or rubber grommet.

After the usual last check of the wiring, connect a power supply and remove the 6AK5 r.f. amplifier from its socket. Listen in on your receiver at the crystal frequency, and if you don't find the crystal signal, adjust L_5 until you do. Then set your receiver on the proper harmonic frequency and peak L_6 for maximum signal, as indicated by your S-meter. When you have done this, you can probably squeeze out a little more by readjustment of L_5. Then back off on L_5 a little, because there is no need to run the crystal at maximum.

COIL TABLE FOR THE CRYSTAL-CONTROLLED CONVERTER

	14 Mc.	21 Mc.	28 Mc.
L_1	23 t. No. 24 ¾-inch diam. (B & W 3012)	9 t. No. 24 1-inch diam. (B & W 3016)	10 t. No. 20 1-inch diam. (B & W 3015)
L_2	21 t. No. 24 ¾-inch diam. (B & W 3012)	10 t. No. 20 1-inch diam. (B & W 3015)	9 t. No. 20 1-inch diam. (B & W 3015)
L_3	38 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)	22 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)	16 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)
L_4	Slug-tuned coil (CTC 1-Mc. LS3 with 200 turns removed) (Coils for L_5 and L_6 are wound on ¼-inch diameter CTC LS3 forms)		
L_5	No. 32 enam., close-wound, ½ inch long	No. 32 enam., close-wound, ½ inch long	30 t. No. 28 enam., close-wound
L_6	22 turns No. 28 enam., close-wound, center-tapped	20 t. No. 20 enam., close-wound, center-tapped	20 t. No. 24 enam., close-wound, center-tapped
C_{16}	75 $\mu\mu$fd.	75 $\mu\mu$fd.	33 $\mu\mu$fd.
C_{19}	0	22 $\mu\mu$fd.	22 $\mu\mu$fd.
Xtal	6000 kc. (triples)	5875 kc. (triples)	12,250 kc. (doubles)

Then tune your receiver — its antenna circuit must complete the cathode circuit of the 6C4 follower — to about 3.8 Mc. and peak L_4 for maximum noise. The adjustment is not sharp, because of the low Q of the circuit. If your receiver has an antenna trimmer, don't forget to peak it, too. Then plug in the 6AK5 r.f. amplifier and, after the tube has warmed up, rock C_2 and C_7. Unless you are very lucky, you will find several settings where you are greeted by birdies and squawks. Through the hole in the bottom plate, use an alignment tool to adjust C_6 a little at a time, until you lose all of the unpleasant sounds with any settings of C_2 and C_7, and you have your r.f. stage neutralized. Connect the antenna, and peak C_2 and C_7 on the first signal you find. Do all of your tuning with your regular receiver, and only use C_2 and C_7 to peak the signal when you make a big frequency excursion. The adjustable antenna coupling provides some measure of gain control for the unit, but it is generally best to use fairly tight coupling and hold the gain down in your regular receiver. The coupling is designed for low-impedance input, and will work well with 50- or 75-ohm line.

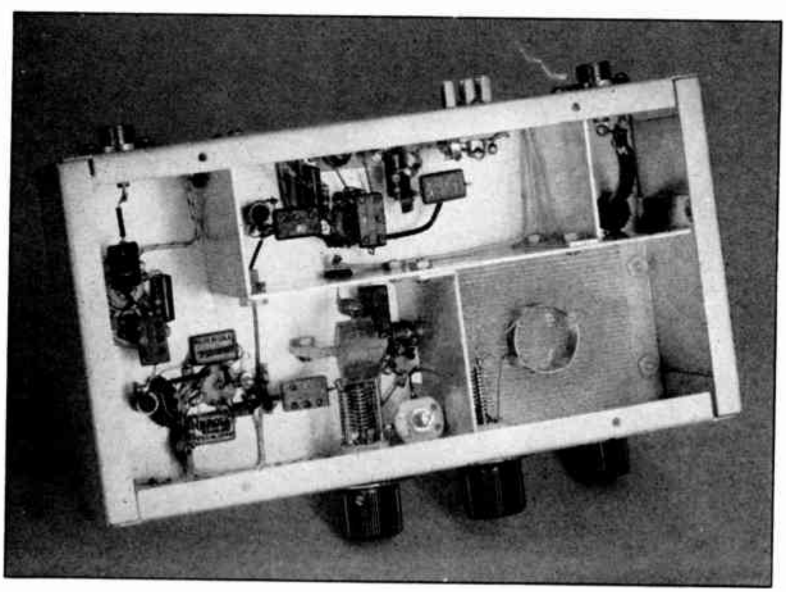


Fig. 5-64 — This view of the underside of the converter with the bottom cover removed shows the Faraday shield at the lower right, the shield straddling the r.f. amplifier socket (lower center) and the shielded oscillator section (top center). The neutralizing condenser for the r.f. stage is adjusted through a hole in the bottom cover.

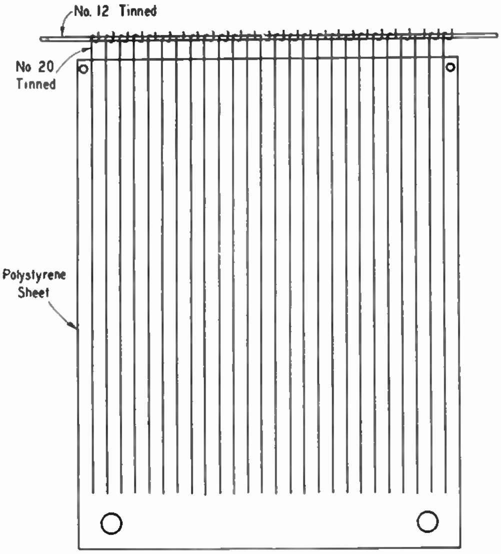


Fig. 5-65 — Constructional details of the Faraday shield, before soldering the ends of the No. 20 wires to the No. 12 wire bus.

coupling. The antenna coupling is designed for low-impedance input, and will work well with 50- or 75-ohm line. If you use 300-ohm Twin-Lead, it is better to leave the short length of coaxial line ungrounded and to use something other than a coaxial fitting for connecting the antenna. If your antenna uses 600-ohm line or tuned feeders, it is best to use a small antenna tuning unit link-coupled through a length of RG-59/U to the converter input.

There is nothing sacred about the crystal frequencies used, other than to be sure that they have no harmonics falling within the signal-frequency range. For the crystals suggested in the coil table, the receiver tunes from 4 to 3.6 to cover 14 to 14.4 Mc. (yes, it tunes backwards!), 3.375 to 3.825 for 21 to 21.45 Mc., and 3.5 to 5.2 for 28 to 29.7 Mc. The 27-Mc. band is also covered by the 10-meter converter, by tuning your receiver below 3.5 Mc.

What first i.f. (tuning range of your receiver) you will use depends on the available crystals and the range your present receiver tunes. Using the second or third harmonic of the crystal should be satisfactory in practically every case. By careful selection of crystal frequencies, you can arrange things so that the band edges start at some even 100-kc. mark on your receiver, thus giving you frequency-calibrated reception (with the necessary mental correction factor). The accuracy of calibration of your receiver on the one tuning range, together with the accuracy of the crystal used in the oscillator portion of the converter, will determine the accuracy of calibration of the receiving system.

Power Supply

The circuit diagram of a suitable power supply for use with the converters is shown in Fig. 5-66, although any source of 6.3 volts a.c. and 105 and 180 volts d.c. will do. One set of connections runs to the converter in use, and the other goes to a small control box located on the operating table. If desired, the a.c. switch can be incorporated in the power supply, but the plate switch, in the 105-volt lead to the r.f. stage, should be handy to the operator. A switch can be provided for shifting the power from one converter to another. Since separate antennas are generally used at these frequencies, the antennas do not require switching.

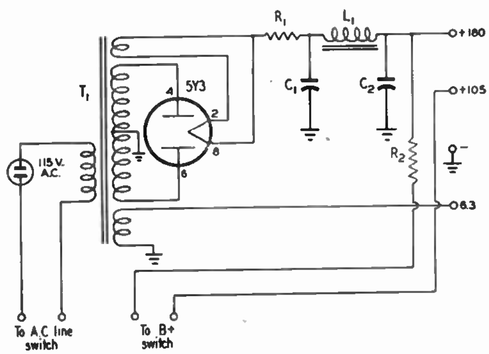


Fig. 5-66 — A power supply for the crystal-controlled converter.

C_1, C_2 — 8-μfd. 450-volt electrolytic.
R_1 — 1500 ohms, 10 watts.
R_2 — 10,000 ohms, 10 watts.
L_1 — 16-hy. 50-ma. choke (Stancor C-1003).
T_1 — 240–0–240 at 40 ma., 5 and 6.3 v. (Stancor P-6297).

A Simple Narrow-Band FM Adapter

Quite a few amateurs are now using NFM transmission, but most of the receivers in current use are of the straight AM type. Reception on a receiver equipped with an FM adapter is quite an improvement, from the standpoint of readability, over the same signal received by detuning the AM receiver to detect the FM signal on the i.f. slope.

With the adapter the "on-signal" noise level from external noise is reduced because of limiter action, and an improvement in readability is immediately noticed when receiving FM signals. Since the adapter allows you to tune to the center of the incoming carrier, any a.v.c. action in the receiver can be used to advantage to hold the "on-signal" noise level down by reducing receiver gain, an advantage that cannot be realized with i.f. slope detection. "Off-signal" noise is somewhat greater than with AM, but this is not too serious since most tuning is done on AM and the adapter is switched in when an FM signal is present.

Basically, the adapter unit consists of a limiter stage followed by a discriminator. The limiter uses a 6SJ7 tube with a 10-μμfd. coupling condenser and a 1-megohm grid leak, as shown in Fig. 5-67. The tube will reach full limiting at about 2.5 microvolts input to the average receiver, and the limiter output is constant over a wide range. Generally, the receiver i.f. system overloads before it is possible to overload the limiter, indicating that little would be gained from the standpoint of maintaining constant output by adding another limiter stage.

The discriminator transformer has two separate low-impedance primary windings, each with a separate secondary winding coupled to it. Each half of the transformer secondary is fixed-tuned with a 510-μμfd. silvered-mica condenser, and variable tuning is accomplished by means of movable iron cores. One of the transformer secondaries is tuned to a frequency approximately 10 kc. higher than the i.f. of receiver to which it is attached. The other is tuned approximately 10 kc. lower than the i.f. The transformer should be used with a receiver having an i.f. of approximately 456 kc. The bandwidth of the transformer is approximately 20 kc. and the characteristic is quite linear over approximately 12 kc. The output of the discriminator is fed directly to the receiver audio system.

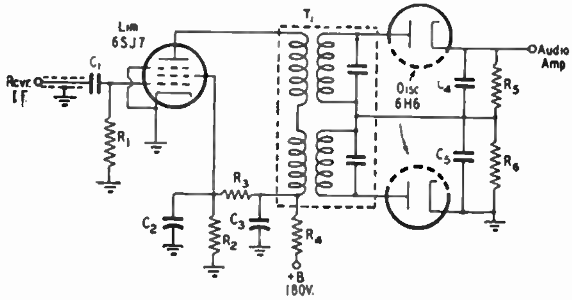


Fig. 5-67 — The NFM-adapter circuit.
C_1 — 10-μμfd. ceramic or mica.
C_2, C_3 — 0.1-μfd. paper.
C_4, C_5 — 100-μμfd. ceramic or mica.
R_1 — 1 megohm, ½ watt.
R_2, R_3 — 33,000 ohms, 1 watt.
R_4, R_5, R_6 — 0.1 megohm, ½ watt.
T_1 — Discriminator transformer (National SA-4842).

Construction of the unit is relatively simple, as will be apparent from reference to Figs. 5-68 and 5-69. It is possible to construct it in an evening. The chassis, which measures 2¼ × 5 × 1 inches, is constructed from a piece of 0.062-inch aluminum sheet measuring 4¼ × 7 inches. Construction can be simplified by

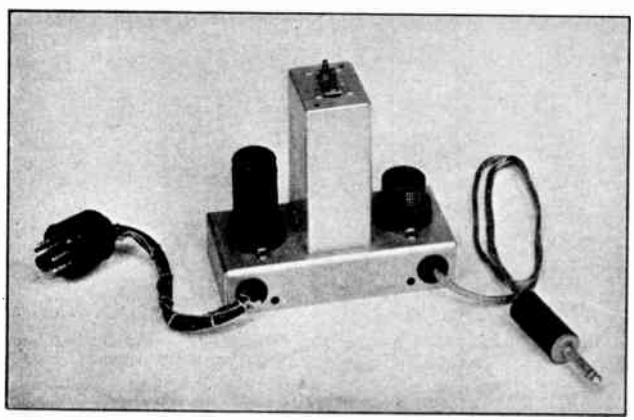


Fig. 5-68 — This two-tube adapter gives NFM reception with a communications receiver having an i.f. in the vicinity of 456 kc. The tubes are a 6SJ7 and a 6H6. The 'phone plug, connected to the audio output terminal in the unit, is plugged into the 'phone jack on the receiver for FM reception, and simply pulled out of the jack for AM.

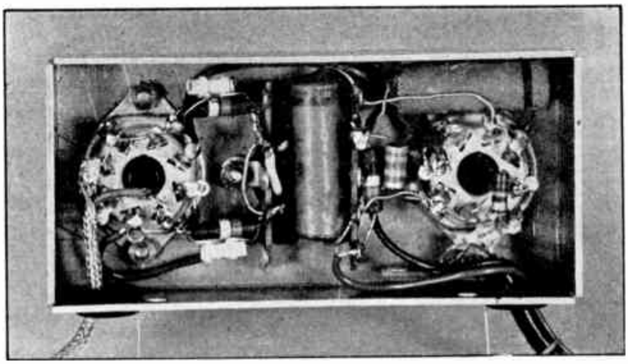


Fig. 5-69 — The simplicity of the wiring is evident in this underneath view of the adapter. The i.f. lead, a piece of small coaxial cable, is laced with the power and audio leads.

using a piece of metal 4¼ by 5 inches and by putting only two bends in the chassis, making it "U"-shaped. Socket holes, as well as the mounting holes for the discriminator transformer, are punched before the chassis is bent to its final shape. Insulated lugs can be mounted on socket screws, as shown, to provide for neat layout of parts.

Some receivers are provided with adapter sockets at the rear into which the adapter may be plugged. This adapter socket provides all voltages necessary to operate the adapter. The audio output can be run through a shielded lead to the phonograph-input jack on the front of the receiver, if the receiver has one, and switching from AM to FM reception is then accomplished by inserting the plug in the phono jack. Simpler methods can be devised, especially if the user has no objection to adding a switch to the front panel of his receiver. In this event it is simply necessary to switch either the AM-detector output or the FM-discriminator output to the audio input of the receiver. It is not necessary to switch off the B-plus of the adapter tubes since interference from cross-talk is negligible.

The i.f. output can be taken from the plate of the second i.f. tube, and there will be some detuning of the detector input transformer. The simplest way to retune the detector transformer, when no signal generator is available, is to set the receiver for maximum background noise with no signal present. When this unit is used in connection with receivers having high-impedance i.f. systems, care must be taken to have the lead from the i.f. tube to the limiter well shielded and as short as possible. Small coaxial cable, such as RG-59/U, should be used for this lead. In the event oscillation troubles are encountered in the receiver with the adapter in place, make sure the adapter unit is well grounded to the receiver and all shielding is attached to a good ground, preferably in the receiver if possible.

High-Frequency Transmitters

Transmitters for the amateur bands lying between 3.5 and 30 Mc. may take a variety of forms, depending primarily upon the frequency bands to be covered and the power output desired. Added to these are such important factors as operating convenience and space restrictions.

The principal requirement that must be met in c.w. transmitters, to which this chapter is limited, is that the output must be confined as closely as the state of the art permits to a single steady frequency free from modulation. A frequency-stable signal is necessary not only to comply with FCC regulations, but also to provide a signal that can be received satisfactorily with a selective receiver, and one that will cause a minimum of interference to amateurs working in the same band. Radiation of signals at harmonic frequencies, or spurious radiations at other frequencies, must be minimized to prevent interference to other radio services.

A simple oscillator may be used as a transmitter, as shown in Fig. 6-1A, but the amount of power obtainable with satisfactory frequency stability is small. Therefore in most transmitters the oscillator is used to feed one or more amplifiers as required to bring the power up to the desired level, as indicated at B, before delivering the power to the antenna system.

An amplifier whose output frequency is the same as the input frequency is called a **straight amplifier.** If such a straight amplifier is placed in an intermediate position between two other transmitter stages it is sometimes called a **buffer amplifier.**

Because it becomes increasingly difficult to maintain oscillator frequency stability as the frequency is increased, it is most usual practice in working at the higher frequencies to operate the oscillator at a low frequency and follow it with one or more **frequency multipliers** as required to arrive at the desired output frequency. A frequency multiplier is an amplifier that delivers output at a multiple of the exciting frequency. A **doubler** is a multiplier that gives output at twice the exciting frequency; a **tripler** multiplies the exciting frequency by three, etc. Although multiplications in a single stage as high as eight or more sometimes are used to reach the bands above 30 Mc., in the majority of low-frequency transmitters, multiplication in a single stage is limited to two or three, since the efficiency of a multiplier decreases rapidly as the order of multiplication increases. Also, it becomes more difficult to keep unwanted harmonics from the output.

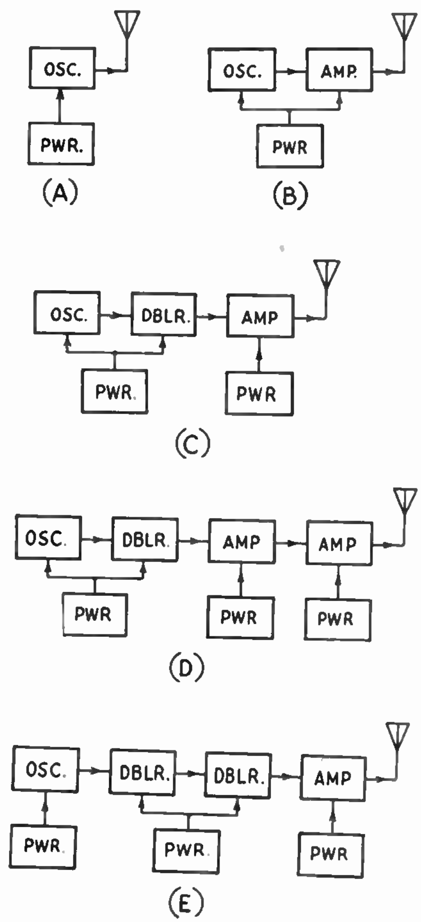


Fig. 6-1 — Block diagrams showing typical combinations of oscillator and amplifiers and power-supply arrangements for transmitters. A wide selection is possible, depending upon the number of bands in which operation is desired and the power output.

Frequency multipliers sometimes are used to feed the antenna system directly, but preferably should feed a straight amplifier which, in turn, feeds the antenna system, as shown in Fig. 1-C, D and E, because it is otherwise difficult to eliminate the multiplier driving frequency in the antenna system. As the diagrams indicate, it is often possible to operate more than one stage from a single power supply.

Variable-Frequency Oscillators

Two general classes of oscillators are used in amateur transmitters. A crystal-controlled oscillator is a fixed-frequency oscillator. The frequency generated is held within very close limits (a few cycles per megacyle) by a quartz crystal. The frequency is determined almost entirely by the thickness of the crystal. Other constants in the circuit have relatively little effect. The frequency of a self-controlled or variable-frequency oscillator (VFO) is determined principally by the values of inductance and capacitance which make up the oscillator tank circuit.

The disadvantage of the crystal type of oscillator is that a different crystal must be used for each frequency desired (or multiples of that frequency). By making the inductance, capacitance, or both, variable in the self-controlled oscillator, it may be operated at any frequency desired within a band at the turn of a dial, in the manner of a receiver. The disadvantage of a VFO is that much care must be exercised in the design and construction if the frequency stability is to approach that of a crystal-controlled oscillator.

Although the trend in recent years has been toward the VFO with its greater flexibility, the crystal oscillator still is widely used by beginners and is preferred by many others because of the comparative ease with which frequency stability and calibration are maintained.

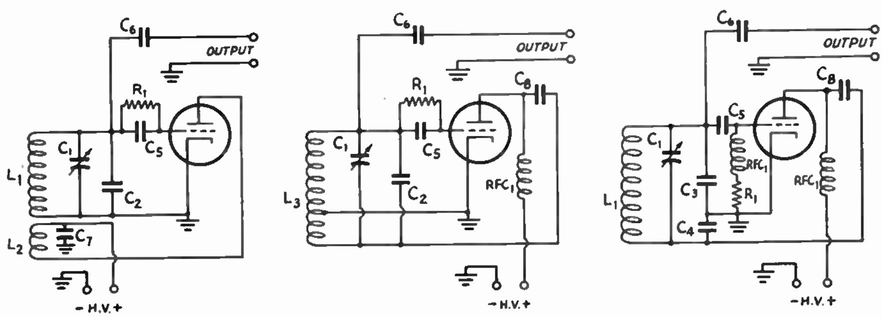
While any of the basic self-controlled oscillator circuits may be used, the prevailing choice lies among those shown in Fig. 6-2, or modifications of these circuits.

To provide satisfactory performance on the air, special attention must be paid to the circuit and mounting of parts. Since the frequency depends upon the L and C in the circuit, anything which operates to change these values will cause a change in frequency. For stability which will approach that of which a crystal oscillator is capable, the values of inductance and capacitance must be held within extremely small tolerances.

It is perhaps not too difficult to provide a satisfactory coil and condenser for the tank circuit. But the tube must be connected across this circuit and its effect upon frequency is by no means negligible nor easily controlled. The tube has the effect of a capacitance which can be made to hold satisfactorily constant only with great care.

Effects of Load

It is obvious too that the connection of any reactive load, such as an antenna or the input of an amplifier stage, will change the frequency, since this load must be connected across the frequency-determining circuit, thereby changing the net value of inductance or capacitance as the case may be. An antenna and feeders cannot be held sufficiently rigid to prevent changes in their capacitances. For this



(A) (B) (C)

Fig. 6-2 — Typical simple VFO circuits. A — Tickler feed-back. B — Hartley. C — Colpitts. Low-μ triodes, such as the 6C5 or 6J5, are preferable. Approximately appropriate values for the 3.5-Mc. band are as follows:

C_1 — Tuning condenser — 150-$\mu\mu$fd. variable.
C_2 — Tank condenser — 500-$\mu\mu$fd. zero-temp. mica.
C_3 — Tank condenser — 700-$\mu\mu$fd. zero-temp. mica.
C_4 — Tank condenser — 0.0021-μfd. zero-temp. mica.
C_5 — Grid condenser — 100-$\mu\mu$fd. zero-temp. mica.
C_6 — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.
C_7 — Plate by-pass condenser — 0.01-μfd. paper.
C_8 — Plate blocking condenser — 0.001-μfd. mica.
R_1 — Grid leak — 50,000 ohms.
L_1 — Tank coil — 4.3μh.
L_2 — Tickler winding — Approximately one-third number of turns on L_1, wound on same form next to L_1 or over ground end of L_1.
L_3 — Same as L_1, tapped approximately one-third from plate end.
RFC_1 — Parallel-feed r.f. choke — 2.5 mh.

reason it is almost universal practice to use an amplifier between the VFO and the antenna system.

Under practical operating conditions the input circuit of an amplifier may develop changes in the reactance which it presents across the oscillator circuit, especially while it is being tuned or alternately connected and disconnected, which it is in effect if the amplifier is keyed. Special oscillator circuits have been developed to minimize this effect. Two forms of the **electron-coupled** oscillator circuit are shown in Fig. 6-3. In circuits of this type a single screen-grid tube performs the functions of both an oscillator and an amplifier. The screen serves as the plate of a triode oscillator, while the power is taken from a separate tuned output-plate tank circuit, the coupling between the two being principally through the common electron stream.

In Fig. 6-3A, the oscillator circuit is a Hartley in which the ground point has been shifted from the cathode to the "plate." Fig. 6-3B shows the Colpitts modified in a similar manner. The choke, RFC_1, is required to provide a d.c. path to the cathode without grounding it for r.f.

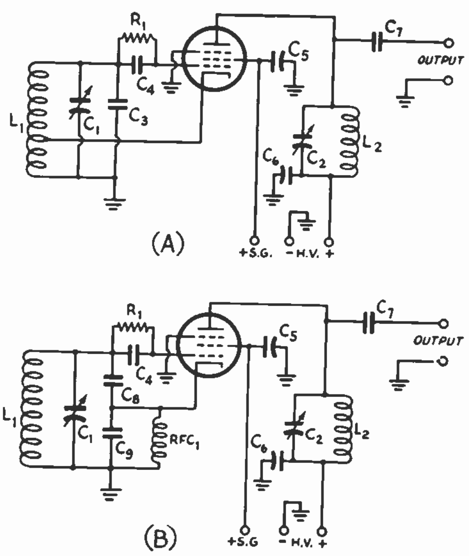


Fig. 6-3 — ECO circuits. A — Hartley. B — Colpitts. Approximate values are as follows:

C_1 — Oscillator tuning condenser — for 3.5 Mc.: 150-$\mu\mu$fd. variable.
C_2 — Output tank condenser — 100-$\mu\mu$fd. variable.
C_3 — Tank condenser — 500-$\mu\mu$fd. zero-temp. mica for 3.5 Mc.
C_4 — Grid condenser — 100 $\mu\mu$fd. or less, zero-temp. mica.
C_5 — Screen by-pass condenser — 0.01-μfd. paper.
C_6 — Plate by-pass condenser — 0.01-μfd. paper.
C_7 — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.
C_8 — 700-$\mu\mu$fd. zero-temp. mica.
C_9 — 0.0021-μfd. zero-temp. mica.
R_1 — Grid leak — 50,000 ohms.
L_1 — Oscillator tank coil — 4.3 μh. tapped approximately one-third from ground end for 3.5 Mc.
L_2 — Output tank coil — 22 μh. for 3.5 Mc., 7.5 μh. for 7 Mc.
RFC_1 — Parallel-feed r.f. choke — 2.5 mh.

In both of these circuits, output at a multiple of the oscillator frequency may be obtained by tuning the output-plate tank circuit to the desired harmonic, although this is seldom done beyond the second harmonic.

The oscillator frequency is not entirely independent of tuning or loading in the output plate circuit. The reaction is less, however, when the output-plate circuit is tuned to a harmonic or replaced by an untuned circuit, such as an r.f. choke, as shown in Fig. 6-4. The power output obtainable with the latter arrangement is much lower, however.

Well-screened tubes are preferable as electron-coupled oscillators. Those commonly used are the 6K7, 6SK7, 6F6 or 6AG7.

Another measure that may be taken to provide isolation between the oscillator and a following tuned amplifier is the use of an untuned amplifier, as shown in Fig. 6-5.

The power gain of an amplifier of this type is quite small, the purpose being almost entirely that of securing isolation between the VFO and tuned power amplifiers whose adjustment might react on the frequency of the oscillator if coupled to it directly. Two amplifier stages of this type usually are necessary before a following amplifier can be tuned or keyed without noticeably affecting the oscillator frequency and stability.

When using such an amplifier following an electron-coupled oscillator, a nonresonant output circuit also is usually used in the ECO. R.f. chokes are used as nonresonant circuits in the outputs of the ECO and in the second amplifier. L_1 in the plate circuit of the first amplifier is a winding that is self-resonant with the tube and circuit capacitances at a frequency near but not in the band of frequencies over which the amplifier is intended to operate. This is to prevent forming a low-frequency t.g.t.p. oscillating circuit which occurs when chokes of approximately the same characteristics are used in both input and output circuits of the amplifier tubes. For the same reason, resistors without chokes are used in the grid circuits.

Regulated voltages for the screens and plates are desirable.

Chirp

Variations in the voltage of the oscillator-tube elements can cause changes of appreciable magnitude in the effective input capacitance of the tube. If the oscillator can be run continuously during transmission, this effect can be made negligible by the use of regulated plate and screen voltages. But if the oscillator must be keyed for break-in work, an objectionable shift in frequency with keying **(chirp)** can be avoided only by reducing the time constant of the keying circuit to the point where the change in frequency between zero voltage, when the key is open, and full voltage, when the key is closed, takes place so rapidly that the ear cannot detect it. The time constant is reduced by minimizing any capacitance which may appear across the key contacts, including by-pass condensers in the transmitter. Unfortunately, as discussed in Chapter Eight, a certain time lag is required to eliminate clicks. Therefore the measures necessary for the elimination of chirps and clicks are in opposition. A compromise is usually necessary, unless the oscillator can be made insensitive to voltage changes by other means. It is possible that the keying of an amplifier may constitute little improvement over oscillator keying, for reasons previously given, unless sufficient isolation is provided between the oscillator and the keyed stage.

Drift

The effects of temperature change are characterized by a slow drift or creep in frequency. Part of this change, especially for the first few minutes after power is applied to the oscillator, may be attributable to change in tube-electrode capacitance as the tube heats up. But over a protracted period of time, drift is a result of small changes with temperature in physical dimensions of the coil and condenser in the tank circuit. Good design dictates that these components be of good construction and isolated as much as possible from the heat developed in the tubes and power-supply equipment. With care, frequency drift can be brought within satisfactory limits by mounting the tubes external to the enclosure surrounding the tank coil and condensers and the use of zero-temperature mica condensers for all tank capacitance other than that required for tuning purposes, by providing ventilation and by keeping the power input to the oscillator at a minimum — not more than a few watts. Where maximum stability with temperature change is desired, temperature-compensating condensers may be used to form part of the tank-circuit capacitance.

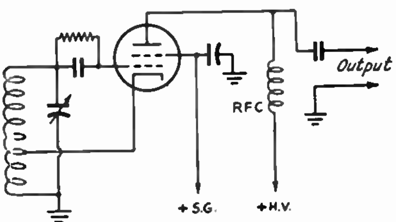


Fig. 6-4 — ECO with an r.f. choke replacing the output tank circuit for the purpose of reducing reaction on the oscillator portion of the circuit.

Mechanical Considerations

Any mechanical vibration which causes a change in the capacitance across the tank circuit, or in dimensions of the coil, will cause a corresponding change in frequency. This should be minimized by solid construction, by secure wiring and by cushioning the mounting of the oscillator unit against shocks. The oscillator should be thoroughly shielded from the strong r.f. fields of the antenna and

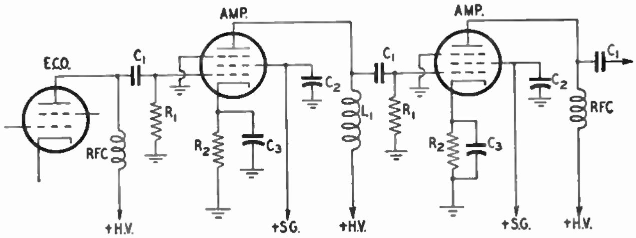


Fig. 6-5 — Diagram showing two isolating-amplifier stages coupled to the output of an ECO. Well-screened tubes are preferable, 6K7s or 6F6s being suitable.

C_1 — Coupling condenser — 100 $\mu\mu$fd. or less, mica.
C_2 — Screen by-pass condenser — 0.01-μfd. paper.
C_3 — Cathode by-pass condenser — 0.01-μfd. paper.
R_1 — Grid leak — 50,000 to 100,000 ohms.
R_2 — Cathode biasing resistor — 200 to 500 ohms.
L_1 — Coupling inductance — see text.
RFC — Plate choke — 2.5 mh.

adjacent high-power amplifier stages which may, through overloading of the oscillator grid, cause roughening of the oscillator signal.

Plug-in coils for changing oscillator frequency ranges are not recommended because experience has shown that the coil contacts may become the source of undesirable frequency instability.

VFO Tank Q

All of the previously-mentioned effects upon the frequency of an oscillator may be minimized by the use of high capacitance in the tank circuit, thus making uncontrollable capacity changes a small percentage of the total circuit capacitance. At 3.5 Mc., a tank capacitance of 500 to 1000 $\mu\mu$fd. is considered adequate, with values increased in proportion if the oscillator is designed to operate at lower frequencies. An increase in Q can be obtained also by tapping the tube across only a portion of the tank circuit. Fig. 6-6A shows the Hartley circuit with the grid and plate tapped across small portions of the tank-coil reactance. An equivalent arrangement for the Colpitts circuit is shown at B. C_1 (and C_2 in parallel) is small compared with C_3 and C_4. Therefore, the reactance across which each tube element is connected is a small portion of the total. C_2, which is the tuning condenser, should be no larger than is necessary to tune across the band so as not to influence the function of C_1 any more than necessary. The tuning condenser should not be connected across the coil, since this reduces the Q of the circuit.

In both of these arrangements, the higher the Q of the coil, the smaller the reactance between tube elements may be without stopping oscillation and, therefore, the greater the stability. Because of the high L/C ratio which results with the circuit of B, greater care must be exercised in the construction and mounting of tank-circuit components.

Any of the bandspread tuning systems used in receivers may be applied to the oscillator circuits which have been under discussion. The parallel-condenser system is used most widely since it lends itself well, particularly to high-C circuits.

Because it is considered easier to maintain percentage stability at lower frequencies, VFOs usually are designed to operate at a frequency not higher than the 3.5-Mc. band, the higher-frequency bands being reached by frequency-multiplier stages.

VFO ADJUSTMENT

Tuning Characteristics

Normally-operating VFO circuits of the types under discussion will function quite uniformly, over the range of an amateur band at least, as soon as plate voltage is applied. If, through incorrect adjustment of excitation or overloading the circuit does not oscillate, the plate current will be the zero-bias value for the tube at the plate voltage at which it is being operated, falling to a lower value when oscillation takes place. If the oscillator is functioning, touching the grid with a grounded prod will cause a variation in plate current. The value of plate current to be expected with a given tube when oscillating depends upon such factors as plate and screen voltages, grid-leak resistance, excitation adjustment and loading. It should remain essentially constant with reasonable changes in tuning capacitance. With normal excitation adjustment, the plate current should show an increase when the load is connected. Excitation and grid-leak resistance should be adjusted for maximum frequency stability — not maximum output.

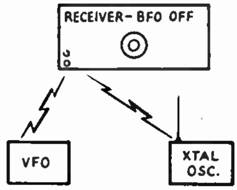


Fig. 6-7 — Set-up for checking VFO stability. The receiver should be tuned preferably to a harmonic of the VFO frequency. The crystal oscillator may operate somewhere in the band in which the VFO is operating. The receiver b.f.o. should be turned off.

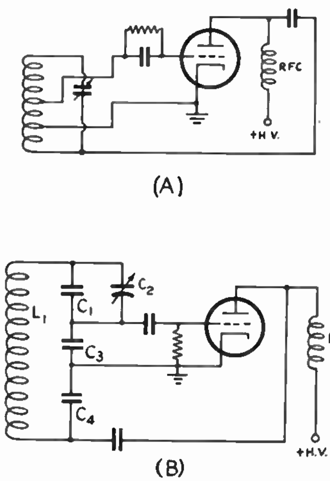


Fig. 6-6 — As an alternative to the use of a high-C tank circuit, oscillator tubes sometimes are connected across only a portion of the tank circuit to increase the Q. In the Hartley circuit of A, the grid and plate connections are made to taps instead of to the ends of the coil. In the Colpitts circuit of B, the division is by capacitive means. For 3.5 Mc., C_3 and C_4 should be about 0.001 μfd. and $C_1 + C_2$ no larger than necessary to maintain oscillation and tune across the band. The Q of L_1 and the G_m of the tube should be as high as possible.

In the circuit of Fig. 6-6A, maximum frequency stability is obtained with the plate and grid taps as close as possible to the cathode tap without stopping oscillation. In the circuit of Fig. 6-6B, maximum stability is obtained when C_3 and C_4 (usually equal) are large and the ratio of $C_1 + C_2$ to C_3 or C_4 is the maximum possible without stopping oscillation. The adjustment in each case will be limited by the Q of the coil. Therefore, the Q must be high for greatest frequency stability.

Checking VFO Stability

A VFO should be checked thoroughly before it is placed in regular operation on the air. Since succeeding amplifier stages may affect the signal characteristics, final tests should be made with the complete transmitter in operation. Almost any VFO will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, well-shielded monitoring arrangements is a receiver combined with a crystal oscillator, as shown in Fig. 6-7. (See "Crystal Oscillators.") The crystal frequency should lie in the band of the lowest frequency to be checked and in the frequency range where its harmonics will fall in the higher-frequency bands. The receiver b.f.o. is turned off and the VFO signal is tuned to beat with the signal from the crystal oscillator instead. In this way any receiver instability caused by overloading of the input circuits, which may result in "pulling" of the h.f. oscillator in the receiver, or by a change in line voltage to the receiver when the transmitter is keyed, will not affect the reliability of the check. Most present-day crystals have a sufficiently-low temperature coefficient to give a satisfactory check on drift as well as on chirp and signal quality if they are not overloaded.

Harmonics of the crystal may be used to beat with the transmitter signal when monitoring at the higher frequencies. Since any chirp at the lower frequencies will be magnified at the higher frequencies, accurate checking can best be done by monitoring at the latter.

The distance between the crystal oscillator and receiver should be adjusted to give a good beat between the crystal oscillator and the transmitter signal. When using harmonics of the crystal oscillator, it may be necessary to attach a piece of wire to the oscillator as an antenna to give sufficient signal in the receiver.

Checks may show that the stability is sufficiently good to permit oscillator keying at the lower frequencies, where break-in operation is of greater value, but that chirp becomes objectionable at the higher frequencies. If further improvement does not seem possible, it would be logical in this case to use oscillator keying at the lower frequencies and amplifier keying at the higher frequencies.

Crystal Oscillators

While crystal-controlled oscillators are much more tolerant than VFOs in respect to temperature changes, the danger of crystal fracture, as well as drift, places a limitation on the amount of power output obtainable. The oscillator normally should be considered as a frequency-generating device only, with power output of secondary importance. The amount of power which may be obtained from a crystal oscillator is limited by the heat it will stand without fracturing. The amount of heating is dependent upon the r.f. crystal current which, in turn, is a function of the amount of feedback required to provide proper excitation. Crystal heating short of the danger point results in frequency drift to an extent depending upon the way the crystal is cut. Excitation should always be adjusted to the minimum necessary for proper operation.

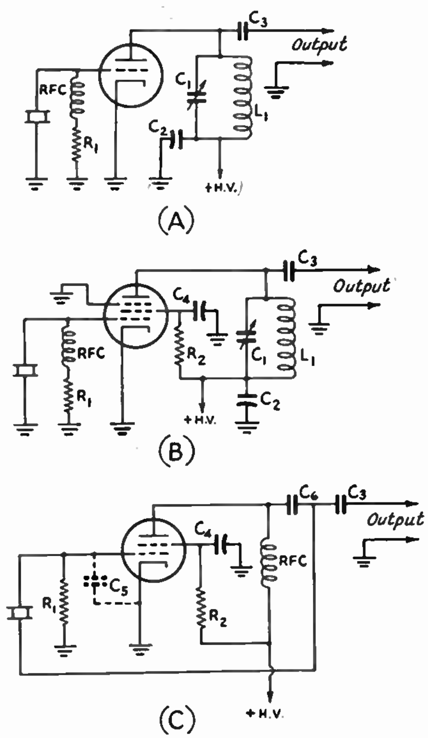


Fig. 6-8 — Simple crystal-oscillator circuits. A — Triode. B — Tetrode or pentode. C — Tetrode Pierce. Approximate values are as follows:

C_1 — Tank condenser — 100-μμfd. variable.
C_2 — Plate by-pass condenser — 0.01-μfd. paper.
C_3 — Output coupling condenser — 100 μμfd. or less, mica.
C_4 — Screen by-pass condenser — 0.01-μfd. paper.
C_5 — Feed-back condenser — 50 to 100 μμfd.
C_6 — Plate blocking condenser — 0.001-μfd. mica.
R_1 — Grid leak — 50,000 ohms.
R_2 — Screen voltage-dropping resistor — 25,000 to 50,000 ohms.
L_1 — Tank coil — 22 μhy. for 3.5 Mc.; 7.5 μhy. for 7 Mc.
RFC — Parallel-feed r.f. choke — 2.5 mh.

SIMPLE CIRCUITS

The basic crystal-controlled oscillator circuits are shown in Fig. 6-8. Since the crystal is the equivalent of a high-Q tuned circuit of fixed frequency, it will be observed that each of the crystal circuits is essentially the equivalent of a self-controlled circuit.

Triode, Tetrode and Pentode Oscillators

The triode crystal circuit of Fig. 6-8A is the equivalent of the t.g.t.p. circuit in which a crystal replaces the tuned grid circuit. The pentode circuit of B is the same except for the substitution of a screen-grid tube for the triode. This circuit sometimes is operated with the suppressor by-passed and raised to a positive voltage of about 50 instead of being grounded as shown. The same circuit is used for tetrodes, such as the 6V6 and 6L6, the suppressor connection being omitted.

With this circuit, oscillation takes place only when the plate tank circuit is tuned to a frequency higher than that of the crystal, and maximum output usually occurs when it is tuned close (but not exactly) to the crystal frequency. If the plate tuning condenser has sufficient range to tune the circuit to a frequency lower than that of the grid circuit, oscillation will cease and the plate current will jump to a relatively high value, as shown at the left in Fig. 6-9. As the plate circuit is tuned past the point of resonance with the crystal in the high-frequency direction, the plate current will drop suddenly (point A) indicating the starting of oscillation, then dip rapidly to a minimum (point C) where the power output is greatest. As the tuning capacitance is decreased further, the plate current will rise gradually to point *B* where it will jump to a higher value, indicating that oscillation has ceased. For maximum frequency stability the circuit should be tuned in the region *D-E*.

When the oscillator is loaded, the characteristic is similar (see dashed curve in Fig. 6-9), but the minimum plate-current dip is much less pronounced and the range of plate tuning over which the circuit will oscillate becomes less as the loading is increased.

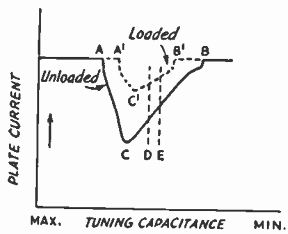


Fig. 6-9 — General tuning characteristic of triode, tetrode or pentode crystal oscillators. As the capacitance of the plate tank condenser is decreased from maximum, oscillation will start suddenly, the plate current dropping abruptly, as shown at point *A*. As the capacitance is decreased further, the plate current will dip to a minimum, *C*, and then gradually rise to point *B* where an abrupt rise in plate current will indicate that oscillation has ceased. Maximum output will be obtained at point *C*, but the oscillator should be adjusted for operation in the *D-E* region for best frequency stability.

With triode, tetrode or pentode circuits, the feed-back may be adjusted by the tuning of the plate tank circuit, as in the t.g.t.p. circuit. However, it is safer to limit the amount of power fed back from the plate circuit by restricting the plate voltage so that the crystal limitation cannot be exceeded inadvertently during normal tuning procedure. With the large prewar-type crystals, triode crystal oscillators may be operated with proper adjustment at plate voltages as high as 200 or 250, but the voltage should be reduced to 100 to 150 for the new-type smaller crystals. Low- or medium-μ tubes are preferable when triodes are used. Beam tetrodes or pentodes, with their high power-sensitivity and reduced grid-plate capacitance, require less voltage across the crystal than a triode for the same amount of output. With screen-grid tubes the larger-type crystals can be operated with plate voltages of 300 or 400 with power output up to 10 or 15 watts if required. The smaller crystals may take plate voltages up to 250 or 300 before showing marked drift in frequency. However, as stated previously, it is always advisable to limit the input to the oscillator and depend upon amplifiers for the desired power.

With the triode, tetrode and pentode crystal-oscillator circuits shown in Fig. 6-8, excitation, and therefore r.f. voltage across the crystal, is greatest when the oscillator is unloaded and it is under this condition that danger to the crystal is greatest.

Pierce Oscillator

The circuit shown in Fig. 6-8C is known as the Pierce circuit. It is the equivalent of the ultraudion variation of the Colpitts in which the grid-cathode and plate-cathode capacitances of the tube form the capacitive divider. The crystal replaces the single tuned circuit and thus this oscillator requires no tuning adjustment and will work without change in values over a wide range of crystal frequencies. Since excitation otherwise is not adjustable, except by change in plate voltage, the condenser C_5 sometimes is required to obtain satisfactory operation. Less power is obtainable from the Pierce circuit than from the preceding oscillators because the crystal is directly in the power-delivering circuit which limits the r.f. voltage that may be developed without danger to the crystal. Triodes also may be used in this circuit.

COMBINATION CRYSTAL OSCILLATORS

Tri-Tet Circuit

Fig. 6-10 shows three crystal-oscillator circuits which operate on principles similar to those of the ECO. Circuits such as these have the additional advantage that they are invariably found to key more reliably than the simple triode or tetrode circuits, and do not incur the considerable loss in efficiency sometimes involved in detuning the plate circuit far to the high-frequency side of resonance for reliable crystal starting under load.

The extent to which the output-plate circuit reacts on the oscillator portion of the circuit, and the output-circuit tuning characteristics, are influenced to a considerable degree by the effectiveness of the screening of the tube selected. Well-screened tubes always are preferable from the standpoints of both isolation and safety to the crystal.

Fig. 6-10A shows the Tri-tet circuit. The oscillator portion is equivalent to that of a triode crystal oscillator, with the screen serving as the "plate" and the ground point being shifted from the cathode to the "plate." Power is taken from a separate output-plate tank circuit. Since the output-plate circuit returns to cathode through the L_1C_1 tank circuit, the plate contributes to the feed-back to a certain extent. In addition, the screen-to-control-grid capacitance is greater than the corresponding plate-grid capacitance of most triodes. Therefore, L_1C_1 should always be tuned well to the high-frequency side of the crystal frequency to prevent excessive feed-back and consequent unnecessary crystal heating. In respect to heating of the crystal, limitation of screen voltage is of greater importance than plate voltage.

As with the ECO, harmonic output may be obtained by tuning the output tank circuit, L_2C_2, to the desired multiple of the crystal frequency.

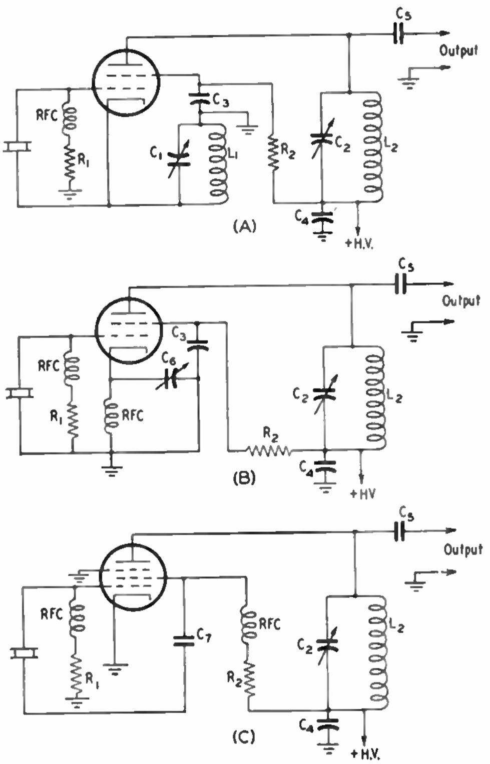


Fig. 6-10 — Crystal-controlled amplifying-oscillator circuits. A — Tri-tet. B — Grid-plate. C — Modified Pierce. Approximate values are as follows:

- C_1 — Cathode-tank tuning condenser — 100-μμfd. variable.
- C_2 — Output-tank tuning condenser — 100-μμfd. variable.
- C_3 — Screen by-pass condenser ("plate" grounding) — 0.01-μfd. paper.
- C_4 — Plate by-pass condenser — 0.01-μfd. paper.
- C_5 — Output coupling condenser — 100 μμfd. or less, mica.
- C_6 — Feed-back control condenser — 100-μμfd. variable.
- C_7 — Parallel-feed blocking condenser — 0.001-μfd. mica.
- R_1 — Grid leak — 50,000 to 150,000 ohms.
- R_2 — Screen voltage-dropping resistor — 25,000 to 100,000 ohms.
- L_1C_1 — Tuned well above crystal frequency (see text).
- L_2C_2 — Tuned to crystal frequency or desired harmonic.
- RFC — Parallel-feed r.f. choke — 2.5 mh.

When operating the Tri-tet circuit at the crystal frequency, L_1C_1 should be tuned no closer to the crystal frequency than is necessary to make the circuit oscillate without output-plate voltage applied. With poorly-screened tubes, it may be advisable to short-circuit L_1C_1 when operating at the crystal fundamental frequency, reverting to the simple tetrode circuit as a measure of safety to the crystal.

With well-screened tubes, such as the 6SK7, 2E25 or 802, the output-plate tuning characteristic is like that shown in Fig. 6-11 at the fundamental as well as at the harmonics, and the circuit will continue to oscillate regardless of the tuning of the output circuit. However, with poorly-screened tubes, such as the 6V6, 6F6 or 6L6, the circuit may stop oscillating abruptly when the output circuit is tuned to a frequency lower than the crystal frequency, more in the manner of a simple triode or tetrode oscillator. With well-screened tubes, feed-back is at a minimum when the output circuit is unloaded, the excitation increasing as load is increased. This characteristic is opposite to that of the triode or tetrode crystal oscillator.

Grid-Plate Oscillator

A less widely-used circuit is the grid-plate crystal-oscillator circuit of Fig. 6-10B. The oscillator portion is similar to the triode Pierce with the screen being used as the "plate," and the ground point shifted to the "plate." L_2C_2 is the output-plate circuit. C_6 adds to the "plate"-cathode capacitance for feed-back-adjustment purposes. As in the Tri-tet circuit, the return for the output-plate circuit to cathode is through C_6 which is here common to the grid return, and therefore the isolation between oscillator and output circuits is incomplete and the plate contributes to the feed-back. Here, too, screen voltage is important in respect to crystal heating. With a well-screened tube, the oscillator will continue to function regardless of the tuning of the output circuit. As with the Tri-tet, harmonic output may be obtained by tuning L_2C_2 to multiples of the crystal frequency.

Modified Pierce

Another version of the Pierce circuit, adaptable to pentodes, is shown in Fig. 6-10C. The oscillator portion of the circuit is the triode Pierce with the cathode grounded and the screen serving as the "plate." In this arrangement the output-plate and grid returns are through independent paths and the suppressor provides screening against capacitive coupling. In theory, at least, this circuit provides better isolation between the oscillator and output portions of the circuit than either of the other two. Pentodes, such as the 6SK7, 6AG7 and 802 are suitable for this circuit.

Variation in screen voltage provides the primary means of adjusting feed-back, although it may also be changed by adding

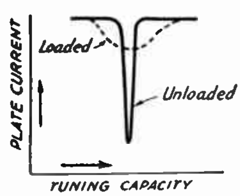


Fig. 6-11 — Plate-tuning characteristic for Tri-tet, grid-plate and modified Pierce crystal-oscillator circuits, using well-screened tubes, for loaded (dashed line) and unloaded (solid line) conditions. In this case the output-plate circuit may be tuned accurately to the minimum plate-current dip for maximum output.

capacitance between either grid and cathode or screen and cathode, as found desirable. Harmonic output may be obtained by tuning the output circuit to a multiple of the crystal frequency.

CRYSTALS

Crystal Characteristics

While crystals are produced for frequencies as high as 50 Mc., by far the majority of those used in amateur high-frequency transmitters are cut for the 3.5- and 7-Mc. bands. With suitable frequency-multiplying stages, this permits the use of a single crystal for operation in the harmonically-related parts of higher-frequency bands, as well as at the crystal fundamental frequency. As an example, a 3501-kc. crystal with appropriate multipliers may be used for frequencies of 7002 kc., 14,004 kc., 28,008 kc., etc.

Crystals vary in characteristics depending upon the manner in which they are cut from the natural quartz crystal, particularly in the thickness-frequency and temperature-frequency relationships. The frequency of crystals of the earlier cuts, designated "X" and "Y," vary appreciably in frequency with changes in temperature. More recently they have been almost entirely superseded by the modern "AT-" and "BT"-type crystals which are cut so as to have very small frequency change with temperature. The temperature-frequency characteristics for various crystal types are summarized in the following tabulation:

X-cut — −10 to −25 cycles per Mc. per degree C.
Y-cut — +100 to −20 cycles per Mc. per degree C.
AT-cut — +10 cycles per Mc. per degree at 0 degrees C.
— 0 cycles per Mc. per degree at 45 degrees C.
— +20 cycles per Mc. per degree at 85 degrees C.
BT-cut — −10 cycles per Mc. per degree at 0 degrees C.
— 0 cycles per Mc. per degree at 30 degrees C.
— −20 cycles per Mc. per degree at 70 degrees C.

The relationship between the thickness of a crystal and its frequency is given by:

$$f_{Mc.} = \frac{k}{t_{mil}}$$

where $f_{Mc.}$ is the frequency in megacycles, t the thickness in thousandths of an inch and k is a constant of the crystal cut approximately as follows:

X-cut — 112.6
Y-cut — 77
AT-cut — 66.2
BT-cut — 102

An AT crystal usually is more active than one of the BT-cut type, but since it is thinner for the same frequency, there is greater danger of fracture in operation. Therefore, AT-cut crystals usually are used for frequencies below 5 Mc., while the BT-cut is used for crystals whose frequencies lie above 5 Mc., although this is not true in all cases.

While crystals are sometimes cut for fundamental frequencies as high as 14 Mc., most crystals used by amateurs for frequencies higher than the 7-Mc. band are "harmonic-type" crystals; that is, the thickness corresponds to a frequency of one-third (sometimes one-fifth) of the normal operating frequency. The other dimensions of the crystal are proportioned so that the mechanical vibration is at three times (or five times) the fundamental frequency.

GRINDING CRYSTALS

Crystal blanks, cut to approximate frequency, are available at very reasonable prices. With proper equipment and a little care, these blanks can be ground to the desired frequency. Complete crystal-grinding equipment includes several components. First necessity is a flat piece of plate glass, about 4 inches square or larger. To hold the crystal flat while grinding a flat "button" (shown in Fig. 6-12), also of plate glass, either round or square and slightly larger than the crystal, is required. Both pieces may be obtained at glass stores. Two grades of abrasive, No. 303 emery for surface grinding and No. 600 Carborundum for edge grinding and beveling are obtainable from hardware stores or opticians' supply houses. A small paint brush is handy for moistening the abrasive and spreading it around the lapping plate. To facilitate frequent checking of frequency during the grinding process, the quick-change holder shown in Fig. 6-13 is desirable. It consists of an FT243 holder with a sliding cover

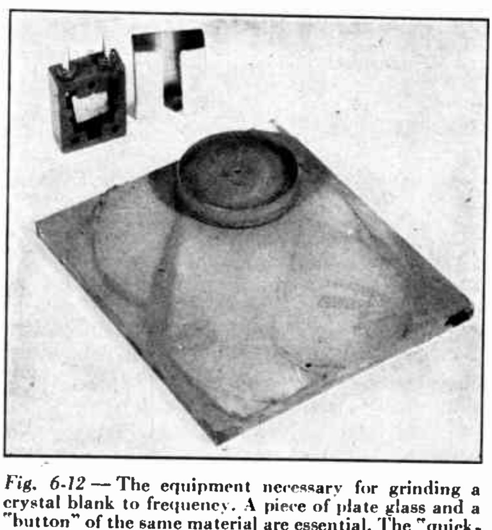


Fig. 6-12 — The equipment necessary for grinding a crystal blank to frequency. A piece of plate glass and a "button" of the same material are essential. The "quick-change" adaptation for the crystal holder is a convenience. Not shown, but also convenient, are a small paint brush for spreading abrasive and a toothbrush for scrubbing.

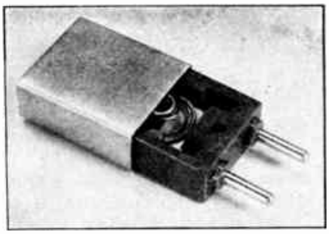


Fig. 6-13 — The quick-change crystal holder with sliding cover.

fashioned from sheet metal. Soap, warm water and a toothbrush are used to clean and rinse the crystal. Lintless cloth from an optician's or a clean towel can be used for drying.

Present-day electrodes have raised lands on each corner, as shown in Fig. 6-14, and the crystal should lie at least halfway across these lands and should not be larger than the electrode. The electrodes should be cleaned as carefully as the crystal. Before final assembly both crystal and electrodes should be handled carefully by the corners or edges after their last good scrubbing.

The actual grinding is done as follows: Spread the 303 abrasive over an area about a half inch square on the lapping plate, wet the brush, mix water into the spot and spread the abrasive over the lapping plate. Always keep the abrasive moist. Take the button, put a drop of water at its center, and press the dry crystal blank over the drop of water. There should be just enough water in the drop so that it squeezes out under the edges of the blank, where it is wiped away. Place the button, blank down, on the emery and put the index finger in the center of the button. Use just enough pressure to move the button in a figure-8 pattern. This motion is used because it helps keep the blank flat.

After grinding through ten or fifteen "8s" the blank should be rechecked for frequency and activity. The frequency change probably will be between 200 and 1000 cycles per "8," using a 7-Mc. crystal. The crystal can be moved along faster as the operator becomes more familiar with the technique, but for the beginner frequent checks of activity are in order.

To grind a crystal successfully, the activity must be good when the crystal is brought to the desired frequency. There are several ways to raise the activity. Assuming that, with careful grinding on a flat plate with a flat button, the two faces of the crystal are parallel, the major cause of low activity will be dirt or moisture on the crystal or electrodes. Before checking activity the crystal should be scrubbed carefully with the toothbrush, using warm water and soap. Wipe the crystal clean and be sure that the electrodes are clean and dry. If the activity is still down the next thing is to bevel all eight edges of the crystal. The beveling can be done with either fine or coarse abrasive, but is usually more effective with the coarse. Beveling, incidentally, will also raise the frequency because of the quartz ground off during the process.

Although beveling will usually improve the activity, another method — and probably the simplest — is to change electrodes. The land heights on the electrodes have a critical effect on activity. If the center of the crystal becomes too high and the lands are so low that the center of the crystal touches the center of the electrodes, the crystal will stop oscillating.

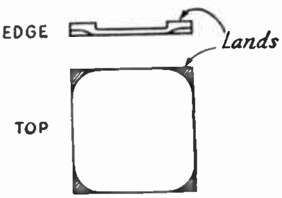


Fig. 6-14 — The $\frac{1}{2} \times \frac{1}{2}$-inch electrodes used in modern crystal holders, showing the lands at the corners between which the crystal is firmly held.

The last step — and the most drastic method of raising activity — is to edge-grind adjacent edges. This grinding is best done with coarse abrasive and should be followed by a slight bevel to remove any chips which may remain. By checking the crystal frequently, a drop in activity can be corrected by the above methods. If the crystal is ground too far and goes completely dead, the frequency may be too high when the crystal is again reactivated.

Most crystals produced in the last five years or so have been brought to the desired frequency by an etching process. This process is not only a convenient means of quantity production, but it also results in a completely clean surface for the crystal, which plays an important part in the activity of the crystal and the maximum power it will handle without overheating. Therefore, regrinding may impair the performance of etched crystals, since regrinding destroys the etched surface.

R.F. Power Amplifier Circuits

The power output from an oscillator is limited for reasons previously stated. Power greater than a few watts usually is obtained by feeding the output of the oscillator into one or more amplifiers as may be required to raise the power level to that desired before feeding it to the antenna.

Fig. 6-15 shows a fundamental amplifier circuit. The oscillator output is fed into the grid circuit of the amplifier. Power output is taken from the plate circuit. Both grid and plate circuits are tuned to the frequency of the oscillator. It will be noticed, however, that this fundamental circuit is the same as the circuit for the tuned-grid tuned-plate oscillator. Therefore the amplifier circuit itself will function as an oscillator, independent of the oscillator feeding it, unless measures are taken to reduce the plate-grid capacitance or nullify its effect.

TRIODE CIRCUITS

Plate Capacitive Neutralizing Systems

The plate-grid capacitance can be neutralized by feeding back to the grid, through an external path, a voltage which at any instant

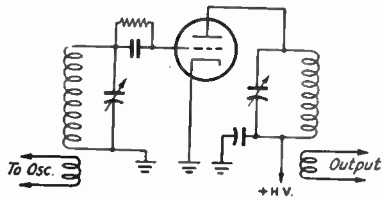


Fig. 6-15 — Fundamental r.f. power-amplifier circuit. Means must be provided to prevent oscillation since the circuit is the same as that for a t.g.t.p. oscillator. See text for discussion.

is equal, but in opposite phase, to the voltage fed back through the tube.

The most commonly-used circuits for this purpose are shown in Fig. 6-16. Amplifiers using these systems of neutralization are known as **plate-neutralized amplifiers.** In each case, the midpoint of the plate tank circuit, either coil or condenser, is grounded. Thus the voltages at opposite ends of the tank are essentially equal, but 180 degrees out of phase.

The neutralizing and feed-back voltages are matched in amplitude by adjusting the capacitance of the neutralizing condenser, C_6.

In Fig. 6-16A, the division of voltage across the tank circuit is dependent upon the ratio of the capacitances of the two sections of the tank condenser. Since these capacitances are equal in a split-stator condenser, the voltages at the ends of the tank circuit in respect to the cathode, which is connected to the center of the tank circuit through ground, are equal. Therefore the neutralizing voltage is the same as the feed-back voltage when the capacitance of the neutralizing condenser is equal to the grid-plate capacitance of the tube, including socket and other external stray capacitances across the elements.

In Fig. 6-16B, the voltage division for neutralization is dependent upon the ratio of inductances in the two sections of the coil. The coil usually is tapped at the center to give equal voltages at the ends of the tank circuit.

Push-Pull Triode Circuits

Fig. 6-16C and D show equivalent push-pull arrangements. In circuit D, better circuit balance can be maintained by using split-stator tank condensers. The rotors in this case should *not* be grounded.

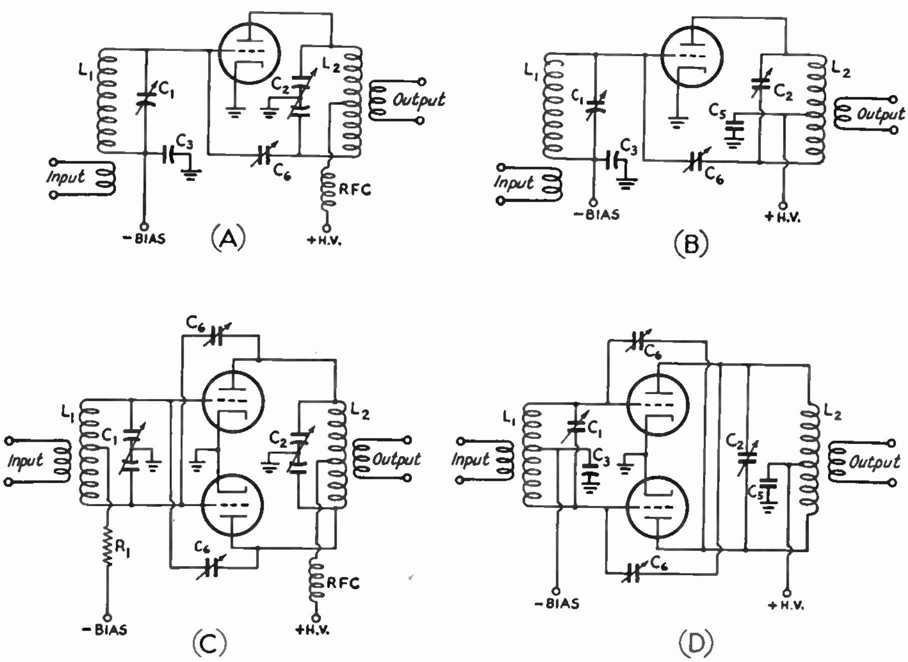


Fig. 6-16 — Neutralized-triode amplifier circuits. A — Single tube with capacitive balance. B — Single tube with inductive balance. C and D show corresponding push-pull arrangements.

C_1L_1 (grid tank) and C_2L_2 (plate tank) are tuned to the frequency fed to the amplifier.

C_3 — Grid by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.

C_5 — Plate by-pass condenser, 0.001-μfd. mica to 0.01-μfd. paper.

C_6 — Neutralizing condenser — approximately same capacitance as tube grid-plate capacitance.

R_1 — Grid-circuit isolating resistor — 100 ohms.

RFC — Plate-circuit isolating radio-frequency choke — 1 to 2.5 mh.

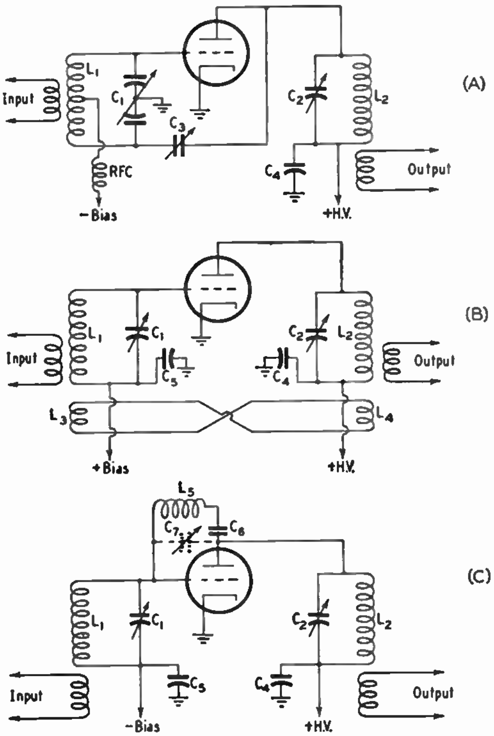


Fig. 6-17 — Additional, but less commonly-used neutralizing circuits. A — Grid neutralizing. B — Link neutralizing. C — Inductive neutralization.

C_1L_1, C_2L_2 — Tank circuits tuned to operating frequency.
C_3 — Neutralizing condenser — approximately same capacitance as grid-plate capacitance of tube.
C_4 — Plate by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
C_5 — Grid by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
C_6 — Voltage-blocking condenser — 0.001-μfd. mica.
C_7 — Variable condenser to tune trap circuit to operating frequency with L_5 and grid-plate capacitance of tube. (See text.)
L_3, L_4 — Neutralizing links — 2 to 10 turns, depending upon frequency.
L_5 — Neutralizing trap coil — to tune to operating frequency with C_7 and grid-plate capacitance of tube. (See text.)

In Fig. 6-16C, the r.f. choke in the plate circuit prevents r.f. grounding of the coil center-tap (through the power supply) and the rotor of the condenser simultaneously. This condition is to be avoided because it sets up three tuned circuits — each half of the tank circuit in addition to the circuit as a whole. The isolating resistor in the grid circuit serves a similar purpose.

Grid Capacitive Neutralizing Systems

Additional, but less widely-used neutralizing circuits are shown in Fig. 6-17. The circuit of Fig. 6-17A is similar to that of Fig. 6-16A, except that the voltage division takes place in the grid circuit instead of the plate circuit. Any voltage which may be fed back to the grid circuit through the grid-plate capacitance of the tube is divided in the grid tank circuit so that half appears at the grid, while the other half is fed, 180 degrees out of phase, back to the plate. In another similar version the grid tank coil, instead of the condenser, is used as the voltage divider, the circuit being comparable to Fig. 6-16B.

Link Neutralizing Circuit

The link neutralizing circuit of Fig. 6-17B sometimes is useful as an expedient to stabilize a screen-grid amplifier which is not sufficiently screened or shielded. It has the advantage that it may be added readily to an already-existing amplifier circuit without the necessity for the major alteration in either grid or plate circuits which would be required to shift the ground point to the center of the tank circuit. The link provides the path for coupling back the neutralizing voltage and proper phasing is dependent upon the polarization of the two link coils. Connections to one of the link coils may be switched to obtain correct polarization.

Inductive Neutralization

The inductive neutralizing arrangement of Fig. 6-17C consists merely of making the plate-grid capacitance of the tube part of a circuit tuned to the frequency at which the amplifier is designed to operate. Since such a circuit presents a high impedance to the flow of current at the frequency to which it is tuned (wavetrap), it prevents feed-back. C_7 may be

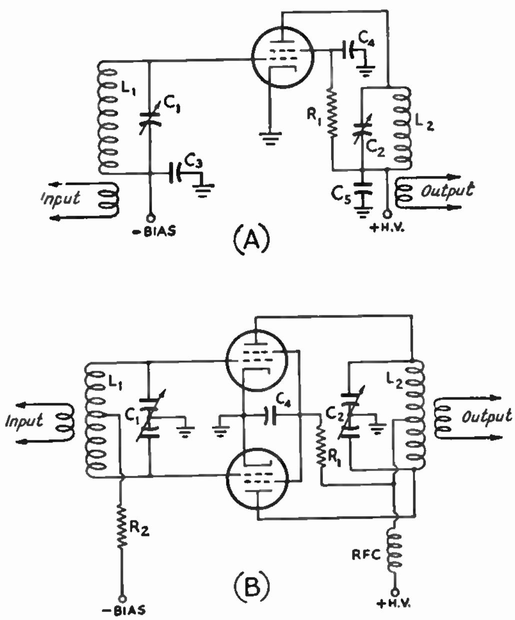


Fig. 6-18 — Screen-grid amplifier circuits. A — Single-tube amplifier. B — Push-pull.

C_4 — Screen by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
R_1 — Screen voltage-dropping resistor.
R_2 — Grid-circuit isolating resistor — 100 ohms.
Other values same as Fig. 6-16.

added for adjustment, although this may decrease the frequency range over which one neutralizing adjustment will hold.

All of the circuits of Fig. 6-17 have disadvantages in amateur practice, particularly in respect to the tuning range over which a single adjustment of neutralization will hold.

SCREEN-GRID AMPLIFIER CIRCUITS

Single-tube and push-pull screen-grid amplifier circuits are shown in Fig. 6-18. The grounded screen in transmitting-type tetrodes and pentodes serves as a shield between the plate and grid to reduce the grid-plate capacitance to the point where feed-back is insufficient to support oscillation. Thus tubes of this type are designed to be operated without neutralization when circuit simplicity is of importance. However, neutralization usually will result in more reliable stability. Poorly-screened audio tetrodes, such as the 6L6 and 6V6, invariably require neutralization.

The power sensitivity of screen-grid tubes is much higher than that of triodes of comparable power rating. Therefore greater care must be exercised in eliminating possible paths for feed-back coupling external to the tube.

Interstage Coupling Systems

Of the various systems that have been devised for feeding the output of one stage into the input of another, the inductive-link and capacitive systems are the most widely used in amateur transmitters. The link system is used principally in cases where there must be appreciable physical separation between stages, where balanced and unbalanced circuits are to be coupled, or when minimum circuit capacitance is desired. The capacitive system has the advantages of simplicity, cheapness and compactness, but it does not lend itself so readily to the conditions listed above.

INDUCTIVE COUPLING SYSTEMS

Link Coupling

The link system, examples of which are shown in Fig. 6-19, consists merely of a two-wire low-impedance line with each end terminated in a coil of a few turns coupled tightly to the low-potential point of the output tank coil of the driver and the input tank coil of the driven stage. This low-potential point occurs at the "ground" end of the tank coil in unbalanced circuits (Fig. 6-19A, B and C) and at the center of the tank coil in balanced arrangements (Fig. 6-19B, C and D).

The coupling between the two stages is largely a matter of the tank-circuit Qs but it can be adjusted within limits either by changing the number of turns in the link windings or by changing the coupling between the links and the tank coils. If increasing the number of link turns does not provide sufficient coupling, the tank-circuit Q must be increased. This system does not upset the symmetry of a balanced circuit through the introduction of unbalancing capacitances of the single-ended circuit coupled to it.

Fig. 6-19A shows the link system coupling two unbalanced circuits. This arrangement would be used, for instance, in coupling an oscillator or a screen-grid driver to the input of a single-tube stage.

The scheme at B would be suitable for coupling a neutralized or push-pull driver to a single-tube amplifier.

Fig. 6-19C shows the method applied in coupling the output of an unneutralized driver to a push-pull amplifier, while D is the circuit to be used in coupling a neutralized or push-pull stage to another push-pull input.

Inductive Coupling

Another system which is used sometimes in coupling between an unbalanced driver and a balanced amplifier is shown in Fig. 6-20. The output coil of the driver stage is designed to resonate, with the driver-tube and stray circuit capacitances, near the desired operating frequency. The amplifier input tank circuit tunes to the operating frequency and serves to a considerable degree also to tune the output circuit of the driver stage, since the two coils are coupled quite tightly. L_1 should be wound centrally over or inside L_2 and the turns of L_1 adjusted experimentally for optimum power transfer. Sometimes both circuits are tuned, in which case the coils need not be coupled so tightly.

CAPACITIVE COUPLING CIRCUITS

In a capacitive coupling system, the output tank circuit of the driver stage serves also as the input tank circuit of the driven stage. Several arrangements for coupling between balanced and unbalanced circuits, depending upon whether series or parallel power feed is desired, are shown in Fig. 6-21.

With capacitive coupling, the two stages cannot be separated physically any appreciable distance without involving loss in transferred power, radiation and the danger of instability because of feed-back which long high-impedance leads may provide. Since both the output capacitance of the driver tube and the input capacitance of the driven tube are lumped across the single tuned circuit, this sometimes makes it difficult, with the high-capacitance of screen-grid tubes, to obtain a tank circuit with a sufficient amount of inductance to provide an efficient circuit for the higher frequencies. Another disadvantage is that it is difficult to preserve circuit balance in

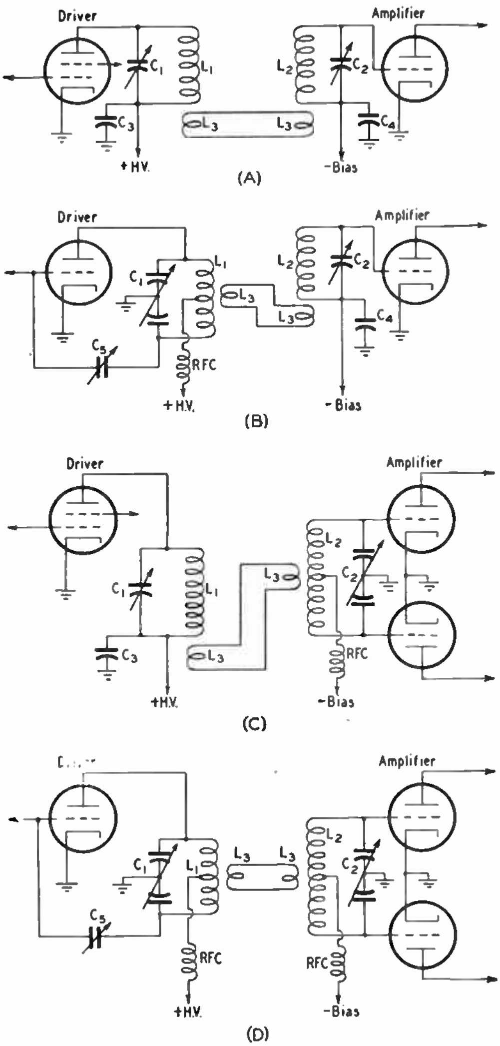


Fig. 6-19 — Link coupling circuits. A — Unbalanced output to unbalanced input. B — Balanced output to unbalanced input. C — Unbalanced output to balanced input. D — Balanced output to balanced input.

C_1 — Driver-stage plate tank condenser.
C_2 — Driven-stage grid tank condenser.
C_3 — Plate by-pass condenser.
C_4 — Grid by-pass condenser.
C_5 — Neutralizing condenser.
L_1 — Driver output tank coil.
L_2 — Driven-stage input tank coil.
L_3 — Link winding.
C_1L_1 and C_2L_2 are always tuned to the same frequency.
RFC — R.f. choke.

coupling from a single-tube stage to a push-pull stage because the circuit tends to become unbalanced by the output capacitance of the driver tube which appears across only one side of the circuit. This does not, however, preclude its use for this purpose, if simplicity in circuit is considered of greater importance, for frequencies below 30 Mc.

The arrangements of Fig. 6-21A and B are most often seen with the plate tap of A and the grid tap of B connected to the top end of the coil. A is used when series driver plate feed is desired; B when series amplifier grid feed is wanted. In the circuit of C, the tank condenser and coil are grounded directly, but parallel power feed is required for the driver plate and usually for the amplifier grid although the grid leak sometimes is placed across the coupling condenser, C_3.

An arrangement which makes possible series feed to both plate and grid is shown at D. L_1 in D is a single coil, opened at the center for feeding in plate and biasing voltages. Since the by-pass condensers, C_2, are directly in the tank circuit, they should be of good-grade mica and capable of handling the r.f. current circulating through the tank circuit. The scheme is practical chiefly in low-power stages. Because it provides a "double-ended" output circuit, it may be used in a neutralized amplifier stage simply by the addition of neutralizing condenser C_5. The grid of the driven tube and the plate of the driver tube being connected across opposite halves of the tank circuit helps to distribute stray capacitances more evenly, thereby preserving a better circuit balance. A still better balance can be achieved by using a split-stator condenser at C_1 and a single mica condenser at C_2, grounding the circuit at the split-stator rotor rather than between the two fixed condensers. Excitation may be adjusted, if necessary, by tapping the grid or plate, as may be required, down on the coil. Such a change, however, will necessitate readjustment of neutralization if the tank is used for neutralizing the driver as suggested.

The circuit of Fig. 6-21E is the preferred arrangement for coupling a neutralized driver to a single-tube amplifier in cases where series feed to the grid of the amplifier is not considered important. F shows the same system feeding a push-pull amplifier. If a more accurate balance is desired, a balancing condenser, C_6, can be used across the other half of the circuit to compensate for the driver-tube output capacitance.

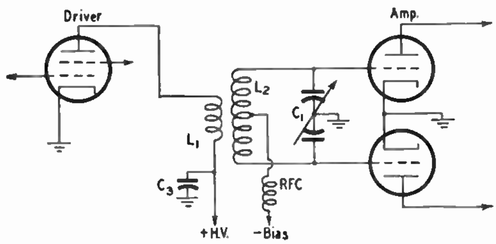


Fig. 6-20 — Inductive coupling from unbalanced output to balanced input.

C_1 — Driven-stage grid tank condenser.
C_3 — Plate by-pass condenser.
L_1 — Self-resonant (approximately) output coil.
L_2 — Driven-stage grid tank coil.
L_1 and L_2 should be coupled tightly.
RFC — R.f. choke.

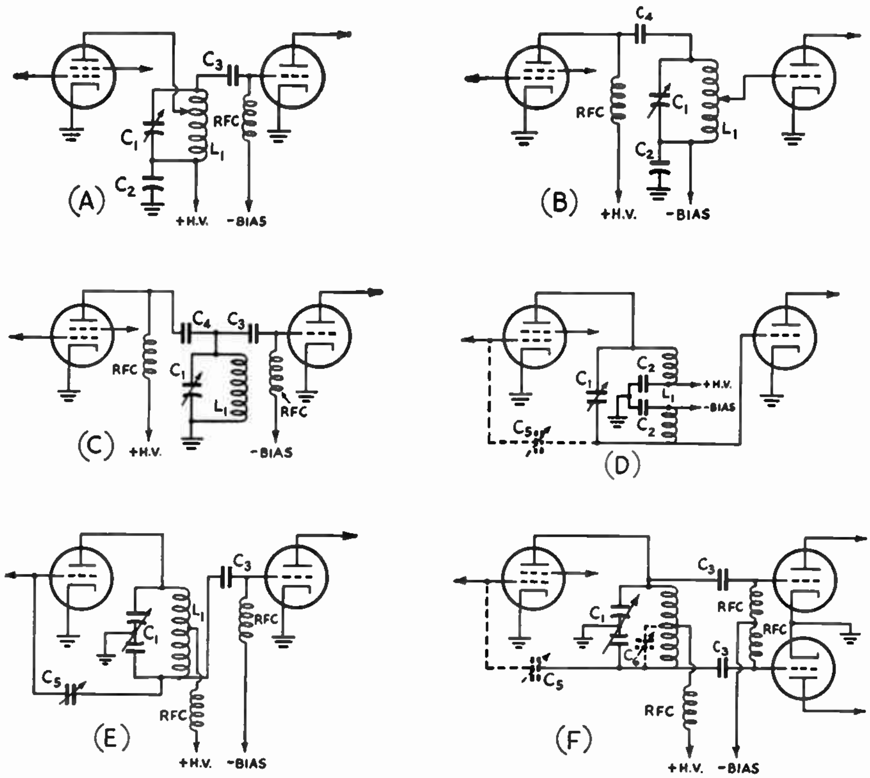


Fig. 6-21 — Examples of capacitive coupling. A — Series plate feed, parallel grid feed. B — Parallel plate feed, series grid feed. C — Parallel feed in both plate and grid. D — Series feed in both plate and grid. E — Balanced output to unbalanced input, series plate feed, parallel grid feed. F — Single tube to push-pull.

C_1 — Tank condenser.
C_2 — By-pass condenser.
C_3 — Coupling condenser.
C_4 — Driver plate blocking condenser.
C_5 — Driver neutralizing condenser.
C_6 — Circuit-balancing condenser.
L_1 — Tank coil.
RFC — R.f. choke.

Capacitive-Coupling Adjustment

Overcoupling can be remedied by reducing the capacitance of the coupling condenser, or by tapping the grid of the driven tube across only a portion of the tank coil, as indicated in Fig. 6-21B. If increasing the capacitance of the coupling condenser does not provide sufficient coupling, the tank-circuit Q must be increased. This can be done by decreasing the L/C ratio of the tank circuit or by tapping the plate of the driver across a portion of the tank coil, as shown in Fig. 6-21A. However, it is preferable and often possible to choose a tank-circuit L/C ratio that will give the desired coupling with both grid and plate connected to the end of the tank circuit.

Coupling Condensers

Coupling condensers should be of the mica type with a voltage rating above the sum of the driver plate and amplifier biasing voltages.

Amplifier Design Considerations

PLATE-CIRCUIT VALUES

Tank-Circuit Q

Power cannot be readily coupled out of a plate tank circuit if the Q is too low. Also, harmonics are more readily coupled out of a tank circuit whose Q is low. On the other hand, a large C/L ratio causes high circulating current in the tank circuit, increasing the losses. Unless one of these factors is considered to be of greater importance than the other, a compromise Q value of 12 usually is selected.

With the conditions under which r.f. power amplifiers in amateur transmitters usually are operated, the L/C ratio for the same Q varies in proportion to the ratio of d.c. plate voltage to plate current with the amplifier in operation and loaded. The chart of Fig. 6-22 shows recommended values of tank capacitance for a Q of 12 for a wide range of plate-voltage/plate-current ratios for each of the low-frequency amateur bands. The values given apply to the type of plate tank circuits shown in Fig. 6-23A and B only. Because the tube is connected across only half of the tank in the remainder of the circuits shown in Fig. 6-23, the total capacitance across the tank coil may be reduced to one-quarter that shown by the graph for the same plate-voltage/plate-current ratio. This means that in circuits in which a split-stator condenser is used, the capacitance of each section of the condenser may be half the value shown in the graph, since the two sections are in series across the coil.

The values shown in Fig. 6-22 are the capacitances which should be in actual use when the circuit is tuned to resonance in the selected band — not the maximum rated capacitance of the tank condenser — including tube and circuit capacitances. They should be considered minimum values for satisfactory operation. They can be exceeded 50 to 100 per cent without involving an appreciable loss in circuit efficiency. The Q can be increased also by tapping the plate down on the tank coil, although this sometimes results in setting up a parasitic oscillatory circuit.

Plate Tank-Condenser Voltage

In selecting a tank condenser with a spacing between plates sufficient to prevent voltage breakdown, the peak r.f. voltage across a tank circuit under load, but without modulation, may be taken conservatively as equal to the d.c. plate voltage. If the d.c. plate voltage also appears across the tank condenser, this must be added to the peak r.f. voltage, making the total peak voltage twice the d.c. plate voltage. If the amplifier is to be plate-modulated, this last value must be doubled to make it four times the d.c. plate voltage, because both d.c. and r.f. voltages double with 100-per-cent plate modulation. At the higher plate voltages, it is desirable to choose a tank circuit in which the d.c. and modulation voltages do not appear across the tank condenser, to permit the use of a smaller condenser with less plate spacing. Fig. 6-23 shows the peak voltage, in terms of d.c. plate voltage, to be expected across the tank condenser in various circuit arrangements. These peak-voltage values are given assuming that the amplifier is loaded to rated plate current. Without load, the peak r.f. voltage will run much higher. Since a c.w. transmitter may be operated without load while adjustments are being made, although a modulated amplifier never should be operated without load, it is sometimes considered logical to select a condenser for a c.w. transmitter with a peak-voltage rating equal to that required for a 'phone transmitter of the same power. However, if minimum cost and space are considerations, a condenser with half the spacing required for 'phone operation can be used in a c.w. transmitter for the same carrier output, as indicated under Fig. 6-23, if power is reduced temporarily while tuning up without load.

In the circuits of Fig. 6-23C, D and E, the rotors are deliberately connected to the positive side of the high-voltage supply, eliminating any difference in d.c. potential between the rotors and stators.

The plate spacing to be used for a given peak voltage will depend upon the design of the variable condenser, influencing factors being the mechanical construction of the unit, the dielectric used and its placement in respect to intense fields, and the condenser-plate shape and degree of polish. Condenser manufacturers usually rate their products in terms of the peak voltage between plates.

Plate Tank Coils

The inductance of a manufactured coil usually is based upon the highest plate-voltage/plate-current ratio likely to be used at the maximum power level for which the coil is designed, following the logical conclusion that it is easier to cut off turns than to add them. Therefore in the majority of cases, the capacitance shown by Fig. 6-22 will be greater than

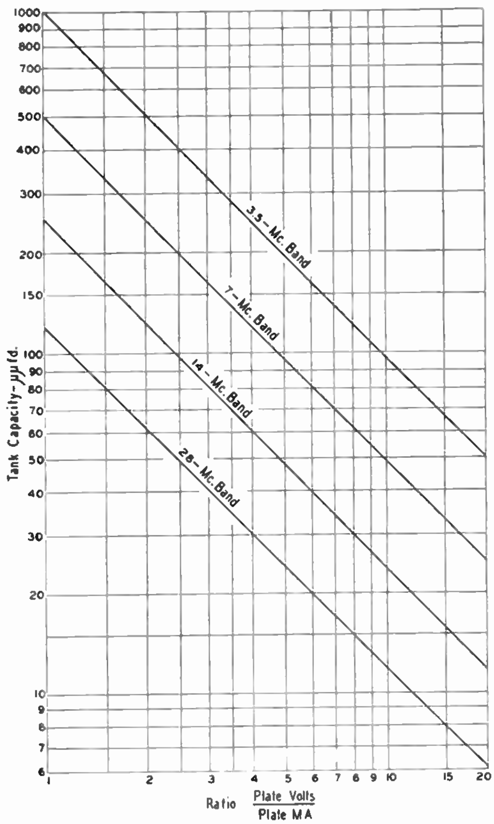
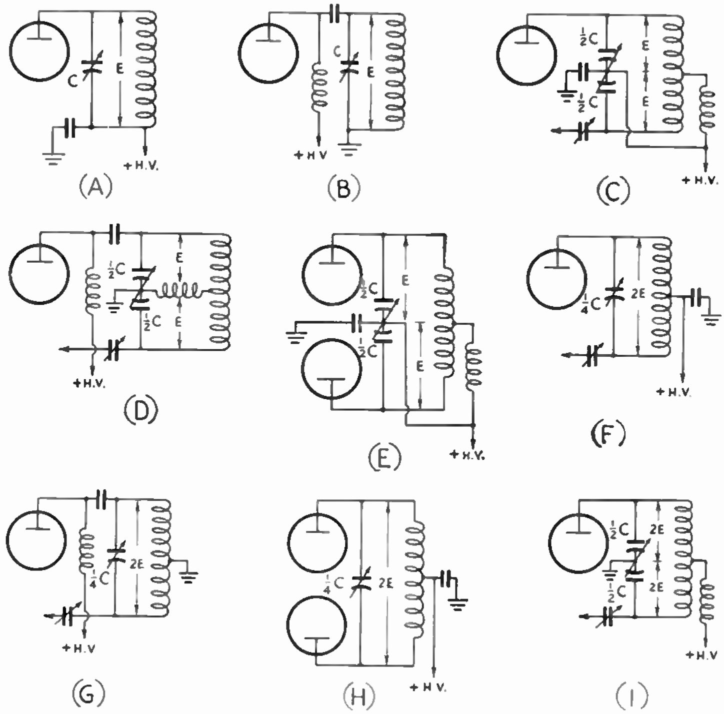


Fig. 6-22 — Chart showing minimum plate tank capacitances recommended with various ratios of plate voltage to plate current, for the four low-frequency amateur bands. In the circuits F, G and H of Fig. 6-23, the values shown by the graph may be divided by four. In circuits C, D, E, I, J and K, the capacitance of each section of the split-stator condenser may be one-half the value shown by the graph. The full graph values should be used for circuits A and B. These values are based on a circuit Q of 12.



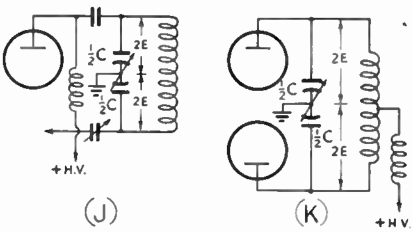


Fig. 6-23 — Diagrams showing the peak voltage for which the plate tank condenser should be rated for c.w. operation with various circuit arrangements. *E* is equal to the d.c. plate voltage. The values should be doubled for plate modulation. The circuit is assumed to be fully loaded. Circuits A, C, E, F and H require that the tank condenser be insulated from chassis or ground.

that for which the coil is designed and turns must be removed to permit the use of the proper value of capacitance. At 28 Mc., and sometimes 14 Mc., the value of capacitance shown by the chart for a high plate-voltage/plate-current ratio will be lower than that attainable in practice with the components available. The design of manufactured coils usually takes this into consideration also and it may be found that values of capacitance greater than those shown in the graph (if stray capacitance is included) are required to tune these coils to the band.

Manufactured coils are rated according to the plate power input to the tube or tubes when the stage is loaded. Since the circulating tank current is much greater when the amplifier is unloaded, care should be taken to operate the amplifier conservatively when unloaded to prevent damage to the coil as a result of excessive heating.

Plate-Blocking and By-Pass Condensers

Plate-blocking condensers should have low inductance; therefore condensers of the mica type are preferred. The capacitance should be large enough to have low reactance at the lowest operating frequency. For frequencies between 3.5 and 30 Mc., a capacitance of 0.001 μfd. is commonly used. The voltage rating should be 25 to 50 per cent above the plate-supply voltage.

By-pass condensers also should have low reactance at the operating frequency. Paper condensers with a capacitance of 0.01 μfd. are satisfactory for supply voltages up to 500 or 600 at frequencies up to at least 7 Mc. Mica condensers, usually 0.001 μfd., are preferable at the higher frequencies and greater plate voltages.

Voltage ratings should be doubled in the case of plate modulation.

R.F. Chokes

Parallel plate feed provides a considerable measure of protection against serious injury to the operator from accidental contact with high-voltage d.c. in the tank circuit. However, the r.f. choke in this case is called upon to present a high impedance at the operating frequency if serious loss of power in the choke is to be avoided. In the design of manufactured r.f. chokes, an attempt is made to make the choke universally satisfactory for several amateur bands. However, when the transmitter is designed to operate on all amateur bands from 28 Mc. to 3.5 Mc., loss in r.f. chokes often occurs on one or more of the bands. There is no simple remedy for this difficulty aside from a shift to series plate feed which, of course, nullifies the safety angle. One possible remedy is the use of different chokes for each band, the chokes being plugged in with the tank coil.

For frequencies between 3.5 and 30 Mc., 2.5-mh. chokes are used where the plate current is 125 ma. or less, and 1 mh. when the plate current is above 125 ma. In the circuit of Fig. 6-23D, the choke does not carry any current, so a low-current choke may be used, regardless of the power. In series-fed circuits in which the choke is used to isolate the coil center-tap from ground, the value of the choke inductance is not critical.

GRID TANK CIRCUITS

The value of capacitance to be used in a grid tank circuit when employing link coupling is not critical so long as the *Q* is high enough to permit satisfactory coupling to the driver stage. A capacitance of 200 μμfd. should be sufficient in most cases for unbalanced grid tank circuits tuned to 3.5 Mc., with the value decreased in proportion as the frequency increases, as given under Fig. 6-24. For unbalanced grid tank circuits, the total condenser capacitance may be cut in half, making the capacitance of each section of a split-stator condenser the same as that of the single condenser used in an unbalanced input grid tank circuit.

The *Q* can also be increased by tapping the grid down on the input coil, at some risk, however, of setting up a parasitic circuit.

Approximate tank-condenser voltage ratings are suggested under Fig. 6-24. Tank coils with a power rating equal to that of the driver plate tank coil should be used in the grid tank circuit.

The resistor *R* in Fig. 6-24C and D is recommended in place of the r.f. choke customarily used in the same position, to eliminate the possibility of forming a low-frequency parasitic t.g.t.p. oscillator in conjunction with the r.f. choke usually used similarly in the plate circuit. A resistance of 100 ohms will be sufficient in most cases. If a grid leak is used, the 100-ohm resistor will not be necessary.

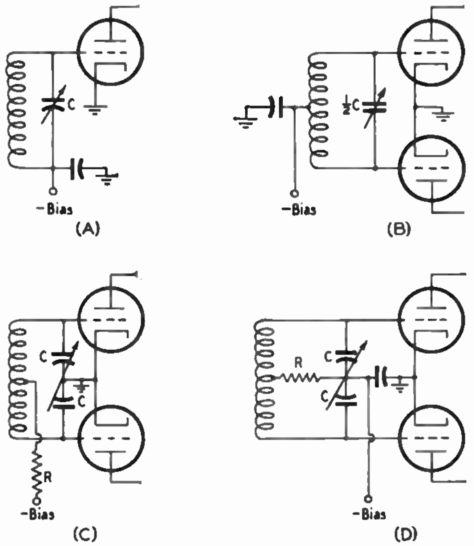


Fig. 6-24 — Diagrams for determining grid tank-condenser capacitance. C should be approximately 200 μμfd. for 3.5 Mc., 100 μμfd. for 7 Mc., 50 μμfd. for 14 Mc. and 25 μμfd. for 28 Mc.

The tank condenser should have a voltage rating approximately equal to the operating bias voltage plus 20 per cent of the plate voltage for circuit A, twice this value for circuit B and each section of the condenser in circuit D, while the biasing voltage must be added to this latter figure in determining the voltage rating of each section of the condenser in circuit C. *R* is an isolating resistor of 100 ohms.

R. F. Power-Amplifier-Tube Operating Factors

R.f. amplifiers in c.w. transmitters invariably are operated as Class C amplifiers. Transmitting-tube instruction sheets and data tables specify the limitations on various electrode voltages and currents which should be observed to insure normal tube life. Included also are sets of recommended operating conditions which may be followed to obtain rated output with good efficiency consistent with reasonable driving power.

GRID-CIRCUIT RATINGS

Grid Bias

Two values of grid-biasing voltage are of interest in the practical operation of r.f. power amplifiers. These are **protective bias** and **operating bias.**

When plate (and screen) voltage is applied, most tubes will draw appreciable plate current in the absence of any grid bias. Therefore protective bias must be used with all but "zero-bias"-type tubes to hold the power input to the tube below the rated dissipation value when excitation is removed without removing plate (and screen) voltage. Without excitation, the amplifier delivers no power. Therefore any power input is dissipated in heat which would ruin the tube in a short length of time. This condition exists when the transmitter is keyed ahead of the amplifier, while tuning adjustments are being made, or through failure of a crystal oscillator to function or other accidental failures.

Operating bias is the value of biasing voltage between grid and cathode when the amplifier is being driven and delivering power. The optimum value of biasing voltage for operating under a given set of conditions is listed in tube tables and manuals, and with triodes is normally two to three times the **cut-off bias** value — the value necessary to reduce the plate current to zero with plate voltage applied.

Protective bias may be any value between that which limits the input to the tube to its rated plate (and screen) dissipation as a minimum, and the operating value as a maximum. It is common practice, however, to set the value at some point between that which is necessary to cut off plate current completely and the operating value. With fixed plate voltage, the cut-off value for a triode can be determined quite closely by dividing the plate voltage by the amplification factor obtained from the tube data sheet. For screen-grid tubes,

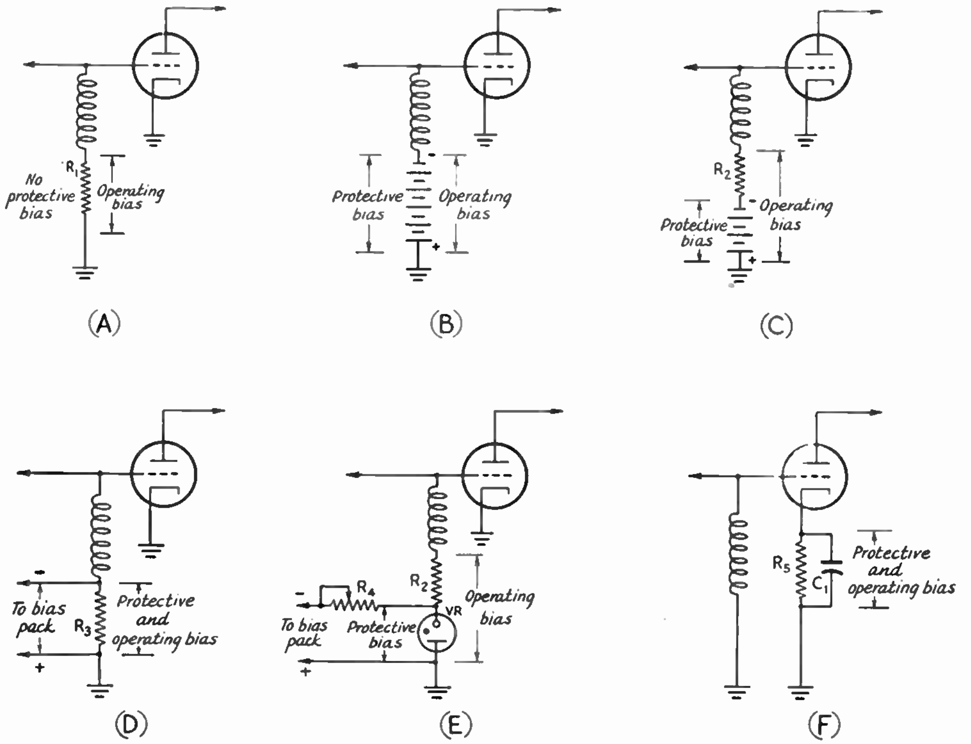


Fig. 6-25 — Various systems for obtaining protective and operating bias for r.f. amplifiers. A — Grid-leak. B — Battery. C — Combination battery and grid leak. D — Grid leak and adjusted-voltage bias pack. E — Combination grid leak and voltage-regulated pack. F — Cathode bias.

the amplification factor and voltage of the screen must be used instead. In cases where this is not included in the operating data, the approximate cut-off value may be obtained from an inspection of the plate-current plate-voltage curves which show the plate current for a wide range of plate, screen and biasing voltages.

A saving in the operation of a c.w. amplifier sometimes can be effected by adjusting the *protective* bias so that the tube (or tubes if more are operated from the same supply) draws the same current as the required bleeder resistance for the power supply (see Chapter Seven), if this can be done without exceeding the dissipation rating of the tube. This saves the cost of the bleeder resistor and some of the power it wastes and also improves the regulation, since the difference between minimum and maximum load as the amplifier is keyed is less.

A factor which must be considered in determining the value of bias which will protect the tube is plate- (and screen-) voltage regulation. If the power-supply regulation is poor, or if the plate or screen is fed from a resistance voltage divider or a voltage-dropping resistor, the electrode voltages will soar as the tube draws less than normal operating current and therefore an increase over the calculated value of cut-off bias will be required to bring the current to zero. This condition is encountered most often in the operation of a screen-grid tube where the screen is not fed from a fixed-voltage source. In such cases, care should be taken to make certain that the proper operating bias is not exceeded when excitation is applied.

Several different systems for obtaining bias are shown in Fig. 6-25. At A, bias is obtained entirely from the voltage drop across the grid leak, R_1, caused by the flow of rectified grid current when the amplifier is being driven. This system has the desirable feature that the biasing voltage, being dependent upon the value of grid current, is kept adjusted close to proper operating value automatically over a considerable range of excitation levels. However, when excitation is removed, grid-current flow ceases and the voltage across R_1 falls to zero and there is no bias. Therefore this system provides no protection for the amplifier tube in case excitation fails or is removed.

A battery delivering the required operating bias is used in the arrangement of Fig. 6-25B. Since the biasing voltage still remains when excitation is removed, plate-current flow ceases and the tube is protected. A factor which must be taken into consideration when dry batteries, such as "B" batteries, are used, is the resistance of the batteries. If the internal resistance is high, the resistance will cause an increase.

by grid-leak action, in the operating bias above that normally delivered by the batteries. Batteries develop internal resistance with age and should be replaced from time to time. Another factor is that the direction of grid-current flow is such as to reverse the normal direction of current through the battery. This acts to charge the battery. A battery which has been in use for some time, particularly if the grid current under excitation is high, will show a considerably higher-than-rated terminal voltage because of the charging action of the grid current. The terminal voltage of a battery used in transmitter bias service where grid current flows cannot be used as an indication of the condition of the battery. Its internal resistance may be high, even though it shows normal or above-normal terminal voltage. If the grid current in a battery-biased stage falls off after a period of operation and no other reason is obvious, it is probable that the biasing battery should be replaced. The battery life which may be expected in bias service with a given value of grid current will be approximately the same as it would be if that same current were being drawn from the battery.

In Fig. 6-25C, the battery voltage is reduced to the protective value. When excitation is applied, grid-leak action through R_2 supplies the additional biasing voltage necessary to bring the total up to the operating value. This combination of fixed and grid-leak bias is the most popular system, since it combines the safety of protective fixed bias and a measure of automatic adjustment of the operating value through grid-leak action.

In Fig. 6-25D, a power pack is used to supply protective bias. The output of the power pack is connected across the grid resistor which is of the normal grid-leak value for the tube. The peak voltage output of the transformer used in the power pack must not exceed the operating-bias value. A bleeder resistance cannot be used across the output of the pack, nor can the output voltage be reduced by means of a voltage divider or series dropping resistor without affecting the biasing voltage when excitation is applied.

These restrictions on the use of a power pack can be avoided by the addition of a voltage-regulator tube across the output of the pack, as shown in Fig. 6-25E. The voltage across the regulator tube remains constant with or without grid current flowing. By making the voltage-regulator series resistor, R_4, of proper value, the output voltage of the pack may be anything within reason above a minimum of approximately twice the voltage rating of the VR tube. These tubes are available for 75, 90, 105 and 150 volts and each tube will handle up to 30 or 40 ma. of grid current. VR tubes may be used in series to obtain regulated voltages above 150, and in parallel for grid currents above 40 ma. It is usual practice to use a VR tube, or combination of VR tubes in series or series-parallel, with the minimum voltage rating which will give plate-current cut-off, and obtain the additional voltage required to bring the total bias up to the operating value by grid-leak action when excitation is applied, as with battery bias in Fig. 6-25C. The use of VR tubes for this purpose is discussed more fully in Chapter Seven.

A single source of fixed biasing voltage, such as batteries or VR tubes in series, may be used to provide protective bias for more than one amplifier stage, tapping the batteries or connecting to the junction of the tubes in the VR series if lower biasing voltages are required for other stages. In this case, the current flowing through the fixed-bias source is the sum of the grid currents of the individual stages obtaining bias from the source.

In Fig. 6-25F, bias is obtained from the voltage drop across a resistor in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across R_5 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cut-off protective bias cannot be obtained by this system. When excitation is applied, plate (and screen) current increases and the grid current also contributes to the drop across R_5, thereby increasing the bias to the operating value. Since the voltage between plate and cathode is reduced by the amount of the voltage drop across R_5, the over-all supply voltage must be the sum of the plate and operating-bias voltages.

The resistance of R_5 should be adjusted to the value which will give the correct operating bias with rated grid, plate and screen currents flowing with the amplifier loaded to rated input. When excitation is removed, the input to most types of tubes will fall to a value that will prevent damage to the tube, at least for the period of time required to remove plate voltage.

Calculating Bias-Resistor Values

The calculation of the required grid-leak and cathode biasing-resistor values is not difficult. For simple grid-leak bias, as shown in Fig. 6-25A, the resistance is obtained by dividing the required operating-bias voltage by the rated grid current.

Example: Required operating bias = 100 volts.
Rated grid current = 20 ma. = 0.02 amp.

$$\text{Grid-leak resistance} = \frac{100}{0.02} = 5000 \text{ ohms.}$$

If a combination of grid-leak and fixed protective bias is used, the amount of protective bias should be subtracted from the required operating-bias voltage before the calculation is made (except in the case of the arrangement of Fig. 6-25D).

Example: Required operating bias = 150 volts.
Protective bias from battery or VR tube = 90 volts.
150—90 = 60 volts = required bias from grid leak.
Rated grid current = 10 ma. = 0.01 amp.

$$\text{Grid-leak resistance} = \frac{60}{0.01} = 6000 \text{ ohms.}$$

In the case of a cathode biasing resistor, the rated grid, screen and plate currents under load are added together. The required operating voltage is then divided by this total current to obtain the resistance.

Example: Rated grid current = 15 ma. = 0.015 amp.
Rated screen current = 20 ma. = 0.02 amp.
Rated plate current = 200 ma. = 0.2 amp.
Total rated cathode current = 235 ma. = 0.235 amp.
Required operating bias = 150 volts.

$$\text{Cathode resistance} = \frac{150}{0.235} = 638 \text{ ohms.}$$

For two tubes in parallel or push-pull that use a single common resistor in examples similar to those above, the calculated value of resistance should be cut in half.

The power rating of the resistor may be determined from Ohm's Law:

$$P = I^2R$$

Example: In the first example above for grid-leak resistance,
I = 20 ma. = 0.02 amp. I^2 = 0.0004
R = 5000 ohms.
P = (0.0004) (5000) = 2 watts.

Example: In the above example for cathode resistor,
I = 235 ma. = 0.235 amp. I^2 = 0.055
R = 638
P = (0.055) (638) = 35.1 watts.

Maximum Grid Current

When a Class C amplifier is properly excited, and the grid is driven positive over part of the cycle, rectification takes place as it does in a diode. The rectified grid current flows between grid and cathode within the tube and thence through the external d.c. circuit which must always be provided, connecting grid and cathode. This external circuit includes the bias source (grid leak or voltage source) and either the grid r.f. choke with parallel feed, or the tank coil in series-feed arrangements. The flow of rectified current causes heating of the grid. As with the plate, there is a limit to the heat which the grid can dissipate safely. This limit is expressed in terms of maximum d.c. grid current which should not be exceeded in regular operation of the amplifier. Efficient operation usually can be attained with grid current below the maximum rated value.

The rated total grid current of two tubes in parallel or push-pull is twice that of a single tube of the same type.

Excitation

Excitation, or driving power, is the r.f. power fed to the grid of the amplifier by a preceding oscillator or amplifier. For efficient operation, a triode amplifier requires a driver capable of delivering 15 to 20 per cent as much power as the rated output of the amplifier. Screen-grid tubes require much less — usually from 5 to 10 per cent of their rated power output. To cover tank-circuit and coupling losses, a driver capable of supplying several times the driving power listed in the tube data should be used.

Two tubes in parallel or push-pull require twice the driving power of a single tube of the same type under similar conditions.

PLATE-CIRCUIT RATINGS

Power Output

The figure for power output given in the tube data is the r.f. power that the tube can be expected to deliver to the tank circuit (not the power output from the tank which is somewhat lower) under the conditions specified, at the fundamental frequency.

Power Input

Power input for both triodes and screen-grid tubes is the d.c. power input to the plate circuit. It is the product of the d.c. plate voltage and plate current.

Example: Plate voltage = 1250 volts.
Plate current = 150 ma. = 0.15 amp.
Power input = (1250) (0.15) = 187.5 watts.

Plate and Screen Dissipation

All of the d.c. power fed to the plate circuit of an amplifier is not converted into r.f. power. Part of it is wasted in heat within the tube. There is a limit to the amount of power that a tube can dissipate in the form of heat without danger of damage to the tube. This is the maximum rated plate dissipation given in tube data. The power dissipated is the difference between the d.c. power input and the r.f. power output.

Since the d.c. power furnished to the screen of a pentode or tetrode does not contribute to the r.f. output, it is entirely dissipated in heating the screen, and the maximum-input rating should be carefully observed.

Plate Efficiency

The efficiency of an amplifier is the ratio of r.f. power output to the d.c. power input.

Example: D.c. power input = 175 watts.
R.f. power output = 125 watts.
Dissipation = 175—125 = 50 watts.

$$\text{Efficiency} = \frac{125}{175} = 0.714 = 71.4 \text{ per cent.}$$

The plate efficiency at which an r.f. power amplifier can be operated depends chiefly upon

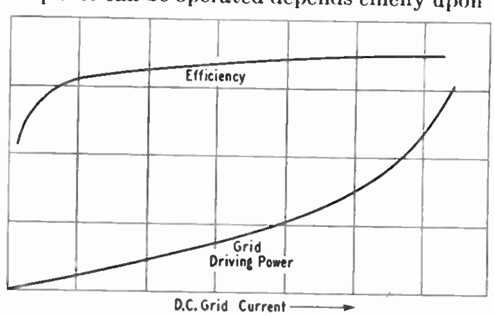


Fig. 6-26 — Curve showing relation between driving power and plate-circuit efficiency of an r.f. power-amplifier stage.

the relative driving power delivered to the input circuit. Fig. 6-26 shows that the driving power must be increased considerably out of proportion to the increase in efficiency at the higher efficiencies. An efficiency of 65 to 75 per cent represents a satisfactory balance between power output and driving power.

Maximum Plate Current and Voltage

All voltage figures given in tube data, unless otherwise specified, refer to the voltage between the electrode mentioned and cathode, or filament center-tap. Included are figures for maximum rated plate voltage and plate current. These are the respective maximum values that should be used under any circumstances. Neither should be exceeded to compensate for a lower-than-maximum value of the other in attempting to bring the power input up to permissible level. These maximum values should not be used simultaneously unless it is possible to do so without exceeding the rated plate dissipation.

OTHER OPERATING FACTORS

Filament Voltage

The filament voltage for the indirectly-heated cathode-type tubes found in low-power classifications may vary 10 per cent above or below rating without seriously reducing the life of the tube. But the voltage of the higher-power filament-type tubes should be held closely between the rated voltage as a minimum and 5 per cent above rating as a maximum. Care should be taken to make sure that the plate power drawn from the power line does not cause a drop in filament voltage below the proper value when plate power is applied. When the filament transformer is found not to deliver the required filament voltage, the voltage may be adjusted by means of a resistor in series with the transformer primary if the transformer voltage is too high, or by one of the line-voltage adjusting schemes described in Chapter Seven that either boosts the voltage or reduces it as necessary.

Thoriated-type filaments lose emission when the tube is overloaded appreciably. If the overload has not been too prolonged, emission sometimes may be restored by operating the filament at rated voltage with all other voltages removed for a period of 10 minutes, or at 20 per cent above rated voltage for a few minutes.

Interelectrode Capacitances

The value given in tube data for grid-plate capacitance is useful in determining the value of capacitance necessary to neutralize a triode. The input- and output-capacity values are helpful in arriving at a figure of minimum circuit capacitance, particularly where capacitive coupling is used.

Adjustment of R. F. Amplifiers

GENERAL TUNING PROCEDURE

Metering

Sets of typical operating conditions for r.f. amplifiers are given in all tube-data sheets and these should be followed closely for maximum output with a good balance between efficiency and required driving power. In amateur service, ICAS (intermittent commercial-amateur service) ratings may be used when this set of ratings is given. When the available plate voltage falls between values given in the data, satisfactory performance may be obtained by using intermediate values for the other voltages and currents listed.

Fig. 6-27 shows the connections for a voltmeter and milliammeter to obtain desired readings. While cathode metering often is used for reasons of safety to the operator and meter insulation, it is frequently difficult to interpret readings that are the resultant of three currents, one of which may be falling while the other two are increasing. Fig. 6-28 shows a commonly-used system for switching a single meter to read current in any of several different circuits. The resistors, R, are connected in the various circuits in place of the milliammeters shown in Fig. 6-27. Since the resistance of R is several times the internal resistance of the milliammeter, it will have no practical effect upon the reading of the meter itself.

When the meter must read currents of widely differing values, a meter with a range sufficiently low to accommodate the lowest values of current to be measured may be selected. In the circuits in which the current will be above the scale of the meter, the resistance of R can be adjusted to a lower value which will give the meter reading a multiplying factor. (See Chapter Sixteen.) Care should be taken to observe proper polarity in making the connections between the resistors and the switch.

Input-Circuit Adjustment

In setting up an r.f. power amplifier for operation, the necessary provisions for grid bias should be made first. ("R.F. Power-Amplifier Tube Operating Factors," this chapter.) The output of the driver (the oscillator and whatever intermediate amplifier stages there may be) should have been checked previously and found to be adequate. The amplifier biasing system should be connected, and if it includes a fixed protective supply, this should be turned on. No plate or screen voltage should be applied to the amplifier, however.

In general, with capacitive coupling, an amplifier grid-current reading should be obtained when the driver is coupled to the amplifier and tuned to resonance. If the driver is

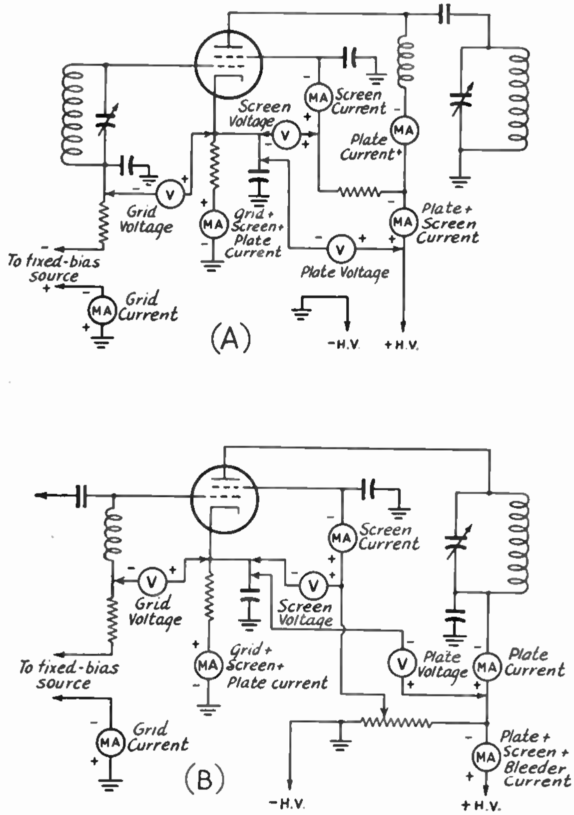


Fig. 6-27 — **Diagrams showing placement of voltmeter and milliammeter to obtain desired measurements. A — Series grid feed, parallel plate feed and series screen voltage-dropping resistor. B — Parallel grid feed, series plate feed and screen voltage divider.**

a simple VFO or a crystal oscillator of the Pierce type, with no separate tuned output-circuit tank, the operation is merely one of adjusting the coupling to the amplifier until rated amplifier grid current, or the maximum consistent with satisfactory oscillator stability, is obtained. If link coupling is used, the grid tank circuit must also be tuned to resonance as indicated by the peak in grid current.

With all capacitive-coupled drivers having a tuned output tank, maximum amplifier grid current should occur at or very close to the point where the driver plate current dips to a minimum. With link coupling, the amplifier grid tank condenser should first be set at minimum or maximum, whichever is judged to be farthest from resonance. The driver output circuit should then be tuned for minimum plate current. Then the grid tank condenser should be swung for maximum grid current. As a final tuning adjustment, the driver plate tank circuit should be retuned to make sure that it is at the minimum point of its plate-current dip. As the coupling is increased, the driver plate-current dip will become less pronounced and may almost disappear altogether if the coupling is increased sufficiently. This, however, usually is an indication of driver overload. Maximum driver output (maximum amplifier grid-current reading) usually will be obtained with the coupling adjusted to the point where there is still a fair amount of dip in plate current. The dip is likely to be less with a fully-loaded screen-grid tube than with a triode. Each time an adjustment in coupling is made, the above tuning process should be repeated.

Proper excitation to an amplifier is indicated when the recommended grid current is obtained simultaneously with recommended grid bias, with the amplifier operating and fully loaded. But here, for preliminary tuning, any grid-current reading approximating the recommended value will suffice.

Output-Circuit Adjustment

At this point, the driver should be turned off and the amplifier checked for parasitic oscillation. (See "Parasitic Oscillations," this chapter.)

The next step in the adjustment of a triode amplifier is that of neutralization. (See "Neutralizing Procedure," this chapter.)

After the amplifier stage has been stabilized, the output circuit may be adjusted. With normal bias and excitation applied again, reduced plate voltage can now be turned on and the plate tank circuit resonated.

Resonance in the plate circuit of an r.f. power amplifier is accompanied by a dip in plate current similar to that shown in Fig. 6-11. This dip is caused by the increase in tank impedance in the plate circuit when the tank is tuned to resonance. When the tank is not at resonance, the plate-circuit impedance is low and therefore the plate current is high. An external load coupled to the tank circuit lowers the impedance and therefore the plate current at resonance increases.

If no other means is available for reducing plate voltage, a 115-volt lamp of 100- to 150-watt size may be connected in series with the primary of the plate transformer, provided it is separate from the transformer supplying filaments. A dummy load (see "Checking Power Output," this chapter) should now be coupled to the output tank circuit and the tank retuned to resonance. The minimum plate current at the dip at resonance should be higher after the load is connected and the dummy load should show an indication of output. Full plate voltage may now be applied and the plate tuning checked carefully for the dip at resonance. When testing at full plate voltage, care should be taken not to operate the amplifier off resonance longer than necessary, because the dissipation runs high and the tube may be damaged.

If the plate current at full voltage is not up to the rated value, the coupling to the load should be increased until the plate current at resonance is the rated value. Under no circumstances should the plate circuit be detuned from resonance to obtain the desired increase in plate current, since this results in a decrease in power output and an increase in dissipation. If the plate current exceeds the rated value at resonance, the coupling to the load should be reduced.

Final Adjustment

The grid current and biasing voltage now should be checked while the amplifier is in operation under load. In a properly-neutralized triode amplifier, the grid current normally will fall off when plate voltage and load are applied. If it does not, it is an indication of regeneration and the amplifier should be checked for feed-back, either through the tube because of incomplete neutralization, or through paths external to the tube.

If the grid current falls below the recommended value when plate voltage and load are applied, the biasing voltage should be checked. If this is found to be above the recommended value, it should be decreased. This decrease in bias should serve to increase the grid current. If the grid current is still too low, or if the biasing voltage also checks low, the excitation must be increased by tightening the coupling to the driver or raising its plate voltage if either or both can be done without exceeding the driver-tube rating.

If the increase in excitation causes an increase in plate current to above the rated value, the coupling to the load should be reduced. The amplifier is correctly adjusted when all of the recommended values are obtained simultaneously.

SPECIAL ADJUSTMENT OF PUSH-PULL AMPLIFIERS

Proper push-pull operation requires an accurate balance between the two sides of the circuit. Otherwise the dissipation will not be distributed evenly between the two tubes, one being overloaded if an attempt is made to operate the amplifier at full rating. The construction and adjustment should be as symmetrical as possible. Particular attention should be paid to lead lengths and the placement of tank-circuit components in respect to grounded metal surfaces. Both ends of a split-stator condenser not mounted directly on a metal chassis should be by-passed to the chassis if the rotor has no center connection.

Serious unbalance is indicated when the plate of one tube shows more color than the other and exchanging tubes does not shift the apparent overload. Balance can be checked most easily in screen-grid amplifiers by comparing the individual screen currents. Both grid and plate currents should be checked with triodes.

Amplifiers using capacitive tank center-taps (split-stator condensers with rotors grounded or by-passed) can be balanced by connecting a small balancing condenser across one section or the other of the grid tank condenser as required. In amplifiers with the coil center-tap by-passed to ground, the center-tap should be shifted to obtain balance. In some cases it may be necessary to balance the plate tank circuit similarly. Neutralization should be checked after each balancing adjustment.

OPERATION OF SCREEN-GRID AMPLIFIERS

Most of the foregoing procedure relating to triodes applies also to screen-grid tubes. However, principally because of the presence of the screen, there are additional factors which must be considered. Most screen-grid transmitting tubes are designed to operate without neutralization. However, this assumes certain further considerations. Because of the high power-sensitivity of such tubes, the feed-back coupling needed for oscillation is very small. Beyond the requirement of a well-screened tube, any possible feed-back coupling external to the tube must be reduced to a minimum. Special care must be used in the construction so that the input and output tank-circuit components and their respective wiring are well isolated from each other through judicious placement,

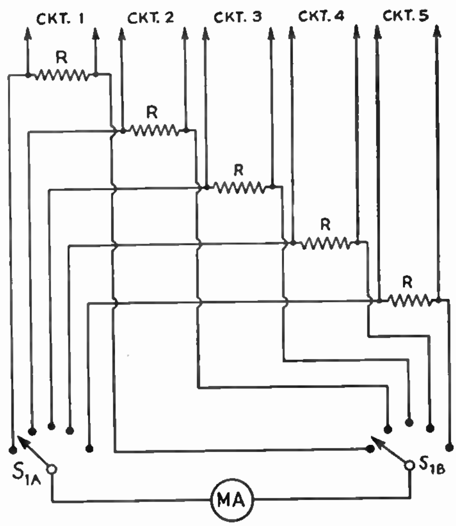


Fig. 6-28 — Method of switching single milliammeter to various circuits with a two-gang switch. The control shaft should be well insulated from the switch contacts, and should be grounded. The resistors, *R*, should have values of resistance ten to twenty times the internal resistance of the meter; 47 ohms will usually be satisfactory. S_1 is a 2-section multiposition rotary switch. Its insulation should be ceramic for high voltages, and a suitable insulating coupling should always be used between shaft and control knob.

and by shielding as completely as possible. Because it is sometimes difficult to eliminate all external capacitive coupling, it may be necessary to neutralize a screen-grid amplifier to eliminate all tendency toward oscillation.

Considerable dependence must be placed also on the fact that, from other considerations, a screen-grid amplifier should always be operated fully loaded, since the loading helps to prevent oscillation. Return leads to cathode,

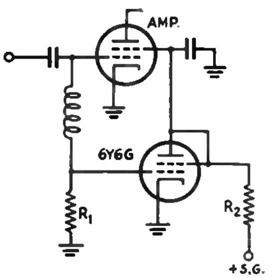


Fig. 6-29 — Screen protective circuit for screen-grid amplifier as an alternative to the use of fixed bias. R_1 is the normal grid leak for the amplifier and R_2 the recommended screen voltage-dropping resistor.

common to both plate and grid circuits, should be avoided. It is particularly important that the cathode be grounded directly or by-passed at the socket terminal and that the screen be by-passed thoroughly to the cathode with a mica condenser and short leads. The use of an un-by-passed parasitic-suppressing resistor at the screen is not recommended, since it aggravates instability at the operating frequency.

An indication of the coupling existing between input and output circuits can be obtained by the use of a sensitive r.f. indicator coupled to the output circuit as mentioned under "Neutralizing Procedure" in this chapter.

Other measures that can be taken to assist in stabilization at the operating frequency are the use of at least partial fixed bias and a nonresonant or detuned input circuit. With sufficient power from the driver, it is possible to secure rated excitation without having the grid circuit tuned close enough to resonance to start oscillation. In such a case the grid circuit should be detuned to the high-frequency side of resonance. Care should be taken that the grid circuit does not become resonant when the transmitter is tuned to another frequency.

Screen Considerations

For greatest protection to the tube, the screen voltage should be supplied from a series voltage-dropping resistor or a "light" voltage divider. When the screen is operated from a fixed-voltage source, the screen current increases rapidly with even slight amounts of overdrive or underloading. Since the series resistor serves to drop the voltage as the screen current increases, it affords a measure of protection. However, this same action may make it necessary to adjust the excitation with more than ordinary care if rated output is to be obtained. When a screen resistor or voltage divider is used, screen voltage should always be checked after each adjustment of excitation and loading to make sure that it is at rated value.

A screen-grid tube should never be operated at full screen voltage without plate voltage and full load. The screen current runs to damaging proportions under such conditions, especially if the screen is operated from a fixed-voltage source.

When plate and screen voltage and load are applied to a screen-grid amplifier, the grid current may increase, decrease or remain about the same, depending largely on the screen-voltage adjustment in relation to excitation.

Aside from the use of fixed bias, a screen-grid tube can be protected against excessive input when excitation is removed by the scheme shown in Fig. 6-29. A 6Y6G tetrode is connected as a low-μ triode. Since it is connected to the same point at the grid leak, the same bias appears at the grid of the protective tube and the grid of the amplifier. So long as excitation is supplied, the bias is sufficient to cut off the protective tube and it has no effect upon the operation of the amplifier. However, when excitation fails, the bias drops to zero and the 6Y6 draws current through the screen resistor, dropping the screen voltage to a point where the input to the amplifier is held within the dissipation rating.

● CHECKING POWER OUTPUT

Dummy Loads

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. 6-30. At A a thermoammeter, *A*, and a noninductive (ordi-

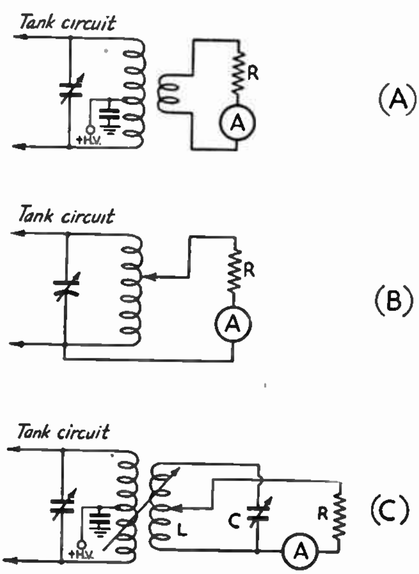


Fig. 6-30 — "Dummy-antenna" circuits for checking power output and making adjustments under load without applying power to the actual antenna.

nary 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.** 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 non-inductive 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. 6-30B 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. 6-30B is used.

Frequency Multiplication

SINGLE-TUBE MULTIPLIER

Output at a multiple of the frequency at which it is being driven may be obtained from an amplifier stage if the output circuit is tuned to a harmonic of the exciting frequency instead of to the fundamental. Thus, when the frequency at the grid is 3.5 Mc., output at 7 Mc., 10.5 Mc., 14 Mc., etc., may be obtained by tuning the plate tank circuit to one of these frequencies. The circuit otherwise remains the same as that for a straight amplifier, although some of the values and operating conditions may require change for maximum multiplier efficiency.

Efficiency in a single- or parallel-tube multiplier comparable with the efficiency obtainable when operating the same tube as a straight amplifier involves decreasing the operating angle in proportion to the increase in the order of frequency multiplication. Obtaining output comparable with that possible from the same tube as a straight amplifier involves greatly increasing the plate voltage. A practical limit as to efficiency and output within normal tube ratings is reached when the multiplier is operated at maximum permissible plate voltage and maximum permissible grid current. The plate current should be reduced as necessary to limit the dissipation to the rated value by increasing the bias. High efficiency in multipliers is not often required in practice, since the purpose is usually served if the frequency multiplication is obtained without an appreciable gain in power in the stage.

Since the input and output circuits are not tuned close to the same frequency, neutralization usually will not be required. Instances may be encountered with tubes of high transconductance, however, when a doubler will oscillate in t.g.t.p. fashion, requiring the introduction of neutralization. The link neutralizing system is convenient in such a contingency.

OTHER MULTIPLIER CIRCUITS

Push-Pull Multiplier

A single- or parallel-tube multiplier will deliver output at either even or odd multiples of

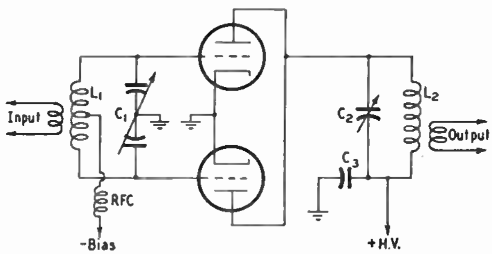


Fig. 6-31 — Circuit of a push-push frequency multiplier for even harmonics. The grid tank circuit, L_1C_1, is tuned to the frequency of the preceding driving stage, while the plate tank circuit, L_2C_2, is tuned to an even multiple of that frequency, usually the second harmonic. C_3 is the plate by-pass capacitor, usually a 0.01-μfd. paper condenser, while RFC is a 2.5-mh. r.f. choke.

the exciting frequency. A push-pull multiplier does not work satisfactorily at even multiples because even harmonics are largely canceled in the output. On the other hand, amplifiers of this type work well as triplers or at other odd harmonics. The operating requirements are similar to those for single-tube multipliers.

Push-Push Multipliers

A two-tube circuit which works well at even harmonics, but not at the fundamental or odd harmonics, is shown in Fig. 6-31. It is known as the **push-push** circuit. The grids are connected in push-pull while the plates are connected in parallel. The efficiency of a doubler using this circuit may approach that of a straight amplifier under similar operating conditions, because there is a plate-current pulse for each cycle of the output frequency.

This arrangement has an advantage in some applications. If the heater of one of the tubes is turned off, making the tube inoperative, its grid-plate capacitance, being the same as that of the remaining tube, serves to neutralize the circuit. Thus provision is made for either straight amplification at the fundamental with a single tube, or doubling frequency with two tubes as desired.

Multiplications of four or five sometimes are used to reach the bands above 28 Mc. from a lower-frequency crystal, but in the majority of lower-frequency transmitters, multiplication in a single stage is limited to a factor of two or three, because of the rapid decline in practicably obtainable efficiency as the multiplication factor is increased. Screen-grid tubes make the best frequency multipliers because their high power sensitivity makes them easier to drive properly than triodes.

Parasitic Oscillations

Before placing the amplifier in operation, measures should be taken to make sure that the amplifier will function in a stable manner. In addition to the possibility of oscillation at or near the operating frequency, r.f. power amplifiers are subject to parasitic oscillation at frequencies far removed from the frequencies to which the amplifier is tuned by the conventional tank circuits. Oscillations of this type not only cause the transmission of illegal spurious signals, but they also impair the efficiency of the amplifier. In fact, they can be so severe as to make operation of the stage as an amplifier impossible and may destroy the tube if they are allowed to persist for any appreciable time. Erratic tuning characteristics invariably are a result of oscillation of one type or another. Parasitic oscillations may not be obvious under normal conditions of bias and load, but may be transient in nature, occurring intermittently during keying or modulation, causing widespread clicks or splatter. They can be treated most successfully only by adjusting the amplifier for conditions favorable toward sustained oscillation when they can be more readily observed and identified.

● V.H.F. PARASITIC OSCILLATION

Parasitic oscillation in the v.h.f. range (usually in the vicinity of 100 to 200 Mc.) almost invariably will take place in an amplifier unless steps are taken to suppress it. In most cases, this sort of oscillation takes place as the result of an unavoidable t.g.t.p. circuit set up by the grid and plate leads tuned by the tank condensers in series, as shown by the heavy lines in Fig. 6-32A. The normal tank coils act only as r.f. chokes or capacitances at this frequency. The same condition holds for balanced or push-pull circuits.

Testing Procedure

To test for this type of oscillation, the 28-Mc. tank coil should be plugged into the grid tank circuit (or the plate tank circuit of the driver stage if capacitance coupling is used) and the 3.5-Mc. coil in the plate tank circuit. This is to prevent any possible t.g.t.p. oscillation at the

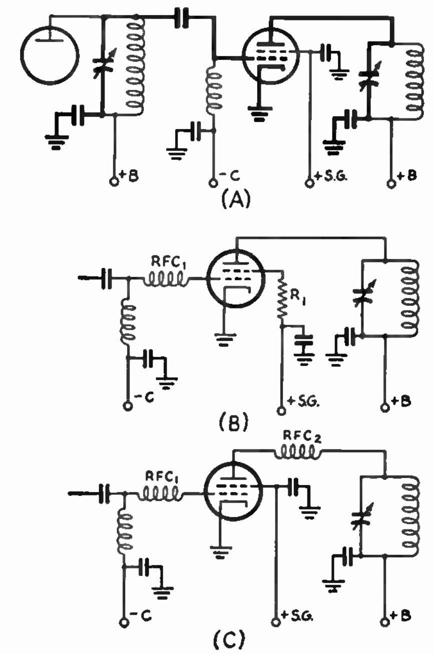


Fig. 6-32 — A — V.h.f. parasitic circuit hidden in high-frequency amplifier. B — Common method of suppressing v.h.f. parasitic with tetrodes; R_1, however, is not recommended. C — Recommended circuit. Suggested dimensions for RFC_1 are 15 turns No. 22 wire, ¼-inch diameter, close-wound; for RFC_2, 8 turns No. 14 wire, ¾ inch long, self-supporting.

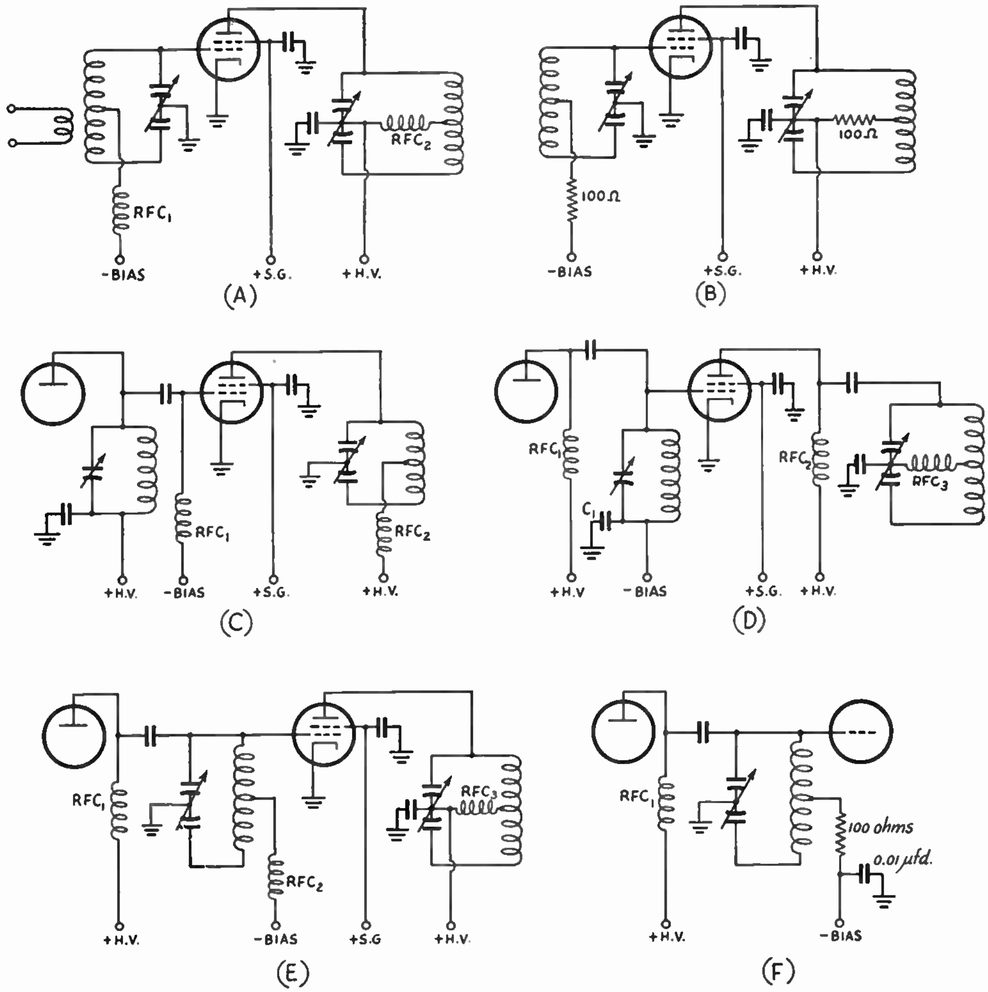


Fig. 6-33 — A, C and E show circuit arrangements to be avoided in eliminating low-frequency parasitic oscillation. The circuits of B, D and F are the recommended alternatives, which apply equally well to push-pull.

operating frequency which might lead to confusion in identifying the parasitic. If either input or output tank circuit employs a split-stator tank condenser with an r.f. choke at the center of the tank coil, the r.f. choke should be short-circuited during the test. Any fixed bias should be replaced with a grid leak of 10,000 to 20,000 ohms. In a capacitive-coupled stage, the driver should be coupled to the amplifier in the normal way, but all load on the output of the amplifier should be disconnected. If the stage is an intermediate amplifier, the tube in the following stage should remain in place, but with its heater or filament turned off. Plate (and screen) voltage should be reduced to the point where the rated dissipation is not exceeded. If a Variac is not available, voltage may be reduced by a 115-volt electric lamp of suitable wattage rating in series with the primary of the plate transformer.

With power applied only to the amplifier under test (not the driver), a careful search should be made by adjusting the input tank condenser to several settings, especially including minimum and maximum, and turning the plate tank condenser through its range for each of the grid-condenser settings. Any grid-current reading, or any dip or slight flicker in plate current at any point, indicates oscillation. This can be confirmed by using an indicating absorption wavemeter (see Chapter Sixteen) tuned to the frequency of the parasitic and held close to the plate lead of the tube.

Remedies

A combination of v.h.f. choke at the grid of the amplifier and a suppressor resistor at the screen, as shown in Fig. 6-32B, is an effective means of suppressing v.h.f. parasitic oscillation in screen-grid amplifiers. But since the screen resistor decreases the effectiveness of the screen in suppressing oscillation at the operating frequency, it is preferable to use other methods. V.h.f. chokes at both grid and plate, as shown in Fig. 6-32C, usually will serve to suppress v.h.f. parasitics in either triode or screen-grid amplifiers. The approximate dimensions for these chokes given under Fig. 6-32 may have to be altered experimentally somewhat, depending upon lead length in individual cases. In some instances it may be necessary to use a trap circuit tuned to the frequency of the parasitic, in the plate circuit, instead of RFC_2. In this case, the grid choke may or may not be required.

In push-pull or parallel-tube amplifiers the same treatment should be given each tube.

LOW-FREQUENCY PARASITICS

Low-frequency parasitic oscillations (which usually lie in the wide range between 100 and 2000 kc.) invariably involve plate- and grid-circuit r.f. chokes in combination with a split-stator tank condenser tuning at least one of them if not both. The normal tank coils have such little reactance at low frequencies that they may be considered merely as long connecting leads.

Although they are not so likely to be encountered in amplifiers using the better-screened transmitting tetrodes and pentodes, low-frequency parasitic oscillations are often found in stages employing triodes and the less effectively-shielded audio tubes, such as the 6L6, 6V6, etc. Even if well-screened tubes are used, it is safer and more convenient to arrange the circuit in advance so that these low-frequency circuits are broken up.

Circuits To Be Avoided

Fig. 6-33 shows several commonly-used circuit arrangements that should be avoided to eliminate the possibility of low-frequency parasitics. In A, either r.f. choke or both may be replaced with a 100-ohm resistor, as shown in B. In a similar circuit, parallel feed can be used in either grid or plate, but not in both.

In Fig. 6-33C, RFC_2 should be replaced by a resistor. If parallel plate feed is desired, series feed should be used in the grid, as shown at D, necessitating parallel feed in the driver-stage plate. If the driver plate tank circuit has a split-stator condenser, as shown in E, the grid choke should be replaced by a 100-ohm resistor by-passed to ground, as shown in F. It is important that the by-pass be fairly large so as to be effective at low frequencies.

A check for low-frequency parasitics should be made after the v.h.f. oscillations have been eliminated. The check is conducted along the lines described for very-high frequencies. Low-frequency oscillation can be detected by coupling the absorption wavemeter closely to the r.f. chokes involved, remembering that the range of frequencies over which this type of parasitic may occur is wide. They can also sometimes be detected by listening on a receiver close to the transmitter, when harmonics, usually rough in character, may be heard at regular intervals that are multiples of the fundamental frequency. On a calibrated receiver, the fundamental frequency can be determined by observing the spacing between adjacent harmonics.

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 immediate objective of the neutralizing process is reducing to a minimum the r.f. driver voltage fed from the input of the amplifier to its output circuit through the grid-plate capacitance of the tube. This is done by adjusting the neutralizing condenser until an r.f. indicator in the output circuit gives minimum response.

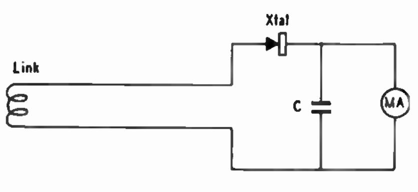


Fig. 6-34 — Circuit of sensitive neutralizing indicator. *Xtal* is a 1N34 crystal detector, *MA* a 0–1 direct-current milliammeter and *C* a 0.001-µfd. mica by-pass condenser.

NEUTRALIZING INDICATORS

Fig. 6-34 shows the diagram of a sensitive neutralizing indicator. By referring to Chapter Sixteen, it will be seen that this forms part of the indicating absorption wavemeter also recommended for checking parasitic oscillation. The link should be coupled to the output tank coil at the low-potential or "ground" point. Care should be taken to make sure that the coupling is loose enough at all times to prevent burning out the meter or the rectifier.

A neon bulb touched to the "hot" end of the tank coil will glow if enough feed-through voltage is developed across the tank, but it is a less-sensitive device. Another disadvantage is that its use introduces capacitance across one side of the circuit which may unbalance the circuit, thus giving an inaccurate indication of neutralization.

A more satisfactory indicator than the neon bulb is a flashlight bulb (the lower the power the more sensitive) connected at the center of a turn or two of wire coupled to the tank coil at the low-potential point. Its sensitivity is poor compared with the milliammeter-rectifier indicator, however.

The grid-current milliammeter may also be used as a neutralizing indicator. If the amplifier is not neutralized, there will be a large dip in grid current as the plate-tank tuning passes through resonance. This dip reduces as neutralization is approached until at exact neutralization all change in grid current should disappear.

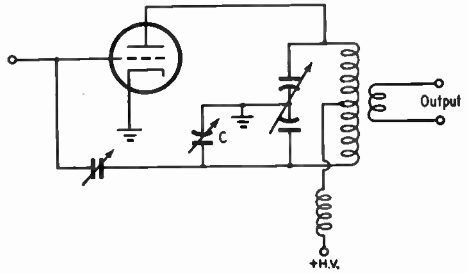


Fig. 6-35 — **In this neutralizing circuit, *C*, which should have the same capacitance as the output capacitance of the tube, has been added to compensate for the tube capacitance across the upper half of the circuit.**

NEUTRALIZING ADJUSTMENTS

The neutralizing condenser should always be adjusted with an insulating rod, not only to protect the operator but also to avoid capacitive effects which might give a false indication.

With excitation applied, the neutralizing adjustment should be started with the neutralizing condenser at minimum capacitance, increasing the capacitance in small steps. At each step, the plate tank should be swung through resonance which will be indicated by maximum deflection of the indicators mentioned above and by the dip in grid current. As the point of neutralization is approached, the indication will become less until it is a minimum when neutralization is reached. If the neutralizing capacitance is increased further, the indication will again increase. If the neutralizing condenser has a proper range of capacitance, it should always be possible to find a point of minimum indication with an increase on either side.

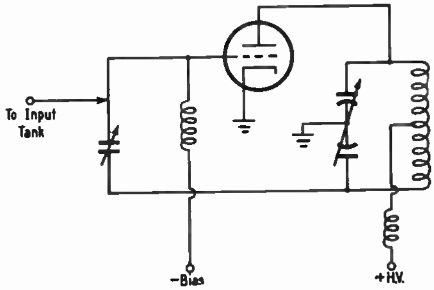
If it is found that neutralization does not hold over the entire range of the tank condenser for any one band in a single- or parallel-tube amplifier, the balancing condenser of Fig. 6-35 should be added and adjusted.

In an amplifier which is to be used on several bands, it should be first neutralized when tuned to the lowest-frequency band. Then the neutralization should be checked at the highest frequency. If it is found that the neutralizing condenser needs readjustment at the higher frequency, the connection between the grid tank circuit (or the plate tank circuit of the driver with capacitance coupling) should be adjusted as indicated in Fig. 6-36 until the neutralizing adjustment is the same for both bands, always neutralizing first on the lowest-frequency band and checking at the highest frequency. If there are parasitic chokes at the grid and plate, connection of the neutralizing condenser to one side and then the other should be tried to determine which connection permits the best neutralizing from band to band.

Any indication remaining at minimum means that coupling between input and output exists external to the tube. The isolation seldom can be made complete, but it should be possible to bring it down to a very low value with proper wiring and shielding. 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 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 as the tank condenser is tuned through resonance. The driver stage must be operating while this is done, of course.

Adjustment of Inductive Neutralizing Systems

With link neutralizing of a single-tube or parallel-tube amplifier, the neutralization can be adjusted by altering the number of turns at either end or by changing the spacing between the link and the tank coil.



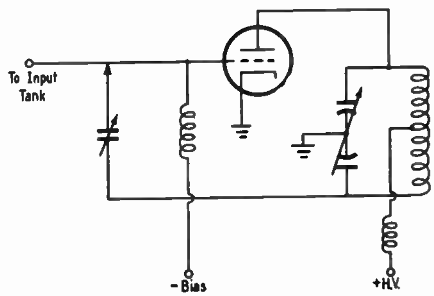


Fig. 6-36 — **If an amplifier fails to remain neutralized on all bands, the condition usually can be remedied by tapping the input-tank lead along the neutralizing-condenser lead (or vice versa), adjusting the position until the amplifier neutralizes at the highest frequency at the same setting of the neutralizing condenser as at the lowest frequency. The same adjustment should be made to both sides of a push-pull circuit.**

The inductive neutralizing system holds neutralization over a wider frequency range if the auxiliary adjusting condenser (Fig. 6-17C) is omitted, adjusting the size of the coil to resonate at the operating frequency with the plate-grid capacitance of the tube only. The use of the auxiliary condenser makes the adjustment more convenient, of course.

Harmonic Suppression

A transmitter may generate and radiate energy at harmonics of the operating frequency. Although the harmonic power seldom is very large in terms of the power at the fundamental, it may be sufficient to cause interference at a distance under favorable conditions of propagation as well as to various amateur and nonamateur services, particularly television, at shorter distances.

The reduction of harmonic energy in the antenna system is treated in Chapter Ten. In this chapter radiation by other means is discussed. Harmonic energy can be radiated from components and wiring in the transmitter itself or by r.f. leaking into power leads and lines external to the transmitter.

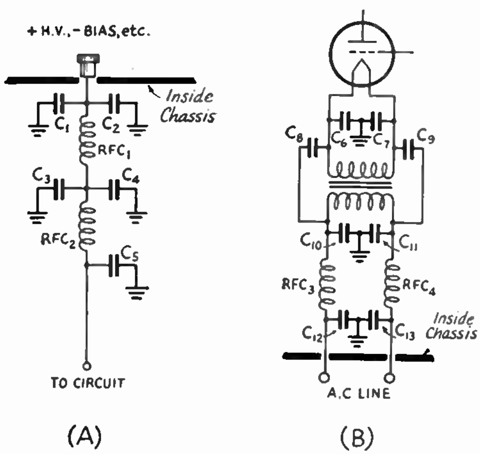


Fig. 6-37 — Typical harmonic-suppression filters. A — For use in plate- and bias-supply, keying and external-meter leads. B — For filament-supply circuit.

C_1, C_6, C_7 — 0.005-μfd. mica.
C_2, C_3, C_{10}, C_{11}, C_{12}, C_{13} — 500-μμfd. mica.
C_4, C_5, C_8, C_9 — 0.001-μfd. mica.
RFC_1 — V.h.f. choke (Ohmite Z-0 or Z-1).
RFC_2 — 2.5-mh. r.f. choke.
RFC_3, RFC_4 — Three inches No. 14, ¾-inch diam.

The best measure that can be taken to eliminate direct radiation from the transmitter itself is to enclose it completely within a metal cabinet or other shielding. To prevent power-lead or power-line radiation, all leads emerging from the shielded enclosure should be filtered. This includes leads to unshielded meters, etc., on the panel.

Harmonic fields around a transmitter and energy in power wiring can be traced with a sensitive indicating absorption wavemeter of the type described in Chapter Sixteen. Checking should be done with the entire transmitter operating and the final loaded with a dummy load. The load should be kept within the transmitter enclosure. In checking power leads, a section of the lead should be looped to couple to the wavemeter coil. When the wavemeter is tuned to the frequency of any harmonic energy flowing in the lead, an indication will be obtained. Even a slight deflection of the meter may indicate sufficient radiation to cause interference in neighboring receivers. Each power lead, as well as any 115-volt lines running into the transmitter enclosure, should be checked.

Fields around a transmitter can be similarly detected by exploring the area around the enclosure while tuning the wavemeter to successive harmonics.

Although each transmitter may require individual treatment, the filtering systems shown in Fig. 6-37 should serve at least as starting points from which additional experimental treatment may begin. The extent to which this additional work must be carried will depend to a great extent upon the distance between the transmitter and the point of interference.

The filter components should be mounted under the chassis, close to the point where the power leads enter and should be shielded from any possible r.f. field.

Keeping connecting leads in the transmitter short should help to discourage the radiation of harmonics by eliminating segments which may resonate at these frequencies. The path for r.f. from the plate terminal of the amplifier to ground in particular should be the shortest possible. In accomplishing this, a vacuum-type condenser may be connected directly from plate to filament. This will, of course, add to the effective tank-circuit capacitance so that the tank-condenser capacitance will have to be reduced accordingly or the size of the coil reduced to compensate. In push-pull or neutralized stages, a similar condenser should be connected across each tube. The same thing may be done in the grid circuit to short-circuit harmonics from the driver stage. At 3.5 Mc., it is feasible to use a fixed capacitance as great as 50 μμfd. in such a manner.

Other measures that can be taken are to do the frequency multiplying in very-low-power stages and build up the power at the operating frequency in stages biased not beyond plate-current cut-off, drive the output stage no harder than is necessary for proper modulation in the case of plate modulation, and install parallel-tuned traps in each plate lead tuned to trap out offending harmonics. Make sure, too, that the transmitter is completely free from parasitic oscillation.

A Simple Single-Tube Transmitter

One of the simplest practical transmitters is shown in the photographs of Figs. 6-38 and 6-39. If the station receiver has a power audio stage which is not required for headphone reception, the tube may be taken from the receiver and used in the transmitter (provided that the tube is a pentode or tetrode as it usually is). A plug inserted in the empty socket in the receiver may be used to obtain power for operating the transmitter.

The schematic diagram is shown in Fig. 6-40. The Tri-tet oscillator circuit is used to permit operation in either the 3.5- or 7-Mc. bands with a single 3.5-Mc. crystal. Series plate feed is used and no means of reducing the voltage of the screen below that of the plate is necessary if the supply potential does not exceed 250 to 300 volts.

The cathode circuit is tuned by a fixed mica condenser, C_1, but if necessary, the tuning of this circuit can be changed by changing the dimensions of the coil, L_1.

No provision is included for tuning the antenna system, for the sake of maximum simplicity. This can be done by selecting the proper feeder length and adjusting the size of the antenna coupling coil, L_3.

Construction

To minimize the tools required for the construction of the transmitter the parts are mounted on a simple chassis of wood finished with clear lacquer or shellac. Two 1¾ × 9¾-inch strips of ¼-inch-thick wood are fastened with screws to the two 4½ × 2½ × ¾-inch end pieces, leaving enough separation between the strips for the Amphenol M1P octal sockets used for holding both the crystal and the tube. Wood screws can be used to mount the sockets, or they can be bolted to the wood strips with 6-32 machine screws. The key of the tube socket should be mounted toward the front of the transmitter for convenience in wiring the plate circuit to the tuning condenser. Because the tuning condenser does not have a long mounting shank, it is necessary to drill a clearance hole for the shank and then dig away — or counterbore — clearance for the nut. The two Fahnestock clips for the antenna are secured under two of the screws used for fastening the wood strips to the right-hand end piece, and the other two clips used for the key leads are held down by machine screws on the left-hand end piece. The r.f. choke is held in place on the left-hand end piece by a machine screw. The four wires used for a power cable are brought out at the rear left under the wood strip — a half-round hole is filed in the end piece to clear the wires.

The plate and antenna coils are held in place on three small sticks set in the top of the chassis — penny suckers are a good source of these sticks. The bottom of the plate coil connects to a brass machine screw soldered to a lug which is sweated to the stator terminal of the tuning condenser, and the screw is built up most of its length by adding nuts or small spacers. The screen end of the coil, the top end of the winding, is fastened to a brass screw that runs through the rear wood strip. The coil ends have lugs soldered to them to facilitate band changing. The antenna-coil ends similarly fasten to two brass screws supported by short lengths of heavy wire and the wire is sweated to the Fahnestock clips and to the heads of the screws.

Wiring

The wiring is done with the same wire that is used for the coils, because a single 50-foot

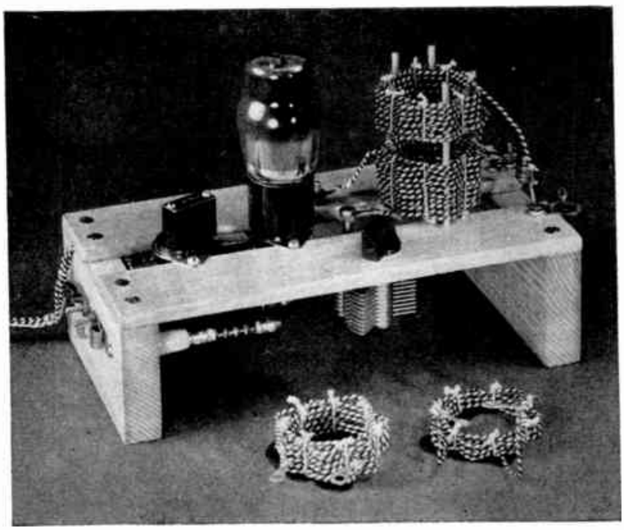


Fig. 6-38 — By using wood for the chassis and simplified construction throughout, this simple oscillator transmitter can be built with very few shop tools. Using a 3.5-Mc. crystal, operation in the 3.5- and 7-Mc. bands is possible by changing the plate and antenna coils. The arrangement is suitable for 6F6, 6V6 or other similar pentodes and tetrodes.

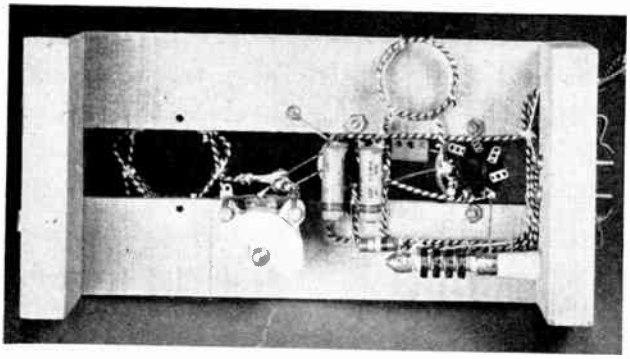


Fig. 6-39 — Bottom view of the simple single-tube transmitter. The cathode coil is between the tube and crystal sockets. The r.f. choke is to the right. C_4 is at left center with the two by-pass condensers, C_2 and C_3, to the right of it.

roll of No. 18 bell wire, available in any "5 & 10" or hardware store, suffices for the whole rig with some to spare. To insure good electrical connection, the wire is soldered at every connection, which means that the wire is soldered to the heads of the brass machine screws used for the key leads and the screen end of L_2 before the screws are put in place. One key lead, one end of R_1, the outer foil connections on C_2 and C_3, and the lead to Pin 1 of the power plug must be connected to Pin 1 of the tube socket. At the crystal socket, two adjacent pins (e.g., 1 and 8) are bonded together for the grid side of the crystal and the next two pins (e.g., 2 and 3) are bonded together for the cathode side. This permits plugging the crystal into either Pins 8 and 2 or 1 and 3. The connection can be elaborated still further by bonding Pins 4 and 5 with 8 and 1 and tying 6 and 7 to 2 and 3, in which case the crystal can be plugged in any way and it will make the proper connection.

The cathode coil, consisting of 5 turns of No. 18 bell wire, is wound on a 1¼-inch diameter form and then removed and tied with string at a number of places. The cathode coil is mounted by its leads only but, being short, they offer adequate support.

The plate and antenna coils are wound by equally spacing seven nails on a 2-inch diameter circle, driving the nails completely through the board used so that the heads are flush against the board. Small spikes can be used, or nails of the "8-penny" size will be satisfactory if a thin board is used. One end of the wire is secured to a nail and the wire is threaded over alternate nails, so that the coil repeats itself every two turns. When the required number of turns has been made, the end of the wire is wrapped around a nail and the coil tied together with string at the seven crossover points. Soldering lugs are soldered to the ends of the coil for ease in changing bands.

The four wires coming out the side of the chassis that go to the power plug are twisted together slightly and cabled with string to form a neat cable, and the cable plug, P_1, is simply the base from an old tube. If the receiver is to be used as a source of power, the base should be one that will fit the power-output tube in the receiver. Break the tube and chew out the glass from inside the base with a pair of pliers, being careful not to break the bakelite of the base. It will help in making connection to the proper pins if a small drill, slightly larger than the diameter of the No. 18 wire, is run through the pins before the wires are inserted and soldered in place.

Tuning

After checking the wiring, plug in a crystal and connect the 7-Mc. coil in place. Place the audio tube from the receiver in the transmitter and plug in the power cable, and connect a key to the clips on the side of the transmitter. If the receiver has push-pull output, it is probably best to remove both power tubes. Set the tuning condenser, C_4, at about 40 per cent meshed and turn on the power to the receiver. When the tube has had time to warm up — about 30 seconds — close the key and touch a neon bulb to the plate end of L_2. Or a small 10-watt electric lamp can be connected to the antenna posts with the 6-turn antenna coil in place. If C_4 is set properly, the

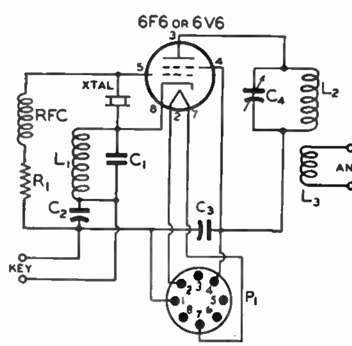


Fig. 6-40 — Wiring diagram of the inexpensive easy-to-build transmitter.

C_1 — 470-$\mu\mu$fd. mica.
C_2, C_3 — 0.01-μfd. 600-volt paper.
C_4 — 140-$\mu\mu$fd. variable (Hammarlund SM-140 or Bud MC-1876).
R_1 — 0.1-megohm 1-watt composition.
L_1 — 5 turns No. 18 d.c.c., 1¼-inch inside diameter, close-wound.
L_2 — 3.5 Mc.: 19 turns. 7 Mc.: 12 turns.
L_3 — 13 turns and 6 turns. Requires experiment — see text. See text for L_2 and L_3 winding instructions.
P_1 — See text.
RFC — 2.5-mh. r.f. choke (National R-100U).

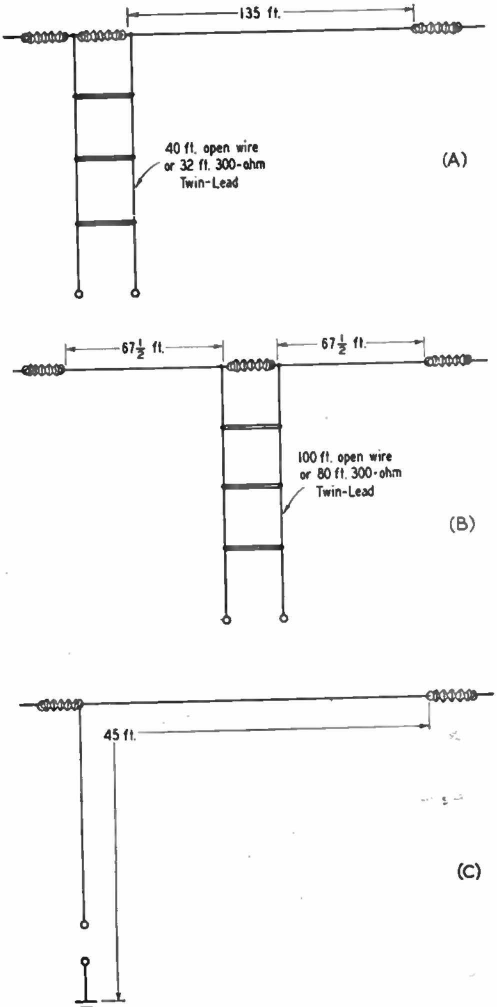


Fig. 6-41 — Suggested antenna dimensions for use with the single-tube transmitter. A — End-fed half-wave or Zepp. B — Center-fed half-wave. C — Quarter-wave grounded antenna.

neon bulb will glow or the lamp will light. If this does not happen, try tuning the plate condenser until signs of output become apparent. The transmitter can then be checked on the 3.5-Mc. band by putting in the proper coils — remembering, however, to turn off the receiver and hold the key closed until the power pack of the receiver has been discharged, to avoid getting a shock when touching the coil terminals. The tuning condenser setting will be about 85 per cent meshed on the lower-frequency band.

It will not be possible in most cases to check the keying on the receiver used to furnish power to the transmitter, and it is highly advisable to check the keying in a monitor or another receiver. If the keying is chirpy, the cathode coil, L_1, should be squeezed out of round to reduce its inductance until the keying is better. On the 3.5-Mc. band, best keying will generally be obtained with slightly less capacity at C_4 than the setting for maximum output. In the oscillator shown in the photographs, a slight key click on "break" was eliminated by connecting a 0.1-μfd. 600-volt paper condenser directly across the key. Some crystals key better than others.

Antennas

A 135-foot piece of wire for the antenna can be fed in several ways to give satisfactory results. It can be fed at one end with about 40 feet of open-wire feeders (about 32 feet of Amphenol 300-ohm Twin-Lead), as shown in Fig. 6-41A or it can be fed in the center with 100 feet of open-wire feedline (about 80 feet of 300-ohm Twin-Lead) as indicated at B. These lengths will enable one to connect the feedline directly to the antenna posts of the transmitter without the necessity for tuning condensers — other lengths may require either series or parallel condensers. Some experiment with the antenna coil may be necessary, but a small flashlight bulb in series with one of the feeders will serve as a good indication of feeder current, and will help in the tune-up process. The lamp need not be shorted during normal operation unless it burns too brightly. A neon bulb will also help in detecting r.f. energy in the transmission line, but it may not always light with this low power.

If room for only a short length of wire is available for the antenna, say 40 or 50 feet, it is best to connect its end to one antenna post and a good ground to the other as shown in Fig. 6-41C. Here again some experimentation will be necessary to determine the optimum size of L_3. The diagram of a suitable alternative power supply is shown in Fig. 6-42.

The power can be increased by substituting a 6L6 for a smaller tube and adding a separate power supply to give 350 volts at 100 ma., but with the newer small crystals it is not advisable to increase the voltage much above this value without keeping the screen voltage down by the addition of a dropping resistor and another by-pass condenser.

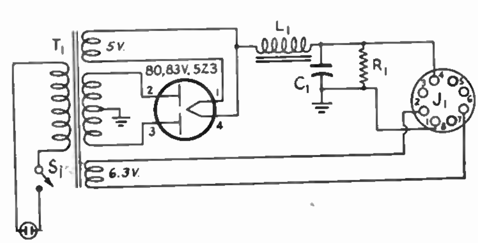


Fig. 6-42 — Circuit diagram of alternative power supply for the simple single-tube transmitter.

C_1 — 8-μfd. 450-volt electrolytic.
R_1 — 25,000 ohms, 10 watts.
L_1 — Filter choke — any receiver replacement type, 15 hy. or more, 50 ma. or more.
J_1 — 8-prong tube socket.
S_1 — S.p.s.t. toggle switch.
T_1 — Power transformer — any receiver replacement type, not over 750 volts c.t., 50 ma. or more.

A Low-Power Two-Stage Transmitter

Figs. 6-43 through 6-47 show two different types of construction for a two-stage 40-watt transmitter. The circuit diagram, which applies to both models, is shown in Fig. 6-45. The oscillator circuit is a modified Pierce with an untuned output circuit (RFC_2). An 807 is used in the amplifier stage, which may be operated as either a straight amplifier at the crystal frequency, or as an output doubler. Thus 7-Mc. output may be obtained with a 3.5-Mc. crystal in the oscillator, or 14-Mc. output with a 7-Mc. crystal. The use of 14-Mc. crystals is not recommended in this circuit.

C_4 serves to adjust the oscillator feed-back for best output consistent with reliable keying, while C_3 provides lag in the keying circuit to soften the keying characteristic. RFC_4 and RFC_5 are parasitic suppressors. In operation, only two tuning adjustments are required, including antenna tuning.

● METAL HORIZONTAL MODEL

The horizontal unit is of all-metal construction. The chassis is 7 × 13 × 2 inches. The two tank condensers, C_8 and C_{10}, are mounted on small cone insulators on either side of the tube sockets at the center. The respective coil sockets are immediately to the rear, at right angles to each other to minimize direct coupling between the coils.

Power-supply components occupy the rear portion of the chassis. The smaller parts — r.f. chokes, by-pass condensers and resistors — are mounted underneath, close to the tube sockets with which they are associated. Each by-pass condenser should be connected directly to the terminal to be by-passed and grounded to the chassis with the shortest practicable lead. The ends of r.f. chokes and resistors which must be insulated from the chassis are mounted on insulated lug strips.

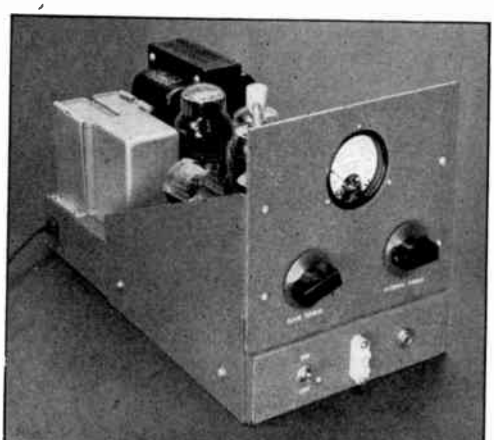


Fig. 6-43 — The metal horizontal model of the low-power two-stage transmitter. The two tuning controls are for the output-stage tank and antenna coupler.

The panel may be of metal or crackle-finished Presdwood. It is 7¼ inches wide and 6¼ inches high and is fastened to the chassis by means of standard 8-inch chassis brackets. The two tank-condenser shafts are extended, with flexible insulating couplings, to the two controls on the panel. The milliammeter is mounted behind a clearance hole above and central between the two tuning controls. The crystal socket, power switch and key jack are mounted along the front edge of the chassis below the panel.

Fig. 6-44 — Panel view of the vertical model of the low-power two-stage transmitter.

● VERTICAL WOOD MODEL

The vertical model is assembled on a frame of wood and is divided into two separate sections — the panel, which holds the components associated with the r.f. circuits, and the supporting base on which the power supply is assembled. The panel, 13 inches high and 7 inches wide, must be of insulating material, since the two tank condensers are mounted directly on it without further insulation. For the same reason, well-insulated tuning dials must be used. Crackled Presdwood is satisfactory as panel material.

The two tubes are supported on brackets near the bottom of the panel. The bracket for the 807 is part of a Millen 80009 shielding assembly. A long lug strip serves as an anchorage for the various by-pass condensers and r.f. chokes grouped about the sockets. All ground

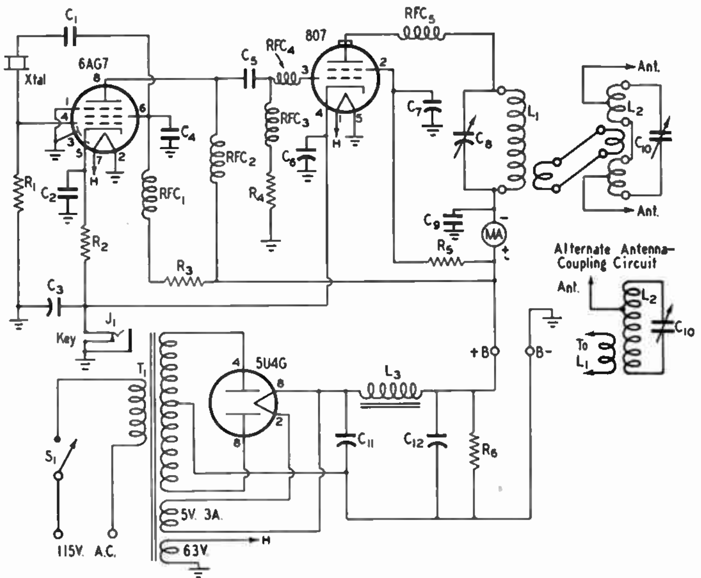


Fig. 6-45 — Circuit diagram of the low-power two-stage transmitter.

C_1 — 680-μμfd. mica.
C_2 — 0.0068-μfd. mica.
C_3 — 8-μfd. 150-volt electrolytic.
C_4 — 100-μμfd. mica (see text).
C_5 — 100-μμfd. mica.
C_6 — 0.01-μfd. 400-volt paper
C_7 — 0.01-μfd. 600-volt paper.
C_8 — 140-μμfd. max. receiving-type variable (Hammarlund MC-140-M).
C_9 — 0.0022-μfd. mica.
C_{10} — 250-μμfd. max. receiving-type variable (Hammarlund MC-250-M).
C_{11}, C_{12} — 4 μfd., 600 volts.
R_1 — 56,000 ohms, ½ watt.
R_2 — 330 ohms, 1 watt.
R_3, R_4 — 22,000 ohms, 1 watt.
R_5 — 20,000 ohms, 5 watts, wire-wound.
R_6 — 50,000 ohms, 20 watts, wire-wound.
L_1 — 3.5 Mc. — National AR-80-E with 20 turns removed — 35 turns No. 24 d.s.c., 1⅛ inches long, 1¼-inch diam., 3-turn link.
— 7 Mc. — National AR-20-E — 14 turns No. 18 enam., 1¼ inches long, 1¼-inch diam., 3-turn link.
— 14 Mc. — National AR-10-E — 8 turns No. 18 enam., 1⅜ inches long, 1¼-inch diam., 3-turn link.
L_2 — 3.5 Mc., 7 Mc. — National AR-40-S, 20 turns No. 20, center-tapped, 1⅝ inches long, 1⅛-inch diam., 5-turn variable center link.
— 14 Mc. — National AR-20-S, 12 turns No. 18 center-tapped, 1⅝ inches long, 1¼-inch diam., 4-turn swinging center link.
L_3 — Receiver-type filter choke, 8 hy. at 150 ma. approx. 150 ohms d.c. resistance.
J_1 — Closed-circuit 'phone jack.
MA — 0–200 d.c. milliammeter.
RFC_1, RFC_2, RFC_3 — 2.5 mh., 100 ma.
RFC_4 — 17 turns No. 18 d.s.c., wound on 1-watt 1-megohm resistor.
RFC_5 — 8 turns No. 14, 5/16-inch inside diam., ¾ inch long.
S_1 — S.p.s.t. toggle switch.
T_1 — Replacement-type power transformer, 400–0–400 v. a.c. at 160 ma., 5 volts at 3 a., 6.3 volts at 2 a. (Stancor P-4081).

connections are made to a metal strip running under the lug strip. The two metal brackets and the plate-circuit by-pass condenser, C_9, should be connected to this strip.

The plate tank coil and condenser are mounted on the panel just above the 807 and the antenna tank components at the top of the panel with the two coils at right angles. A hole for the milliammeter is cut in the panel above the 6AG7. The crystal socket is mounted on the front of the panel so that its terminals come out at the rear just below the 6AG7 socket, while the key jack and power switch are placed so as not to interfere with the tubes behind the panel.

The frame that holds the panel and the base on which the power-supply components are mounted are made of sections of 1 × 2 soft pine, finished in grey enamel. Triangular pieces of the panel material are used as braces for the uprights. Strips of the same material spanning the top of the base are spaced to leave openings below for the filter-condenser, power-transformer and choke terminals and the rectifier-tube socket. Thus the power supply may be wired up easily from the bottom without the necessity for cutting holes.

Adjustment

The most important thing to remember in

TABLE 6-I
Representative Voltage and Current Measurements

Conditions: 3.5-Mc. crystal, 7-Mc. output, 25-watt lamp used as dummy antenna, 807 stage loaded to 100 ma.

6AG7 stage:	*807 stage:*
Plate volts: 420	Plate volts: 420
Plate ma.: 35	Plate ma.: 100
Screen volts: 260	Screen volts: 310
	Grid volts: −65
	Grid ma.: 2.9

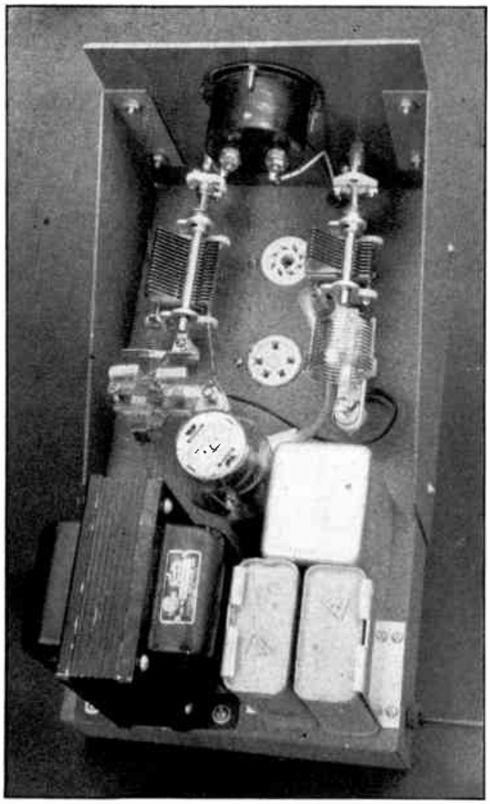


Fig. 6-46 — Top view of the horizontal low-power two-stage transmitter. The power supply occupies the rear end of the chassis.

adjusting the transmitter is that the power switch *must* be turned off whenever the amplifier plug-in coils are to be changed, since the coil carries the full supply voltage.

With the coils and crystal for the desired band in place, the power supply may be turned on. Several seconds should be allowed for the heaters to warm up. With the antenna taps at the ends of the coil, C_{10} should be turned to minimum capacitance and the variable link swung part way out. When the key is closed, the milliammeter should read between 100 and 150 ma., depending upon the voltage delivered by the supply. The plate tank circuit should be quickly tuned to resonance as indicated by the dip in plate current. The key should be closed for only short periods at a time until the amplifier is loaded, to protect the screen.

With the tank tuned to resonance, C_{10} should be swung through its range to the point where it causes a rise in plate current. C_{10} should be tuned accurately to the point where the plate current is greatest. If there is no increase in plate current within the range of the antenna tuner, the taps must be adjusted until an increase is obtained. If the plate current does not increase to 90 or 100 ma. when the antenna tuning is at resonance, the coupling link should be adjusted for tighter coupling and the entire process repeated, starting with C_{10} set at minimum capacitance as before. After each time C_{10} is set at the point where the plate current is maximum, C_8 should be retuned carefully to the point where the plate current is minimum. This minimum should be 90 to 100 ma. when the amplifier is fully loaded. It will be found that the tuning adjustments of C_8 and C_{10} are not independent and that a certain amount of juggling back and forth between the two may be necessary to bring the tuning to the point where any further adjustment of C_{10} causes a decrease in plate current, while any adjustment of C_8 causes an increase. Table 6-I shows a typical set of operating currents and voltages.

With the power supply described, it should be possible to obtain a power output of 25 watts at the fundamental with either 3.5- or 7-Mc. crystals, 20 watts at 7 Mc. with a 3.5-Mc. crystal, or 10 watts at 14 Mc. with a 7-Mc. crystal.

Fig. 6-45 shows also the coupling arrangement for a single-wire antenna. The adjustment precedure is similar to that outlined above.

In case the keying is found to be chirpy, the value of C_4 should be changed in small steps until the chirp disappears.

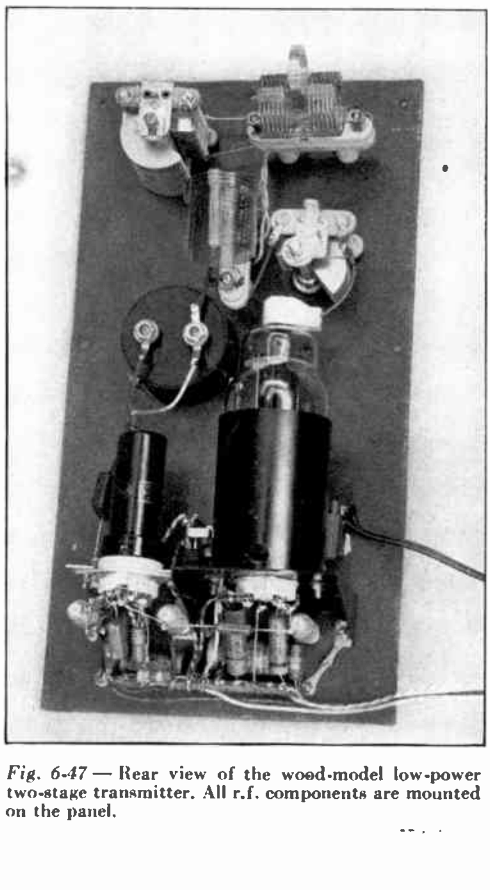


Fig. 6-47 — Rear view of the wood-model low-power two-stage transmitter. All r.f. components are mounted on the panel.

An Enclosed 150-Watt Push-Pull Transmitter

Figs. 6-48 through 6-53 show diagrams and various views of an enclosed transmitter with push-pull 807s in the output stage. Provision is made to cover the bands from 3.5 to 28 Mc. inclusive, 7-Mc. crystals being required for 28 Mc. only.

The oscillator circuit is a modified Pierce using a 6AG7 with a separate output-plate tank which may be tuned to the second or third harmonic of the crystal as well as to its fundamental frequency. The oscillator feeds a push-push 6V6 stage arranged so that there is a choice between using a single tube as a neutralized straight amplifier, or two tubes to double frequency, by operating S_2 which controls the heater of one of the 6V6s. C_7 and C_{15} are balancing condensers adjusted to compensate for the driver-tube output capacitance across the other half of the circuit. RFC_3, RFC_4, RFC_5, RFC_9, RFC_{10}, RFC_{11} and RFC_{12} are v.h.f. parasitic suppressors. RFC_{13}-C_{25}, $RFC_{14}C_{26}$ and $C_{27}RFC_{15}C_{28}RFC_{16}$ are harmonic filters to reduce the possibility of TVI. C_{29} is for the same purpose.

Two power supplies are required, one delivering up to 750 volts, 200 ma. for the output stage and the other 300 to 350 volts, 100 ma. for the exciter stages (see Fig. 6-53). Screen voltage for the 6AG7 is supplied from a voltage divider, while series voltage-dropping resistors are used in the amplifier stages. The 6Y6G is a protective tube. When excitation is removed from the output stage, the 6Y6G draws current through the screen resistor, R_{12}, dropping the 807 screen voltage to the point where the input falls below the rated dissipation level. A filament transformer is included in the transmitter.

A meter-switching system is provided so that all significant current values may be checked with the single milliammeter. All meter shunts except R_{13} are of sufficient resistance to have negligible effect upon the meter reading. R_{13} is adjusted to multiply the scale reading by 10, making the full-scale reading 250 ma.

Included also is a control system for the two power supplies whose input circuits may be plugged into two a.c. outlets at the rear of the enclosure. S_3 and S_4 are interlock switches which operate on opening or closing either of the two hinged compartments. These switches control the relay, Ry_1, whose contacts are in series with the a.c. line to the power-supply outlets. With the relay closed, the low-voltage supply is controlled by S_5 and the high-voltage supply by S_6, S_5 controlling both supplies so long as S_6 is closed. As in any protective system involving a relay, dependence should not be placed upon the relay for safety. Switch S_5 should *always* be turned off before opening the enclosure, the relay serving only in cases of forgetfulness.

Construction

Aside from the factor of appearance, the completely-enclosed construction provides safety in operation and shielding against direct radiation of harmonics. A standard 10 × 17 × 3-inch amplifier foundation makes up the two top sections. A second chassis is used inverted as a bottom deck to house the control equipment, harmonic filters, the filament transformer, etc. It also permits hinging the transmitter proper for easy inspection or servicing. The top cover is hinged, too, so that it may be swung open quickly for changing plug-in coils. Piano hinges extending the full length of the chassis make a good solid job.

The components on the upper deck are laid out to provide as much isolation as possible between input and output circuits. Thus the first and last tank condensers appear above the chassis, while the middle one is mounted below. The two exciter coils, both of which are on top for convenience in changing bands, are mounted in shielded plug-in units. The 807 sockets are submounted to bring the lower edge of the internal shields of the tubes level with the chassis. The sockets are fastened to a

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Fig. 6-48 — Front view of the push-pull 807 transmitter showing the arrangement of controls.

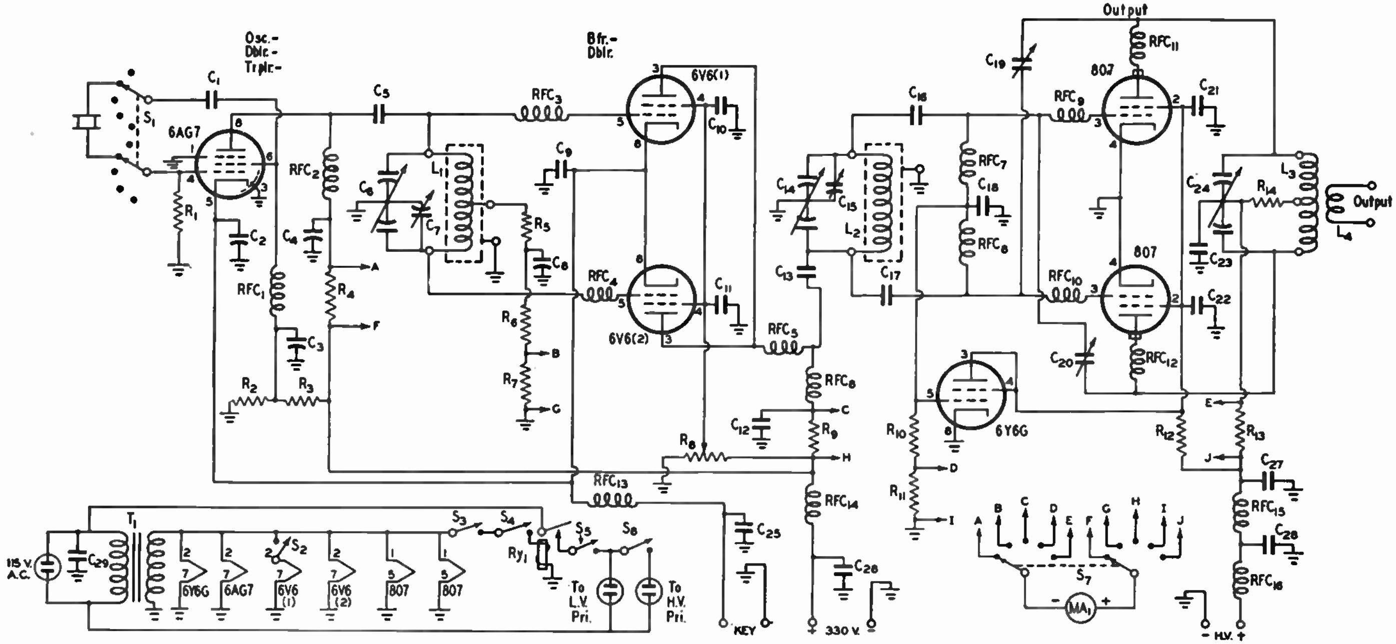

Fig. 6-49 — Circuit diagram of the enclosed 150-watt transmitter.

C_1, C_5, C_9, C_{10}, C_{11}, C_{12}, C_{13}, C_{18}, C_{25} — 0.0015-μfd. mica.
C_2, C_3, C_4, C_8 — 0.01-μfd. paper.
C_6, C_{14} — 100-μμfd.-per-section variable (Hammarlund HFD100).
C_7 — 30-μμfd. mica trimmer.
C_{15} — 50-μμfd. variable (Hammarlund HF50).
C_{16}, C_{17} — 47-μμfd. mica.
C_{19}, C_{20} — Neutralizing condensers (see text).
C_{21}, C_{22} — 0.0047-μfd. mica.
C_{23} — 0.001-μfd. 5000-volt mica.
C_{24} — 190-μμfd.-per-section variable, 0.05-inch plate spacing (Cardwell MO-180 BD).
C_{26}, C_{28}, C_{29} — 470-μμfd. mica.
C_{27} — 220-μμfd. mica.
R_1 — 12,000 ohms, ½ watt.
R_2 — 10,000 ohms, 1 watt.
R_3 — 20,000 ohms, 2 watts.
R_4, R_5, R_7, R_{11} — 220 ohms, 1 watt.
R_6 — 82,000 ohms, ½ watt.
R_8 — 25,000-ohm 7-watt potentiometer.
R_9 — Four-times shunt, wound with No. 30 copper wire.
R_{10} — 11,200 ohms, 2 watts.
R_{12} — 25,000 ohms, 10 watts.
R_{13} — 10-times shunt, wound with No. 30 copper wire.
R_{14} — 100 ohms, 5 watts.
L_1 — 3.5 Mc. — 50 turns No. 24 d.s.c., close-wound.
— 7 Mc. — 26 turns No. 24, 1¼ inches long.
— 14 Mc. — 16 turns No. 22, 1¼ inches long.
— 21 Mc. — 10 turns No. 22, 1¼ inches long.
All above coils are tapped at center; 1-inch diam.
L_2 — 3.5 Mc. — 36 turns No. 24 d.s.c., close-wound.
— 7 Mc. — 20 turns No. 24, 1¼ inches long.
— 14 Mc. — 12 turns No. 22, 1¼ inches long.
— 21 Mc. — 8 turns No. 22, 1¼ inches long.
— 28 Mc. — 6 turns No. 18, 1 inch long.
All above coils are 1-inch diameter.
L_3 — 3.5 Mc. — 34 turns No. 16.
— 7 Mc. — 20 turns No. 16.
— 14 Mc. — 12 turns No. 14.
— 21–28 Mc. — 6 turns No. 14.
All above coils are 1⅞ inches diameter.
L_4 — Link-coupling coil to suit requirements.
MA_1 — 25-ma. d.c. meter.
RFC_1, RFC_2, RFC_6, RFC_7, RFC_8 — 2.5-mh. r.f. choke.
RFC_3, RFC_4, RFC_5, RFC_9, RFC_{10}, RFC_{11}, RFC_{12} — 15 t. No. 20 d.s.c., ¼-inch diam., close-wound.
RFC_{13}, RFC_{14}, RFC_{16} — Ohmite Z-1 r.f. choke.
RFC_{15} — Ohmite Z-0 r.f. choke.
Ry_1 — 6.3-volt a.c. relay, 5-amp. contacts.
S_1 — 2-section ceramic rotary switch.
S_2, S_5, S_6 — S.p.s.t. toggle switch.
S_3, S_4 — Interlock switches (see text).
S_7 — 2-section bakelite rotary switch.
T_1 — 6.3 volts, 5 amp.

Fig. 6-50 — Top view of the push-pull 807 transmitter, showing the arrangement of components on top of the upper chassis.

2½-inch strip of aluminum spanning the bottom of the chassis. This strip is braced between the sockets by a metal spacing pillar running between the strip and the chassis.

Because of the hinged cover, all tuning controls must be brought out below the top of the chassis. In the case of the two tuning condensers that are mounted on top, the shafts are operated from controls below by means of pulleys.

Care should be used to keep parts on top of the chassis back far enough so as not to interfere with opening and closing of the cover. The crystal switch and sockets are grouped together at the left-hand end of the chassis, close to the front where they are readily accessible. There is room for several more crystals than the five shown, if desired. To the rear of the crystals are the 6AG7 and its output tank circuit with the 6V6s close to the tuning condenser, C_6, which is mounted directly on the chassis. The parasitic chokes, RFC_3 and RFC_4, are mounted in grommets set in the chassis alongside C_6, forming the connection between the stators of C_6 and the grids of the 6V6s. The compensating condenser, C_7, is supported underneath at the L_1 coil socket, between the prong that connects to the proper end of the coil and the prong that grounds the shield.

Underneath, the buffer tank condenser, C_{14}, is mounted on metal spacers to bring its shaft in line with the shaft of the crystal switch. The aluminum-strip bracket that supports the bearing for the tuning control for C_6 is then moved around until the shaft is central and in line between the shafts of the crystal switch and C_{14}. Holes are drilled in the chassis for the string connecting the two pulleys.

The balancing condenser, C_{15}, is mounted close to the right of C_{14} with its shaft pointing toward the rear so that it may be adjusted with a screwdriver through a hole from the back. National 2½-inch isolantite pillars (GS-2) are used as supports and junction points for RFC_7, RFC_8, RFC_9, RFC_{10}, C_{16}, C_{17} and the neutralizing leads. The neutralizing condensers, C_{19} and C_{20}, are made by first twisting the wire conductors out of a pair of National TPB polystyrene feed-through bushings. The holes are drilled out with a No. 35 drill and tapped for 6-32 machine screws. The bushings are then set in the chassis close to the central stator terminals of the plate tank condenser, C_{24}. Flat-head 6-32 screws, 2 inches long, are threaded into the bushings, the flat heads serving as the movable plates of the neutralizing condensers. A slot is cut in the end of the screws so that they may be adjusted with a screwdriver from the top. Connections are made by means of a soldering lug under a locking nut on top of the bushing. The necessary crossover connection is made above the chassis

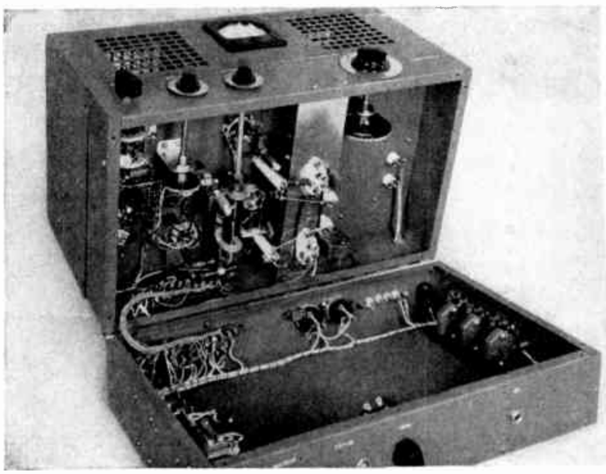

Fig. 6-51 — Bottom view of the enclosed transmitter. The lower chassis, which serves as a base, houses the filament transformer, relay, v.h.f. filter components and controls.

between the neutralizing condensers and the tank-condenser stator sections. The stationary plates of the neutralizing condensers are ½-inch washers (the top washers from the GS-2 insulators) mounted on ½-inch feed-through insulators set in the aluminum strip holding the 807 sockets.

The final-amplifier plate tank condenser is insulated from the chassis on ¾-inch cone pillars at all of the four corners except the left rear. Here a feed-through insulator, topped by a spacing washer of proper thickness, is used to provide a means of feeding the high-voltage line to the rotor of the tank condenser. C_{23} is placed immediately below, fastening it to the rear inside edge of the chassis on a metal

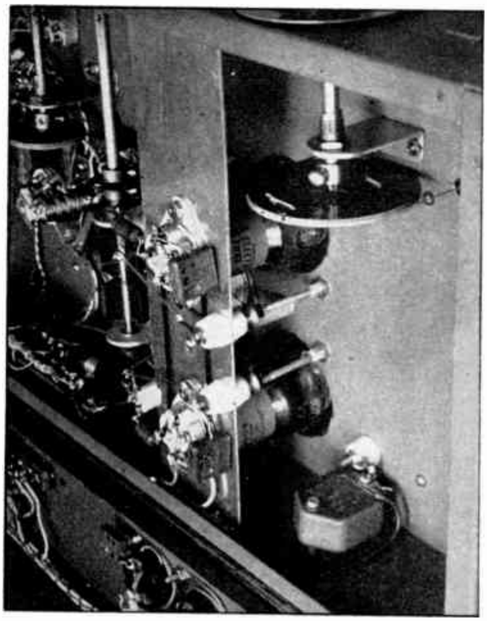

Fig. 6-52 — **Detail view of the push-pull 807s showing the neutralizing-condenser construction.**

spacer at its ground side. The link output is brought down through the chassis on feed-through insulators and then to a coaxial fitting at the rear.

All power terminations from the upper chassis are brought to a terminal board at the left. This terminal board is duplicated in the lower chassis and the corresponding points tied together through a cable, thus bridging the hinge. Along the front edge of the lower chassis, from left to right, are the excitation control, R_8, the doubler switch, S_2, the plate-supply switches, S_5 and S_6, the meter switch, S_7, and the key jack. The filament transformer and the safety-interlock relay, Ry_1, are fastened along the left-hand edge. One of the interlock switches, S_4, is mounted in the front left-hand corner. It is made from the "works" of a leaf-type open-circuit jack. When the upper chassis is closed, the jack is closed by a small cone insulator fastened to the side of the upper chassis just below the crystal switch. Both sides of this jack must be insulated. The other interlock switch operates when the cover is raised and lowered. A long screw projecting on the inside of the cover at the rear makes contact with a leaf from a 'phone jack mounted on a stand-off insulator in the rear left-hand corner of the upper chassis. Thus both the cover and the upper chassis must be shut, closing the relay, before the power-supply switches will operate.

The power-input plug for the 115-volt line is at the left-rear corner of the lower chassis, below the terminal board. C_{29} by-passes the ungrounded side of the line to the chassis. To the right are the two outlets for the high- and low-voltage transformer primaries. D.c. input connections from the power supplies are made at the Millen terminal strip and safety terminal at the right. Components of the harmonic filters are mounted on a terminal board fastened to the right-hand edge of the chassis. The meter is connected to the switch by means of long cabled flexible leads passing to the terminal board through a rubber grommet in the rear left corner of the upper chassis. The metering resistors are mounted directly on the switch.

Coils

The exciter coils are wound on Millen 1-inch bakelite forms mounted in Millen octal-base shielded plug-in units, Type 64400. The forms are fastened to the base with a machine screw after drilling and tapping the octal base at the center through the locating plug. The shield can is wired to one of the prongs so that it is connected to ground when the unit is plugged in.

The output-stage tank coils are wound on Millen 44001 polystyrene forms. The link winding, L_4, is wound with No. 14 wire covered with small spaghetti, on a diameter which will fit inside the Millen coil form where it is centered and fastened with Duco cement. On the higher-frequency coils, the link leads may be brought out between the turns of the tank coil near the center. But on the lower-frequency coils, where the turns are too close together, it is necessary to bring the leads out through the ends of the form.

Adjustment

Eighty-meter output is obtained, of course, with a 3.5-Mc. crystal and all tank circuits tuned to this band. S_2 is open, since only one of the two 6V6s is used unless the stage is doubling frequency. Seven-megacycle excitation for the final may be obtained in any of three different ways. With a 3.5-Mc. crystal, L_1C_6 may be tuned to the fundamental, doubling taking place in the buffer stage with L_2C_{14} tuned to 7 Mc. and both 6V6s in use. Equivalent results should be obtained by tuning L_1C_6 to 7 Mc. with either a 3.5- or 7-Mc. crystal and amplifying straight through with

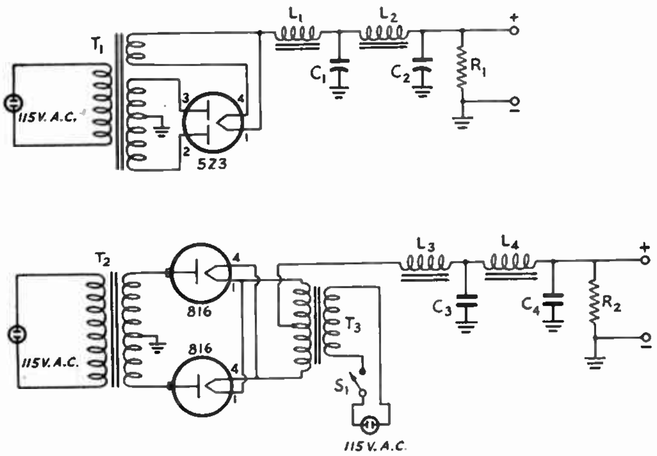

Fig. 6-53 — Wiring diagram of a suitable power supply for the push-pull 807 transmitter.

C_1, C_2 — 8-μfd. 600-volt wkg. electrolytic.
C_3, C_4 — 4-μfd. 1000-volt oil-filled.
R_1 — 20,000 ohms, 25 watts.
R_2 — 25,000 ohms, 50 watts.
L_1, L_2 — 20-hy. 100-ma. filter choke.
L_3 — 5/25-hy. 250-ma. swinging choke.
L_4 — 10-hy. 250-ma. smoothing choke.
S_1 — S.p.s.t. toggle switch, 3 amp.
T_1 — Power transformer: 350 volts d.c.,100 ma.; 5 volts, 3 amp.
T_2 — Plate transformer: 750 volts d.c., 250 ma.
T_3 — Filament transformer: 2.5 volts, 4 amp.

a single tube in the push-push buffer stage.

Fourteen-megacycle output may be obtained from a 3.5-Mc. crystal by doubling to 7 Mc. in the oscillator and doubling again in the buffer stage. With a 7-Mc. crystal, the doubling may take place in either oscillator or buffer as desired. Twenty-one-megacycle operation requires tripling frequency — from a 7-Mc. crystal — in the output circuit of the oscillator and amplifying straight through with a single tube in the buffer-amplifier, since the push-push arrangement cannot be used for tripling. Ten-meter drive for the final is obtained from a 7-Mc. crystal by doubling frequency in both oscillator and buffer.

It is best to adjust the transmitter initially at the highest frequency at which operation is desired. With a plate supply delivering between 325 and 350 volts, the oscillator plate should draw about 15 ma., kicking upward a milliampere or two at resonance. This current remains about the same regardless of whether the oscillator is doubling or working at the crystal fundamental. Resonance in the oscillator output circuit is best determined by tuning the circuit for maximum grid current to the buffer stage. This grid current will run between 1 and 2 ma.

As soon as the buffer stage has been tuned up, the screen leads should be opened up and the individual screen currents checked for balance. C_7 should be adjusted carefully until the screen currents match when the plate tank circuit is tuned to resonance. The buffer plate current at resonance normally should run between 10 and 30 ma., depending upon the band of operation and whether one or two tubes are in use, when R_8 is adjusted to deliver required drive to the final stage. If the wiring to the buffer tubes is kept closely symmetrical and the tubes themselves do not differ appreciably, no neutralizing adjustment should be necessary when working the stage as a straight amplifier with one tube inactive, since the grid-plate capacitance of the inactive tube acts as the neutralizing condenser for the active tube. However, if self-oscillation should show up, the stage can be stabilized by introducing a small amount of capacitance, such as provided by spaced pieces of wire, between the grid and plate terminals of the socket of one tube or the other, bending as required to make the neutralization complete.

Neutralization of the final amplifier is best checked with an indicator of the rectifier-milliammeter type (see "Neutralizing Procedure," this chapter). Unless the stage is already close to neutralization, a considerable reading on the indicator will be obtained when excitation is applied and the output tank circuit tuned to resonance with no screen or plate voltage applied. The neutralizing condensers should be kept at equal settings and adjusted for minimum feed-through of exciter energy as shown by the indicator. The output tank circuit should be kept tuned to resonance during the neutralizing procedure. A point in the adjustment of the neutralizing condensers should be found where both decreasing and increasing the capacitance causes an increase in the indicator reading.

When neutralizing is complete, the plate and screen voltages may be applied and the amplifier loaded up to rated input with a dummy load (a 100-watt lamp makes a good load). The next step is to balance the 807 screen currents by inserting milliammeters (or switching a single meter) in the two screen leads and adjusting C_{15} until the screen currents are equal when the plate tank circuit is tuned to resonance.

By adjusting R_8, it should be possible to set the 807 grid current to the recommended value of 8 ma. for 'phone operation. Higher grid current than this should not be permitted, since it drives the screen dissipation up unnecessarily. For c.w. operation, a total grid current of 4 to 6 ma. is sufficient for good efficiency.

The size of the output link winding will have to be determined experimentally to give proper loading with the antenna system in use.

A Three-Stage 250-Watt Transmitter

The three-stage transmitter illustrated in Figs. 6-54 through 6-58 uses a single Hytron 5514 in the output stage. A 6AG7 in a modified Pierce circuit is the crystal oscillator, arranged to deliver output at the crystal fundamental, or its second or third harmonic, depending upon the frequency to which the plate circuit is tuned. This stage drives a pair of 6L6s that can be operated as a push-push doubler, or as a neutralized single-tube amplifier.

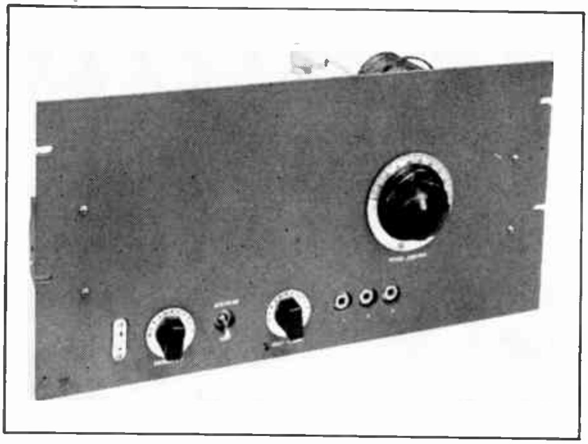

Fig. 6-54 — Front view of the 250-watt transmitter. The panel is 10¾ inches high and the chassis 8 × 17 × 3 inches. The crystal socket is to the left followed by the two exciter-stage tuning controls with the 6L6 heater switch between. The dial to the right is the control for the output-stage tank condenser. The three jacks are for keying or cathode-metering any stage.

Referring to the circuit diagram of Fig. 6-56, a combination of fixed and grid-leak bias is used in all stages. C_{18} is used to compensate for the output capacitance of the 6AG7 which is connected across the other half of the circuit.

One of the 6L6s is made inoperative when the stage is not doubling by turning off its filament by means of S_1. The plate-grid capacitance of the inoperative tube serves as the neutralizing capacitance for the remaining active tube. A split-stator tank condenser, C_9, is used in the plate circuit of the doubler stage to reduce circuit minimum capacitances.

Because both tubes are not always in operation simultaneously, separate screen voltage-dropping resistors, R_5 and R_6, are used.

The circuit diagram of a power supply capable of operating the 5514 at maximum ratings is shown in Fig. 6-57.

Construction

In Fig. 6-55, the exciter stages occupy the right-hand side of the chassis while the output-stage components are grouped to the left. In line, from front to back along the right-hand edge, are the 6AG7 and a removable shield can which covers L_1. Immediately to the left are the 6L6s and their plate tank coil, L_2.

The neutralizing condenser, C_{15}, is mounted on small stand-off insulators in front of the 5514. The lead between C_{15} and the grid terminal passes through a clearance hole in the chassis. The plate tank condenser, C_{14}, is elevated on ½-inch cone insulators to bring its terminals closer to those of the tank coil at the left.

In the bottom view of Fig. 6-58, the oscillator tank condenser, C_4, is mounted between the two rows of sockets on metal spacers to bring its shaft level with that of the doubler tank condenser, C_9, which is mounted on small stand-off insulators to the right. The 6.3-volt filament transformer, T_1, and the 7.5-volt filament transformer, T_2, are fastened to the rear and side of the chassis. The three jacks are set in a row in the front edge.

Adjustment

The oscillator should function regardless of the tuning of the plate tank circuit. If the plate coils, L_1, are made closely to the dimensions

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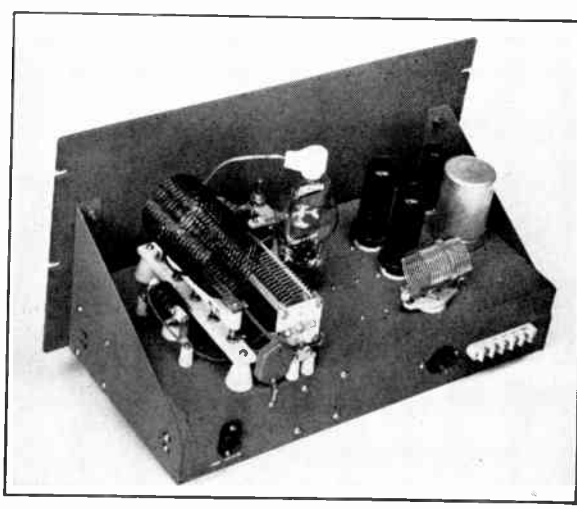

Fig. 6-55 — Rear view of the 250-watt transmitter. The plate-tank components are to the left. The tank condenser must be insulated from the chassis and provided with an insulating shaft coupling. The plate choke is to the left of the tank coil. The neutralizing condenser is mounted on small cone insulators in front of the 5514. C_{16} is fastened to the rear of the tank condenser.

The exciter tubes and tank coils are to the right with the oscillator coil shielded by a can with a removable cover. Power terminals are arranged along the rear.

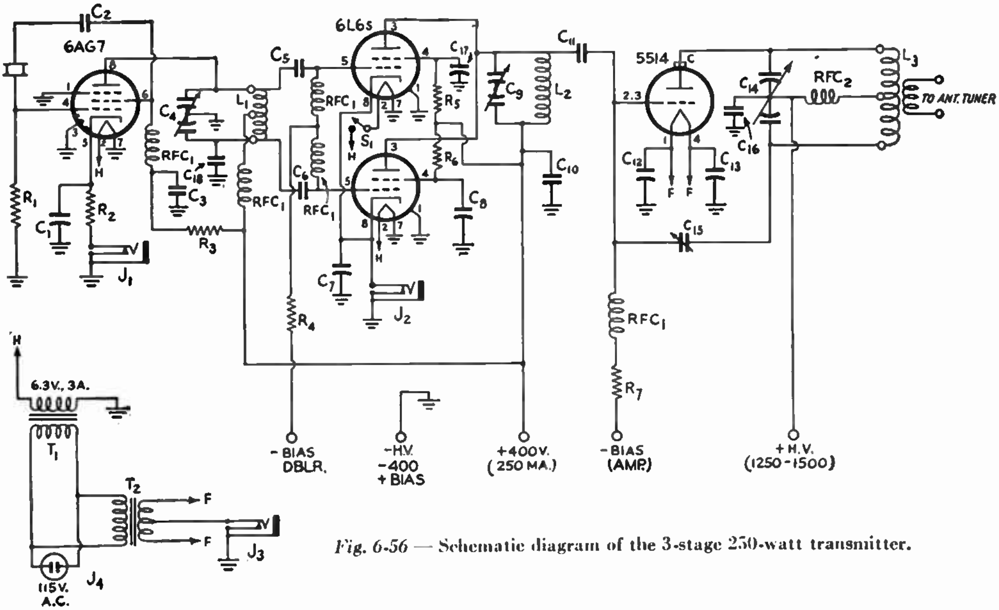

Fig. 6-56 — Schematic diagram of the 3-stage 250-watt transmitter.

C_1, C_7, C_8, C_{12}, C_{13}, C_{17} — 0.01-μfd. paper.
C_2, C_3 — 0.0047-μfd. mica.
C_4 — 140-μμfd.-per-section dual variable (Hammarlund HFD-140).
C_5, C_6 — 47-μμfd. mica.
C_9 — 100-μμfd.-per-section dual variable (Cardwell EU-100-AD).
C_{10} — 0.001-μfd. mica.
C_{11} — 47-μμfd. 1000-volt mica.
C_{14} — 100-μμfd.-per-section transmitting variable (Hammarlund HFBD-100-E).
C_{15} — Neutralizing condenser (National STN).
C_{16} — 0.001-μfd. 5000-volt mica.
C_{18} — 7.5-μμfd. ceramic (two 15-μμfd. Erie Ceramicons in series).
R_1 — 47,000 ohms, ½ watt.
R_2 — 330 ohms, ½ watt.
R_3 — 68,000 ohms, 1 watt.
R_4 — 4700 ohms, 1 watt.
R_5, R_6 — 25,000 ohms, 5 watts.
R_7 — Bias resistor (see text).
L_1 — 3.5 and 7 Mc. — 44 turns No. 22 d.s.c., close-wound, 1-inch diam., center-tapped.
— 7, 10.5 and 14 Mc. — 20 turns No. 22 d.s.c., 1-inch diam., 1⅛ inches long, center-tapped.
— 14 and 21 Mc. — 10 turns No. 22 d.s.c., 1-inch diam., 1⅛ inches long, center-tapped.
Above coils wound on Millen Type 45000 forms.
L_2 — 3.5 Mc. — B & W 80 JCL with 12 turns removed.
— 7 Mc. — B & W 20 JCL.
— 14 Mc. — B & W 10 JCL.
— 21 Mc. — B & W 10 JCL with 4 turns removed.
— 28 Mc. — Bud OCL-5.
(The above coils are supplied with center links. These link windings are unused, but need not be removed.)
L_3 — B & W TL series.
— 3.5 Mc. — 80 TL; one turn added each side.
— 7 Mc. — 40 TL.
— 14 Mc. — 20 TL.
— 21 Mc. and 28 Mc. — 10 TL.
J_1, J_2, J_3 — Closed-circuit jack.
J_4 — 115-v. a.c. male plug.
RFC_1 — 2.5-mh. 100-ma. r.f. choke.
RFC_2 — Transmitting r.f. choke (Millen 34140).
S_1 — S.p.s.t. toggle switch.
T_1 — Filament transformer, 6.3 volts, 3 amp.
T_2 — Filament transformer, 7.5 volts, 4 amp.

given, it should be possible to tune to both 3.5 and 7 Mc. with one coil, to 7, 10.5 and 14 Mc. with the second, and 14 and 21 Mc. with the third. Thus, depending on the frequency of the crystal in use at the time, the oscillator may be used to deliver output at the crystal fundamental, the second harmonic, or the third harmonic. The plate-current dip indicating resonance in the output circuit will not be pronounced, so it must be watched for carefully. Output in the 21-Mc. band may be obtained in either of two ways. In the first, a 3.5-Mc. crystal is used, and the oscillator plate circuit is tuned to 10.5 Mc. The 6L6s are then used as doublers from 10.5 Mc. to 21 Mc. If a 7-Mc. crystal is used, the plate circuit of the oscillator is tuned to the third harmonic of the crystal, and the 6L6 stage is used as a neutralized straight-through amplifier with S_1 turned off. Either method will produce adequate drive to the 5514 grid. The plate coil used for 28-Mc. operation may also be used to cover the 21-Mc. range. When the complete transmitter is in operation, the grid current to the final amplifier will be the best indicator of maximum oscillator and buffer-doubler output. With a 400-volt supply, the oscillator screen potential should be about 180 volts and the cathode current approximately 18 milliamperes, whether or not the oscillator stage is doubling frequency.

The buffer-doubler should show a cathode current of about 200 ma. with both tubes operating as doublers or 100 ma. with only the single tube in operation as a buffer amplifier. In this stage, also, the dip in cathode current at resonance is slight. Grid current to the 6L6s under load should be about 5 ma. per tube. If the difference in grid currents to the two tubes is more than 1 ma., the value of the

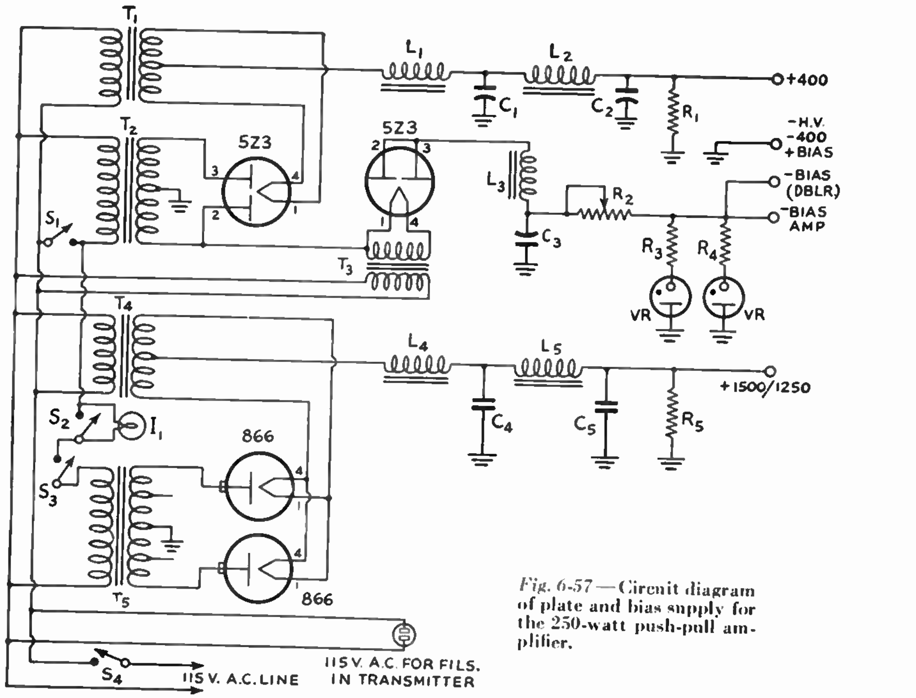

Fig. 6-57 — Circuit diagram of plate and bias supply for the 250-watt push-pull amplifier.

C_1, C_2, C_3 — 8-μfd. 600-volt-wkg. electrolytic.
C_4, C_5 — 4-μfd. 2000-volt oil-filled.
R_1 — 25,000 ohms, 25 watts.
R_2 — 30,000 ohms, 10 watts, with slider.
R_3, R_4 — 47 ohms, 1 watt.
R_5 — 25,000 ohms, 150 watts.
L_1 — 5/25-hy. 225-ma. swinging choke.
L_2 — 20-hy. 225-ma. smoothing choke.
L_3 — 30-hy. 75-ma. filter choke.
L_4 — 5/25-hy. 175-ma. swinging choke.
L_5 — 10-hy. 175-ma. smoothing choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_3, S_4 — 10-amp. toggle switch.
S_2 — 5-amp. toggle switch.
T_1, T_3 — 5-volt 3-amp. filament transformer.
T_2 — 400-v. d.c. 225-ma. plate transformer.
T_4 — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T_5 — 1500/1250-v. d.c. 175-ma.-or-more plate trans.
VR — VR-75 voltage-regulator tube.

balancing condenser, C_{18}, should be changed until balanced grid current is obtained.

With fixed bias applied, a final-amplifier grid current of 100 ma. or more should be obtained before plate voltage is applied to the 5514.

After the amplifier has been neutralized (see "Neutralizing Procedure," this chapter), plate voltage may be applied and the amplifier coupled to an antenna system and loaded. For maximum c.w. ratings, the plate voltage should not exceed 1500 volts and the plate current 175 ma., with R_7 500 ohms. Similar ratings with 100-per-cent plate modulation are 1250 volts and 142 ma., with R_7 150 ohms. The excitation should be adjusted to obtain the rated value of 60 ma. under actual operating conditions with load.

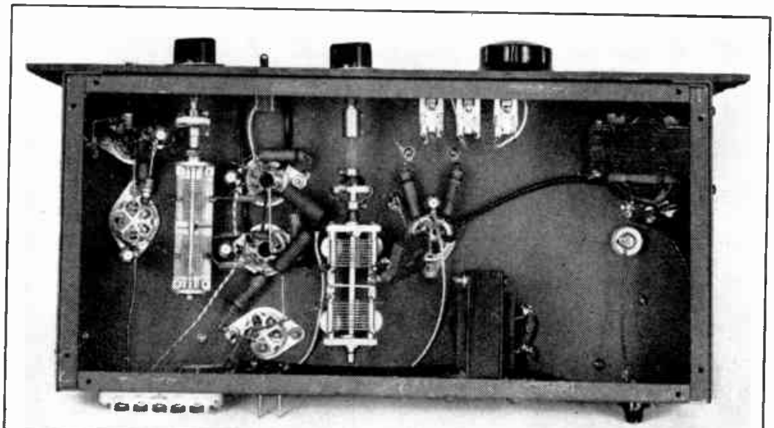

Fig. 6-58 — Bottom view of the 3-stage 250-watt transmitter. The shafts of C_4 to the left and C_9 at the center are brought to the same level by mounting spacers of suitable height. The frame and shaft of C_9 must be insulated. The two filament transformers are to the right.

An Enclosed 1-Kw. Transmitter

Figs. 6-59, 6-61, 6-62 and 6-64 show different views of an enclosed three-stage transmitter which will handle an input of 1 kw., with c.w. operation on all bands from 3.5 or 7 Mc. to 28 Mc. using either 3.5- or 7-Mc. crystals. Push-pull 813s are used in the final amplifier.

Circuits

Fig. 6-60 shows the circuit of the exciter. Two 807s are used in an arrangement that permits either of the two stages to be used as a Tri-tet oscillator. The second stage may be operated also as a frequency doubler.

The arrangement shown has a number of advantages. The two 807 stages need not be operated at the same frequency for output on any band, thereby avoiding stabilization difficulties sometimes encountered with tubes of this type when operating as straight amplifiers. It provides for oscillator keying for break-in work at 3.5 and 7 Mc. and amplifier keying for 14 and 28 Mc. where chirp with oscillator keying might become objectionable. Shielding problems are reduced because the first 807 never is called upon to operate at the same frequency as either of the two following stages. A single set of coils suffices for both stages of the exciter.

When both 807 stages are in use, the crystal and cathode coils are plugged into the first stage, while a jumper closes the cathode circuit of the second stage. When the second 807 is used as an oscillator, crystal and cathode coils are transferred to the second stage and removal of the coil in the plate circuit of the first stage breaks the connection between the two stages.

Parallel plate and grid feed is used in both exciter stages, while the screens are fed through series voltage-dropping resistors. The value of the resistor for the screen of the second tube may be varied in steps by S_1 and this provides a means of adjusting the excitation to the final amplifier. Jacks are provided for shifting a milliammeter from one circuit to another to obtain readings of plate current in either exciter stage or grid or screen current to the final amplifier. RFC_2, RFC_5, R_2 and R_5 are inserted to prevent v.h.f. parasitic oscillation. The key is in the cathode of the second 807 stage.

The circuit of the push-pull 813 final amplifier is shown in Fig. 6-63. Series feed is used in both grid and plate circuits. Although the 813 is a screened tube, experience has shown that neutralization is necessary to prevent self-oscillation. A single set of coils suffices for L_2 in both the amplifier section and the exciter section. Variable-link output is provided for coupling to the antenna system.

Construction

The r.f. circuit components are assembled on a panel of ¼-inch crackle-finished tempered Presdwood, 18 inches wide and 19 inches high, backed with copper screening. The panel is hinged at the bottom so that it may be tipped outward, as shown in Fig. 6-59, for changing plug-in coils, etc. Suspended from the back of the panel is a vertical partition of ⅝-inch plywood, 13¼ inches wide and 18 inches high. This partition also is covered on both sides with copper screening from the bottom edge up to within 6 inches of the top. The lower edge of the partition comes ⅝ inch above the lower edge of the panel and is placed 11¼ inches from the right-hand edge of the panel as viewed from the front. A strip of ⅝-inch plywood 1¼ inches wide runs across the top of the panel, as shown in Fig. 6-64, and a similar strip ⅝ inch wide runs across the bottom of the panel, flush

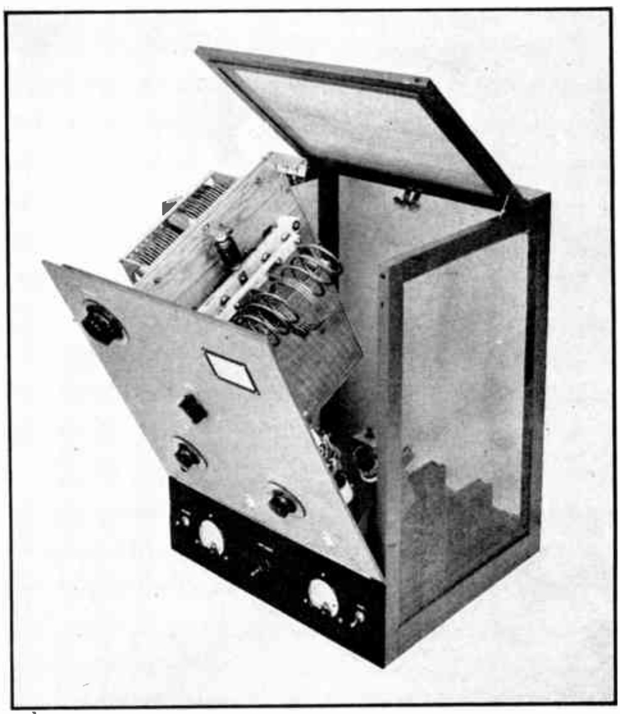

Fig. 6-59 — A three-stage 3.5–30-Mc. kilowatt transmitter using push-pull 813s. Measuring 18 inches wide, 24 inches high, and 16 inches deep, it fits readily on most operating tables. The hinged front panel drops down for coil changing. Interlock switches combined with complete enclosure ensure against accidental contact with any high-voltage circuits when the power is on.

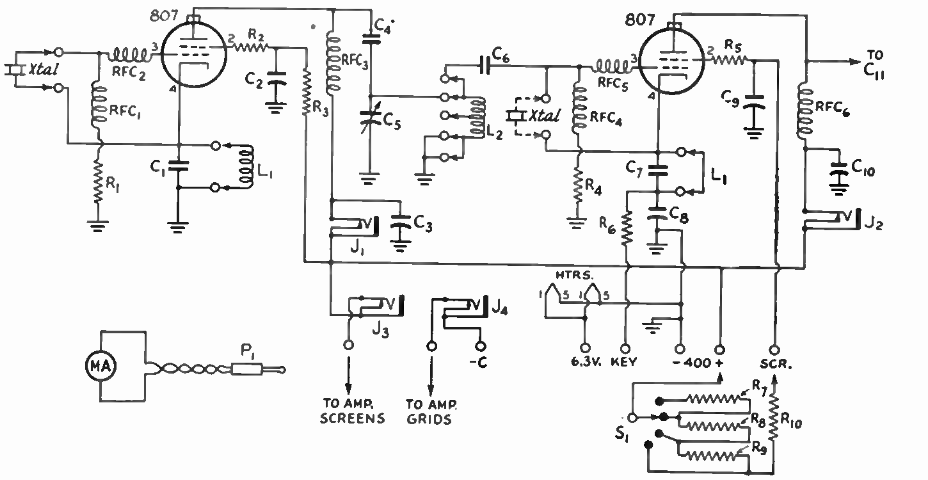

Fig. 6-60 — Circuit diagram of the exciter section of the enclosed 1-kw. transmitter.

C_1, C_7 — 150-μμfd. mica.
C_2, C_3, C_9, C_{10} — 0.01-μfd. paper, 600 volts.
C_4 — 470-μμfd. mica, 1000 volts.
C_5 — 50-μμfd. variable (National ST-50).
C_6 — 100-μμfd. mica, 1000 volts.
C_8 — 0.0047-μfd. mica.
R_1 — 0.33 megohm, 1 watt.
R_2, R_5 — 50 to 100 ohms, ½-watt carbon.
R_6 — 330 ohms, 1 watt.
R_3 — 25,000 ohms, 10 watts.
R_4, R_7 — 47,000 ohms, 1 watt.
R_{10} — 10,000 ohms, 1 watt.
R_8, R_9 — 22,000 ohms, 1 watt.
L_1 — Tri-tet cathode coil: for 3.5-Mc. crystals, 13 turns No. 22 d.s.c., close-wound, diameter 1 inch, shunted by extra 75-μμfd. mica condenser inserted inside form. For 7-Mc. crystals: 3 turns No. 22 d.s.c., close-wound on 1-inch diameter form. (Cathode coils wound on Millen No. 45004 forms.)
L_2 — Plate coil; for 3.5 Mc., 36 turns No. 22 d.s.c., close-wound; 7 Mc., 16 turns No. 18 bare, length 2 inches; 14 Mc., 8 turns No. 18 bare, length 2 inches; 28 Mc., 4 turns No. 18 bare, length 1 inch. All coils wound on 1⅛-inch diameter forms (Millen No. 44001) and tapped at center.
J_1, J_2, J_3, J_4 — Closed-circuit jack.
MA — 0–200 d.c. milliammeter.
P_1 — Insulated plug.
RFC_1, RFC_3, RFC_4, RFC_6 — 2.5-mh. r.f. choke.
RFC_2, RFC_5 — Parasitic chokes; 18 turns No. 20 d.c.c. on ¼-inch diameter form (a high-value 1-watt resistor is suitable as a form).
S_1 — 4-position single-pole rotary switch.

with the lower edge. The bottom strip provides a means of fastening the hinge to the panel. Thin strips of wood along the vertical edges of the panel serve to hold the copper screening in place, while a metal strip is used to bind the edges of the partition.

The portion of the r.f. circuit shown in Fig. 6-60 is built as a subassembly on a 5 × 10 × 3-inch chassis, fastened to the panel in the lower left-hand corner and braced by the partition, as shown in Figs. 6-61 and 6-62. As viewed from the rear, the first 807 is placed near the left-hand edge of the chassis with its cathode-coil socket behind. The tuning condenser, C_5, is placed between the 807 and the first-stage tank coil, 4 inches from the outside end of the chassis. The socket for the cathode coil for the second 807 is in the rear corner to the right. The socket for the second 807 is set in the right-hand edge of the chassis and a clearance hole for the base of the tube is cut in the partition so that the tube protrudes horizontally as shown in Fig. 6-62. The two crystal sockets are fastened to the front edge of this chassis and protrude through holes cut in the panel.

In the rear edge of the chassis are the four meter jacks and strips bearing the terminals indicated in Fig. 6-60.

The 813s and their input-circuit components are made up as another subassembly on a 5½ × 9½ × 1½-inch chassis which is fastened centrally on the partition with its bottom edge 3½ inches up from the bottom edge of the partition, as shown in Fig. 6-64. The two tube sockets are submounted as close as possible to the ends of the chassis with their centers 2½ inches in from the outside edge. Underneath, as shown in Fig. 6-62, the coil socket is

TABLE 6-II

Comb.	Xtal *f*	Output *f*	1st 807 Plate	2nd 807 Plate
A	3.5	3.5	—	3.5
B	3.5	7	—	7
C	7	7	—	7
D	3.5	7	3.5	7
E	3.5	14	7	14
F	7	14	7	14
G	3.5	28	14	28
H	7	28	14	28

centered on the tube sockets and the tank condenser, C_{13}, is to the left. It is placed so that its dial will balance the dial of C_5 on the panel. The strips which form the neutralizing condensers are mounted on feed-through insulators set in the chassis about 1⅛ inches from the bases of the tubes. A ceramic strip set in the rear edge of the chassis bears the power-supply terminals indicated in Fig. 6-63.

The plate tank condenser for the 813s is fastened to the partition above the screening, the wood serving to insulate the frame and rotors from ground as required. The output tank-coil jack bar is mounted on short stand-off insulators on the opposite side of the partition, with its center 4 inches below the top.

The link control shaft is coupled to a knob centered on the panel by means of a pair of Millen universal-joint type shaft couplings. Leads from the link terminals are brought to the upper rear corner of the partition where they connect to a pair of banana plugs which slide into jacks connected to feed-through insulators set in the back of the enclosure. These serve as the output terminals.

Below the main panel is a smaller panel 5¼ inches high which is used for the control switches and meters.

Plenty of room is available within the enclosure to the rear of the panel assembly for filament transformers and the additional terminals, indicated in Fig. 6-65, which serve as junctions between the external and internal power leads. One of the two interlock switches is of the push-button type. This is mounted on an aluminum bracket fastened to the floor of the enclosure so that the rear end of the lower edge of the partition closes the switch when the panel is hinged back into place. The circuit is broken as soon as the panel is tipped forward. The second interlock is fastened to the side of the enclosure where it is operated by the top lid.

The frame of the enclosure is made up of 1¼-inch strips cut from ⅝-inch plywood. It is 18 inches wide, 24½ inches high and 16 inches deep overall, and has a solid bottom of the same plywood material. The copper screening is tacked over the outside of the panels formed by the framework and the edges of the screening are covered with thin strips of wood. The woodwork is finished off in gray enamel.

Power Supply

The diagram of a suitable power supply for operating the 813s at maximum rated input is shown in Fig. 6-66. At current tube prices, however, the 813s are an economical proposition for operation at considerably less than maximum ratings if power-supply considerations make this desirable.

TABLE 6-III

Comb.	*1st 807*	*2nd 807*	*Final Amp.*	*Undesired Harm., 2nd 807*
A	—	75	45	None
B	—	15	25	3rd–90
C	—	15	25	None
D	50	15	25	3rd–90
E	60	20	75	5th–60/6th–85
F	60	20	75	3rd–90
G	70	80	85	6th–20/7th–53
H	70	80	85	3rd–23

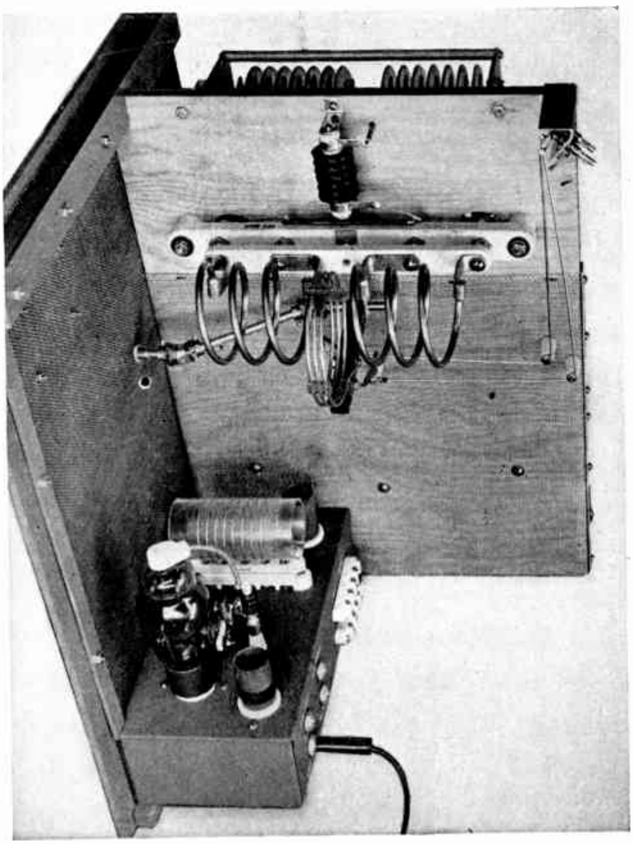

Fig. 6-61 — The exciter unit of the enclosed 1-kw. transmitter is mounted between the vertical partition and the panel, and serves as additional support for both. This view shows the first 807 with its Tri-tet cathode coil in the foreground. The plate tuning condenser is partly concealed by the tube, and to its right is the plate choke. The small bakelite coil form behind the plate coil is the jumper for the second-tube cathode circuit; when the latter tube is used as the oscillator a Tri-tet cathode coil goes in this socket.

The final tank-coil assembly and plate choke are at the top. The banana plugs at the right automatically engage a pair of jacks on the rear of the case when the panel is in position, providing connection to the antenna binding posts.

Adjustment

Since there is only a maximum of three tuned circuits to adjust, tuning the transmitter for any desired output frequency is a relatively simple job. Undesired harmonic responses, always to be found in a multiband rig, are quite readily identifiable as such, so there is little danger of tuning up on the wrong harmonic if a little care is used in watching the dial readings. For reasons given previously, the design does not provide for operating the first 807 output at the same frequency as that of the final amplifier and this sort of operation should not be attempted, even if the constructor is willing to make the extra coils, since the simplified construction does not provide the necessary shielding between the input and output circuits of the final amplifier.

The accompanying Table 6-II shows the correct coils to be used in the 807 tank circuits, depending upon the crystal frequency and the desired output frequency. Combinations A, B and C, for 3.5- and 7-Mc. output, permit break-in operation in these two bands, since in each of these cases the keyed stage (the second 807) is operating as the oscillator. In the remainder of the combinations, the keyed stage operates as a frequency multiplier, the first 807 becoming the oscillator, which runs continuously.

Table 6-III gives the approximate dial settings for resonance in the three tank circuits. Variations in wiring or coil dimensions will, of course, alter these readings, but they may be used as a guide. The last column to the right shows the dial setting where undesired crystal-harmonic responses may be expected in the multiplier circuit, the unwanted harmonics being identified.

It is advisable initially to choose one of the first three combinations from Table 6-II — one that requires the use of only two stages. With the proper coils and crystal plugged in (be sure to plug the crystal in the right socket!), the low-voltage and bias supplies may be turned on and the key closed. At some point within the range of the 807 tank condenser the plate current should dip to a minimum, rising on either side. If the tank circuit is tuned to the crystal frequency (3.5 Mc. with a 3.5-Mc. crystal or 7 Mc. with a 7-Mc. crystal) the crystal will usually stop oscillating entirely, as indicated by a sudden increase in plate current

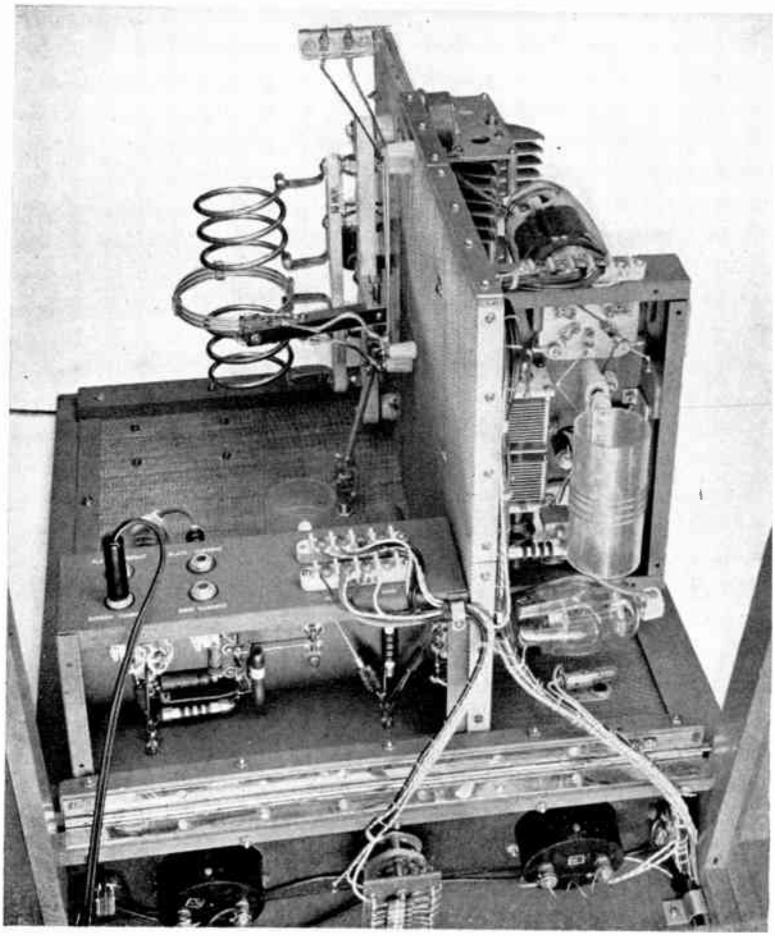

Fig. 6-62 — View from the back of the enclosed 1-kw. transmitter with the front panel dropped down. The second 807 projects through the partition to bring its plate near the tuned circuit that feeds the amplifier grids. The 807 plate choke, visible just below the amplifier grid-tuning condenser, is supported by a feed-through insulator mounted on the end of the exciter chassis. The four metering jacks are accessible only when the case and interlock switches are open, providing for safety.

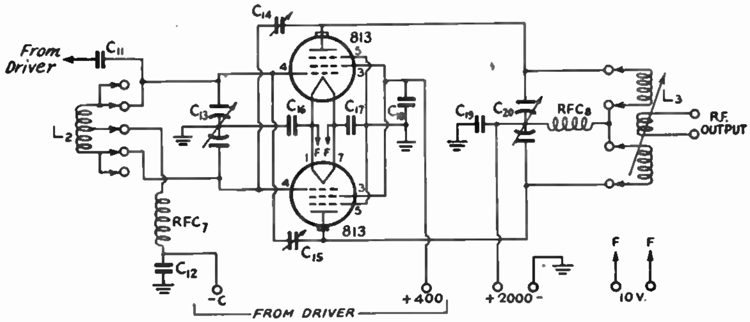

Fig. 6-63 — Circuit of the final amplifier in the enclosed 1-kw. transmitter.

C_{11} — 470-μμfd. mica, 1000 volts.
C_{12}, C_{16}, C_{17} — 0.0047-μfd. mica.
C_{13} — 100-μμfd.-per-section variable (Cardwell ER-100-AD).
C_{14}, C_{15} — Neutralizing condensers; see text.
C_{18} — 0.001-μfd. mica, 1000 volts.
C_{19} — 0.001-μfd. mica, 5000 volts working.
C_{20} — 100-μμfd.-per-section variable (National TMA-100DA).
L_2 — Same as L_2 in Fig. 6-60.
L_3 — Amplifier plate tank coils, Barker & Williamson HDVL series with following modifications:
— 3.5 Mc.: 9 turns shorted out at each outer end.
— 7 Mc.: 2 turns shorted out at each outer end.
— 14 Mc.: 1 turn shorted out at each outer end.
— 28 Mc.: No modification.
RFC_7 — 2.5-mh. r.f. choke.
RFC_8 — 2.5-mh. r.f. choke, 500 ma. (Hammarlund CH-500).

to a high value when the tank circuit is tuned to the high-capacity side of the dip in plate current. The best tuning adjustment under these circumstances is a bit to the low-capacity side of the dip. When the oscillator tank circuit is tuned to a harmonic of the crystal frequency, the circuit normally will continue to oscillate regardless of the setting of the tank condenser. Tuning is then merely a matter of adjusting the tank circuit to resonance as indicated by the plate-current dip or by maximum final-amplifier grid current. Tuning the tank circuit to resonance should result in a reading of grid current to the final amplifier. The oscillator tank circuit can then be adjusted to a point that will give maximum amplifier grid current consistent with reliable keying. During the adjustment the key should not be held closed longer than is absolutely necessary, since the amplifier screen current will run to a considerably higher-than-normal value without plate voltage and load.

Neutralizing

The amplifier should be neutralized at this point. At the start the aluminum strips (shown in Fig. 6-64) should be exactly alike — about ½ inch by 3 inches, with the centers of the feed-through insulators on which they are mounted 1⅛ inches from the tube bases. The adjustment consists of clipping off the ends of the strips about 1/16 inch at a time until the amplifier is neutralized. Any of the usual indicators of neutralization may be employed.

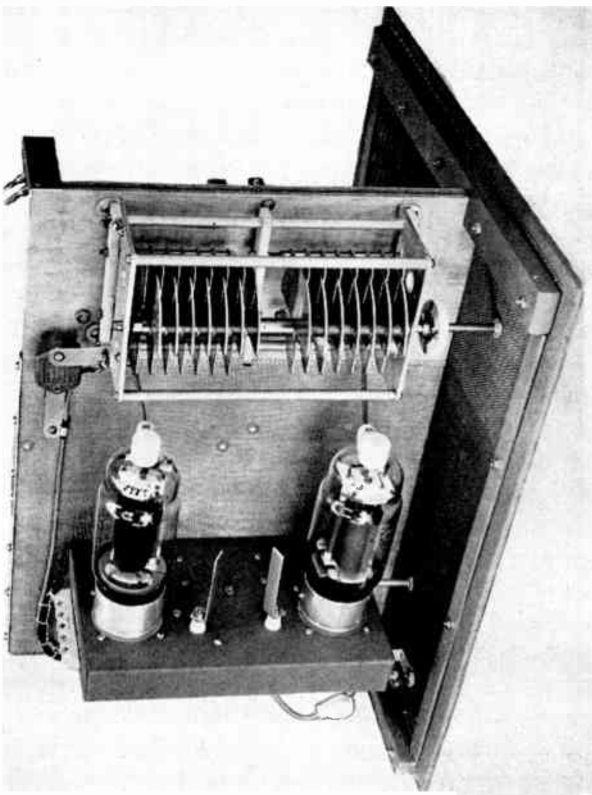

Fig. 6-64 — The amplifier side of the partition in the enclosed 1-kw. transmitter. This view shows the plate tank condenser and the plate by-pass. The aluminum strips between the two 813s, mounted on stand-off insulators, are the neutralizing condensers.

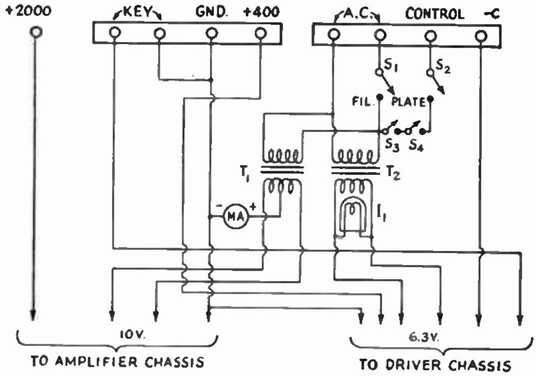

Fig. 6-65 — Schematic of the filament-power and power-terminal wiring of the enclosed 1-kw. transmitter. Two interlocks are included in the control circuit, to turn off the plate power whenever the enclosure is opened.

I_1 — 6.3-volt pilot-lamp assembly.
MA — 0–100 milliammeter, shunted to read 1000 ma.
S_1 — 10-amp. toggle switch.
S_2 — 15-amp. toggle switch.
S_3, S_4 — Interlock switches. See text.
T_1 — 10-volt 10-ampere filament transformer.
T_2 — 6.3-volt 3-ampere filament transformer.

After neutralizing, reduced high voltage may be applied to the amplifier and its plate tank circuit tuned to resonance as indicated by a dip in plate current. The antenna should not be coupled to the output stage until it has been tuned as described, the link being swung as far out as possible during preliminary adjustment. When the point of resonance has been found, full voltage may be applied and the antenna circuit coupled and tuned in the manner proper for the type of antenna system and antenna tuning arrangement used. Regardless of the system employed, it should be remembered that the last adjustment in coupling and adjusting the antenna to the transmitter is that of tuning the amplifier tank circuit for minimum plate current. This should always be done after every adjustment of antenna coupling or tuning. Otherwise, the amplifier may be operating very inefficiently off resonance and exceeding the dissipation rating of the tubes. Tuning the first 807 as an oscillator is similar to the method outlined for the second tube.

The adjustment of the second tube as a multiplier is simply the selection of the proper coil and tuning to resonance, making certain that it is not tuned to an undesired harmonic.

Table 6-IV shows typical current values. They were taken at 400 volts and may vary somewhat because of differences in crystal response and length of leads in the oscillator circuits which may affect feed-back.

The Tri-tet oscillator may be expected to self-oscillate with the crystal removed. With the crystal operating normally, however, no trouble of this sort should be experienced.

At the maximum rated input of just under 1 kw., this transmitter should deliver 600 watts or better on all bands.

TABLE 6-IV

Combination	*1st 807 at Resonance — Key Open — Ma.*	*1st 807 at Resonance — Key Closed — Ma.*	*2nd 807 at Resonance — Ma.*	*Final Grid — Ma. No Plate Voltage*	** Final Grid — Full Load — Ma.*
A	—	—	40–60	40–50	40–50
B	—	—	75	40	42
C	—	—	65–70	43	46
D	46	48	80	38	39
E	15	20	64	30	30
F	16	100	78	30	32
G	40	50	86	18	25
H	73	55	81	18	28

* With excitation control at maximum.

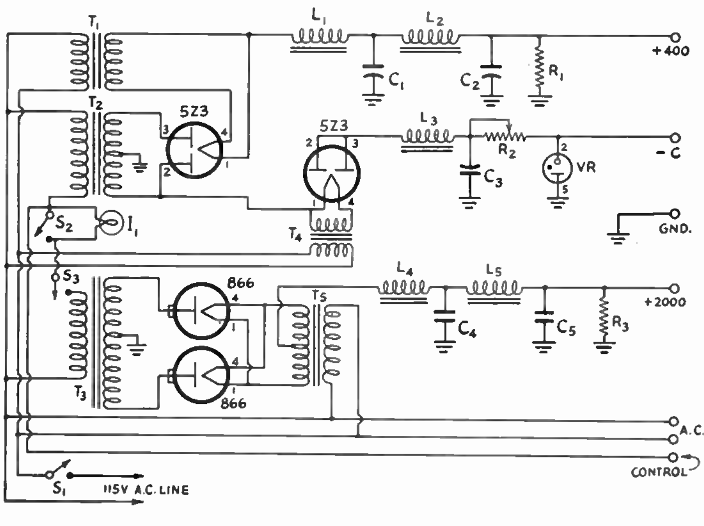

Fig. 6-66 — Circuit diagram of a power supply for the enclosed 1-kw. transmitter.

C_1, C_2, C_3 — 8-μfd. 600-volt-wkg. electrolytic.
C_4 — 2-μfd. 2500-volt oil.
C_5 — 4-μfd. 2500-volt oil.
R_1 — 20,000 ohms, 25 watts.
R_2 — 50,000 ohms, 25 watts.
R_3 — 50,000 ohms, 200 watts.
L_1 — 5/25-hy. 300-ma. choke.
L_2 — 20-hy. 300-ma. choke.
L_3 — 30-hy. 50-ma. choke.
L_4 — 5/25-hy. 500-ma. choke.
L_5 — 20-hy. 500-ma. choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_3 — 10-amp. toggle.
S_2 — 5-amp. toggle.
T_1, T_4 — 5-v. 3-a. fil. trans.
T_2 — 400 v. d.c., 250–300 ma.
T_3 — 2000 v. d.c., 500 ma.
T_5 — 2.5 volts, 10-amp., 10,000-volt insulation.
VR — Voltage-regulator tube.

A High-Stability VFO

Figs. 6-67 through 6-73 illustrate the construction of a well-stabilized VFO delivering sufficient output at 1.75, 3.5 or 7 Mc. to drive any small triode or tetrode stage up to and including one or two 807s. The two output frequences are chosen so that if the VFO is used to drive a crystal-oscillator stage in the transmitter, the oscillator stage may be operated as a doubler to avoid the possibility of oscillation in the crystal stage.

Referring to the diagram of Fig. 6-68, a 6AG7 pentode is used in the electron-coupled series-tuned Colpitts oscillator circuit. C_1 is the bandspread tuning condenser which covers the fundamental range of 1750 to 2000 kc. (or 3500 to 4000 kc.). C_2 is a padder to provide a fixed minimum circuit capacitance. C_3 and C_4 are the tube-shunting capacitances. Since the screen, which serves as the plate in the oscillating circuit, is grounded, the cathode is above ground and therefore must be returned to ground for d.c. through a choke.

The output circuit (RFC_2) is nonresonant and is capacity coupled to a 6L6 output stage fitted with plug-in coils so that it may be operated at either 1.75, 3.5 or 7 Mc. The tuning condensers of the oscillator and amplifier are ganged.

A power supply is included in the unit. Screen and plate voltages for both stages are taken from a VR-tube voltage divider. The regulator tubes are used both as a convenient voltage-divider arrangement and to limit the shaping of the keying characteristic entirely to any key-click filter that may be used with the unit.

Construction

In the unit shown in the photographs, the frequency-determining tank is isolated from the rest of the circuit by enclosing it in a standard steel box 5 × 6 × 9 inches. The tuning condenser is mounted on the top plate of a 4 × 4 × 2-inch steel box with metal brackets that space the bottom edges of the condenser end plates 1/8 inch from the plate.

The coil is removed from its original mounting, the link removed, and the coil remounted on a 3/4-inch cone insulator at the forward end and a small feed-through insulator at the rear. The first quarter turn at the front end of the coil is broken loose and a short connection between the adjacent condenser terminal and the coil at this point is made with a piece of heavy wire. This serves as a brace for the coil against vibration. Another short piece of heavy wire goes from this point to a small feed-through insulator set directly below in the top plate. This feed-through insulator and the one at the rear end of the coil serve in making connections to the condensers on the underside of the plate.

The adjustable padder, C_2, is mounted centrally on the underside of the plate with its shaft pointed toward the right. The end of the shaft is slotted for a screwdriver and holes

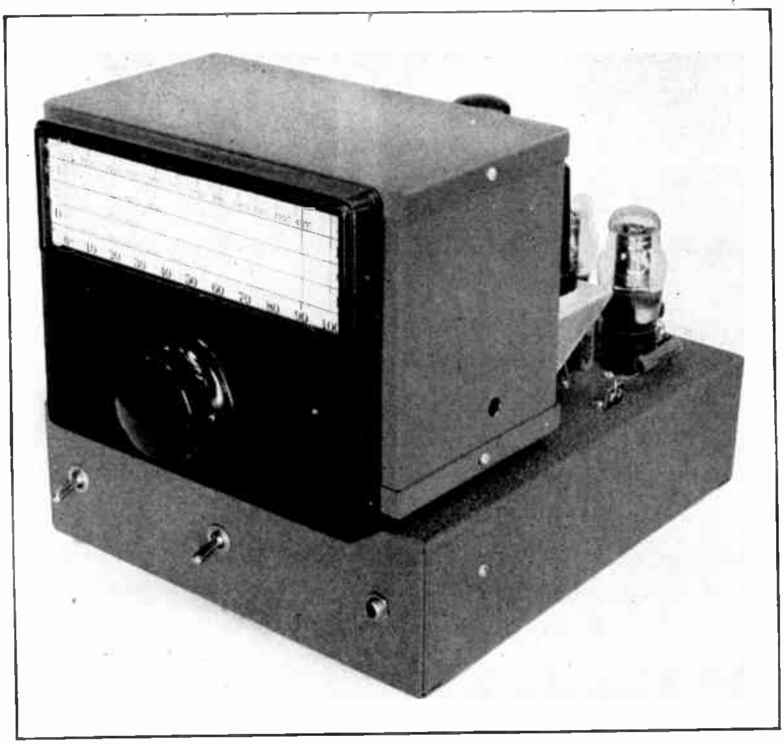

Fig. 6-67 — The completed VFO. The entire r.f. section is floating on an anti-shock mounting. The hole in the side of the box is to permit adjustment of the frequency range

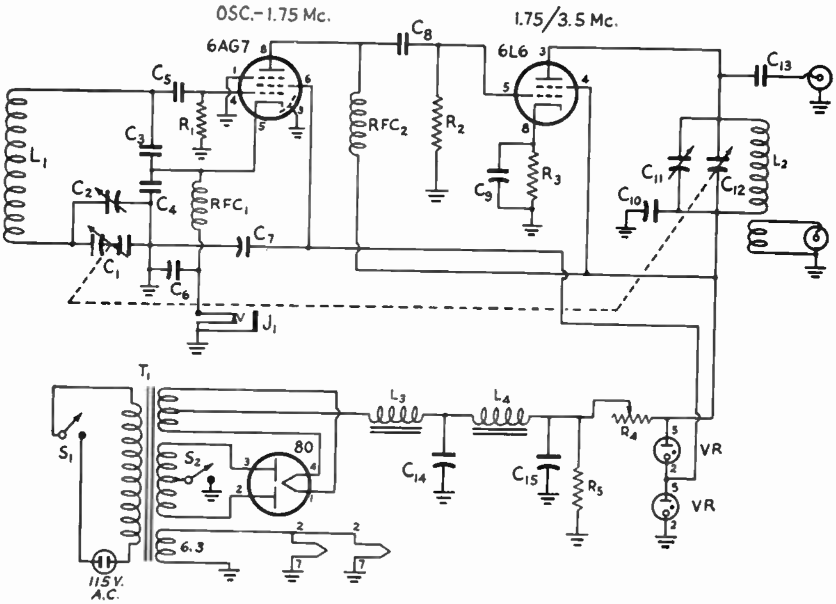

Fig. 6-68 — Circuit diagram of the series-tuned VFO.

C_1 — 50-μμfd.-per-section variable (Millen 23050).
C_2 — 100-μμfd. variable (Millen 19100).
C_3, C_4 — 0.001-μfd. zero-temp. mica.
C_5, C_8 — 100-μμfd. mica.
C_6, C_7, C_9, C_{10} — 0.01-μfd. paper.
C_{11} — 1.75 and 3.5 Mc. — 45-260-μμfd. mica trimmer. 7 Mc. — 100-μμfd. air trimmer (Hammarlund APC-100).
C_{12} — Approx. 75-μμfd. variable (Millen 22100 with 3 stator plates removed).
C_{13} — 220-μμfd. mica.
C_{14}, C_{15} — 16-μfd. 450-volt electrolytic.
R_1 — 47,000 ohms, ½ watt.
R_2 — 0.1 megohm, 1 watt.
R_3 — 470 ohms, 1 watt.
R_4 — 1000 ohms, 10 watts, adjustable.
R_5 — 50,000 ohms, 10 watts.
L_1 — 1.75 Mc. — 140 μh. (National AR-160).
3.5 Mc. — 35 μh. (B & W JEL-80).
L_2 — 1.75 Mc. — 37 turns No. 22 d.c.c., 1½ inches diam., close-wound.
3.5 Mc. — 16 turns No. 22, 1½ inches diam., ⅞ inch long.
7 Mc. — 14 turns No. 20, 1½ inches diam., 1½ inches long, tapped 5½ turns from ground end for C_{11}.
L_3, L_4 — 14-h. 100-ma. filter choke (UTC R-19).
J_1 — Closed-circuit jack.
RFC_1, RFC_2 — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — Power transformer: 350-350 r.m.s., 90 ma.; 5 volts, 3 amp.; 6.3 volts, 3.5 amp.

are drilled in the sides of both inner and outer boxes so that the padder may be adjusted from the outside after the unit has been assembled. The mica condensers, C_3 and C_4, are fastened alongside the padding condenser by cementing them to the plate with Duco cement to eliminate movement. The top lip of the small box may have to be notched out in a few places before the top plate will fit in place.

Discarding the bottom plate of the small box, the height

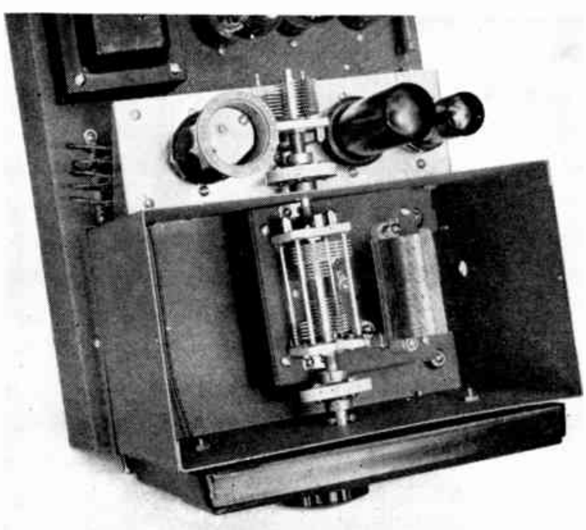

Fig. 6-69 — The oscillator tank circuit is isolated from the remainder of the circuit by enclosing it in a shock-proof metal box.

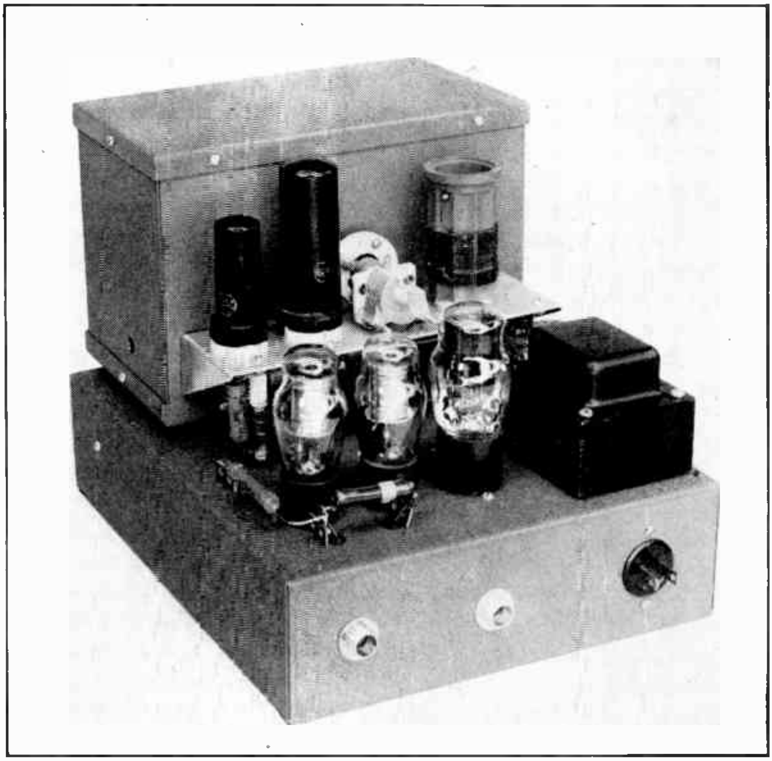

Fig. 6-70 — Rear view of the VFO. The oscillator tube and amplifier components are mounted on a shelf fastened to the floating box. Power-supply components occupy the rear portion of the chassis.

of the tuning-condenser shaft above the lower edge of the box should be measured carefully and 1-inch clearance holes cut centrally in the outer box at this same level. Placing the smaller box inside, with its rear face against the back wall of the outer box and with the tuning-condenser shaft lined up with the shaft holes, the position of the smaller box should be marked on the rear wall. Then the top plate should be removed, the small box replaced, and holes marked in the bottom of the outer box so that the smaller box can be fastened in place with screws up through the bottom. With this done, a grommet hole for the leads to the oscillator tube should be drilled through the rear of both boxes simultaneously near the oscillator-tube socket. Three leads — connections to the grid condenser, C_5, to the cathode, and to the ground point of the cathode by-pass condenser of the oscillator tube — are bunched together and brought out through this hole.

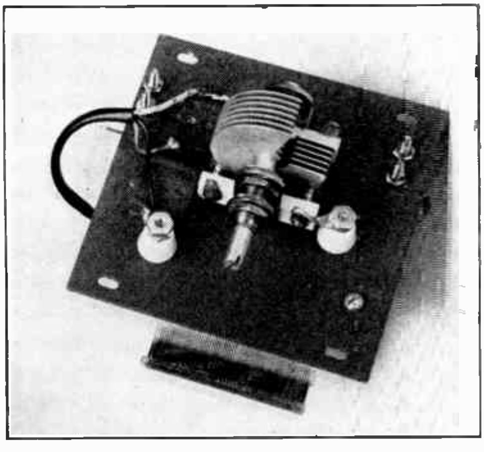

Fig. 6-71 — Bottom view of the oscillator-tank unit, showing the placement of the oscillator tuning padder and the tube-shunting condensers.

With the oscillator-tank unit fastened in place within the large box, and flexible insulated couplings on each end of the tuning-condenser shaft, the dial can be lined up and its mounting holes marked on the front of the outer box. The lower edge of the dial plate will overhang a half inch or so at the bottom.

The remainder of the r.f.-circuit components are assembled on a 2½ × 8-inch aluminum shelf fastened to the rear of the box to isolate the tank components from the heating of the tubes. The amplifier tuning condenser, C_{12}, must be insulated from the shelf. The height of the shelf is adjusted, after the condenser has been mounted, so that its shaft lines up with the tail shaft of C_2. Wiring and associated small parts are placed under the shelf. All power-supply connections and the key connection are made to a 5-point lug strip at the left-hand end of the shelf.

The entire unit is guarded against mechanical vibration by mounting the box on rubber grommets. A grommet is placed in each of the four corners of the bottom of the

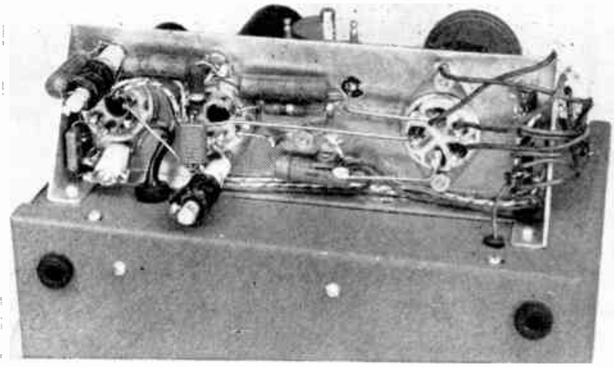

Fig. 6-72 — Wiring underneath the shelf. The loose flexible leads at the right are anchored to a lug strip on the chassis after assembly.

box. These are duplicated in the top of the 10 × 12 × 3-inch chassis which serves as a base. Machine screws with washers at either end are passed through both sets of grommets to fasten the floating box to the chassis. Care should be taken in locating the grommet holes in the chassis to provide 1/16 inch or so of clearance between the lower overhanging edge of the dial plate and the chassis, so that the dial is free from contact with the chassis.

A duplicate lug strip is fastened to the chassis directly below the terminal strip on the shelf. The two strips are then connected together with highly-flexible wire bent to form half loops between the terminal strips. This is done to minimize any vibration that might be transmitted from the base chassis to the box through the connecting leads. Similar flexible connections are made to anchorages on the chassis for the output leads.

The output coil, L_2, is wound on a standard 1½-inch diameter 5-prong plug-in form (Bud). The padder condenser, C_{11}, in each case is mounted inside the form where it may be adjusted with a screwdriver.

The power transformer, rectifier, the two VR tubes and their voltage-dropping resistor, R_4, as well as the bleeder resistor, R_5, are mounted along the rear edge of the chassis. The filter chokes and condensers are placed underneath, since they develop no appreciable heat. A 115-volt connector and two coaxial output connectors are mounted in the rear edge of the chassis. The output may be either capacity or link coupled to a following stage. The two power switches and the key jack are set in the front edge of the chassis.

Adjustment

The adjustment of the unit is very simple. The VR resistor, R_4, should first be set so that the VR tubes stay ignited with the key closed. VR75s or VR90s may be used, the higher voltage giving somewhat greater output from the unit. Then, the tuning condenser C_1 should be set at maximum capacitance. Listening on a receiver tuned to about 3490 kc., the oscillator padder, C_2, should be adjusted until the oscillator signal is heard at that frequency. The oscillator tuning should then cover the range up to a frequency slightly above 4000 kc.

The amplifier padder is adjusted by tuning the oscillator to the approximate center of the band and adjusting C_{11} for maximum grid current to the following stage. If the coil dimensions have been followed carefully, the output should then be substantially constant over the entire band. The 1750-kc. or 3.5-Mc. output should be used in feeding a crystal stage normally using 3.5-Mc. crystals, while 3.5- or 7-Mc. output should be used in cases where 7-Mc. crystals are normally employed.

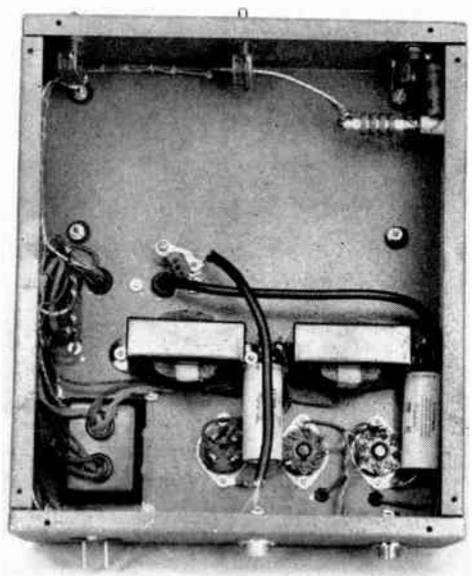

Fig. 6-73 — Bottom view of the completed series-tuned VFO unit. Power leads are cabled. The coaxial cables go to the r.f. output connectors.

A 75/100-Watt Bandswitching Transmitter

Figs. 6-74 through 6-78 show the diagram and photographs of a bandswitching unit covering all bands from 3.5 to 28 Mc. inclusive. Either one or two 807s may be used in the output stage.

Referring to the circuit of Fig. 6-75, the oscillator circuit is a modified Pierce with crystal switching and provision for capacitive coupling to an external VFO. The output circuit of the oscillator (RFC_2) is nonresonant. The second stage, employing a 6V6 tetrode, may be used as a straight amplifier at the crystal frequency or as a frequency doubler for 3.5-Mc. crystals. C_7 has sufficient capacitance range to cover both 3.5 and 7 Mc.

The 6V6 in the third stage is connected as a high-μ triode and is used as a doubler to 14 Mc. with 7-Mc. input from the buffer, or as a tripler to 10.5 Mc. with 3.5-Mc. input. The 10.5-Mc. frequency is used in arriving at 21 Mc. in the following stage. In the fourth stage, another triode-connected 6V6 is used as a doubler, covering both 21 and 28 Mc.

When two 807s are used in the output stage, they are connected in parallel. RFC_7, RFC_8, RFC_9 and RFC_{10} are v.h.f. parasitic-suppressor chokes. The 6Y6G is a protective tube, used in lieu of fixed bias for the 807s.

Bandswitching System

The first two sections of the bandswitch, S_{2A} and S_{2B}, in the cathodes of the last two doubler stages, serve to disable these stages when they are not required. S_{2C} connects the input of the 807s to the output of the exciter stage delivering the desired output frequency. All bands are covered in the output tank circuit with two tapped coils, S_{2D} selecting the proper coil and tap. The two output links are connected in series to avoid the necessity for switching. C_{21} bypasses the link on L_4 for the frequencies covered by L_5, but is small enough in capacitance to have no detrimental effect on the operation of the link at 7 and 3.5 Mc.

I_1, I_2 and I_3 are resonance-indicator lamps coupled to the multiplier tank circuits.

Provision is made for keying either the oscillator (J_1) or the first amplifier stage (J_2). Excitation to the output stage is adjusted by means of R_4 which varies the screen voltage of the first 6V6 buffer-doubler.

Construction

The unit is assembled on an 8 × 17 × 3-inch chassis with a 7-inch rack panel. Referring to the photograph of Fig. 6-76, the crystal sockets are lined up along the right-hand end of the chassis, centered 1¼ inches from the edge. The 6AG7 oscillator tube is in line to the rear, 5 inches behind the panel. The following two 6V6s are to the left in line with the 6AG7, with the 6Y6G to the rear. The last 6V6 is 3½ inches from the panel.

The output-stage tank condenser, C_{20}, is insulated from the chassis by mounting it on National polystyrene button-type insulators. Its shaft, broken with an insulating coupling, is 6¼ inches from the left-hand end of the chassis with the two tank coils, at right angles, to the left. The plug bases are removed and the coils mounted on 1-inch stand-off insulators. The last section of the bandswitch, S_{2D}, is mounted with its shaft vertical, 4 inches from the left-hand end of the chassis and 4 inches behind the panel.

The 807s, centered back of the tank condenser, are 6½ inches behind the panel. Their sockets are submounted 2½ inches below the top of the chassis on a 5¼ × 2¼-inch piece of aluminum. A lip on the rear end of the piece is fastened to the rear edge of the chassis and the plate is braced at the front with a metal post between the tubes.

Underneath (Fig. 6-77), the bandswitch control is at the left, 1¼ inches from the end of the chassis. The control operates two Millen right-angle drives in tandem, one shaft of the second unit driving the output-tank switch, S_{2D}, through a hole in the chassis, while the other shaft is extended down the center line of

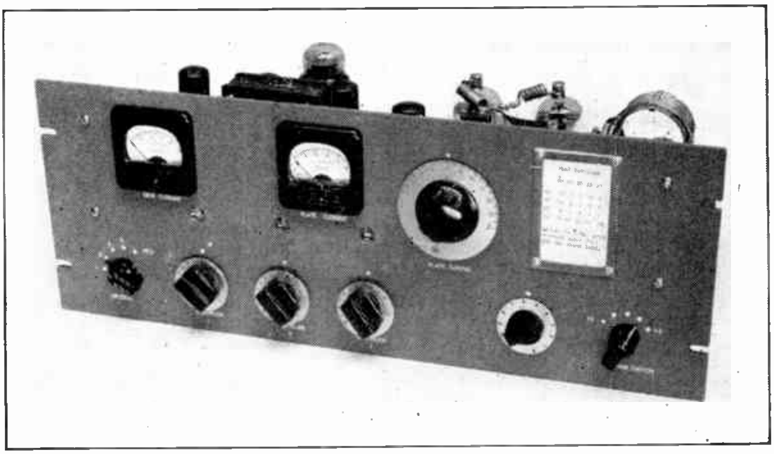

Fig. 6-74 — Front view of the bandswitching-transmitter unit. Along the bottom of the panel, the crystal switch is to the left followed by the tuning controls for the three driver stages and their indicator lamps. The excitation control and bandswitch are to the right. The large dial is the output-tank control.

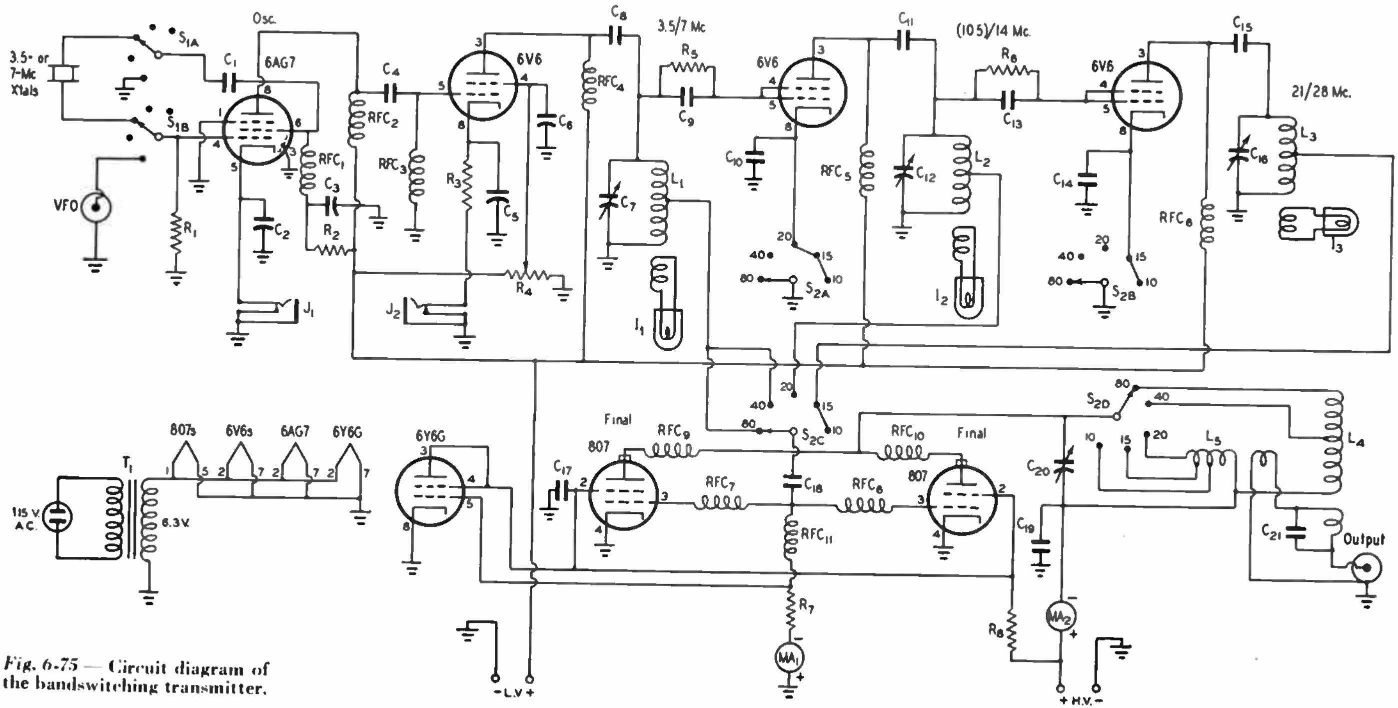

Fig. 6-75 — Circuit diagram of the bandswitching transmitter.

C_1, C_8, C_{10}, C_{11}, C_{14}, C_{15}, C_{17} — 0.001-μfd. mica.
C_2, C_3, C_5, C_6 — 0.01-μfd. paper.
C_4, C_9, C_{13}, C_{18} — 100-μμfd. mica.
C_7 — 250-μμfd. variable (National ST-250).
C_{12} — 100-μμfd. variable (National ST-100).
C_{16} — 50-μμfd. variable (National ST-50).
C_{19} — 0.001-μfd. 2500-volt mica.
C_{20} — 300-μμfd. variable (National TMS-300).
R_1 — 12,000 ohms, ½ watt.
R_2 — 39,000 ohms, 1 watt.
R_3 — 1200 ohms, 1 watt.
R_4 — 25,000-ohm 7-watt potentiometer (Mallory E25MP).
R_5 — 15,600 ohms, 3 watts (three 47,000 ohms, 1 watt in parallel).
R_6 — 2200 ohms, 1 watt.
R_7 — 12,000 ohms, 1 watt.
R_8 — 25,000 ohms, 25 watts.
L_1 — 16 turns No. 22 d.c.c., 1-inch diam., tapped at 13 turns from ground end.
L_2 — 9 turns No. 18, 1-inch diam., ⅝ inch long, tapped 5½ turns from ground end (B & W 3015).
L_3 — 5 turns No. 16, 1-inch diam., ¾ inch long, tapped 2¾ turns from ground end (B & W 3014).
L_4 — 14 turns No. 16, 2 inches diam., 1¾ inches long, tapped at center, 2-turn link (B & W BEL-40 with 5 turns removed).
L_5 — 8 turns No. 14, 2 inches diam., 2 inches long, tapped at 3¼ and 2 turns from ground end, 2-turn link (B & W BEL-10).*
I_1, I_2, I_3 — 2.5-volt 60-ma. dial lamp.
J_1, J_2 — Closed-circuit jack.
MA_1 — D.c. milliammeter, 25-ma. scale.
MA_2 — D.c. milliammeter, 500-ma. scale.
RFC_1, RFC_2, RFC_3, RFC_4, RFC_5, RFC_6, RFC_{11} — 2.5-mh. r.f. choke.
RFC_7, RFC_8 — 15 turns No. 20 d.c.c., ¼-inch diam.
RFC_9 — 16 turns No. 14, 5/16-inch inside diam., 1⅜ inches long.
RFC_{10} — 8 turns No. 14, 5/16-inch inside diam., ¾ inch long.

* Add 1 turn if single 807 used in output stage.

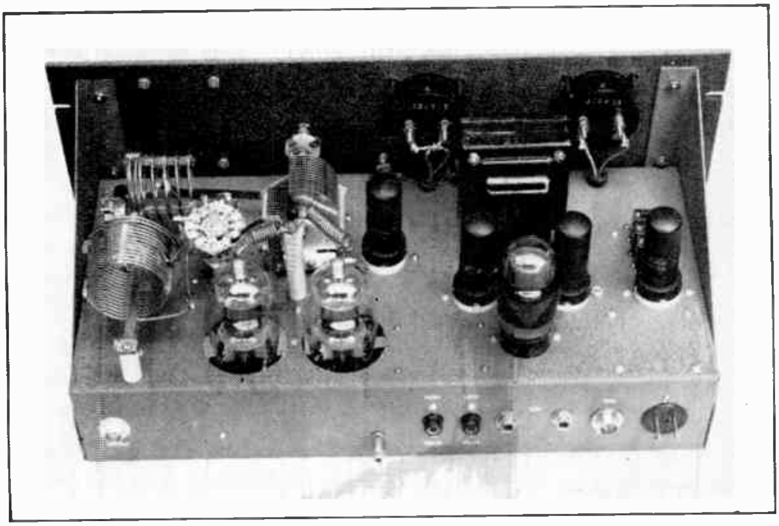

Fig. 6-76 — Rear view of the bandswitching transmitter-exciter.

the chassis to operate the other three sections, S_{2A}, S_{2B} and S_{2C}. The last three sections are mounted as a unit on a metal bracket, 7¼ inches from the left-hand end of the chassis.

Along the front edge of the chassis are the crystal switch, 1¼ inches from the right-hand end, followed by the three driver-stage tuning condensers with the shafts spaced 2½ inches apart. The excitation control, R_4, is to the right of the bandswitch.

L_1 is wound on a Millen 1-inch coil form fastened to the chassis behind C_7. The other two exciter tank coils are sections of self-supporting strip coils fastened directly to the rear ends of their respective tank condensers.

The resonance-indicator lamps are mounted in rubber grommets directly above the tuning controls with which they are associated. The links each consist of a single loop at the end of a twisted pair, connecting to the lamp socket on the panel through a hole in the chassis. Each link is adjusted to the minimum coupling that will give a satisfactory indication. This can be done by simply twisting the loop to change its diameter.

Adjustment

The unit is designed to be operated from two plate supplies, one delivering 400 volts at 150 ma. for the driver stages and the other 750/600 volts 100/200 ma. for the output stage. The diagram of a suitable power unit is shown in Fig. 6-78.

The output stage should first be tested for parasitic oscillation, following the procedure outlined under "Parasitic Oscillation," this chapter. In this case, the test may be conducted at full plate-supply voltage without load on the final amplifier, since the 6Y6G protective tube will hold the input down to a safe value. The lead lengths are such in the model shown that RFC_9 must be connected in the plate lead of the 807 to the left and RFC_{10} in the lead to the 807 to the right as viewed from the front. Slight changes in the dimensions of these chokes may be required to compensate for other lead lengths. The test should be made for each position of the bandswitch. If the construction pictured is followed closely, there should be no tendency toward oscillation at the operating frequency so long as the output stage is loaded.

In tuning up with a dummy load, the operation on each band should be checked progressively from 3.5 Mc. (or 7 Mc. in the case of 7-Mc. crystals only). L_1C_7 will show two responses — 3.5 Mc. near maximum capacitance and 7 Mc. near minimum — regardless of the

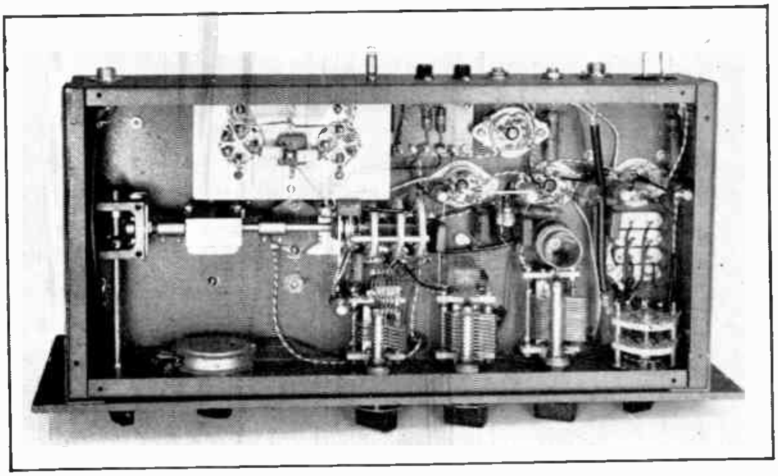

Fig. 6-77 — Bottom view of the bandswitching unit. The two right-angle drives at the left drive the band-changing switches. The bottoms of the 807 sockets are at the upper left. The excitation control is to the left and the crystal switch to the right.

position of the bandswitch. With the switch in the 14-Mc. position, L_2C_{12} will also show two responses — at 10.5 and 14 Mc. For 14-Mc. output, L_1C_7 should be tuned to 7 Mc. and L_2C_{12} to 14 Mc. With the bandswitch in the 21- or 28-Mc. positions, L_3C_{16} may show three responses — at 21, 24.5 and 28 Mc. The unwanted response at 24.5 Mc. should be weaker than the other two, however, and the coupling to the indicator lamp can be made loose enough so that it does not show up on the lamp. For 21-Mc. output, a 3.5-Mc. crystal is required. L_1C_7 is tuned to 3.5 Mc., L_2C_{12} to 10.5 Mc. and L_3C_{16} to 21 Mc. With the switch in the 21-Mc. position, a response may be found at 17.5 Mc. at the minimum of C_{16}. Here, too, the coupling to the indicator lamp should be reduced to eliminate the indication at this frequency.

When 28-Mc. output is desired, L_1C_7 should be tuned to 7 Mc., L_2C_{12} to 14 Mc. and L_3C_{16} to 28 Mc. Care should be taken to check the responses in each stage with an absorption wavemeter and to record them on the chart.

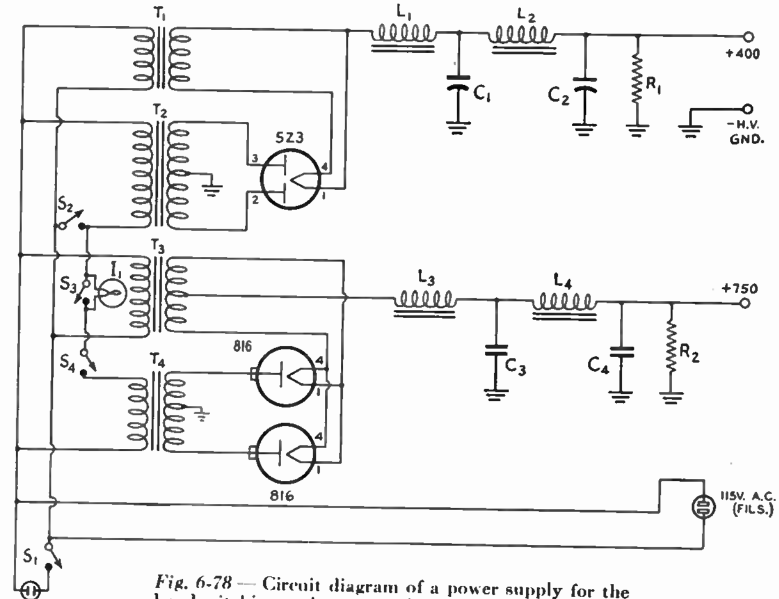

Fig. 6-78 — Circuit diagram of a power supply for the bandswitching exciter-transmitter.

C_1, C_2 — 16-μfd. 450-volt electrolytic.
C_3, C_4 — 4-μfd. 1000-volt oil-filled.
R_1 — 15,000 ohms, 25 watts.
R_2 — 25,000 ohms, 50 watts.
L_1, L_2 — 20-hy. 150-ma. filter choke.
L_3 — 5/25-hy. 250-ma. swinging choke.
L_4 — 20-hy. 250-ma. smoothing choke.
I_1 — 100-watt 115-volt lamp.
S_1, S_2 — 10-amp. toggle switch.
S_3, S_4 — 5-amp. toggle switch.
T_1 — 5-volt 3-amp. filament transformer.
T_2 — Plate trans., 400 v. d.c., 150 ma.
T_3 — 2.5-volt 4-amp. filament transformer.
T_4 — Plate transformer, 750 v. d.c., 250 ma.

S_1 turns on all filaments in the power supply and transmitter and sets up circuit for S_2. S_2 turns on low-voltage supply and sets up circuit for S_4 which turns on high-voltage supply. When S_3 is open, high voltage is reduced by I_1 in series with the primary of T_4. In operating the transmitter, all switches except S_2 are closed. S_2, then, is the stand-by switch controlling both plate supplies simultaneously.

In each case, grid current to the 807s should be limited to 4 ma. per tube fully loaded, adjusting the excitation control R_4. The load should be adjusted until each 807 draws 100 ma.

Table 6-V shows various typical voltages and currents taken with a 400-volt exciter supply, with the excitation control adjusted for a final-amplifier grid current of 4 ma. per tube, fully loaded at a plate voltage of 600.

TABLE 6-V

		* Parallel 807s							* Single 807						
		1st 6V6			2nd 6V6		3rd 6V6		1st 6V6			2nd 6V6		3rd 6V6	
Output Mc.	Crystal Mc.	Bias Volts	Cathode Ma.	Screen Volts **	Bias Volts ***	Plate Ma.	Bias Volts ***	Plate Ma.	Bias Volts ***	Cathode Ma.	Screen Volts **	Bias Volts ***	Plate Ma.	Bias Volts ***	Plate Ma.
3.5	3.5	30	25	6	—	—	—	—	21	25	5	—	—	—	—
7	3.5	35	29	20	—	—	—	—	24	29	9	—	—	—	—
14	3.5	40	34	80	165	15	—	—	24	29	12	80	16	—	—
21	3.5	36	30	74	165	22	30	29	23	28	30	105	15	21	26
28	3.5	40	34	95	135	20	27	37	23	27	5.5	60	15	20	25
7	7	18	15	14	—	—	—	—	15	12	12	—	—	—	—
14	7	40	34	80	165	15	—	—	25	21	15	105	15	—	—
28	7	15	13	25	92	18	30	31	12	10	10	50	14	22	20

* Exciter supply — 400 volts. 807 grid current 4 ma. per tube, 600 volts plate, fully loaded. Oscillator cathode current 35 ma. all bands with 3.5-Mc. crystal.
** Measured cathode to screen.
*** Measured grid to ground with 2.5-mh. r.f. choke in meter lead.

A 500-Watt Beam-Tube Amplifier

The photographs of Figs. 6-79, 6-81, 6-82 and 6-84 show the construction of a single-tube screen-grid amplifier using a Type 813 which will handle an input of up to 500 watts. The circuit is shown in Fig. 6-80. The amplifier is designed for link coupling in both input and output circuits. Bandswitching is employed in the grid circuit principally because of the problem of providing plug-in coils with satisfactory shielding. To assure good stability, the amplifier is neutralized. The triode-connected 6Y6 protects the 813 in case of removal of excitation. (See "Operation of Screen-Grid Amplifiers," this chapter.)

The single milliammeter may be switched by S_2 to read grid current, screen current or cathode current. A multiplier shunt increases the range of the 50-ma. meter to 500 ma. when it is switched to read cathode current. A jack, J_2, is provided so that the amplifier may be keyed in the center-tap.

Construction

The entire amplifier unit is suspended from a standard 19 × 8¾-inch panel. The grid-circuit components are housed in the 6 × 6 × 6-inch steel box to the left in Fig. 6-81, while the plate tank condenser and coil are mounted on the panel to the right. The 813 protrudes horizontally from the right-hand side of the box, its socket being submounted on brackets inside the box so that that portion of the tube which extends above its internal shielding plate is exposed. The 6Y6 is mounted alongside the 813.

The r.f. input jack, the grid coils, grid tuning condenser, bandswitch, key jack, and the meter switch are built as one assembly constructed on one of the cover plates of the box as shown in Fig. 6-84. All ground connections in this assembly are made to soldering lugs slipped under the screws which mount the coil forms. The coils themselves are held away from the chassis by National GS-10 stand-off insulators. Care should be taken in locating the mounting holes for the coils and the bandswitch to be sure that they will clear the lip of the utility box when the time comes for final assembly. If required, additional clearance may be obtained by filing semicircular notches in the lips of the box. The coils should be connected to the bandswitch before the coupling links are wound. This keeps the assembly clear of obstructions and makes wiring easier. Padding condensers C_1 and C_2, used with the 40- and 80-meter coils respectively, are connected from the grid end of the coil to the same soldering lugs used for grounding the cold ends of the coils. When the links are wound on later, their ground connections are also made to these same lugs.

The meter switch provides mounting terminals for meter shunts R_2, R_3 and R_6, and for the grid resistor, R_1. The leads to the meter itself are passed through the top of the box through a grommet-lined hole after assembly. The other leads to the metered circuits are cabled and are run along the top of the box.

All of the by-pass condensers associated with the screen and filament circuits are mounted on the socket and should be grounded at a common point. All of this wiring, together with the filament and screen leads, should be soldered in place before the 813 socket bracket is mounted within the box. The grid blocking condenser, C_4, may also be soldered in position, leaving one end free to be connected to the stator plates of the grid tuning condenser after assembly. The mounting of the screen dropping resistors is shown in Fig. 6-82. Both

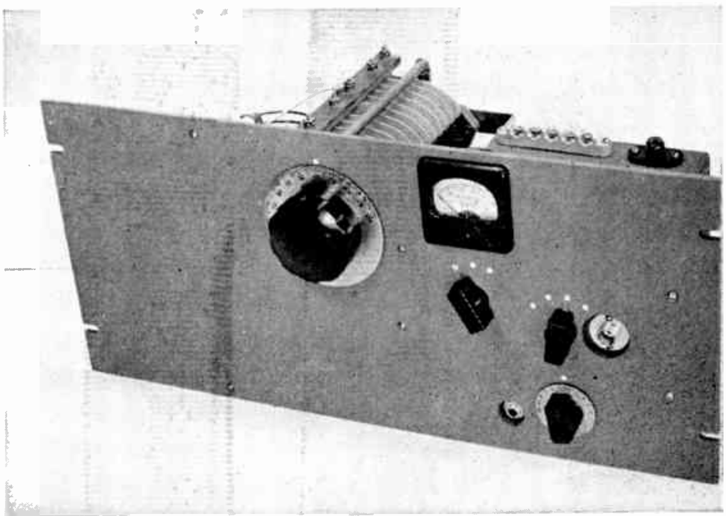

Fig. 6-79 — **Front-panel view of the stabilized 813 amplifier. In addition to the meter and the plate and grid tuning controls, the panel contains the r.f. input jack, the key jack, a three-position meter switch, and a four-position bandswitch for the grid circuit.**

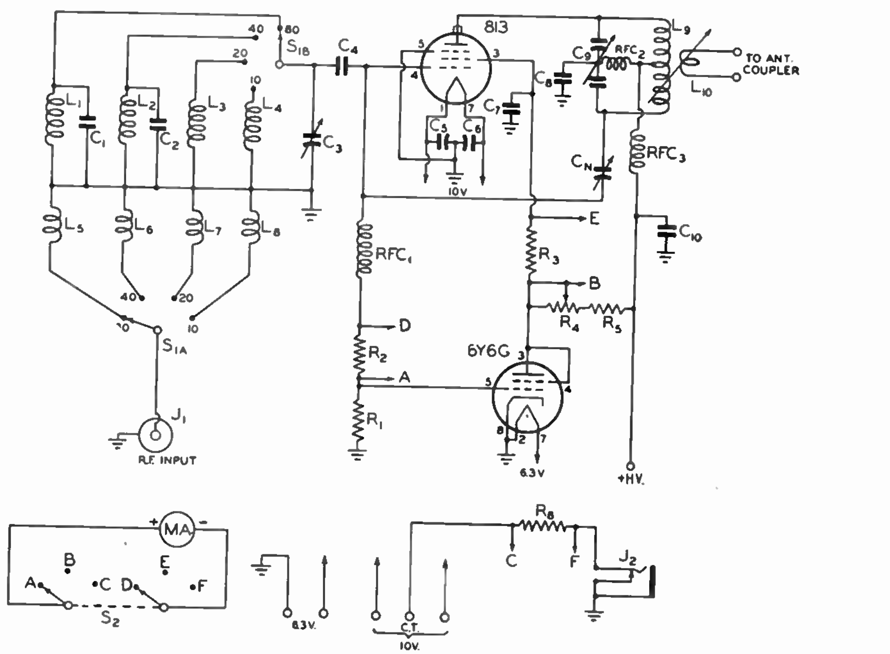

Fig. 6-80 — Schematic diagram of the 813 amplifier.

C_1 — 100-μμfd. mica.
C_2 — 68-μμfd. mica.
C_3 — 50-μμfd. receiving-type variable (Millen Type 19050).
C_4 — 0.0022-μfd. mica.
C_5, C_6 — 0.01-μfd. paper.
C_7 — 0.001-μfd. 2500-volt mica.
C_8 — 0.001-μfd. 5000-volt mica.
C_9 — 50-μμfd.-per-section dual transmitting type, 0.171-inch spacing (Cardwell XG-50-XD).
C_{10} — 0.001-μfd. 5000-volt mica.
C_N — See text.
R_1 — 10,000 ohms, 5 watts. (See text.)
R_2, R_3 — 100 ohms, ½ watt.
R_4 — 35,000 ohms, 50 watts, with slider.
R_5 — 15,000 ohms, 50 watts.
R_6 — Meter shunt. Wound with No. 30 d.s.c. wire, length as required to multiply meter scale by ten.
L_1 — 26 turns No. 22 d.s.c. spaced to occupy 1¼ inches on a 1-inch diam. form.
L_2 — 15 turns No. 18 d.c.c. spaced to occupy 1¼ inches on a 1-inch diam. form.
L_3 — 10 turns No. 18 d.c.c. spaced to occupy 1¼ inches on a 1-inch diam. form.
L_4 — 5 turns No. 18 d.c.c. spaced to occupy 1¼ inches on a 1-inch diam. form.
L_5, L_6, L_7, L_8 — Two-turn links, No. 18 insulated stranded wire, wound over ground ends of L_1 through L_4 inclusive.
L_9 — Note: These coils are B & W TVH series, for use with the B & W TVH swinging-link assembly. The coils are modified as described below.
80 meters: B & W 160-TVH with 4 turns removed from each end. (54 turns No. 18 enameled, 2½ inches diameter, winding length 4⅞ inches.)
40 meters: B & W 80-TVH with 8 turns removed from each end. (22 turns No. 14 enameled, 2½ inches diameter, winding length 3⅜ inches.)
20 meters: B & W 20-TVH with 1 turn removed from each end. (12 turns No. 12 enameled, 2½ inches diameter, winding length 4½ inches.)
10 meters: B & W 10-TVH with 1 turn removed from each end. (6 turns ⅛-inch copper tubing, 2½ inches diameter, winding length 5¼ inches.) (Winding lengths specified above include ⅝-inch separation between halves of the coil for entrance of swinging link coil.)
L_{10} — 3-turn link assembly, part of B & W TVH swinging-link assembly.
J_1 — Coaxial connector.
J_2 — Closed-circuit jack.
MA — 0–50 d.c. milliammeter.
RFC_1 — 2.5 mh. (Millen Type 34104).
RFC_2 — 2.5 mh. (National R-100).
RFC_3 — 2.5 mh. (National R-300).
S_1 — Single-gang 2-pole 5-position ceramic water switch. (Centralab S-2505.)
S_2 — Two-gang 2-pole 5-position ceramic wafer switch. (Centralab S-2511.)

are supported by small ceramic stand-off insulators. High voltage for the plate-and-screen supply enters the top of the box through a Millen safety connector, and is passed through the side of the box to the plate coil through a ceramic bushing. A bushing requiring a ¾-inch hole was used to provide maximum insulation. The fixed plate of the neutralizing condenser is mounted on a similar bushing just above the socket for the 6Y6G. The exact location of this hole should be determined after temporarily assembling the panel, the plate tuning condenser, and the box, because the fixed plate must be aligned with the variable plate, which is supported by the plate tuning condenser.

The plate tuning condenser is mounted on the front panel by three ceramic stand-off insulators. This is necessary because the condenser rotor is at full plate potential above ground. The rotor shaft is cut off about ¼ inch from the rotor bushing, to permit the insertion of a high-voltage type shaft coupling. An insulated shaft made of ¼-inch bakelite rod couples the rotor of the condenser to the dial. Both r.f. chokes used in the plate tank circuit are mounted on the jack bar into which the coils plug. The rear of the plate tuning condenser is held to the rear of the box by a

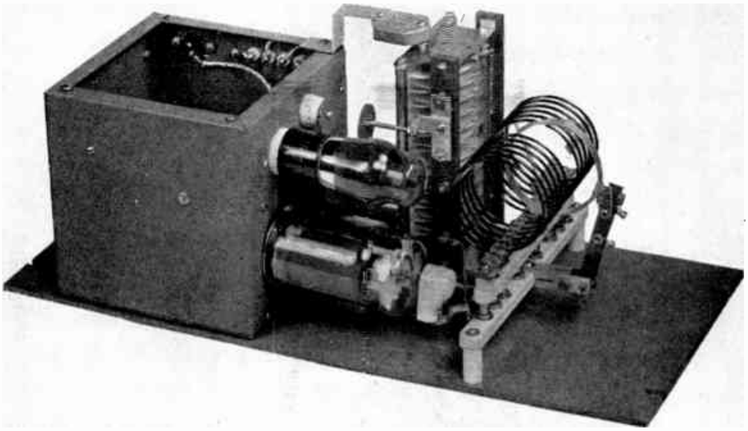

Fig. 6-81 — Rear view of the 813 amplifier. The steel utility box used to shield the input circuits is bolted to the rear of the panel. The home-built neutralizing condenser is below the mica by-pass condenser which forms a part of the rear support for the plate condenser. The plate condenser and the swinging-link assembly are mounted directly on the panel with ceramic stand-off insulators.

small aluminum bracket, bent to provide adequate clearance between itself and the rotor. Blocking condenser C_8 is made a part of this bracket.

The variable plate of the neutralizing condenser is supported by a small bracket bolted to the stator connectors of the tuning condenser. The copper disks used are each 1 inch in diameter. A hole is drilled in the center of each disk to pass a mounting screw. The "stator" disk is bolted to the ceramic feed-through bushing, and is held away from it by a ¼-inch spacer. The other end of the screw which goes through the bushing is fitted with a soldering lug to which the grid connection is soldered. The "rotor" disk is fastened to a 2-inch machine screw with a nut. The threaded end of the screw is then passed through the mounting bracket and is held in position firmly by two nuts, one on each surface of the bracket. This plate should be put in position first, after which the location of the hole for the bushing can be determined to provide proper alignment of the two plates.

Power Supply

The circuit diagram of a plate and filament supply and a suggested control system are shown in Fig. 6-85. The control system is arranged to take care of the exciter as well as the amplifier. S_1 turns on all filaments in the exciter, amplifier and power supply, and sets up circuit for S_2 which turns on the exciter plate supply. Closing S_2 also sets up circuit for S_3 which turns on the high-voltage supply. When S_4 is open, I_1 is in series with the

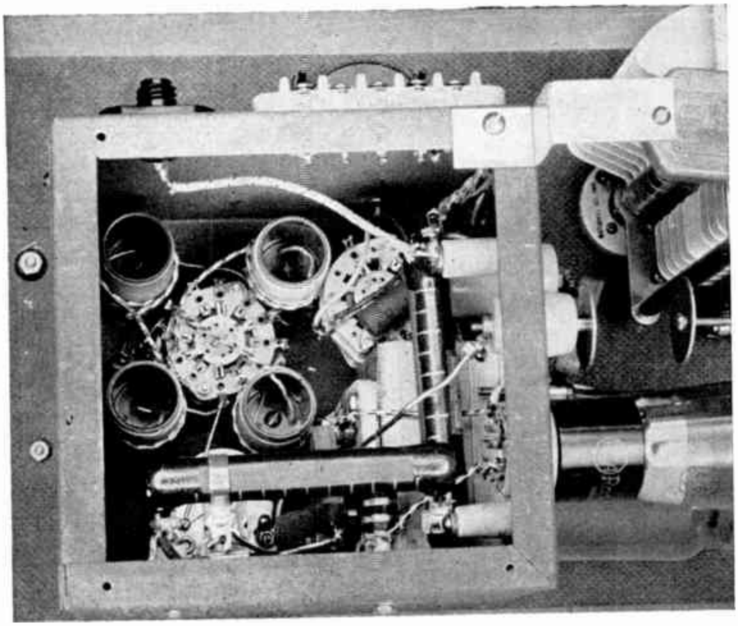

Fig. 6-82 — The interior of the shield box. The socket for the 813 is mounted on a bracket visible to the right of the four grid coils. The two large resistors are the screen-dropping units. This view also shows the construction of the neutralizing condenser, one plate of which mounts on a ceramic bushing on the wall of the box, the other being supported by a bracket from the plate tuning condenser.

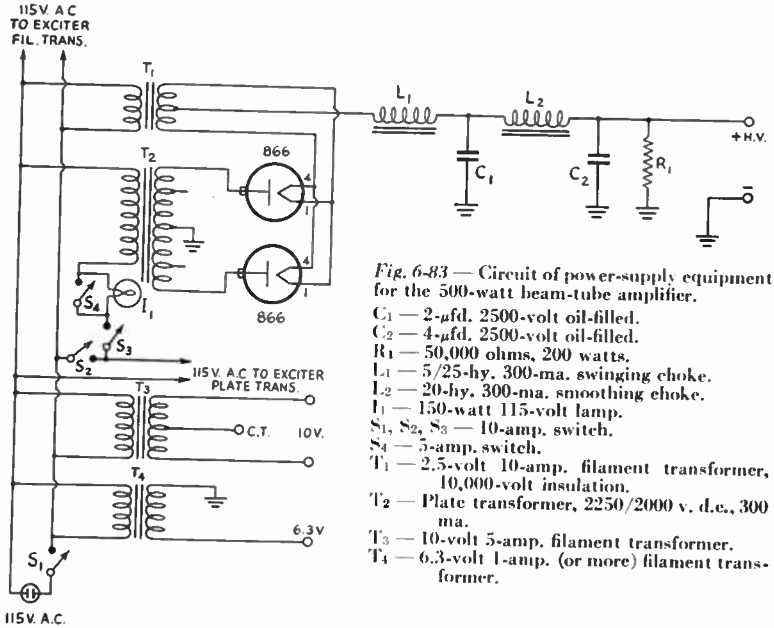

Fig. 6-83 — **Circuit of power-supply equipment for the 500-watt beam-tube amplifier.**

C_1 — 2-μfd. 2500-volt oil-filled.
C_2 — 4-μfd. 2500-volt oil-filled.
R_1 — 50,000 ohms, 200 watts.
L_1 — 5/25-hy. 300-ma. swinging choke.
L_2 — 20-hy. 300-ma. smoothing choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_2, S_3 — 10-amp. switch.
S_4 — 5-amp. switch.
T_1 — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T_2 — Plate transformer, 2250/2000 v. d.c., 300 ma.
T_3 — 10-volt 5-amp. filament transformer.
T_4 — 6.3-volt 1-amp. (or more) filament transformer.

primary of T_2 to reduce voltage while adjustments are being made. When the transmitter is in operation, all switches except S_2 are closed. S_2 then serves as the stand-by switch, controlling both exciter and amplifier plate-voltage supplies simultaneously.

Adjustment

The amplifier should first be neutralized following the method suggested under "Neutralizing Procedure," this chapter.

Tube data sheets furnish proper operating values for a choice of several plate voltages and these should be followed closely. For operating at maximum c.w. ratings, the plate voltage should be 2250. A screen resistor of 46,000 ohms and a grid leak of 10,000 ohms should be used. When the amplifier is loaded to a plate current of 220 ma. and the excitation adjusted to give a grid current of 15 ma., the d.c. grid voltage should be −155 volts and the screen voltage 400 at a screen current of 40 ma. For maximum plate/screen-modulated ratings, the plate voltage should be 2000, the screen resistor 41,000 ohms and the grid leak 11,000 ohms. When the amplifier is loaded to the maximum rated plate current of 200 ma. and the grid current adjusted to 16 ma., the bias should be 175 volts and the screen voltage 350 at 40 ma.

The driver should be capable of an output of 15 to 20 watts to allow for coupling losses. For c.w. operation with a 1500-volt supply and a plate current of 180 ma., the recommended screen voltage is 300. The screen current should be 30 ma. and the required screen voltage-dropping resistor 40,000 ohms. The grid current under load should be 12 ma. through a 7500-ohm grid leak. For 'phone operation at 1600 volts, the following values are recommended: plate current 150 ma., screen voltage 400 at 20 ma., grid bias 130 volts at 6 ma., screen voltage-dropping resistor 60,000 ohms, grid leak 22,000 ohms.

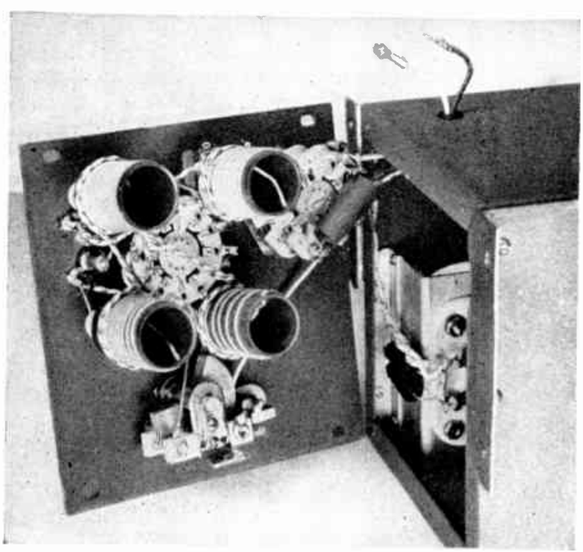

Fig. 6-84 — **The grid coils are grouped around their bandswitch on one of the covers of the utility box. The coaxial input jack is mounted between the two coils at the left. The meter switch is in the upper right-hand corner, and the grid tuning condenser is at the bottom. The two leads coming through the top of the box run from the meter switch to the meter.**

Push-Pull Triode Amplifiers

Figs. 6-85 through 6-93 show diagrams and various views of two triode push-pull amplifiers in the nominal 500-watt class. They illustrate two different approaches, principally in layout and mechanical design. In the first one, shown in Fig. 6-85, the primary feature is that coils may be removed by pulling toward the rear, rather than in an upward direction as is necessary in more conventional arrangements. The latter is sometimes an inconvenient operation if the rack is enclosed, or when the antenna-tuning unit is immediately above in the rack.

In the second one, the upper portion of the panel is hinged so that the coils may be changed from the front of the transmitter rack. An interlock on the panel is an added safety precaution.

● REAR-ACCESS AMPLIFIER

The circuit diagram of the open-back model is shown in Fig. 6-86. Both input and output circuits are designed for link coupling. The amateur bands lying in the frequency range of 3.5 to 30 Mc. are covered by means of plug-in coils.

R_1 is a 100-ohm resistor preventing direct r.f. grounding of the coil center-tap. It is used in preference to an r.f. choke to avoid setting up a low-frequency parasitic circuit in conjunction with RFC_3 which serves the same purpose in the plate circuit. RFC_1 and RFC_2 are v.h.f. parasitic suppressors. Two plate by-pass condensers, C_7 and C_8, are used — one at each end of the tank-condenser rotor — to assist in preserving balance in the push-pull amplifier. All power-supply lines — bias, plate voltage, and both sides of the filament transformer — are filtered for v.h.f. harmonics to reduce TVI.

Although 812-As are shown, the design is suitable for any triodes operating at plate potentials up to 1750 volts with or without plate modulation. Tubes of other types may require a slightly different treatment, however, in suppressing v.h.f. parasitic oscillation.

Construction

The plate tank condenser is mounted in the approximate center of the 10½ × 19-inch rack panel, and is insulated from it by four ¾-inch ceramic cones. Its shaft runs parallel to the length of the panel, and is connected to the dial through a Millen right-angle drive and a high-voltage insulated flexible coupling. The latter is required because the shaft is at full plate voltage above ground. The right-angle drive is fastened to the bottom of the plate tank-coil mounting base and the base is brought to the correct level behind the panel by mounting it on a pair of 2-inch cone insulators. Shim washers between the cones and the base may be used to line up accurately the shafts of the drive unit and the tuning condenser. The plate r.f. choke also is fastened to the bottom of the mounting base.

The tube sockets, neutralizing condensers and grid tuning condenser all are mounted horizontally on the inside of an aluminum shielding partition extending 7 inches behind the panel and 9 inches high. The tube sockets are positioned so that their centers are 3 inches behind the panel and 1¼ inches from the top and bottom edges of the partition. The neutralizing condensers are placed 2½ inches behind the centers of the tube sockets.

The grid tank condenser, C_1, is fastened to the outside of the shielding partition and the partition is placed on the panel so that the two dials are symmetrical. The grid coils plug into a 5-prong ceramic tube socket mounted on a small sheet-aluminum bracket, 2¾ inches wide, which surrounds the grid tank condenser. One end is anchored to the partition while the other end is fastened to the panel. R_1, C_9, the r.f. coaxial input connector and the filament-

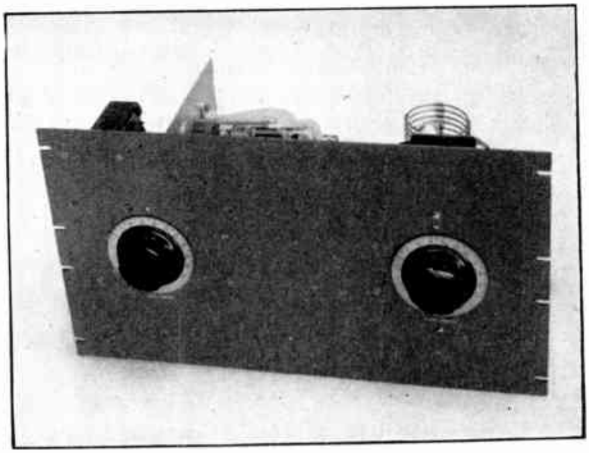

Fig. 6-85 — Front view of the rear-access push-pull amplifier. All components are mounted on a single 10½-inch-deep steel rack panel. The grid tank control is on the left, plate tuning at the right.

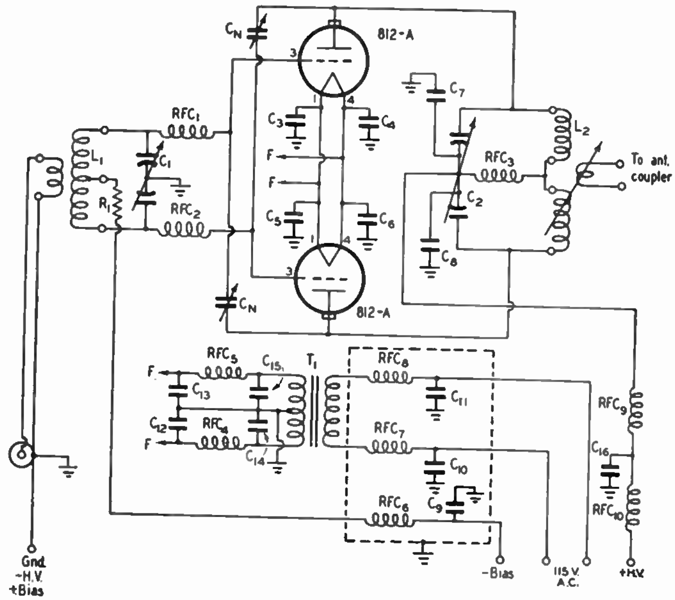

Fig. 6-86 — Diagram of the rear-access push-pull amplifier.

C_1 — 100 μμfd. per section, 0.03-inch plate spacing (Cardwell MR-100BD).
C_2 — 100 μμfd. per section, 0.125-inch plate spacing (B & W JCX-100-E).
C_3, C_4, C_5, C_6 — 0.006-μfd. mica.
C_7, C_8 — 0.001-μfd. 2500-volt mica.
C_9, C_{10}, C_{11}, C_{12}, C_{13}, C_{14}, C_{15} — 47-μμfd. mica.
C_{16} — 470-μμfd. 2500-volt mica.
C_N — Neutralizing condenser (National NC-800-A).
R_1 — 100 ohms, 5 watts.
L_1 — B & W JCL series coils. Two turns removed from each end of JCL-10. 3.5 Mc. — 10 μh.; 7 Mc. — 15 μh.; 14 Mc. — 8 μh.; 21 Mc. — 3.5 μh.; 28 Mc. — 1.4 μh.
L_2 — B & W TVL series coils. 3.5 Mc. — 32 μh.; 7 Mc. — 17 μh.; 14 Mc. — 5.8 μh.; 21 Mc. — 2.6 μh.; 28 Mc. — 1.8 μh.
RFC_1, RFC_2 — 5 turns No. 14, ½-inch diam., ½ inch long.
RFC_3 — 1-mh. 600-ma. r.f. choke (National R-154).
RFC_4, RFC_5 — 7 turns No. 12, ½-inch diam., 1 inch long.
RFC_6 — 2-μh. r.f. choke (National R-60).
RFC_7, RFC_8, RFC_9 — V.h.f. choke (Ohmite Z-0).
RFC_{10} — V.h.f. choke (Ohmite Z-1).
T_1 — Filament transformer; 6.3 volts, 8 amp.

circuit filter components are mounted on the side of this bracket.

The filament transformer is fastened to the panel. The parts forming the bias-lead and a.c.-line filters are enclosed in the small aluminum shield that holds the terminal strip for bias and a.c. line connections. A Millen high-voltage connector is mounted on a small angle bracket fastened to the panel near the tube bases on the plate side of the aluminum shielding partition.

A "T"-shaped ground strip of copper is run along the panel from a point under the plate tank condenser, passing under the partition and thence branching out to the two tube sockets. All by-pass condensers are grounded to this strip. Plate-circuit connections also are made of copper strip. Neutralizing leads are brought through the partition in small ceramic bushings.

Fig. 6-87 shows how a lead from the grid tank (which includes the v.h.f. parasitic choke) is brought out so that it can be tapped along the wire running from the tube-socket grid terminal to the neutralizing condenser for the purpose of equalizing the neutralization on all bands (see "Neutralizing Procedure," this chapter). The opposite grid-tank lead is brought out in similar fashion.

The amplifier should be checked and adjusted following the procedure outlined under "Adjustment of R.F. Amplifiers" in this chapter. The tube tables in this book, or tube-data sheets, should be consulted for recommended operating conditions for the tubes selected.

The diagram of a power supply suitable for use with 812-As in this amplifier at maximum rating is shown in Fig. 6-89. Three VR-75s should be used in the bias supply. R_2 should be adjusted until the VR tubes just ignite without excitation. R_3 should be set at 1250 ohms.

FRONT-ACCESS PUSH-PULL AMPLIFIER

The circuit of the second model, shown in Fig. 6-91, is essentially the same as that of the previous amplifier. Separate filament transformers are used for each tube to permit individual metering in the center-tap. The only other essential difference is that because of differences in tube characteristics and arrangement of components, traps in the plate circuit, instead of chokes in the grid circuit, are used to suppress v.h.f. parasitic oscillation. The amplifier is built around a pair of V-70-Ds, although this model, too, is adaptable to other triodes in the same general power class operating at plate potentials up to 1750 volts with plate modulation.

Construction

Various views of the second model are shown in Figs. 6-90, 6-92 and 6-93. The amplifier is constructed on a standard 10 × 12 × 3-inch steel chassis which is fastened to a standard 5½-inch steel panel so that its upper edge is flush with the top of the panel. The lower edge of a 17-inch strip of piano hinge is inserted between the panel and the chassis, 6-32 machine screws clamping all three pieces together. The upper portion of the panel is a standard Presdwood panel cut down to a height of 6⅝ inches and fastened to the upper leaf of the

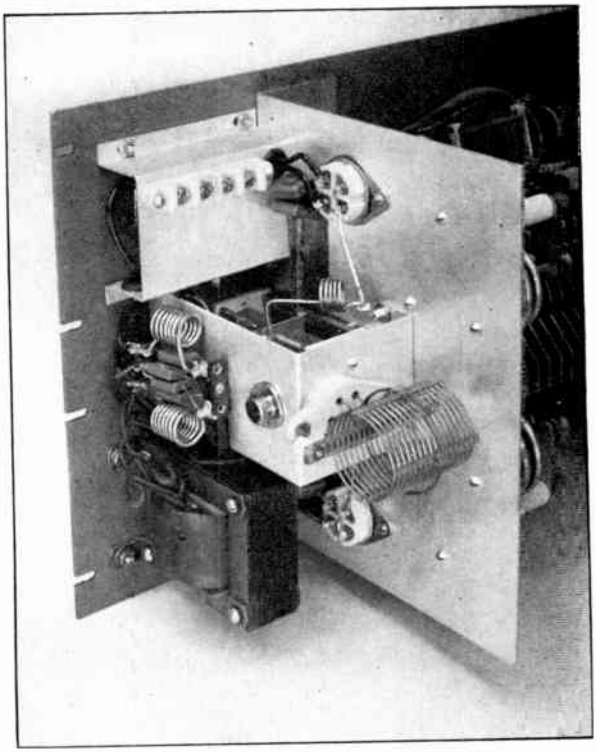

Fig. 6-87 — Detailed view of the grid tank circuit of the rear-access amplifier. The lead from the neutralizing condenser to the grid terminal on the tube socket on each side passes through the shielding partition in a ceramic bushing. The lead from the grid tank condenser to the neutralizing lead includes a parasitic choke and is tapped on the lead at a point that permits neutralization to hold over all bands. Also shown are two of the filters used in the filament leads for harmonic suppression, and the coaxial input connector, mounted on the small bracket that supports the grid coil.

piano hinge in such fashion that it may be hinged forward and down. It is necessary to trim the panel to allow space for the hinge. The hinged portion is backed with copper screening held in place by a strip of aluminum along the top edge and inserted underneath the piano hinge at the bottom. A window 3½ inches by 2½ inches is cut in the center of the panel to provide free air circulation. The front of the window is bordered by a National chart frame. When the amplifier is installed in the rack, only the bottom of the panel is fastened to the rack frame permanently, the upper portion being held closed with thumb nuts to permit easy access to the coils.

The two tank condensers are mounted below deck inside the chassis. The plate condenser is insulated from the chassis on four ¾-inch ceramic cones. Its shaft carries the full plate voltage, requiring a high-voltage shaft coupling to insulate the dial. The grid tank condenser is mounted on metal spacers so that its rotor shaft is level with that of the plate condenser. Small notches must be cut in the edge of the chassis at the points where the condenser shafts pass through the panel.

The jack bar for the plate tank coils is mounted on 1-inch ceramic cones on top of the chassis. One-inch clearance holes are punched in the chassis beneath the terminals of the jack bar to pass the plate tank leads through to the tuning condenser. These leads are of ⅜-inch copper strip cut from flashing

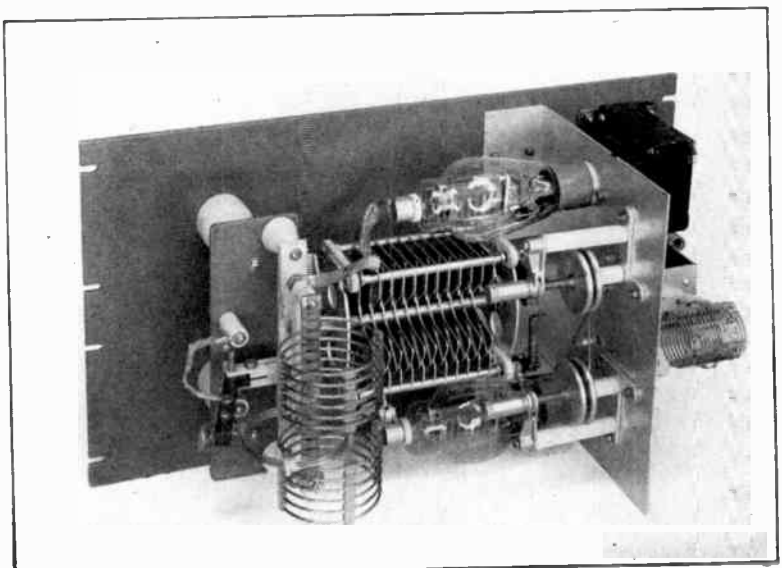

Fig. 6-88 — Plate-tank end of the open-back amplifier. The plate-tank assembly is supported on large ceramic stand-off insulators at the left. The butterfly-type plate tank condenser is mounted in the center, flanked by the horizontally-mounted tubes, which are supported by a shielding partition at the right. The neutralizing condensers are mounted between the tubes.

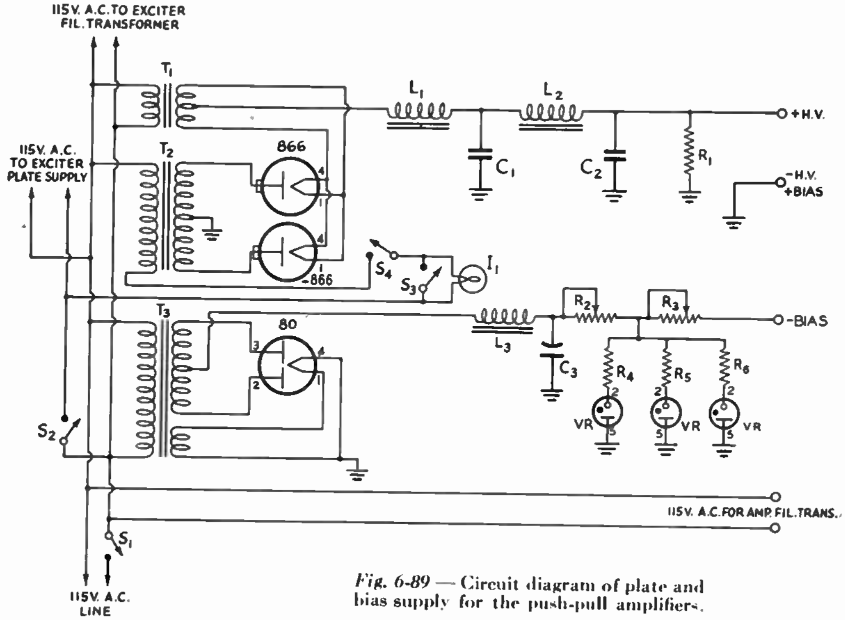

Fig. 6-89 — Circuit diagram of plate and bias supply for the push-pull amplifiers.

C_1, C_2 — 4-μfd. 2000-volt oil-filled.
C_3 — 8-μfd. 450-volt electrolytic.
R_1 — 25,000 ohms, 100 watts.
R_2 — 25,000 ohms, 25 watts, with slider.
R_3 — 5000 ohms, 25 watts, with slider.
R_4, R_5, R_6 — 100 ohms, 1 watt.
L_1 — 5/25-hy. 400-ma. swinging choke.
L_2 — 20-hy. 400-ma. smoothing choke.
L_3 — 30-hy. 50-ma. filter choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_2, S_3 — 10-amp. s.p.s.t. switch.
S_4 — 5-amp. s.p.s.t. switch.
T_1 — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T_2 — 1500/1250-v. d.c. 400-ma. plate transformer.
T_3 — Power transformer, 650 v. a.c., c.t., 50 ma. or more; 5 volts, 2 amp.
VR — Voltage-regulator tubes (see text).

copper sheet. The same type of connections is used throughout the plate circuit, including the leads from the tube caps to the neutralizing condensers.

The tubes are countersunk to a depth of 2⅜ inches in 2¼-inch holes. Their sockets are mounted between two sections of aluminum angle that span the bottom of the chassis as shown in Fig. 6-93. These strips are fastened inside of the bottom lip of the chassis and are held at the correct height by metal spacers on the lip. The exact position of the strips and of the mounting holes for the tube sockets is determined by holding the tubes in position and moving both the strips and the tube sockets into correct alignment, then marking the location of the required screw holes.

A 1-inch copper ground strap joins the centers of the two aluminum strips, and all ground connections in the plate, grid and filament circuits are then made to the strips. Both ends of the rotor of the grid tank condenser are

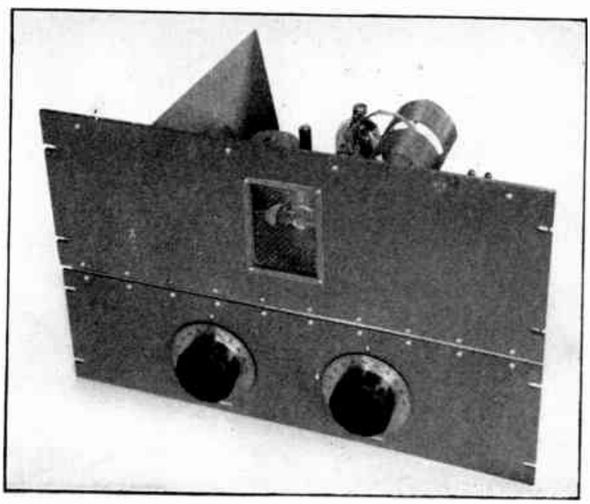

Fig. 6-90 — In this amplifier construction, the two tank condensers are mounted under the chassis to permit the upper portion of the panel to be hinged to provide access to the coils from the front of the rack.

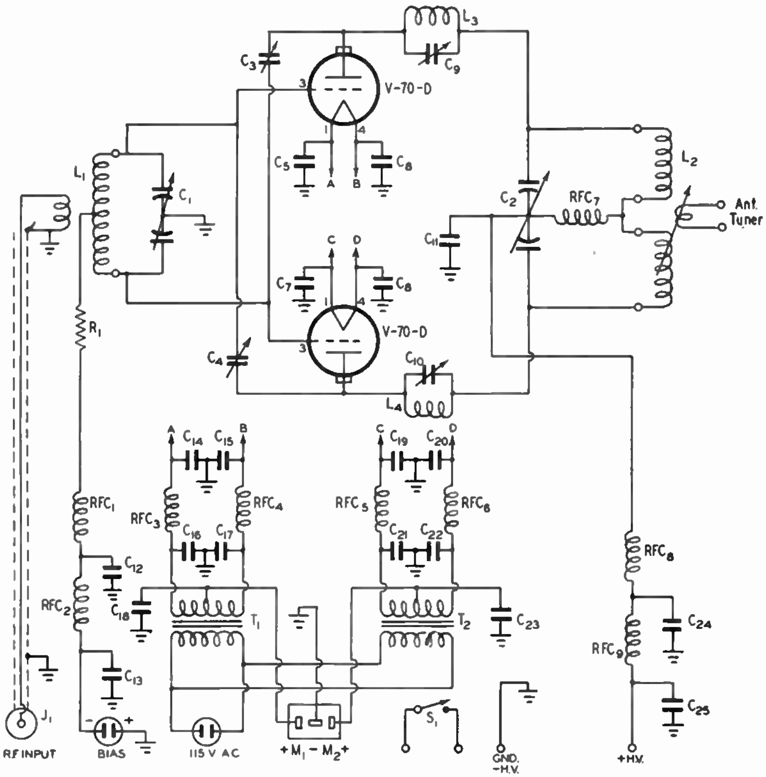

Fig. 6-91 — Circuit diagram of the front-access push-pull amplifier.

C_1 — 100 μμfd. per section, 0.03-inch spacing (National TMS-100-D).
C_2 — 100 μμfd. per section, 0.077-inch spacing (National TMC-100-D).
C_3, C_4 — Neutralizing condensers (National NC-800-A).
C_5, C_6, C_7, C_8, C_{18}, C_{23} — 0.0068-μfd. mica.
C_9, C_{10} — 30-μμfd. mica trimmer.
C_{11} — 0.001-μfd. 2500-volt mica.
C_{12}, C_{13}, C_{14}, C_{15}, C_{16}, C_{17}, C_{19}, C_{20}, C_{21}, C_{22} — 47-μμfd. mica.
C_{24}, C_{25} — 470-μμfd. 2500-volt mica.
R_1 — 100 ohms, 5 watts.
L_1 — B & W JCL series coils. 3.5 Mc. — 40 μh.; 7 Mc. — 15 μh.; 14 Mc. — 8 μh.; 21 Mc. — 3.5 μh.; 28 Mc. — 2.5 μh.
L_2 — Millen 500-watt series coils, variable link. 3.5 Mc. — 52 μh.; 7 Mc. — 17 μh.; 14 Mc. — 5.8 μh.; 21 Mc. — 2.6 μh.; 28 Mc. — 1.8 μh.
L_3, L_4 — 4 turns No. 14, ½-inch diam., ½ inch long.
J_1 — Coaxial input connector.
RFC_1, RFC_8 — V.h.f. choke (Ohmite Z-0).
RFC_2, RFC_9 — V h.f. choke (Ohmite Z-1).
RFC_3, RFC_4, RFC_5, RFC_6 — 7 turns No. 12, ½-inch diam., 1¼ inches long.
RFC_7 — 1-mh. r.f. choke (Millen 14140).
S_1 — Push-button interlock switch (Microswitch).
T_1, T_2 — Filament transformer; 7.5 volts, 3 amp. Connect positive terminals of milliammeters to $+M_1$, $+M_2$. Connect negative terminals to "–".

strapped to the copper bonding strap. Small two-terminal tie-points are supported by the aluminum strips, and are used to mount the components of the harmonic filters in the filament leads.

The neutralizing condensers are mounted on top of the chassis, slightly behind the tubes and between them, to provide short leads between the plate caps of the tubes and the movable plates of the neutralizing condensers. The leads from the fixed plates of the neutralizing condensers are passed through the chassis in ceramic bushings. These leads, which are of heavy copper wire, are crossed over below the chassis and run directly to the grid terminals on the tube sockets. The leads from the grid tank circuit to the grids of the tubes are then tapped along the neutralizing leads to a point determined during the neutralizing process (see "Neutralizing Procedure," this chapter).

The two filament transformers are mounted along the edge of the chassis, next to the aluminum partition that serves as a baffle shield between the tank coils. The partition is 7 inches high, extending the full depth of the chassis. An interlock switch, S_1, is mounted on the forward edge of the partition in such a position that it is actuated by the panel. The switch is held closed only when the hinged panel is closed. It is wired in series with the primary of the high-voltage transformer in the power supply. Thus, whenever the plate circuit is exposed to the front, high voltage is turned off automatically.

The grid coils plug into a 5-prong ceramic tube socket mounted at the center of the right-hand end of the chassis (Fig. 6-92) to provide short leads to the grid tuning condenser.

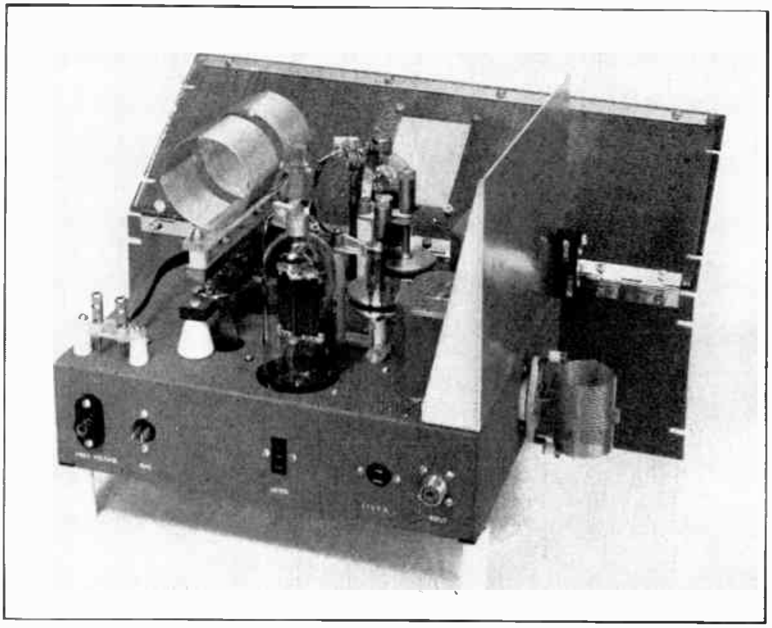

Fig. 6-92 — **Rear view of the hinged-panel arrangement, showing the placement of the tank coils, tubes and neutralizing condensers. The panel interlock switch is at the right, fastened to the shielding partition.**

The plate r.f. choke is fastened to the inside of the chassis just beneath one of the aluminum strips supporting the tube sockets. Terminals for bias, plate voltage, r.f. input, the a.c. line and for connection of the meters are mounted through the rear edge of the chassis. The v.h.f.-filter components in the bias and plate-supply leads are mounted close to their respective terminals inside the chassis. The output terminals are in the rear corner near the plate tank coil.

The amplifier should be checked and tuned following the procedure recommended under "Adjustment of R.F. Amplifiers," this chapter. Proper operating conditions for V-70-Ds, as well as other tubes in the same general class, will be found in the tables in this book or may be taken from tube-data sheets. When two milliammeters are connected as indicated under Fig. 6-91, each will read grid current when plate voltage is not applied and total cathode (the sum of plate and grid currents) when plate voltage is applied.

The circuit of a suitable power supply for operating this amplifier at full rating is shown in Fig. 6-89. Two VR75s should be used in the bias supply for plate-current cut-off. With these in use, R_3 should be set at 600 ohms. R_2 should be set at the highest resistance at which the VR tubes will ignite without amplifier excitation. The voltage of T_2 may be increased to 1500/1750.

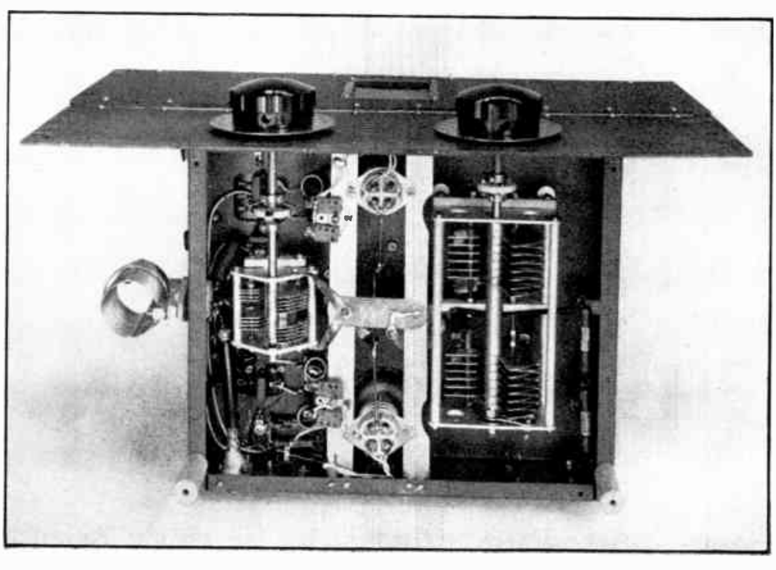

Fig. 6-93 — **Bottom view of the front-access amplifier, showing the two tank condensers and the sub-mounting of the tube sockets.**

A 1-Kw. Push-Pull Amplifier

The push-pull amplifier shown in the photographs of Figs. 6-94, 6-96 and 6-98 is built around a pair of Eimac 250TH triodes. It will handle a full kw. input at a plate voltage of 2000 or less, although the plate tank-condenser spacing is sufficient for 3000-volt operation with plate modulation. The driving stage should be capable of delivering approximately 100 watts. The amplifier may be shifted to any amateur band by a system of plug-in coils.

The circuit, shown in Fig. 6-95, is standard for a push-pull link-coupled neutralized amplifier. The only departure from strict conventionality is the use of the fixed vacuum-type padding condenser (C_9) across the plate tank coil when operating at 3.5 Mc. A filament transformer is included on the chassis to permit short leads which must carry the high heating current.

The components are mounted on a standard 10 × 17 × 3-inch chassis, with the 10-inch side against the panel to provide the necessary depth. The B & W "butterfly"-type plate tank condenser is mounted on heavy 2-inch stand-off insulators, with its shaft along the center line of the chassis and its front mounting feet centered 2 inches from the panel. Since its rotor is connected to the high-voltage supply, use of a good insulating shaft coupling is of utmost importance as a safety measure. The output tank-coil base assembly, with its adjustable link, is fastened to the two upper-rear stator nuts of the condenser by means of a pair of aluminum angle pieces. Similarly, the clips for the 3.5-Mc. vacuum-type padding condenser are mounted at the front of the condenser. Link output terminals are provided by the large stand-off insulators fastened to the rear of the panel near the top.

The neutralizing condensers are special units designed as accessories to the tank condenser. Each consists of a single disk connected to the grids, the rear stator plates of the plate tank condenser serving as the other side of the neutralizing condenser, for a compact unit. The by-pass condenser, C_7, is located under the rear end of the tank condenser and is fastened to the chassis with a small metal angle piece which makes the ground connection.

The sockets for the 250THs are submounted. They are spaced 5 inches, center to center, and 4 inches in from the rear edge of the chassis. The grid tank condenser is mounted between the tubes with an extension shaft to the front of the panel. The rotor plates are grounded to the chassis. The high-voltage line to the plate tank condenser and the plate r.f. choke is brought up through the chassis via a large ceramic feed-through insulator.

Underneath, the jack bar for the grid coil is centered between the tube sockets. Connections between this coil mounting and the condenser on top are made through large clearance holes lined with rubber grommets. Short, direct leads connect the tank circuit to the grid terminals of the tubes.

The filament transformer is mounted directly underneath the plate tank condenser. Since this transformer, as well as the grid coil, protrudes from the underside of the chassis, the chassis is set with its bottom edge 2½ inches above the bottom edge of the panel. The transformer shown in the photographs, and listed under Fig. 6-95, is one designed for rectifier service and has high-voltage insulation. If one with 1600- or 2000-volt insulation is available it may be substituted, of course. A Millen safety terminal for the positive high-voltage connection, a three-terminal ceramic strip for bias and ground connections, and a male power plug for the 115-volt connection to

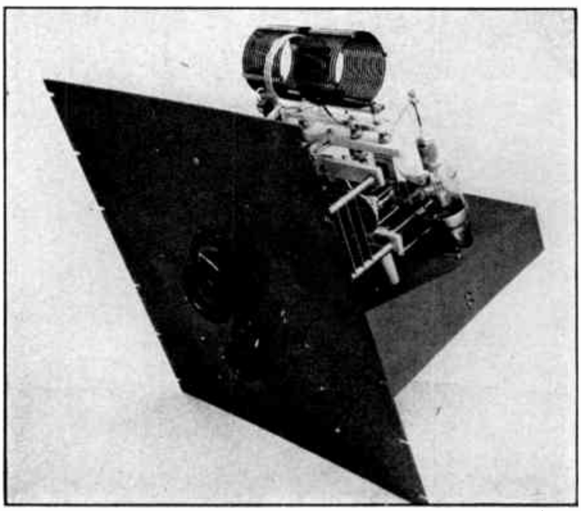

Fig. 6-94 — Front view of the kilowatt amplifier. The panel is 21 inches high and of standard 19-inch width.

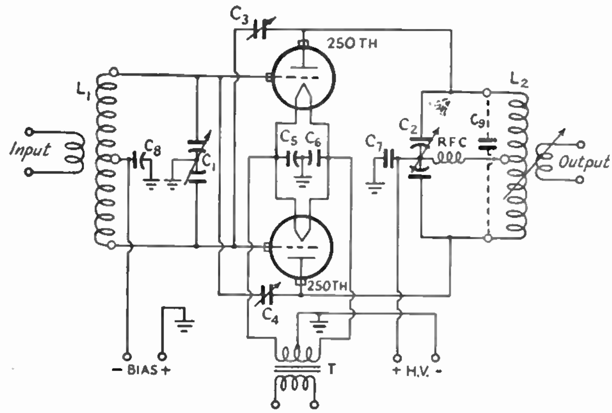

Fig. 6-95 — Circuit diagram of the high-power push-pull amplifier.
C_1 — 100 $\mu\mu$fd. per section, 0.05-inch spacing (Hammarlund HFBD-100-C).
C_2 — 60 $\mu\mu$fd. per section, 0.25-inch spacing (B & W CX62-C).
C_3, C_4 — Disk-type neutralizing condenser (B & W N-3).
C_5, C_6, C_8 — 0.01-μfd. paper, 600 volts.
C_7 — 0.001-μfd. mica, 10,000 volts.
C_9 — 25 $\mu\mu$fd., 16,000 volts (GE GL122).
L_1 — B & W BCL coils.
L_2 — B & W HDVL coils.
RFC — 1-mh. r.f. choke (Hammarlund CH-500).
T — 5 volts, 22 amperes (Stancor P6302: see text).

the filament transformer are set in the rear edge of the chassis while a pair of insulated terminals in the left rear corner is for the excitation input.

Power Supply

Fig. 6-97 shows the details of a suitable high-voltage plate and biasing supply for this amplifier. For a plate voltage of 2500, VR-90s in the bias supply will provide adequate voltage for plate-current cut-off. Five of them in parallel should be used to handle the necessary grid current. R_2 should be adjusted so that the tubes just ignite without excitation to the 250THs. For an operating bias of 150 volts, 60 volts must be obtained from the grid leak, R_3. At a grid current of 150 ma. under operating conditions, this will require a resistance of 400 ohms for R_3. The control switching system operates as described in connection with previously-mentioned supplies. S_1 turns on all filaments and the bias supply. S_2 turns on the exciter plate supply and sets up the circuit for S_4 which controls the amplifier plate supply.

Adjustment

When the amplifier is completed and ready for operation, the first step in adjustment is the neutralization. This may be done with the amplifier set up with all external connections made, except for the antenna and high voltage.

With the coils for the desired band plugged in, the tuning of the grid tank circuit should be adjusted until a grid-current reading is obtained. Then the neutralizing condensers should be adjusted simultaneously, bit by bit, keeping the spacing equal. When the amplifier is not neutralized, a dip in grid current will be found as the plate tank condenser is tuned through resonance. The neutralizing condensers should be adjusted until no change in grid current occurs as the plate tank condenser is swung through its range. This should occur with the adjustable plates of the neutralizing condensers spaced about 1$\frac{3}{16}$ inches away from the rear stator plates of the tank condenser.

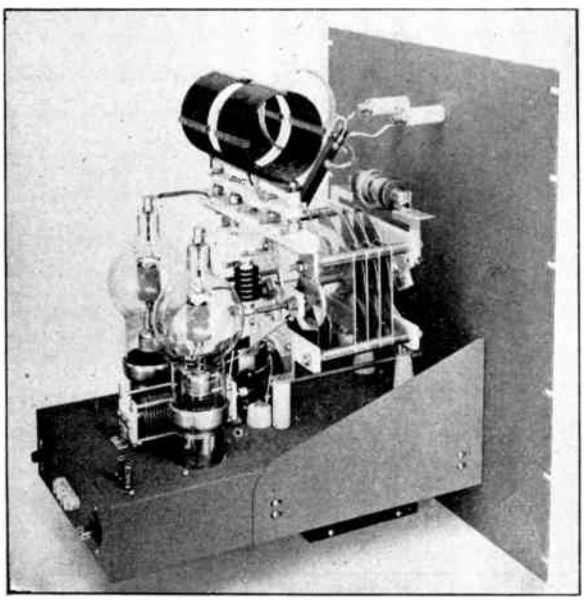

Fig. 6-96 — Rear view of the push-pull 250TH amplifier showing the mounting of the plate tank coil and 3.5-Mc. vacuum-type padding condenser.

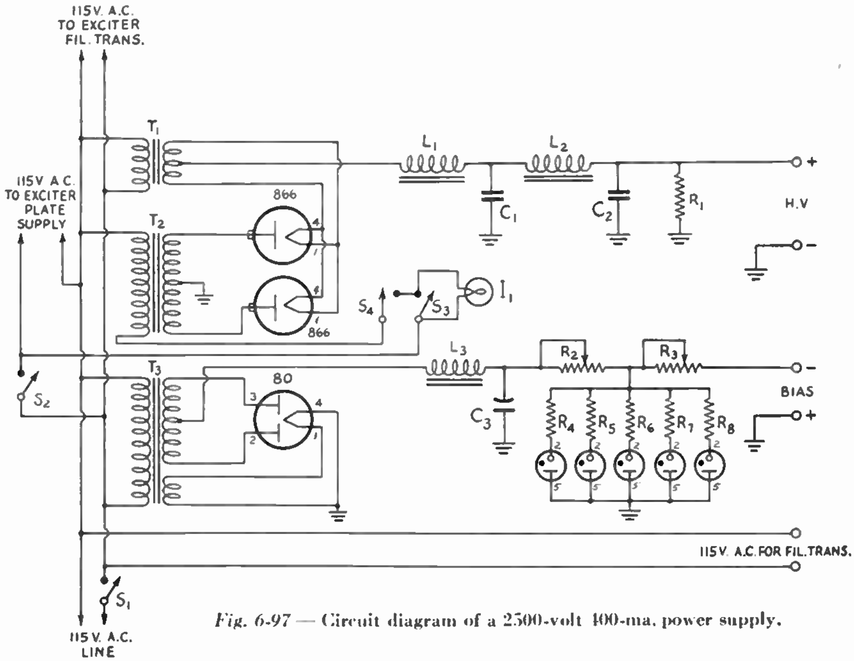

Fig. 6-97 — Circuit diagram of a 2500-volt 400-ma. power supply.

C_1, C_2 — 4-μfd. 3000-volt oil-filled.
C_3 — 8-μfd. 450-volt electrolytic.
R_1 — 50,000 ohms, 200 watts.
R_2 — 10,000 ohms, 25 watts, with slider.
R_3 — 1000 ohms, 25 watts, with slider.
R_4, R_5, R_6, R_7, R_8 — 100 ohms, 1 watt.
L_1 — 5/25-hy. 500-ma. swinging choke.
L_2 — 20-hy. 500-ma. smoothing choke.
L_3 — 30-hy. 50-ma. filter choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_2, S_4 — 15-amp. switch.
S_3 — 5-amp. switch.
T_1 — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T_2 — 2500-v. d.c. 500-ma. plate transformer.
T_3 — Power transformer, 650 v. a.c., c.t., 50 ma. or more; 5 volts, 2 amp.
See text for voltage-regulator tube data.

Although plenty of plate dissipation is available, it is desirable to do the preliminary tuning and loading of the amplifier at reduced plate voltage. Before plate voltage is applied, a grid-current reading of at least 150 to 200 ma. should be possible. The antenna link should be swung out to the minimum-coupling position. As soon as plate voltage and excitation are applied, the plate tank condenser should be adjusted for minimum plate current. Grid current still should be above 150 ma. When the excitation is removed, there should be no indication of oscillation at any setting of the grid- or plate-tank condenser.

The output link may be connected directly to a properly-terminated low-impedance line, or through a link-coupled antenna tuner to the feeders of any antenna system. With excitation and plate power applied, the plate current should increase as the link coupling is tightened and the antenna system tuned to resonance. With each adjustment of coupling or antenna tuning, the plate tank condenser should be retuned for minimum plate current. The minimum reading will increase as the coupling is tightened with the antenna tuned to resonance. The loading may be increased up to the point where the minimum reading is 400 ma., when the input will be 1 kw. at 2500 volts. With the amplifier loaded, the excitation should be adjusted to about 150 ma. for the two tubes.

Fig. 6-98 — The filament transformer and grid coil are mounted underneath the chassis.

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. 6-99. 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.

Each of the main vertical supporting members of the wooden rack is comprised of two pieces (*A* and *B*, and *I* and *J*) joined together at right angles. Each pair of these members is fastened together by No. 8 flat-head 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 $\frac{1}{4}$ inch plus the sum of the thickness and width of the material from either end of pieces *A* and *J*. This distance will be $3\frac{1}{4}$ inches for stock exactly 1 × 2 inches. The second hole will come $1\frac{1}{4}$ inches from the first, the third $\frac{1}{2}$ inch from the second, the fourth $1\frac{1}{4}$ inches from the third and so on, alternating spacings between $\frac{1}{2}$ inch and $1\frac{1}{4}$ inches (see detail drawing Fig. 6-100). All holes should be placed $\frac{3}{8}$ inch from the inside edges of the vertical members. Accompanying Table 6-VI shows standard panel heights and drilling dimensions.

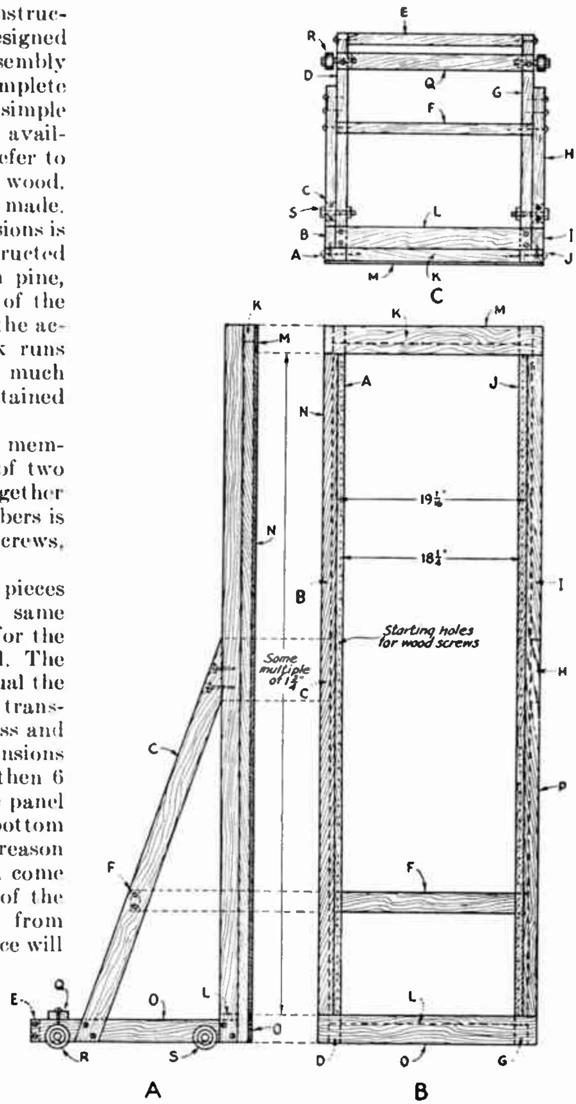

Fig. 6-99 — Detail drawing of a standard rack made of wood. A — Side view. B — Front view. C — Top view.

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 $17\frac{1}{2}$ inches apart at all points. This will bring the lines of mounting holes $18\frac{1}{4}$ 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 $\frac{1}{4}$-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 $\frac{1}{4}$ inch from the first mounting holes, and the distance between the inside edges of the vertical strips, *N* and *P*, $19\frac{1}{16}$ inches. To prevent the screw holes from wearing out when panels are changed frequently, ½ × 1/16 or 1/32-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 Fig. 6-100.) In a panel higher than two or three rack units (1¾ inches 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.

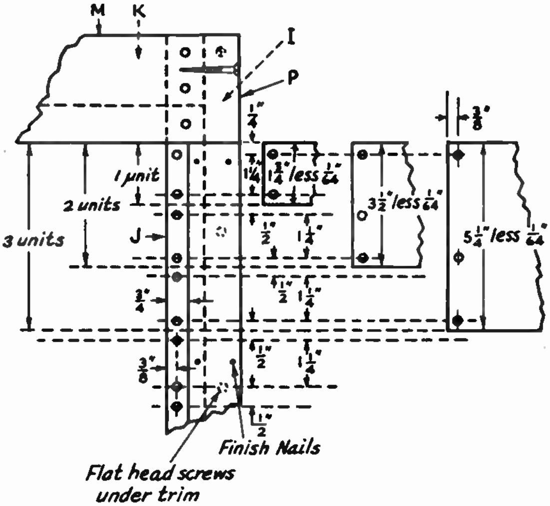

Fig. 6-100 — Detail sketch showing proper drilling for standard rack and panels. As shown for the 3½- and 5¼-inch panels, only sufficient holes are drilled in the panel to provide the necessary strength. When the panels are drilled as shown, they may be moved up and down in steps of 1¾ inches and the holes will always match.

If desired, the rack may be enclosed by completing a framework of one-by-two strip, using ¼-inch plywood for the panels. The panels may be hinged so that three sides are made accessible for servicing. If the transmitter is to be operated in an enclosure, provision should be made for a small amount of forced-air ventilation; otherwise the panels should be open while the transmitter is in operation.

TABLE 6-VI

Panel Height (Inches)	1¾	3½	5¼	7	8¾	10½	12¼	14	15¾
Panel Drilling (Inches)*	¼ 1½	2 3¼	3¾ 5	5½ 6¾	7¼ 8½	9 10¼	10¾ 12	12½ 13¾	14¼ 15½
Panel Height (Inches)	17½	19¼	21	22¾	24½	26¼	28	29¾	31½
Panel Drilling (Inches)*	16 17¼	17¾ 19	19½ 20¾	21¼ 22½	23 24¼	24¾ 26	26½ 27¾	28¼ 29½	30 31¼

* Additional holes for this size panel. Any or all holes given for panels smaller than this size may be added, as required for support.

Power Supplies

Essentially pure direct-current plate supply is required for receivers to prevent hum in the output. Government regulations require the use of d.c. plate supply for transmitters to prevent modulation of the carrier by the supply, which would result in undesired hum in the case of voice transmissions and an unnecessarily broad c.w. signal.

The filaments of tubes in a transmitter may be operated from a.c. Those in a receiver, excepting the power audio tubes, may be a.c. operated only if the cathodes are indirectly heated.

The comparative high cost and inconvenience of batteries and d.c. generators preclude their use except where commercial a.c. lines are not available. Wherever such lines are available, it is universal practice to obtain low a.c. voltage for filaments and heaters from a step-down transformer, and the required high-voltage d.c. by means of a transformer-rectifier-filter system. Such a system is shown in the block diagram of Fig. 7-1. Power from the a.c. line is fed to a transformer which steps the voltage up to that required. The stepped-up voltage is changed to pulsating d.c. by passing through a rectifier — usually of the vacuum-tube type. The pulsations then are smoothed out to the required extent by a filtering system.

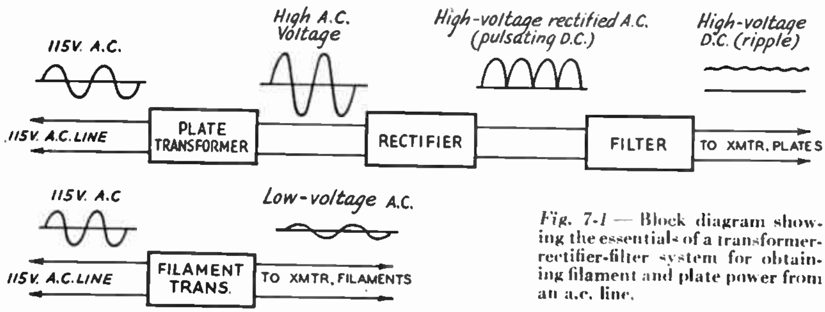

Fig. 7-1 — Block diagram showing the essentials of a transformer-rectifier-filter system for obtaining filament and plate power from an a.c. line.

Rectifier Circuits

Half-Wave Rectifier

Fig. 7-2 shows three rectifier circuits covering most of the common applications in amateur equipment. Fig. 7-2A is the circuit of a half-wave rectifier. During that half of the a.c. cycle when the rectifier plate is positive with respect to the cathode, current will flow through the rectifier and load. But during the other half of the cycle, when the plate is negative with respect to the cathode, no current can flow. The shape of the output wave is shown at the right. It shows that the current always flows in the same direction but that the flow of current is not continuous and is pulsating in amplitude.

The average output voltage — the voltage read by the usual d.c. voltmeter — with this circuit is 0.45 times the r.m.s. value of the a.c. voltage delivered by the transformer secondary. Because the frequency of the pulses in the output wave is relatively low, considerable filtering is required to provide adequately smooth d.c. output, and for this reason this circuit is usually limited to applications where the current involved is small, such as in supplies for cathode-ray tubes and for protective bias in a transmitter.

Full-Wave Center-Tap Rectifier

The most universally-used rectifier circuit is shown in Fig. 7-2B. Being essentially an arrangement in which the outputs of two half-wave rectifiers are combined, it makes use of both halves of the a.c. cycle. A transformer with a center-tapped secondary, or two identical transformers with their secondaries connected in series (with proper polarization), is required with the circuit. When the plate of rectifier No. 1 is positive, current flows through the load to the center-tap. Current cannot flow through rectifier No. 2 because at this instant its cathode is positive in respect to its plate. When the polarity reverses, rectifier No. 2 conducts and current again flows through the

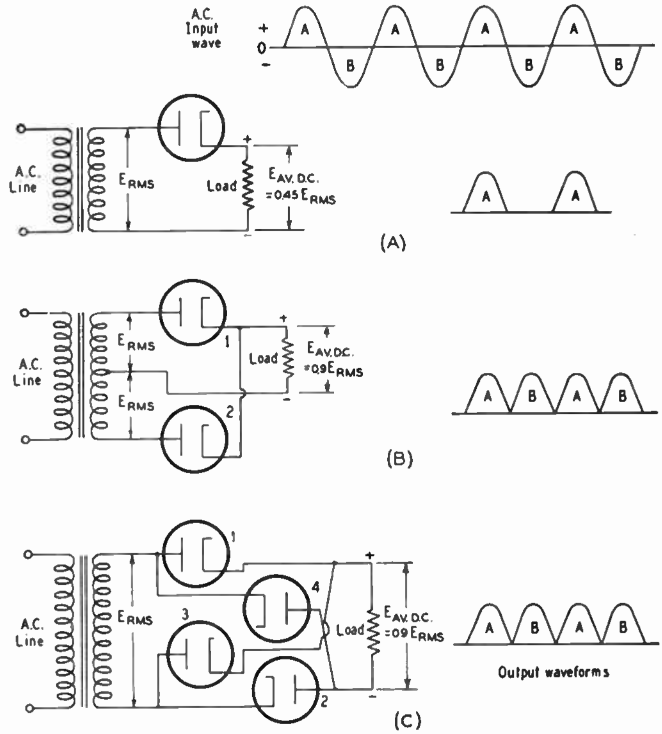

Fig. 7-2 — **Fundamental vacuum-tube rectifier circuits. A — Half-wave. B — Full-wave. C — Bridge.**

load to the center-tap, this time through rectifier No. 2.

The average output voltage is 0.9 times the r.m.s. value of the voltage across *half* of the transformer secondary. For the same *total* secondary voltage, the average output voltage will be the same as that delivered with a half-wave rectifier. However, as can be seen from the sketch of the output waveform, the frequency of the output pulses is twice that of the half-wave rectifier. Therefore much less filtering is required. Since the rectifiers work alternately, each handles half of the average load current. Therefore the load current which may be drawn from this circuit is twice the rated load current of a single rectifier.

Full-Wave Bridge Rectifier

Another full-wave rectifier circuit is shown in Fig. 7-2C. In this arrangement, two rectifiers operate in series on each half of the cycle, one rectifier being in the lead to the load, the other being in the return lead. Over that portion of the cycle when the upper end of the transformer secondary is positive with respect to the other end, current flows through rectifier No. 1, through the load and thence through rectifier No. 2. During this period current cannot flow through rectifier No. 4 because its plate is negative with respect to its cathode. Over the other half of the cycle, current flows through rectifier No. 3, through the load and thence through rectifier No. 4. The crossover connection keeps the current flowing in the same direction through the load. The output waveshape is the same as that from the simple center-tap rectifier circuit. The output voltage obtainable with this circuit is 0.9 times the r.m.s. voltage delivered by the transformer secondary. For the same total transformer-secondary voltage, the average output voltage when using the bridge rectifier will be twice that obtainable with the center-tap rectifier circuit. However, when comparing rectifier circuits for use *with the same transformer*, it should be remembered that the *power* which a given transformer will handle remains the same regardless of the rectifier circuit used. If the output voltage is doubled by substituting the bridge circuit for the center-tap rectifier circuit, only half the rated load current can be taken from the transformer without exceeding its normal rating. The value of load current which may be drawn from the bridge rectifier circuit is twice the rated d.c. load current of a single rectifier.

Rectifiers

Cold-Cathode Rectifiers

Tube rectifiers fall into three general classifications as to type. The cold-cathode type of rectifier is a diode which requires no cathode heating. Certain types will handle up to 350 ma. at 200 volts d.c. output. The internal voltage drop in most types lies between 60 and 90 volts. Rectifiers of this kind are produced in both half-wave (single-diode) and full-wave (double-diode) types.

High-Vacuum Rectifiers

High-vacuum rectifiers depend entirely upon the thermionic emission from a heated cathode and are characterized by a relatively high internal resistance. For this reason, their application usually is limited to low power, although there are a few types designed for medium and high power in cases where the relatively high internal voltage drop may be tolerated. This high internal resistance makes them less susceptible to damage from temporary overload and they are free from the bothersome electrical noise sometimes associated with other types of rectifiers.

Some rectifiers of the high-vacuum full-wave type in the so-called receiver-tube classification will handle up to 250 ma. at 400 to 500 volts d.c. output. Those in the higher-power class can be used to handle up to 500 ma. at 2000 volts d.c. in full-wave circuits. Most low-power high-vacuum rectifiers are produced in the full-wave type, while those for greater power are invariably of the half-wave type.

Mercury-Vapor Rectifiers

In mercury-vapor rectifiers the internal resistance is reduced by the introduction of a small amount of mercury which vaporizes under the heat of the filament, the vapor ionizing upon the application of voltage. The voltage drop through a rectifier of this type is practically constant at approximately 15 volts regardless of the load current. Tubes of this type are produced in sizes that will handle any voltage or current likely to be encountered in amateur transmitters. For high power they have the advantage of cheapness. Rectifiers of this type, however, have a tendency toward a certain type of oscillation which produces noise in near-by receivers. When encountered, this can usually be eliminated by suitable filtering.

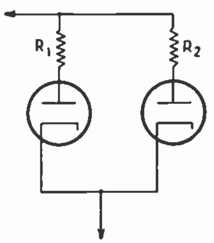

Fig. 7-3 — Connecting rectifiers in parallel for heavier currents. R_1 and R_2 should have the same value, between 50 and 100 ohms.

Selenium Rectifiers

Selenium rectifiers for power applications are a comparatively recent development. Units are now available with which it is possible to design a power supply capable of delivering up to 400 or 450 volts, 200 ma. These units have the advantage of compactness as well as low internal voltage drop (about 5 volts). Since they develop little heat if operated within their ratings, they are especially suitable for use in equipment requiring minimum temperature variation. Electrical noise filtering sometimes is required.

Rectifier Ratings

Vacuum-tube rectifiers are subject to limitations as to breakdown voltage and current-handling capability. Some types are rated in terms of the maximum r.m.s. voltage which should be applied to the rectifier plate, while others, particularly mercury-vapor types, are rated according to maximum inverse peak voltage — the peak voltage which appears between plate and cathode during the time the tube is not conducting. In all of the circuits shown in Fig. 7-2, the inverse peak voltage across each rectifier is 1.4 times the r.m.s. value of the voltage delivered by the entire transformer secondary.

The maximum d.c. output current is the maximum load current that can be drawn safely from the output of the filter. The value listed in tube tables is the value considered to be the safe maximum under average conditions. The exact value is dependent to a considerable extent, however, upon the nature of the filter that follows the rectifier.

A more significant rating is the maximum peak plate current. It is the peak value of the current pulses passing through the rectifier. This peak value can be much greater than the load current, especially if a large condenser is placed across the output of the rectifier as part of the filtering system, because of the large instantaneous charging current drawn by the condenser if there is no impedance between the rectifier and the condenser. These peaks do not run as high with high-vacuum-type rectifiers as they do with rectifiers of the mercury-vapor type because of the relatively high series resistance of the former.

Rectifiers may be connected in parallel for current higher than the rated current of a single unit. This includes the use of the sections of a double diode for this purpose. Equalizing resistors of 50 to 100 ohms should be connected in series with each plate, as shown in Fig. 7-3, as a measure toward maintaining an equal division of current between the two rectifiers.

Operation of Rectifiers

In operating rectifiers requiring filament or cathode heating, care should be taken to provide the correct filament voltage at the tube terminals. Low filament voltage can cause excessive voltage drop in high-vacuum rectifiers and a considerable reduction in the inverse peak voltage which a mercury-vapor tube will withstand without breakdown. Filament connections to the rectifier socket should be firmly soldered, particularly in the case of the larger mercury-vapor tubes whose filaments operate at low voltage and high current. The socket should be selected with care, not only as to contact surface but also as to insulation, since the filament usually is at full output voltage to ground. Bakelite sockets will serve at voltages up to 500 or so, but ceramic sockets, well spaced from the chassis, always should be used at the higher voltages. Special filament transformers with high-voltage insulation between primary and secondary are required for rectifiers operating at potentials in excess of 1000 volts inverse peak.

The rectifier tubes should be placed in the equipment with adequate free space surrounding them to provide for proper ventilation. When mercury-vapor tubes are first placed in service, they should be allowed to run only with filament voltage for ten minutes before applying high voltage.

Filters

The pulsating d.c. wave shown in Fig. 7-2 is not sufficiently smooth to prevent modulation. A filter consisting of chokes and condensers, as shown in Fig. 7-4, is connected between the rectifier output and the load circuit (transmitter or receiver) to smooth out the wave to the required degree.

The filter makes use of the energy-storage properties of the inductance of the choke and the capacitance of the condenser, energy being stored over the period during which the voltage and current are rising and releasing it to the load circuit during the period when the amplitude of the pulse is falling, thus leveling off the output by both lopping off the peaks and filling in the valleys.

Ripple Frequency and Voltage

The pulsations in the output of the rectifier can be considered to be the resultant of an alternating current superimposed upon a steady direct current. From this viewpoint, the filter may be considered to consist of shunting condensers which short-circuit the a.c. component while not interfering with the flow of the d.c. component, and series chokes which pass d.c. readily but which impede the flow of the a.c. component.

The alternating component is called the ripple. The effectiveness of the filter can be expressed in terms of per cent ripple which is the ratio of the r.m.s. value of the ripple to the d.c. value in terms of percentage. For c.w. transmitters, a reduction of the ripple to 5 per cent is considered adequate. The ripple in the output of power supplies for voice transmitters and VFOs should be reduced to 0.25 per cent or less. High-gain speech amplifiers and receivers may require a reduction to as low as 0.1 per cent to avoid objectionable hum.

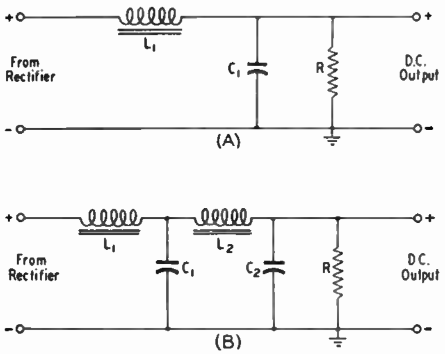

Fig. 7-4 — Choke-input filter circuits. A — Single-section. B — Double-section.

Ripple frequency is the frequency of the pulsations in the rectifier output wave — the number of pulsations per second. The frequency of the ripple with half-wave rectifiers is the same as the frequency of the line supply — 60 cycles with 60-cycle supply. Since the output pulses are doubled with a full-wave rectifier, the ripple frequency is doubled — to 120 cycles with 60-cycle supply.

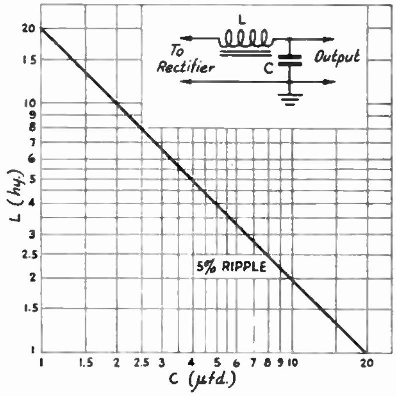

Fig. 7-5 — Graph showing combinations of inductance and capacitance that may be used to reduce ripple to 5 per cent with a single-section choke-input filter.

The amount of filtering (values of inductance and capacitance) required to give adequate smoothing depends upon the ripple frequency, more filtering being required as the ripple frequency is lower.

CHOKE-INPUT FILTERS

The filters shown in Fig. 7-4 are known as choke-input filters because the first element in the filter is a choke. This term is used in contrast to a condenser-input filter in which the first element is a condenser.

The percentage ripple output from a single-section filter (Fig. 7-4A) made up of any values of inductance and capacitance may be determined to a close approximation, for a ripple frequency of 120 cycles, from the following formula:

$$\left.\begin{array}{l}\text{Single-}\\\text{Section}\\\text{Filter}\end{array}\right\}\ \textit{Percentage ripple} = \frac{100}{LC}$$

where L is in hy. and C in μfd.

Example: $L = 5$ hy., $C = 4$ μfd.

$$\text{Percentage ripple} = \frac{100}{(5)\,(4)} = \frac{100}{20} = 5 \text{ per cent.}$$

Fig. 7-5 shows various other combinations of inductance and capacitance which will reduce the ripple to 5 per cent — the required minimum reduction for a supply for a c.w. transmitter.

Example: With a 10-hy. choke, what capacitance is required to reduce the ripple to 5 per cent?

Referring to Fig. 7-5, following the 10-hy. line horizontally, it intersects the ripple line at the 2-μfd. vertical line. Therefore the filter capacitance should be 2 μfd.

Example: With a 4-μfd. condenser, what choke inductance is required to reduce the ripple to 5 per cent?

Follow the vertical $C = 4$-μfd. line to the point where it intersects the ripple line; then follow the horizontal line at that point to read 5 hy., the required inductance.

In the case of a half-wave rectifier, the values of inductance and capacitance in the filter arrived at on the basis of a ripple frequency of 120 cycles must be doubled. It requires twice as much inductance and capacitance for the same degree of filtering with the half-wave circuit.

From the consideration of ripple reduction, any combination of inductances and capacitances which will give the required product and sum respectively will give the same ripple reduction. However, two other factors must be taken into consideration in the design of the filter. These are the peak rectifier current and voltage regulation.

Voltage Regulation

Unless the power supply is designed to prevent it, there may be a considerable difference between the output-terminal voltage of the supply when it is running free without an external load and the value when the external load is connected. Application of the load usually will be accompanied by a reduction in terminal voltage and this must be taken into consideration in the design of the supply. Regulation is commonly expressed as the percentage change in output voltage between no-load and full-load conditions in relation to the full-load voltage.

$$\textit{Per cent regulation} = \frac{100\,(E_1 - E_2)}{E_2}.$$

Example: No-load voltage $= E_1 = 1550$ volts.
Full-load voltage $= E_2 = 1230$ volts.

$$\text{Percentage regulation} = \frac{100\,(1550 - 1230)}{1230} = \frac{32{,}000}{1230} = 26 \text{ per cent.}$$

With proper design and the use of conservatively-rated components, a regulation of 10 per cent or less at the output terminals of the supply unit is possible with a choke-input filter. Good voltage regulation may or may not be of primary importance depending upon the nature of the load. If the load is constant, as in the case of a receiver, speech amplifier or the stages of a transmitter which are not keyed, voltage regulation, so far as that contributed by filter design is concerned, may be of secondary importance. The highly-stabilized voltage desirable for high frequency-stability of oscillators in receivers and transmitters is obtained by other means. Power supplies for the keyed stage of a c.w. transmitter and the stages following, and for Class B modulators, should have good regulation.

Bleeder Resistor

In general, a bleeder resistor is a resistance connected across the output of a filter to supply a minimum load (see R, Fig. 7-4). It also serves as a safety measure to discharge the filter condensers when the supply is turned off. The bleeder resistance need not be composed entirely of a resistor. Any constant load on the supply may serve the same purpose. In this case, a resistor of a high value should be used as a protective device to discharge the filter condensers.

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. 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 of the a.c. voltage applied to the rectifier. For both purposes, its impedance to the flow of the a.c. component 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:

$$L_{\text{crit.}} = \frac{\textit{Load resistance}\text{ (ohms)}}{1000}.$$

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. 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 minimum inductance required varies widely with the load resistance. In the case where there is no load except the bleeder on the power supply, the critical inductance required is the 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, it is possible to effect an economy in materials by designing the choke to have a "swinging" characteristic so 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. With any given input choke, the bleeder resistance (or other steady minimum load) should be 1000 times the maximum inductance of the choke in henrys.

Example: With a swinging choke of 5 to 20 hy., the bleeder resistance (or the resultant of the bleeder plus other steady load in parallel) should not exceed (20) (1000) = 20,000 ohms.

Output Condenser

If the supply is intended for use with an audio-frequency amplifier, the reactance of the last filter condenser should be small (20 per cent or less) compared with the other a.f. resistance or impedance in the circuit, usually the tube plate resistance and load resistance. On the basis of a lower a.f. limit of 100 cycles for speech amplification, this condition usually is satisfied when the output capacitance (last filter capacitor) of the filter is 4 to 8 μfd., the higher value of capacitance being used in the case of lower tube and load resistances.

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. 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, and resonance will occur when the product of choke inductance in henrys times condenser capacitance 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.

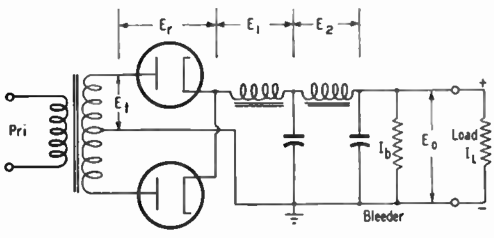

Fig. 7-6 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

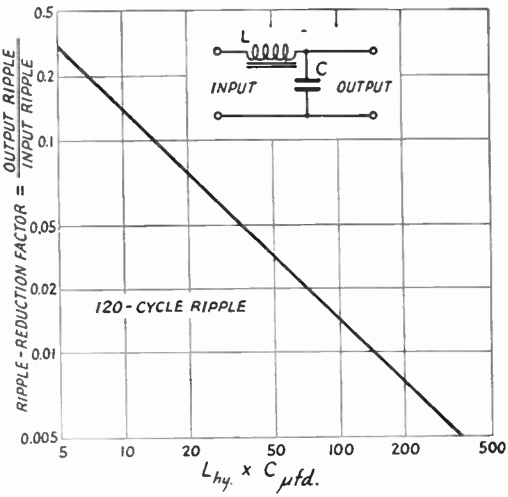

Fig. 7-7 — Ripple-reduction factor for various values of L and C in filter section. *Output ripple = input ripple × ripple factor.*

Output Voltage

Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by the following 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. These voltage drops are shown in Fig. 7-6. 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.

Additional Filtering

The graph of Fig. 7-7 shows the factor by which the ripple percentage may be reduced by the addition of one or more sections of filter, each similar in configuration to the first.

Example:

Ripple after first section = 5 per cent.
L in second section = 10 hy.
C in second section = 8 μfd.
$L \times C$ = 80.

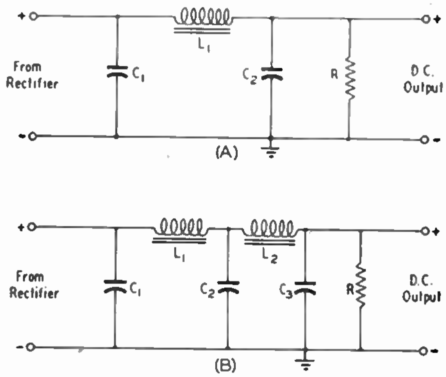

Fig. 7-8 — Condenser-input filter circuits. A — Single-section. B — Double-section.

From Fig. 7-7, the reduction factor is approximately 0.019. Therefore the ripple after the second section will be $5 \times 0.019 = 0.095$ per cent.

CONDENSER-INPUT FILTERS

Condenser-input filters are shown in Fig. 7-8. In comparison with a properly-designed input-choke filter, the d.c. output voltage is higher for most values of load, the ratio of peak rectifier plate current to d.c. output current is greater and the voltage regulation is considerably poorer.

The approximate performance of a filter consisting of the input condenser only is indicated in Figs. 7-9, 7-10 and 7-11. Fig. 7-9 shows the relationship between rectifier peak plate current and d.c. load current for various values of load and input resistance. Input resistance is the sum of transformer and rectifier resistances. In each case a capacitance of 4 μfd. or greater is assumed, since the characteristics change relatively little with higher values of capacitance.

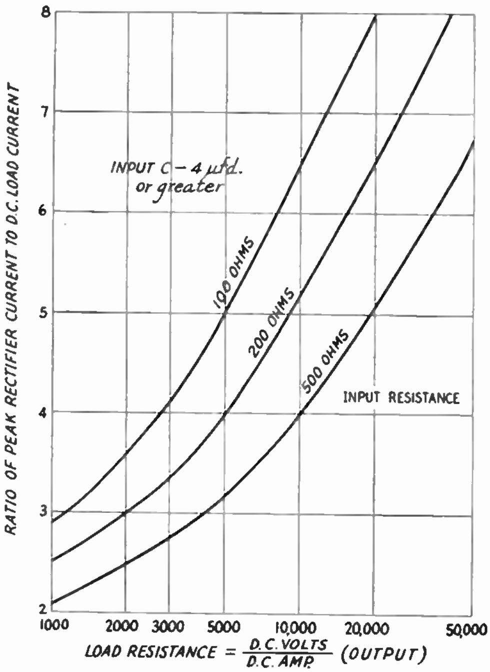

Fig. 7-9 — Graph showing relationship between d.c. load current and rectifier peak plate current with condenser input for various load and input resistances.

Fig. 7-10 shows the ratio of d.c. output voltage to the transformer r.m.s. voltage. In this respect too, the change with higher capacitance values is small.

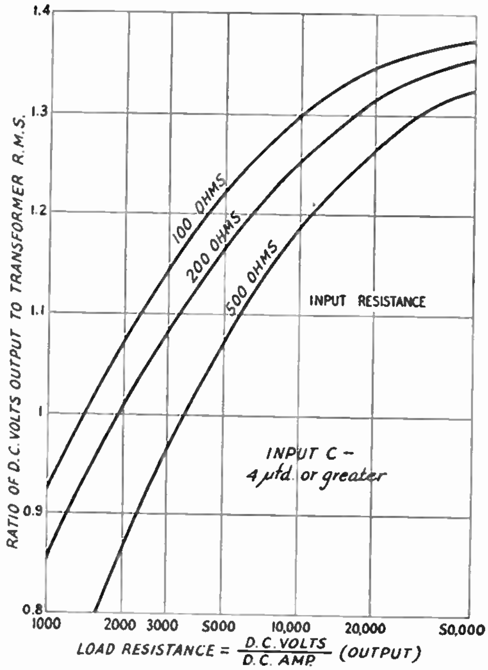

Fig. 7-10 — Chart showing approximate ratio of d.c. output voltage across filter input condenser to transformer r.m.s. secondary voltage for different load and input resistances.

Fig. 7-11 shows the approximate percentage ripple across the input condenser for capacitances of 4 and 8 μfd. The change in ripple voltage with normal differences in input resistance is relatively slight.

Further reduction in ripple may be obtained by adding sections of series inductance and parallel capacitance, as shown in Fig. 7-8. The reduction factor from Fig. 7-7 applies in this case also.

Example:

Input condenser — 4 μfd.
Output condenser — 8 μfd.
Input resistance — 200 ohms.
Transformer r.m.s. voltage — 400.
Load resistance (including resistance of filter choke) — 5000 ohms.

From Fig. 7-10, $\frac{\text{D.c. volts output}}{\text{Transformer r.m.s.}} = 1.17.$

D.c. volts output $= 400 \times 1.17 = 468$ volts.

From Fig. 7-9, $\frac{\text{Peak rectifier current}}{\text{D.c. load current}} = 4.$

D.c. load current $= \frac{468}{5000} = 93.6$ ma.

Peak rectifier current $= 93.5 \times 4 = 374$ ma.
From Fig. 7-11, ripple percentage across input condenser = approximately 8 per cent.
$L \times C = 8 \times 20 = 160.$
From Fig. 7-7, reduction factor = 0.009.
Output ripple percentage $= 8 \times 0.009 =$ 0.072 per cent.

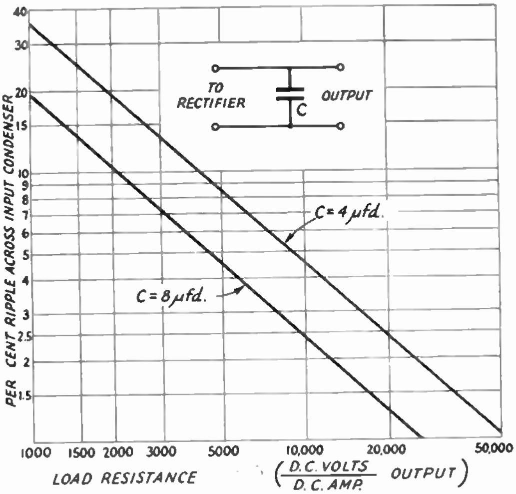

Fig. 7-11 — Chart showing approximate 120-cycle percentage ripple across filter input condenser for various loads.

RATINGS OF FILTER COMPONENTS

Although filter condensers in a choke-input filter are subjected to smaller variations in d.c. voltage than in the condenser-input filter, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no load on the power supply, since the voltage then will rise to the same maximum value as with a condenser-input filter.

In a condenser-input filter, the condensers should have a working-voltage rating 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 condenser should be used, or preferably a 1000-volt unit to allow a margin of safety.

Filter condensers are made in several different types. Electrolytic condensers, which are available for voltages up to about 800, combine high capacitance 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 that it will withstand continuously.

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 and be capable of handling the required load current.

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 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 value when full load current is flowing.

The Plate Transformer

Output Voltage

The output voltage which the plate transformer must deliver 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, the *total* secondary voltage being twice the desired d.c. output voltage.

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 ma., R_1 and R_2 are the resistances of the chokes, and E_r is the voltage drop in the rectifier. E_t is the full-load r.m.s. secondary voltage; the open-circuit voltage usually will be 5 to 10 per cent higher than the full-load value.

Volt-Ampere Rating

The volt-ampere rating 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, the secondary volt-amperes can be calculated quite closely by the equation:

$$Sec.\ V.A. = 0.00075EI$$

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.

Building Small Transformers

Power transformers for both filament heating and plate supply for all transmitting and rectifying tubes are available commercially, but occasionally the amateur wishes to build a transformer for some special purpose or has a core from a burned-out transformer on which he wishes to put new windings.

Most transformers that amateurs build are for use on 115-volt 60-cycle supplies. The number of turns necessary on the 115-volt winding depends on the kind of iron used in the core and on the cross-sectional area of the core. Silicon steel is best, and a flux density of about 50,000 lines per square inch can be used. This is the basis of the table of cross-sections given.

An average value for the number of primary turns to be used is 7.5 turns per volt per square inch of cross-sectional area. This relation may be expressed as follows:

$$No.\ primary\ turns = 7.5\left(\frac{E}{A}\right)$$

where E is the primary voltage and A the number of square inches of cross-sectional area of the core. For 115-volt primary transformers the equation becomes:

$$No.\ primary\ turns = \frac{863}{A}.$$

When a small transformer is built to handle a continuous load, the copper wire in the windings should have an area of 1500 circular mils for each ampere carried. (See Wire Table in Chapter Twenty-Four.) For intermittent use, 1000 circular mils per ampere is permissible.

The primary wire size is given in Table 7-I; the secondary wire size should be chosen according to the current to be carried, as previously described. The Wire Table in

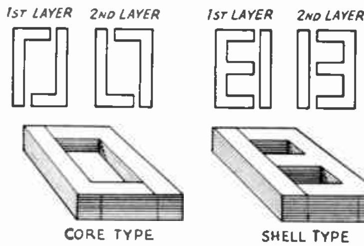

Fig. 7-12 — **Two different types of transformer cores and their laminations.**

Chapter Twenty-Four shows how many turns of each wire size can be wound into a square inch of window area, assuming that the turns are wound regularly and that no insulation is used between layers. The primary winding of a 200-watt transformer, which has 270 turns of No. 17 wire, would occupy 270/329 or 0.82 square inch if wound with double-cotton-covered wire, for example. This makes no allowance for a layer of insulation between the windings (in general, it is good practice to wind a strip of paper between each layer) so that the winding-area allowance should be increased if layer insulation is to be used. The figures also are based on accurate winding such as is done by machines; with hand-winding it is probable that somewhat more area would be required. An increase of 50 per cent should take care of both hand winding and layer thickness. The area to be taken by the secondary winding should be estimated, as should also the area likely to be occupied by the insulation between the core and windings and between the primary and secondary windings themselves. When the total window area required has been figured —

TABLE 7-I
Transformer Design

Input (Watts)	*Full-Load Efficiency*	*Size of Primary Wire*	*No. of Primary Turns*	*Turns Per Volt*	*Cross-Section Through Core*
50	75%	23	528	4.80	1¼" × 1¼"
75	85	21	437	3.95	1⅜ × 1⅜
100	90	20	367	3.33	1½ × 1½
150	90	18	313	2.84	1⅝ × 1⅝
200	90	17	270	2.45	1¾ × 1¾
250	90	16	248	2.25	1⅞ × 1⅞
300	90	15	248	2.25	1⅞ × 1⅞
400	90	14	206	1.87	2 × 2
500	95	13	183	1.66	2⅛ × 2⅛
750	95	11	146	1.33	2⅜ × 2⅜
1000	95	10	132	1.20	2½ × 2½
1500	95	9	109	0.99	2¾ × 2¾

(β = 32,000 lines per inch)

TABLE 7-II
Filter-Choke Design

L (Hy.)	*Ma.*	*Stack Size (Inches)*	*Core Length*		*Gap (Inches)*	*Winding Form*		*Turns*	*Wire Size*	*Feet*
			Long Piece	*Short Piece*		*b*	*c*			
15	50	½ × ½	½ × 2.2	½ × 0.85	0.035	1	0.68	9500	33	3500
10	100	¾ × ¾	¾ × 2.6	¾ × 0.95	0.03	1	0.67	5000	30	2250
15	100	1 × 1	1 × 3.1	1 × 0.9	0.035	0.96	0.65	4800	30	2550
10	250	2 × 2	2 × 5.2	2 × 1	0.4	1.05	0.68	2000	26	1750
20	250	2 × 2	2 × 5.6	2 × 1.2	0.28	1.43	0.95	4000	26	3820
5	500	2 × 2	2 × 5.5	2 × 1.15	0.17	1.35	0.9	1800	23	1700
10	500	2 × 2	2 × 6.2	2 × 1.5	0.4	2	1.3	3800	23	4100

allowing a little extra for contingencies — laminations having the desired leg-width and window area should be purchased. It may not be possible to get laminations having exactly the dimensions wanted, in which case the nearest size should be chosen. The cross-section of the core need not be square but can be rectangular in shape so long as the core area is great enough. It is easier to wind coils for a core of square cross-section, however.

Transformer cores are of two types, "core" and "shell." In the core type, the core is simply a hollow rectangle formed from two "L"-shaped laminations, as shown in Fig. 7-12. Shell-type laminations are "E"- and "I"-shaped, the transformer windings being placed on the center leg. Since the magnetic path divides between the outer legs of the "E," these legs are each half the width of the center leg. The cross-sectional area of a shell-type core is the cross-sectional area of the center leg. The shell-type core makes a better transformer than the core type, because it tends to prevent leakage of the magnetic flux. Calculations are the same for both types.

Fig. 7-13 shows the method of putting the windings on a shell-type core. The primary is usually wound on the inside — next to the core — on a form made of fiber or several layers of cardboard. This form should be slightly larger than the core leg on which it is to fit so that it will be an easy matter to slip in the laminations after the coils are completed and ready for mounting. The terminals are brought out to the side. After the primary is finished, the secondary is wound over it, several layers of insulating material being put between. If the transformer is for high voltages, the high-voltage winding should be carefully insulated from the primary and core by a few layers of Empire Cloth or tape. A protective covering of heavy cardboard or thin fiber should be put over the outside of the secondary to protect it from damage and to prevent the core from rubbing through the insulation. Square-shaped end pieces of fiber or cardboard usually are provided to protect the sides of the windings and to hold the terminal leads in place. High-voltage terminal leads should be enclosed in Empire Cloth tubing or spaghetti.

After the windings are finished the core should be inserted, one lamination at a time. Fig. 7-12 shows the method of building up the core. Alternate "E"-shaped laminations are pushed through the core opening from opposite sides. The "I"-shaped laminations are used to fill the end spaces, butting against the open ends of the "E"-shaped pieces. This method of building up the core ensures a good

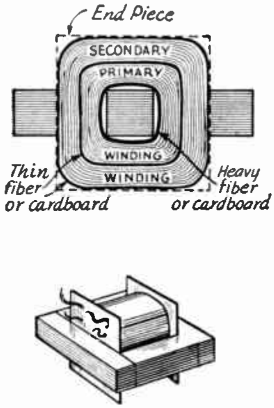

Fig. 7-13 — A convenient method of assembling the windings of a shell-type core. Windings can be similarly mounted on core-type cores, in which case the coils are placed on one of the sides. High-voltage core-type transformers sometimes are made with the primary on one core leg and the secondary on the opposite.

magnetic path of low reluctance. All laminations should be insulated from each other to prevent eddy currents from flowing. If there is iron rust or scale on the core material, that will serve the purpose very well — otherwise one side of each piece can be coated with thin shellac. It is essential that the joints in the core be well-made and be square and even. After the transformer is assembled, the joints can be hammered up tight using a block of wood between the hammer and the core to prevent damaging the laminations. If the winding form does not fit tightly on the core, small wooden wedges may be driven between it and the core to prevent vibration. Transformers built by the amateur can be painted with insulating varnish or waxed to make them rigid and moistureproof. A mixture of melted beeswax and rosin makes a good impregnating mixture. Melted paraffin should not be used because it has too low a melting point. Double-cotton-covered wire can be coated with shellac as each layer is put on. However, enameled wire should never be treated with shellac as it may dissolve the enamel and hurt the insulation, and it will not dry because the moisture in the shellac will not be absorbed by the insulation. Small transformers can be treated

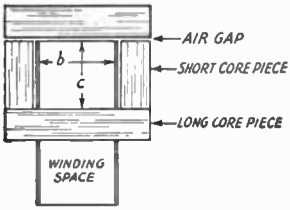

Fig. 7-14 — Core arrangement for filter choke coils. The dimensions *b* and *c* refer to Table 7-II.

with battery compound after they are wound and assembled. Strips of thin paper between layers of small enameled wire are necessary to keep each layer even and to give added insulation. Thick paper must be avoided since it keeps in the heat generated in the winding so that the temperature may become dangerously high.

Keep watch for shorted turns and layers. If just a single turn should become shorted in the entire winding, the voltage set up in it would cause a heavy current to flow which would burn it up, making the whole transformer useless.

Taps can be taken off as the windings are made if it is desired to have a transformer giving several voltages. Taps should be arranged, whenever possible, so that they come at the ends of the layers.

After leaving the primary winding connected to the line for several hours it should be only slightly warm. If it draws much current or gets hot there is something wrong. Some short-circuited turns are probably responsible and will continue to cause overheating.

Making Filter Chokes

Filter choke coils may be either of the core or shell type. The laminations should not be interleaved, a butt joint being used instead. An air gap must be provided at some point in the core circuit to prevent magnetic saturation by the d.c. flowing through the winding.

Table 7-II may be used as an approximate guide in winding choke coils. For the same core size, air gap and ampere turns, the inductance will vary approximately as the square of the number of turns. The arrangement of the core is shown in Fig. 7-14, and the dimensions *b* and *c* in the table refer to this sketch. The core may be built from straight pieces as shown or with "L"-shaped laminations.

Voltage Dropping

Series Voltage-Dropping Resistor

Certain plates and screens of the various tubes in a transmitter or receiver often require a variety of operating voltages differing from the output voltage of available power supplies. In most cases, it is not economically feasible to provide a separate power supply for each of the required voltages. If the current drawn by an electrode, or combination of electrodes operating at the same voltage, is reasonably constant under normal operating conditions, the required voltage may be obtained from a supply of higher voltage by means of a voltage-dropping resistor in series, as shown in Fig. 7-15A. The value of the series resistor, R_1, may be obtained from Ohm's Law, $R = \frac{E_d}{I}$, where E_d is the voltage *drop* required from the supply voltage to the desired voltage and I is the total rated current of the load.

Example: The plate of the tube in one stage and the screens of the tubes in two other stages require an operating voltage of 250. The nearest available supply voltage is 400 and the total of the rated plate and screen currents is 75 ma. The required resistance is

$$R = \frac{400 - 250}{0.075} = \frac{150}{0.075} = 2000 \text{ ohms.}$$

The power rating of the resistor is obtained from P (watts) $= I^2R = (0.075)^2\,(2000) = 11.2$ watts. A 25-watt resistor is the nearest safe rating to be used.

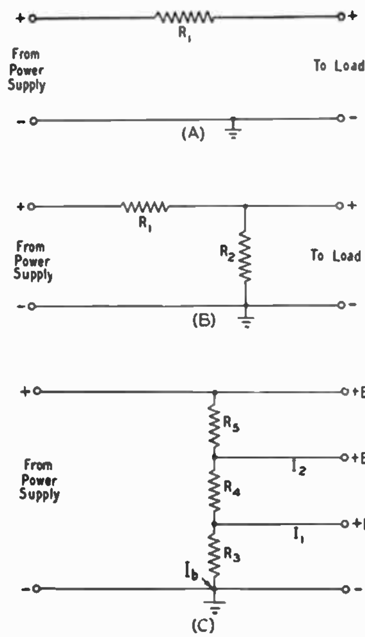

Fig. 7-15 — A — Series voltage-dropping resistor. B — Simple voltage divider. C — Multiple divider circuit.

$$R_3 = \frac{E_1}{I_b};\ R_4 = \frac{E_2 - E_1}{I_b + I_1};\ R_5 = \frac{E - E_2}{I_b + I_1 + I_2}$$

Voltage Dividers

The regulation of the voltage obtained in this manner obviously is poor, since any change in current through the resistor will cause a directly-proportional change in the voltage drop across the resistor. The regulation can be improved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 7-15B. Such an arrangement constitutes a voltage divider. The second resistor, R_2, acts as a constant load for the first, R_1, so that any variation in current from the tap becomes a smaller percentage of the total current through R_1. The heavier the current drawn by the resistors when they alone are connected across the supply, the better will be the voltage regulation at the tap.

Such a voltage divider may have more than a single tap for the purpose of obtaining more than one value of voltage. A typical arrangement is shown in Fig. 7-15C. 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_3, R_4, R_5, between taps. R_3 carries only the bleeder current, I_b; R_4 carries I_1 in addition to I_b; R_5 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 with the total load current (10 per cent or so). Then the required values can be calculated as shown in Fig. 7-15C, I being in amperes.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law using the voltage drop across it and the total current through it. The power dissipated by each section may be calculated either by multiplying I and E or I^2 and R.

Voltage Stabilization

Gaseous Regulator Tubes

There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit at a practically constant value, regardless of the voltage regulation of the power

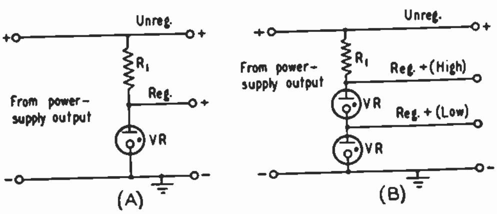

Fig. 7-16 — Voltage-stabilizing circuits using VR tubes.

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. Tubes are available for regulated voltages of 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 7-16A. The tube is connected in series with a limiting resistor, R_1, across a source of voltage that must be higher than the starting voltage. 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 40 ma.; consequently, the load current cannot exceed 30 to 35 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,

Fig. 7-17 — A receiver-type supply that delivers 250 volts at 35 ma. and a regulated potential of 150 volts at 15 ma. The amount of current which can be drawn from the 150-volt tap can be made higher or lower by selecting a suitable limiting resistor for the regulator tube; the current output will increase as the resistance value is reduced. The total current drain imposed on the supply should not exceed 50 ma. unless a transformer of greater current capacity is used. The four octal tube sockets on top of the chassis are wired in parallel with screw-type terminals and pin jacks at the rear to provide an assortment of terminals to which external circuits may be connected. The wiring diagram for the supply is shown in Fig. 7-18.

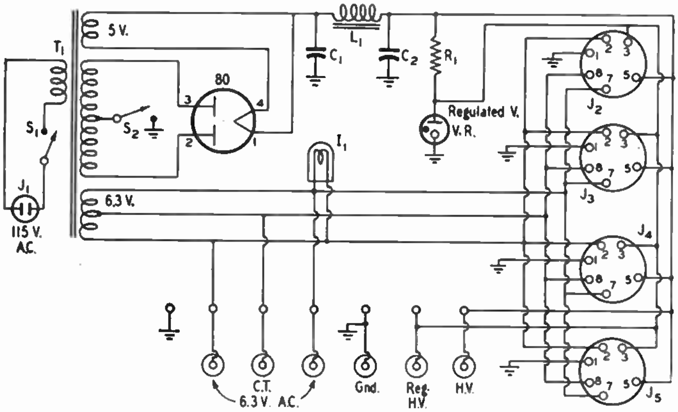

Fig. 7-18 — Circuit diagram of the receiver-type power supply.
C_1, C_2 — 8-μfd. 450-volt electrolytic.
R_1 — 15,000 ohms, 10 watts.
L_1 — 10-hy. 130-ma. 100-ohm filter choke.
I_1 — 6.3-volt pilot lamp.
J_1 — Panel-mounting a.c. plug (Amphenol 61M1).
J_2, J_3, J_4, J_5 — Octal socket.
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — Replacement-type power transformer: 290 volts each side of center-tap, 50 ma.; 5 volts, 3 amp.; 6.3 volts c.t., 2 amp.
A dual-unit electrolytic condenser may be used. The filter choke should have a fairly-high current rating as suggested above in order that the output voltage of the supply will not be reduced because of high resistance in the filter. Most available low-current chokes have a d.c. resistance of 500 ohms or more.

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 40 ma.).

Fig. 7-16B 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 30 to 35 milliamperes.

Voltage regulation of the order of 1 per cent can be obtained with regulator circuits of this type.

A small receiver-type power supply with a regulated voltage tap is illustrated in Fig. 7-17 and the circuit diagram appears in Fig. 7-18.

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. 7-19. A high-gain voltage-amplifier tube, usually a sharp cut-off pentode, 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 current and over a wide range of supply.

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

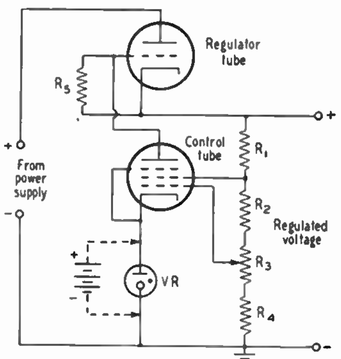

Fig. 7-19 — Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type regulator tube is used. Typical values: R_1, 10,000 ohms; R_2, 22,000 ohms; R_3, 10,000-ohm potentiometer; R_4, 4700 ohms; R_5, 0.47 megohm.

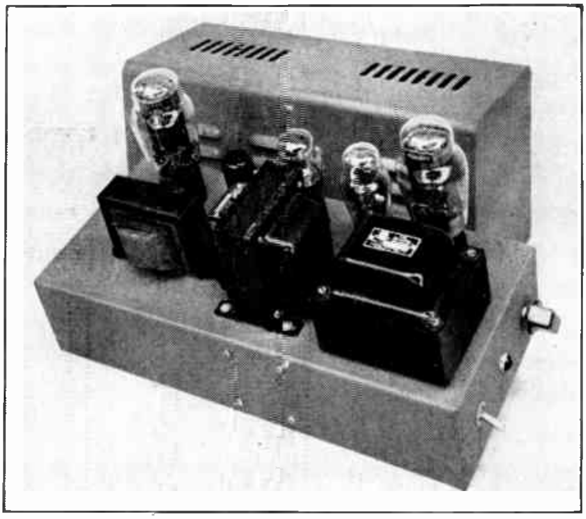

Fig. 7-20 — A heavy-duty electronically-regulated power supply. The unit is assembled on a 6 × 14 × 3-inch chassis fitted with an enclosing cover. The five tubes across the rear, left to right, are the 6AS7G regulator, the 6SJ7 control tube, the VR-105 bias regulator, the 1-V bias rectifier and the 5U4G power rectifier. In the foreground are the two filter chokes and the power transformer. The remainder of the components are mounted underneath.

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instead. If the gas tube or neon bulb is used, a negative-resistance type of oscillation 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, without need for changes in the circuit.

A heavy-duty regulated supply of this type is shown in Fig. 7-20. The circuit is shown in Fig. 7-21. A 6AS7G dual power triode is used as the regulator which is controlled by a 6SJ7. Reference bias is furnished by means of a 1-V half-wave rectifier whose output is regulated by a VR-105 regulator tube. The supply is capable of delivering 150 ma. over a range of 120 to 340 volts. Filament voltage and an external connection from the bias supply are also brought out to the output socket.

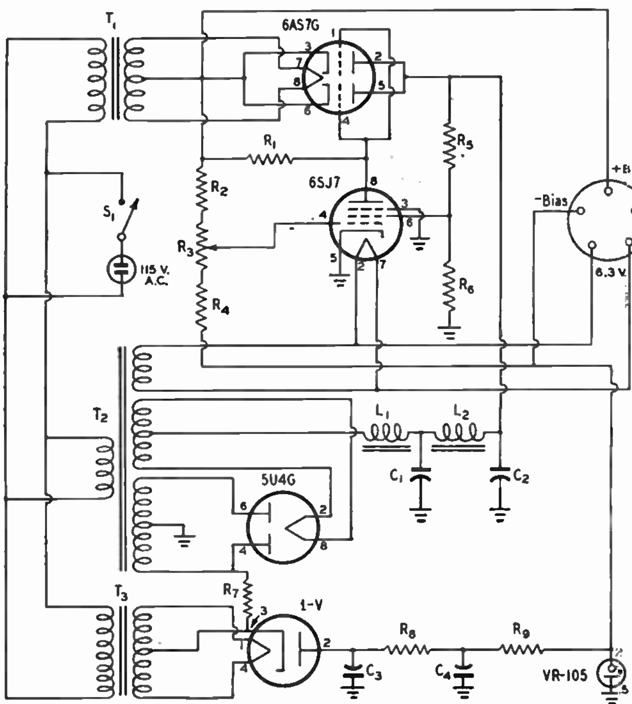

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Fig. 7-21 — Circuit diagram of the electronically-regulated power supply.

C_1, C_2, C_3, C_4 — 16-μfd. 450-volt electrolytic.
R_1 — 0.47 megohm, ½ watt.
R_2 — 0.18 megohm, ½ watt.
R_3 — 75,000-ohm potentiometer.
R_4 — 0.1 megohm, ½ watt.
R_5 — 50,000 ohms, 10 watts.
R_6 — 24,000 ohms, 2 watts.
R_7, R_8, R_9 — 2500 ohms, 10 watts.
L_1 — 8/30-hy. 150-ma. filter choke (Stancor C1718).
L_2 — 30-hy. 110-ma. filter choke (Stancor C1001).
S_1 — S.p.s.t. toggle switch.
T_1 — Filament transformer: 6.3 volts, 3 amp. (Stancor P-5014).
T_2 — Power transformer: 375-0-375 volts, 150 ma.; 5 volts, 3 amp.; 6.3 volts, 5 amp. (Stancor P-6014).
T_3 — Filament transformer: 6.3 volts, 1.2 amp. (Stancor P-6134).

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Miscellaneous Power-Supply Circuits

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. 7-22 and 7-23.

In Fig. 7-22, 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.

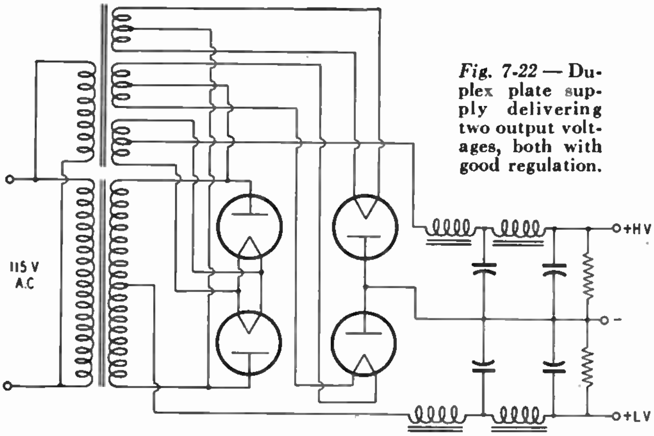


Fig. 7-22 — Duplex plate supply delivering two output voltages, both with good regulation.

Fig. 7-23 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 pair of taps. The filter chokes are placed in the common negative lead, but separate filter condensers are required. The sum of the currents drawn from each pair of taps must not exceed the transformer rating, and the chokes must carry the total load current. Each bleeder should have a value in ohms 1000 times the maximum rated inductance in henrys of the swinging choke, L_1, for best regulation. A power supply of this type is shown in Figs. 7-24 and 7-25. In this case two sets of chokes are used to divide the load current.

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. 7-26. 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 heaters and/or a resistor, *R*, of suitable value to limit the heater current to the rated value for the tubes.

The half-wave circuit shown has a fundamental ripple frequency equal to the line frequency and hence requires more inductance and capacitance in the filter for a given ripple percentage 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 is required to provide additional smoothing.

No ground connection can be used on the 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 rectifiers associated with the condensers, provided they are connected in the proper relationship.

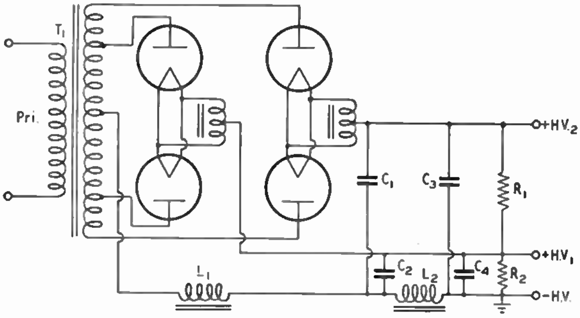


Fig. 7-23 — Power supply in which a single transformer and set of chokes serve for two different output voltages.

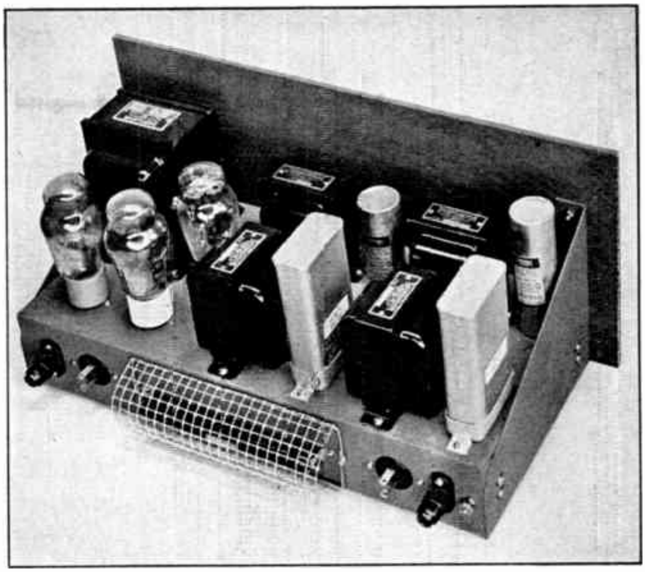


Fig. 7-24 — **This power supply makes use of a combination transformer and a dual filter system, delivering 1000 volts at 125 ma. and 400 volts at 150 ma., or 400 volts and 750 volts simultaneously, depending upon the transformer selected. The circuit diagram is given in Fig. 7-25. The 1000-volt bleeder 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 half-wave voltage doubler is shown in Fig. 7-27A. 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

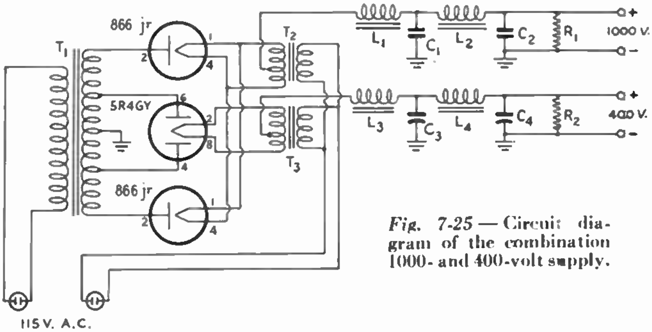


Fig. 7-25 — **Circuit diagram of the combination 1000- and 400-volt supply.**

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).
R_1 — 20,000 ohms, 75 watts.
R_2 — 20,000 ohms, 25 watts.
L_1, L_3 — 5/20-hy. swinging choke, 150 ma. (Thordarson T-19C39).
L_2, L_4 — 12-hy. smoothing choke, 150 ma. (Thordarson T-19C46).
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 amp. (Thordarson T-19F88).
T_3 — 5 volts, 4 amp. (Thordarson T-63F99).

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_2. C_2, however, does not receive its full charge because it begins 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_2 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. The voltage regulation is inherently poor and depends upon the capacities of C_1 and C_2, being better as these condensers are made larger. A 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. 7-28.

The voltage tripler in Fig. 7-27C comprises four diodes in a full-wave doubler and half-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 voltage of the combination is very nearly three times the line voltage, and the regulation is better than in other voltage-multiplier arrangements, as shown in Fig. 7-28.

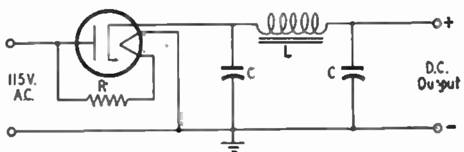


Fig. 7-26 — **Transformerless plate supply with half-wave rectifier. Other heaters are connected in series with *R*.**

Fig. 7-27D 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.

Selenium rectifiers can be used in these circuits to arrive at a very compact and lightweight power unit for portable work.

In the circuits of Fig. 7-27, C_2 should have a working voltage rating of 350 volts and C_1 of 250 volts for a 115-volt line. Their capacitances 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.

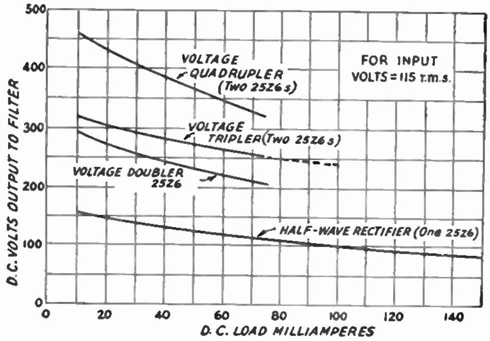


Fig. 7-28 — Curves showing the d.c. output voltage and the regulation under load for voltage-multiplier circuits.

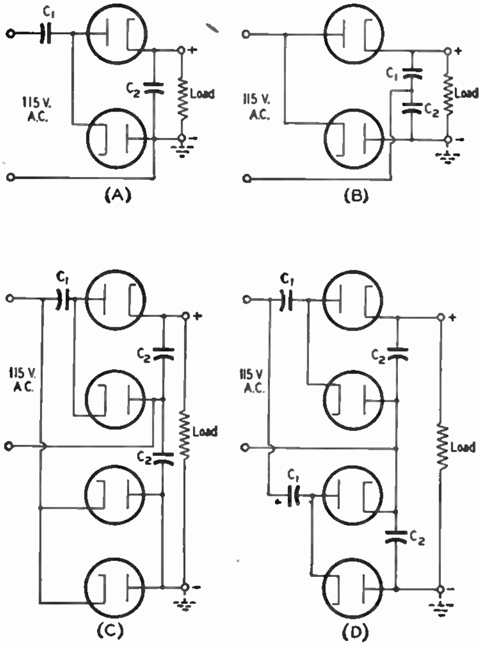


Fig. 7-27 — Voltage-multiplier circuits. A, half-wave voltage doubler. B, full-wave doubler. C, tripler. D, quadrupler. Dual-diode rectifier tubes may be used.

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 capacitance should be small (0.05 μfd.), since it is in shunt from plate to cathode of one rectifier. In addition to the fact that care must be exercised in avoiding direct ground connection or observation of proper line polarization to prevent short-circuiting the power line, transformerless supplies frequently give rise to other difficulties. For this reason their application is recommended only where economy or space is a prime consideration. A regenerative receiver operating from a transformerless supply has a greater tendency toward "tunable hum" than when operating from a supply equipped with a transformer. Apparatus operating from a transformerless supply often is the source of a rough hum when a near-by broadcast receiver is tuned to a carrier. A line filter in the supply, or a switch of line polarization, when this is permissible, sometimes will eliminate trouble of this type, but sometimes only the use of a transformer will be effective.

Bias Supplies

As discussed in Chapter Six, the chief function of a bias supply for the r.f. stages of a transmitter is that of providing protective bias, although under certain circumstances, a bias supply, or pack, as it is sometimes called, can provide the operating bias if desired.

Simple Bias Packs

Fig. 7-29A shows the diagram of a simple bias supply. R_1 should be the recommended grid leak for the amplifier tube. No grid leak should be used in the transmitter with this type of supply. The output voltage of the supply, when amplifier grid current is not flowing, should be some value between the bias required for plate-current cut-off and the recommended operating bias for the amplifier tube. The transformer peak voltage (1.4 times the r.m.s. value) should not exceed the recommended operating-bias value, otherwise the output voltage of the pack will soar above the operating-bias value when rated grid current flows.

This soaring can be reduced to a considerable extent by the use of a voltage divider across the transformer secondary, as shown at B. Such a system can be used when the transformer voltage is higher than the operating-bias value. The tap on R_2 should be adjusted to give amplifier cut-off bias at the output terminals. The lower the total value of R_2, the less the soaring will be when grid current flows.

A full-wave circuit is shown in Fig. 7-29C. R_3 and R_4 should have the same total resistance and the taps should be adjusted symmetrically. In all cases, the transformer must be designed to furnish the current drawn by these resistors plus the current drawn by R_1.

Regulated Bias Supplies

The inconvenience of the circuits shown in Fig. 7-29 and the difficulty of predicting values in practical application can be avoided in most cases by the use of gaseous voltage-regulator tubes across the output of the bias supply, as shown in Fig. 7-30A. A VR tube with a voltage rating anywhere between the biasing-voltage value which will reduce the input to the amplifier to a safe level when excitation is removed, and the operating value of bias, should be chosen. R_1 is adjusted, without amplifier excitation, until the VR tube ignites and draws about 5 ma. Any additional voltage to bring the bias up to the operating value when excitation is applied can be obtained from a grid leak, as discussed in Chapter Six.

Each VR tube will handle 40 ma. of grid current. If the grid current exceeds this value under any condition, similar VR tubes should be added in parallel, as shown in Fig. 7-30B, for each 40 ma., or less, of additional grid current. The resistors R_2 are for the purpose of helping to maintain equal currents through each VR tube.

If the voltage rating of a single VR tube is not sufficiently high for the purpose, other VR tubes may be used in series (or series-parallel if required to satisfy grid-current requirements) as shown in Fig. 7-30C and D.

If a single value of fixed bias will serve for more than one stage, the biasing terminal of each such stage may be connected to a single supply of this type, provided only that the total grid current of all stages so connected does not exceed the current rating of the VR tube or tubes. Alternatively, other separate VR-tube branches may be added in any desired combination to the same supply, as shown in Fig. 7-30E, to suit the needs of each stage.

Providing the VR-tube current rating is not exceeded, a series arrangement may be tapped for lower voltage, as shown at F.

Other Sources of Biasing Voltage

In some cases, it may be convenient to obtain the biasing voltage from a source other than a separate supply. A half-wave rectifier may be connected with reversed polarization to obtain biasing voltage from a low-voltage plate supply, as shown in Fig. 7-31A. In another arrangement, shown at B, a spare filament winding can be used to operate a filament transformer of similar voltage rating in reverse to obtain a voltage of about 130 from the winding that is customarily the primary. This will be sufficient to operate a VR75 or VR90.

A bias supply of any of the types discussed requires relatively little filtering, if the output-terminal peak voltage does not approach the operating-bias value, because the effect of the supply is entirely or largely "washed out" when grid current flows.

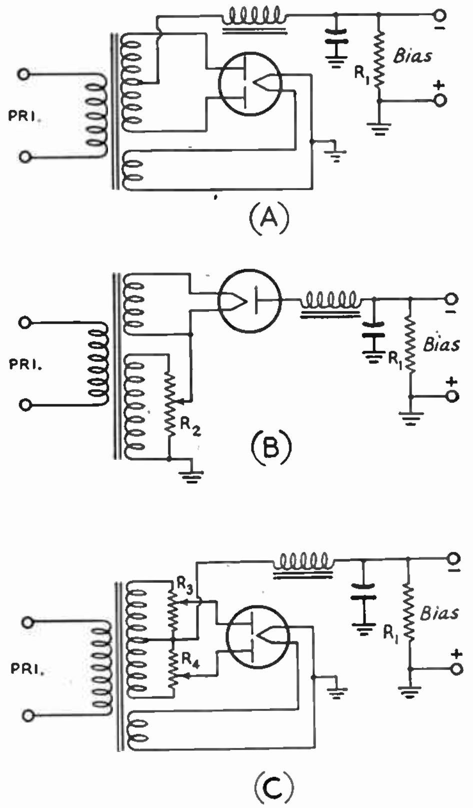


Fig. 7-29 — Simple bias-supply circuits. In A, the peak transformer voltage must not exceed the operating value of bias. The circuits of B (half-wave) and C (full-wave) may be used to reduce transformer voltage to the rectifier. R_1 is the recommended grid-leak resistance.

Other Power Considerations

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 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 across it. The filament or heater transformer generally is center-tapped, to provide a balanced circuit for eliminating hum.

For medium- and high-power r.f. stages of

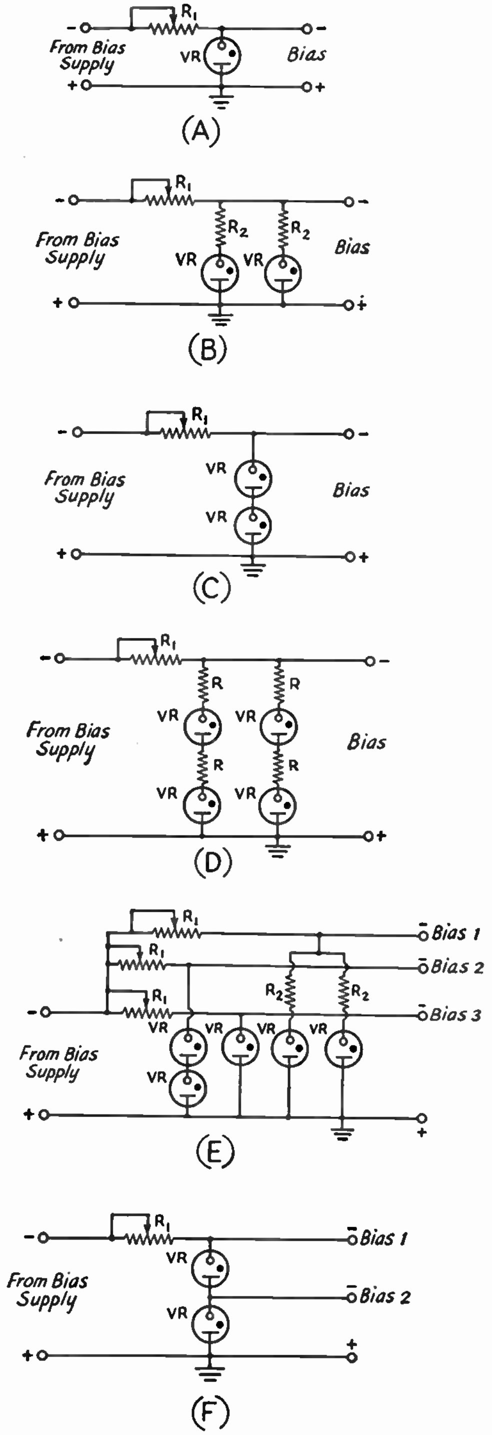


Fig. 7-30 — Illustrating the use of VR tubes in stabilizing protective-bias supplies. R_1 is a resistor whose value is adjusted to limit the current through each VR tube to 5 ma. before amplifier excitation is applied. R and R_2 are current-equalizing resistors of 50 to 100 ohms.

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.

LINE-VOLTAGE ADJUSTMENT

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line and, since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on and off at evening, they may be taken care of by the use of a manually-operated compensating device. A simple arrangement is shown in Fig. 7-32A. A toy transformer is used to boost or buck the line voltage as required. The transformer should have a tapped secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current of the entire transmitter, or that portion of it fed by the toy transformer.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 115 volts by setting the toy-transformer tap switch on the right tap. If the phasing of the two windings of the toy transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 115 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously. The circuit of 7-32B illustrates the use of a variable transformer (Variac) for adjusting line voltage to the desired value.

Another scheme by which the primary voltage of each transformer in the transmitter may be adjusted to deliver the desired secondary voltage, with a master control for compensating for changes in line voltage, is described in Fig. 7-33.

This arrangement has the following features:

1) Adjustment of the switch S_1 to make the voltmeter read 105 volts automatically adjusts all transformer primaries to the predetermined correct voltage.

2) The necessity for having all primaries work at the same voltage is eliminated. Thus, 110 volts can be applied to the primary of one transformer, 115 to another, etc.

3) Independent control of the plate transformer is afforded by the tap switch S_2. This permits power-input control and does not require an extra autotransformer.

CONSTRUCTION OF POWER SUPPLIES

The length of most leads in a power supply is unimportant, so that the arrangement of components from this consideration is not a factor in construction. More important are the points of good high-voltage insulation, adequate conductor size for filament wiring, proper ventilation for rectifier tubes and — most important of all — safety to the operator. Exposed high-voltage terminals or wiring which might be bumped into accidentally should not be permitted to exist. They should be covered with adequate insulation or placed inaccessible to contact during normal operation and adjustment of the transmitter.

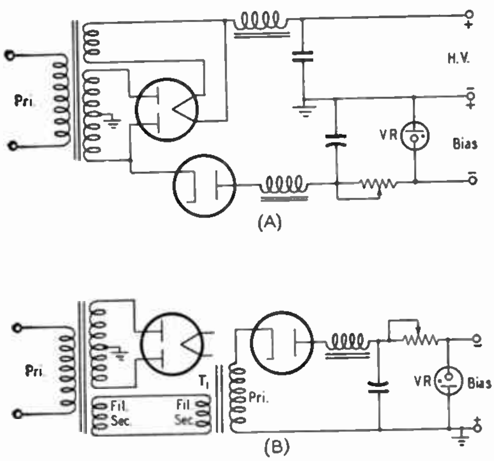


Fig. 7-31 — Convenient means of obtaining biasing voltage. A — From a low-voltage plate supply. B — From spare filament winding. T_1 is a filament transformer, of a voltage output similar to that of the spare filament winding, connected in reverse to give 115 volts r.m.s. output. If cold-cathode or selenium rectifiers are used, no additional filament supply is required.

Rectifier filament leads should be kept short to assure proper voltage at the rectifier socket, and the sockets should have good insulation and adequate contact surface. Plate leads to mercury-vapor tubes should be kept short to minimize the radiation of noise.

Where high-voltage wiring must pass through a metal chassis, grommet-lined clearance holes will serve for voltages up to 500 or 750, but ceramic feed-through insulators should be used for higher voltages. Bleeder and voltage-dropping resistors should be placed where they are open to air circulation. Placing them in confined space reduces the power rating.

It is highly preferable from the standpoint of operating convenience to have separate filament transformers for the rectifier tubes, rather than to use combination transformers, such as those used in receivers. This permits the plate voltage to be switched on without the necessity for waiting for rectifier filaments to come up to temperature after each time the high voltage has been turned off.

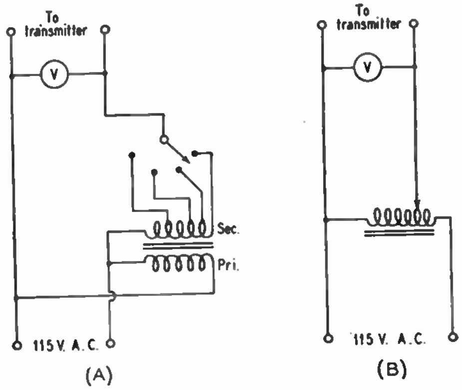


Fig. 7-32 — Two methods of transformer primary control. At A is a tapped toy transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) in series with the transformer primaries.

A bleeder resistor with a power rating giving a considerable margin of safety should be used across the output of all transmitter power supplies so that the filter condensers will be discharged when the high-voltage transformer is turned off. To guard against the possibility of danger to the operator should the bleeder resistor burn out without his knowledge, a relay with its winding connected in parallel with the high-voltage transformer primary and its contacts in series with a 1000-ohm resistor across the output of the power supply sometimes is used. The relay should be arranged so that the contacts open when the relay is energized.

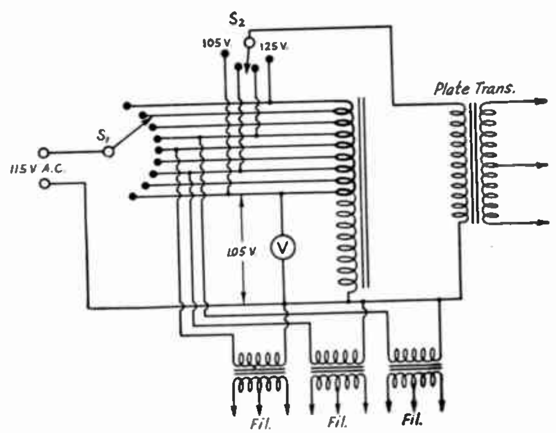


Fig. 7-33 — With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter. Information on constructing a suitable autotransformer at negligible cost is contained in the text. The light winding represents the regular primary winding of a revamped transformer, the heavy winding the voltage-adjusting section.

Control Systems

A well-planned system of controlling power-supply equipment is not only a matter of safety to the operator but also a factor in the convenient and efficient operation of the station while on the air.

The diagrams of power supplies suggested for use with transmitters described in Chapter Six include a suitable control system for each. In general principle they are the same, varying only in the details of special considerations for the specific case at hand.

As a minimum, except possibly in the case of simple transmitters employing a single power supply, there should be a filament switch that controls simultaneously all filaments in the power supplies as well as in the transmitter. This switch sometimes also controls the bias supply if one is used. There should be a separate switch for each plate-voltage supply and one that controls all plate supplies simultaneously. This latter switch is the "stand-by" switch by which power to the transmitter may be turned off quickly during receiving periods. The switches should be arranged in series, so that the plate voltage cannot be applied before filament and bias voltages have been turned on.

Figs. 7-34 and 7-35 show a complete control system for a multistage c.w. and 'phone transmitter. Indicator lamps, proper line fusing and automatic protective features are included. These circuits are more or less basic and will cover all requirements of most transmitters. They are similar except that the circuit of Fig. 7-34 is for use with a 115-volt line, while that of Fig. 7-35 is suitable for a 3-wire 220-volt line.

The system starts out with a polarized plug, P_1, for the line connection. The side of the line indicated should be the grounded side. One or more utility outlets which are not affected by the switching may be connected at J_1. The line-fuse indicator lamp, I_1, should not light unless the line fuse, F_1, is blown.

Turning on S_1 at the transmitter or S_2 at the operating position turns on all r.f. and r.f. power-supply filament transformers, which are connected in parallel at T_1, and the indicator lamp, I_2, lights. If the 'phone-c.w. switch, S_3, is thrown to the 'phone position, all audio and a.f.-supply filament transformers, which are connected in parallel at T_2, will also be turned on by S_1 and the 'phone indicator lamp, I_3, will light. If S_3 is in the c.w. position, the c.w. indicator lamp, I_4, will light, and the a.f. power supplies will be cut off. S_{3B} short-circuits the modulation-transformer secondary.

If the safety interlock switch, S_4, is closed, the bias-supply plate and filament voltages (T_3) will be turned on. As soon as the rectifier of this supply (an indirectly-heated rectifier such as a 6X5G) warms up, and the supply delivers full voltage, the delay relay, Ry_2, will close, extinguishing the bias-indicator lamp, I_7, and setting up the circuit for the plate-supply relay, Ry_3. The time which the bias rectifier takes to come up to temperature provides the required delay between the application of filament voltage and the time when it becomes possible to turn on the plate voltages on the r.f. and a.f. tubes.

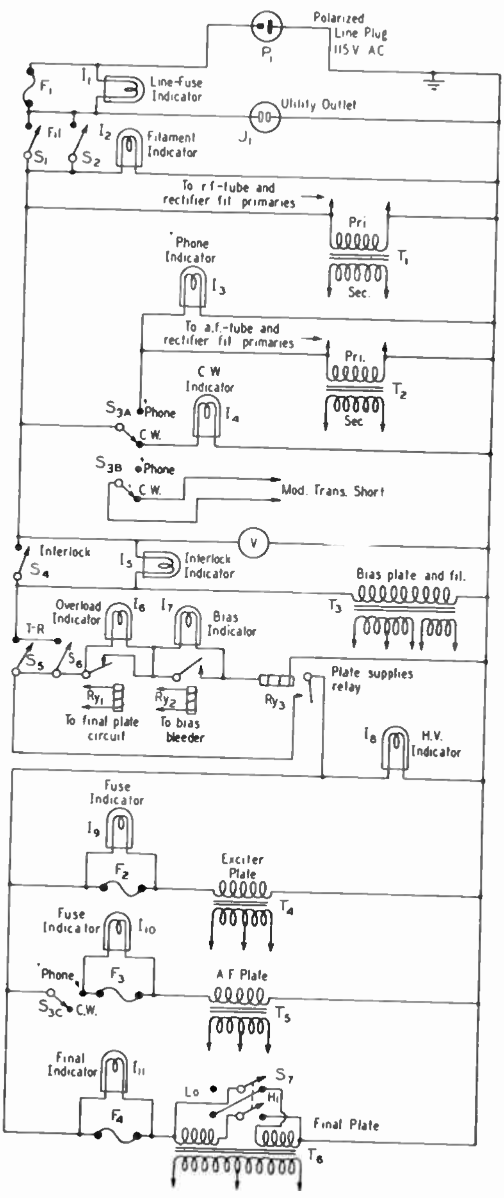


Fig. 7-34 — 115-volt control circuit used at W4DCW. All switches, except S_3, may be 5 amp. S_3 should be a ceramic rotary switch. The indicator lamps are ⅝-inch panel type.

With the contacts of Ry_2 closed, the plate-supply relay, Ry_3, can be operated by closing the transmit-receive switch, S_5, or its extension, S_6, at the operating position. Ry_3 turns on all plate voltages, lights the high-voltage indicator, I_8, and the transmitter is ready for operation.

Should the interlock switch S_4 be open, the indicator lamp, I_5, will light. This lamp, in series with the primary of the bias-supply transformer, has sufficient resistance to prevent lighting of the rectifier filament and thus voltage output from the bias pack, and therefore Ry_2 does not close so that Ry_3 cannot be operated and high voltage cannot be applied, making the transmitter safe so long as the interlock switch is open.

Ry_1 is an overload breaker which breaks off the line to the plate-supply relay whenever the plate current to the final amplifier exceeds a value to which it has been set. The winding of this relay is in the filament center-tap of the final-amplifier tubes. It should be of the reset type so that it will not continue to close and open repeatedly until S_5 is opened as it would do if it were not of the reset type. I_9, I_{10} and I_{11} are fuse-indicator lamps which light when their associated fuses blow. S_7 is a switch for changing to low power for tune-up. This system is, of course, applicable only to transformers with dual primaries. With single-primary transformers, a switch can be arranged to short-circuit a 150- to 200-watt lamp connected in series with the primary winding for reducing power.

The only switch which need be thrown for stand-by is S_5. Only S_3 need be thrown in changing from 'phone to c.w. No other switching is necessary.

The only difference in Fig. 7-35 is that the filament and bias transformers are operated from one side of the line, while the plate supplies are operated from the other. This connection is preferable whenever it can be applied, since it helps to equalize the loads drawn from each side of a 3-wire line. For high power, or in cases where light blinking is experienced, the plate transformer, T_6, should have a 220-volt primary connected to the two outside wires.

All indicator lamps and panel switches should be marked plainly so that there will be no question as to which circuit each belongs, to facilitate switching and localizing of trouble.

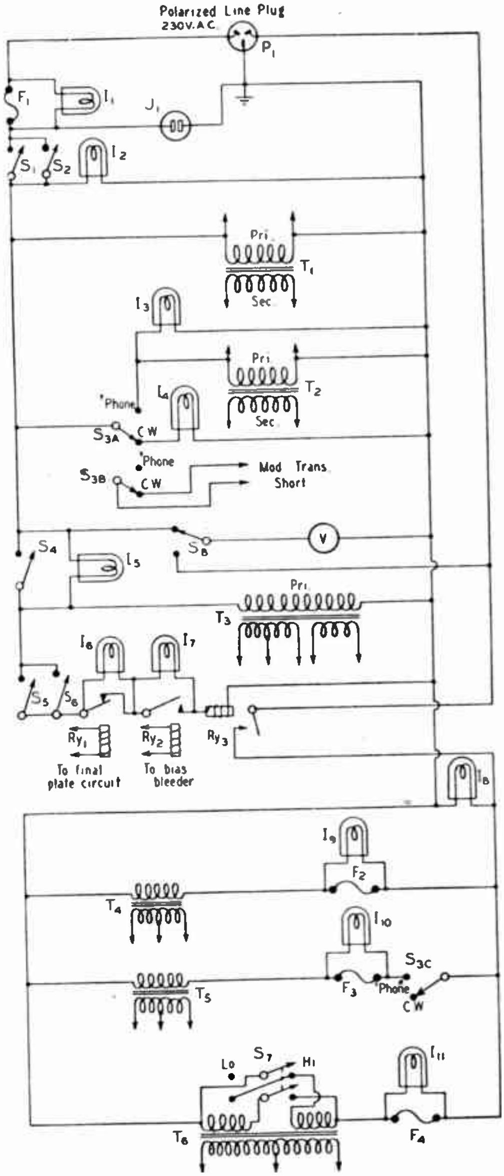


Fig. 7-35 — 230-volt control circuit. Ry_1 is an overload type, Ry_2 is a light-current relay and Ry_3 is a 115-volt a.c. relay with heavy contacts.

Emergency and Independent Power Sources

Emergency power supply which operates independently of a.c. lines is available, or can be built in a number of different forms, depending upon the requirements of the service for which it is intended.

The most practical supply for the average individual amateur is one that operates from a 6-volt car storage battery. Such a supply may take the form of a small motor generator (often called a genemotor), a rotary converter or a vibrator-transformer-rectifier combination.

Dynamotors

A dynamotor differs from a motor generator in that it is a single unit having a double armature winding. One winding serves for the driving motor, while the output voltage is taken from the other. Dynamotors usually are operated from 6-, 12-, 28- or 32-volt storage batteries and deliver from 300 to 1000 volts or more at various current ratings.

Genemotor is a term popularly used when making reference to a 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 600 volts at 300 ma. 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 to make certain that no looseness has developed.

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 bypassed 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 and antenna lead.

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 to the receiver power terminals through a smoothing filter using 4- to 8-μfd. condensers and 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 115-volt 60-cycle a.c. Such converter units are built to deliver outputs ranging from 40 to 300 watts, depending upon the battery power available.

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 (which is supplied directly from the battery in the case of the genemotor) as well as plate power.

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. 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. The smoothing filter can be a single-section affair, but the filter output capacitance should be fairly large — 16 to 32 μfd.

Fig. 7-36 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. 7-36B is

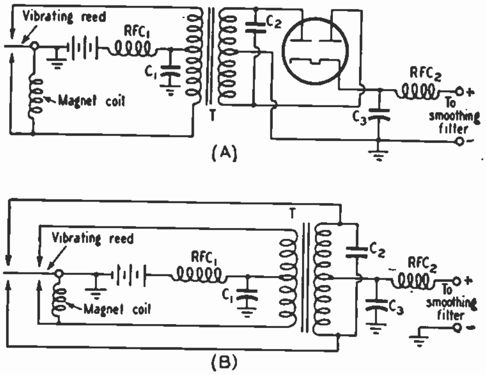


Fig. 7-36 — Basic types of vibrator power-supply circuits. A—Nonsynchronous. B—Synchronous.

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 that occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes very high voltages to be induced in the secondary. 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 capacitance 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 capacitance 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 that 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. 7-37C shows an idealized trace of the optimum waveform when the buffer capacitor 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 condenser 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. 7-37D). Trace E shows the result of insufficient buffering capacitance, while too much buffering capacitance will show a slow build-up in voltage with rounded corners evident in the trace. Figs. 7-37G and H indicate a worn or improperly-adjusted vibrator.

"Hash" Elimination

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.

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.

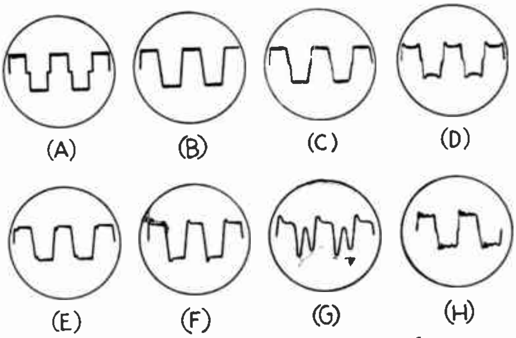


Fig. 7-37 — Characteristic vibrator waveforms as viewed on the oscilloscope. A, ideal theoretical trace for resistive load; current flow stops instantly when vibrator contacts open and resumes approximately 1 microsecond later (for standard 115-cycle vibration frequency) after interrupter arm moves across for the next half-cycle. B, ideal practical waveform for inductive load (transformer primary) with correct buffer capacitance. C, practical approximation of B for loaded nonsynchronous vibrator. D, satisfactory practical trace for synchronous (self-rectifying) vibrator under load; the peaks result from voltage drop in the primary when the secondary load is connected, not from faulty operation.

Faulty operation is indicated in E through H: E, effect of insufficient buffering capacitance (not to be mistaken for "bouncing" of contacts). The opposite condition — excessive buffering capacitance — 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.

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.

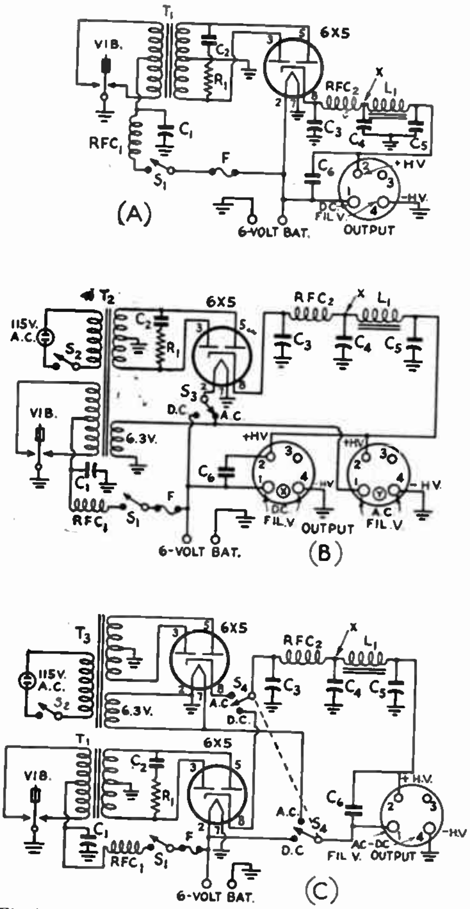


Fig. 7-38 — Typical vibrator-transformer power-supply circuits. The circuit at A shows a simple arrangement for 6-volt d.c. input; the one at B illustrates the use of a combination transformer for operation from either 6 volts d.c. or 115 volts. a.c. The circuit of C is similar to that of B but uses separate transformers.

C_1 — 0.5-μfd. paper, 50-volt rating or higher.
C_2 — 0.005 to 0.01 μfd., 1600 volts.
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-$\mu\mu$fd. mica.
R_1 — 4700 ohms, ½ or 1 watt.
L_1 — 10/12-henry 100-ma. filter choke, not over 100 ohms (Stancor C-2303 or equivalent).
F — 15-ampere fuse.
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 — battery switch.
S_2 — S.p.s.t. toggle — a.c. power switch.
S_3 — S.p.d.t. toggle — rectifier-heater change-over switch.
S_4 — D.p.d.t. toggle — a.c.-d.c. switch.
T_1 — Vibrator transformer.
T_2 — 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_3 — A.c. transformer, 275 to 300 volts each side of center-tap, 100 to 150 ma.; 6.3-volt filament.
VIB — Vibrator unit (Mallory 500P, 294, etc.)
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.

PRACTICAL VIBRATOR-SUPPLY CIRCUITS

A vibrator-type power supply may be designed to operate from a six-volt storage battery only, or in a combination unit which may be operated interchangeably from either battery or 115 volts a.c.

Typical circuits are shown in Fig. 7-38. The one shown at A is the simplest, although it operates from a 6-volt d.c. source only. S_1 turns the high voltage on and off.

The circuit of B provides for either 6-volt d.c. or 115-volt a.c. operation with a dual-primary transformer. S_2 is the a.c. on-off switch while S_3 switches the heater of the 6X5 rectifier from the storage battery to the 6.3-volt winding on the transformer. Filament supply for the transmitter or receiver is switched by shifting the power plug to the correct output socket, *X* when operating from a 6-volt d.c. source, and *Y* when 115-volt a.c. input is used.

The circuit of Fig. 7-38C may be used when a dual-primary transformer is not available. The filter is switched from one rectifier output to the other by means of the d.p.d.t. switch, S_4, which also shifts filament connections from a.c. to d.c. The filter section of the switch could be eliminated if desired by connecting the filtering circuit permanently to the output terminals of both rectifiers and removing the unused rectifier tube from its socket. Similarly, the filament section of S_4 could be dispensed with by providing two output sockets as in the circuit at B. If a separate rectifier filament winding is available on T_3, directly-heated rectifier types may be substituted for the 6X5 in the a.c. supply. In some cases where the required filament windings are not available, a rectifier of the cold-cathode type, such as the 0Z4, which requires no heater voltage, sometimes may be used to advantage.

If suitable filament windings are available, a regular a.c. transformer will make an acceptable substitute for a vibrator transformer. If the a.c. transformer has two 6.3-volt windings, they may be connected in series, their junction forming the required center-tap. A 6.3-volt and a 5-volt winding may be used in a similar manner even though the junction of the two windings does not provide an accurate center-tap. A better center-tap may be obtained if a 2.5-volt winding also is available, since half of this winding may be connected in series with the 5-volt winding to give 6.25 volts.

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-capacitance 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, or 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 capacitance may help in bad cases of hash.

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 capacitance is fairly large — 16 to 32 μfd.

The compactness of selenium rectifiers and the fact that they do not require filament voltage make them particularly suited to compact light-weight power supplies for portable-emergency work.

Fig. 7-39 shows the circuit of a vibrator pack that will deliver an output voltage of 400 at 200 ma. It will work with either 115-volt ac. or 6-volt battery input. The circuit is that of the familiar voltage tripler whose d.c. output voltage is, as a rough approximation, three times the peak voltage delivered by the transformer or line. An interesting feature of the circuit is the fact that the single transformer serves as the vibrator transformer when operating from 6-volt d.c. supply and as the filament transformer when operating from an a.c. line. This is accomplished without complicated switching.

The vibrator transformer, T_1, is a dual-secondary 6.3-volt filament transformer connected in reverse. It may also consist of two single transformers of the same type with their

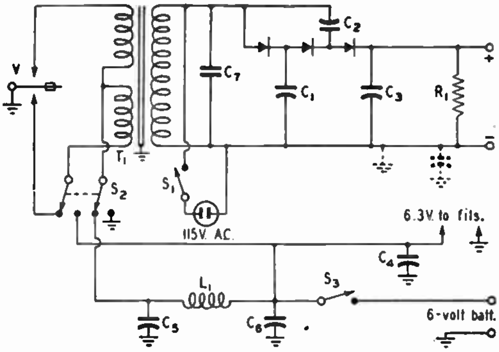


Fig. 7-39 — Circuit diagram of a compact vibrator-a.c. portable power supply suggested by W9CO.

C_1 — 60-μfd. 200-volt electrolytic.
C_2 — 60-μfd. 400-volt electrolytic.
C_3 — 60-μfd. 600-volt electrolytic.
C_4 — 25-μfd. 25-volt electrolytic.
C_5, C_6 — 0.5-μfd. 25-volt paper.
C_7 — 0.007-μfd. 1500-volt paper.
R_1 — 25,000 ohms, 10 watts.
L_1 — 25-μhy. 20-amp. choke.
S_1 — 115-volt toggle switch.
S_2 — D.p.d.t. heavy-duty knife switch.
S_3 — 25-amp. switch.
T_1 — See text.
V — Heavy-duty vibrator.

primaries connected in series and secondaries in parallel, both windings being properly polarized. In either event, the filament windings must have a rating of 10 amperes if the full load current of 200 ma. is to be used. Some excellent surplus transformers that will handle the required current are now available on the surplus market. The vibrator also must be capable of handling the current. The hash-filter choke, L_1, must carry a current of 20 amperes.

The following table shows the output voltage to be expected at various load currents, depending upon the size of condensers used at C_1, C_2 and C_3.

C_1, C_2, C_3 (μfd.)	Output Voltage at 50 ma.	100 ma.	150 ma.	200 ma.
60	455	430	415	395
40	425	390	360	330
20	400	340	285	225

In operating the supply from an a.c. line, it is always wise to determine the plug polarity with respect to ground. Otherwise the rectifier part of the circuit and the transformer circuit cannot be connected to actual ground except through by-pass condensers.

Vibrator-Supply Construction

A typical example of vibrator-supply construction is shown in the photograph of Fig. 7-40.

This model makes use of separate power transformers for 115-volt a.c. and 6-volt d.c. operation, the single rectifier tube being shifted from one octal socket to the other when the change from a.c. to d.c. operation is made. The components are assembled on a 5 × 10 × 3-inch steel chassis. The two transformers are flush-mounting type requiring cut-outs in the chassis. Three socket holes are required — one for the 4-prong socket for the vibrator and two octal sockets for the rectifier. The a.c. line cord and battery and power-output leads are brought out at the rear.

GASOLINE-ENGINE DRIVEN GENERATORS

For higher-power installations, such as for communications control centers during emergencies, the most practical form of independent power supply is the gasoline-engine driven generator which provides standard 115-volt 60-cycle supply.

Such generators are ordinarily rated at a minimum of 250 or 300 watts. They are available up to two kilowatts, or big enough to handle the highest-power amateur rig. Most are arranged to charge automatically an auxiliary 6- or 12-volt battery used in starting. Fitted with self-starters and adequate mufflers and filters, they represent a high order of performance and efficiency. Many of the larger models are liquid-cooled, and they will operate continuously at full load.

A variant on the generator idea is the use of

TABLE 7-III — PLATE-BATTERY SERVICE HOURS

Estimated to 34-volt end-point per nominal 45-volt section.
Based on intermittent use of 3 to 4 hours daily at room temp. of 70° F.
(For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.		Weight		Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.	5	10	15	20	25	30	40	50	60	75	100	150
—	758	14	8	Suggested current range = 7 to 12 Ma.											
21308	—	12	8	1600	1100	690	490	—	300	200	—	130	—	60	30
10308	—	11	4	1300	750	520	350	—	—	130	—	90	—	45	22
—	754	6	8	Suggested current range = 5 to 15 Ma.											
2308	—	8	3	1100	500	330	200	—	150	80	—	43	—	—	—
—	487	4	2	Suggested current range = 7 to 12 Ma.											
B30	—	2	8	350	170	90	50	—	21	17	—	—	—	—	—
A30	—	2	—	260	100	48	28	—	17	7	—	—	—	—	—
—	482	1	14	400	208	122	80	—	—	—	—	—	—	—	—
Z30N	—	1	4	155	70	30	20	15	9.5	—	—	—	—	—	—
—	467	—	12	82	30	—	—	—	—	—	—	—	—	—	—
—	738	1	2	160	70	30	20	10	7	—	—	—	—	—	—
W30FL	—	—	11	70	20	12	7	—	3.5	—	—	—	—	—	—
—	455	—	8	82	30	—	—	—	—	—	—	—	—	—	—
XX30	—	—	9	70	20	12	7	—	3.5	—	—	—	—	—	—

fan-belt drive. The disadvantage of requiring that the automobile must be running throughout the operating period has not led to general popularity of this idea among amateurs. Such generators are similar in construction and capacity to the small gas-driven units.

The output frequency of an engine-driven generator must fall between the relatively narrow limits of 50 to 60 cycles if standard 60-cycle transformers are to operate efficiently from this source. A 60-cycle electric clock provides a means of checking the output frequency with a fair degree of accuracy. The clock is connected across the output of the generator and the second hand is checked closely against the second hand of a watch. The speed of the engine is adjusted until the two second hands are in synchronism. If a 50-cycle clock is used to check a 60-cycle generator, it should be remembered that one revolution of the second hand will be made in 50 seconds and the clock will gain 4.8 hours in each 24 hours.

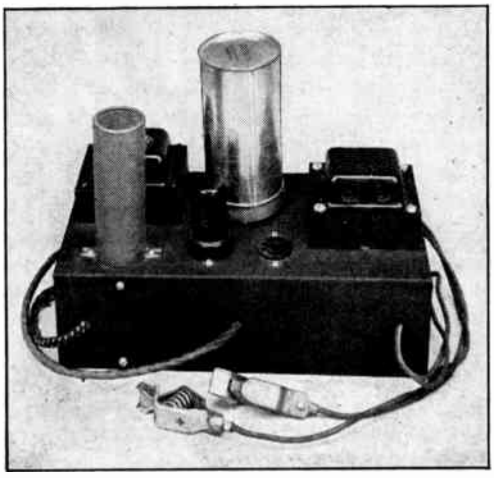


Fig. 7-40 — A typical combination a.c.-d.c. power pack for low-power emergency work. The two transformers are flush-mounted at either end of the chassis. The filter condenser is at the left, the two rectifier sockets at the center and the vibrator to the rear.

Output voltage should be checked with a voltmeter since a standard 115-volt lamp bulb, which is sometimes used for this purpose, is very inaccurate. Tests have shown that what appears to be normal brilliance in the lamp may occur at voltages as high as 150 if the check is made in bright sunlight.

Noise Elimination

Electrical noise which may interfere with receivers operating from engine-driven a.c. generators may be reduced or eliminated by taking proper precautions. The most important point is that of grounding the frame of the generator *and* one side of the output. The ground lead should be short to be effective, otherwise grounding may actually increase the noise. A water pipe may be used if a short connection can be made near the point where the pipe enters the ground, otherwise a good separate ground should be provided.

The next step is to loosen the brush-holder locks and slowly shift the position of the brushes while checking for noise with the receiver. Usually a point will be found (almost always different from the factory setting) where there is a marked decrease in noise.

From this point on, if necessary, by-pass condensers from various brush holders to the frame, as shown in Fig. 7-41, will bring the hash down to within 10 to 15 per cent of its original intensity, if not entirely eliminating it. Most of the remaining noise will be reduced still further if the high-power audio stages are cut out and a pair of headphones is connected into the second detector.

TABLE 7-IV — FILAMENT-BATTERY SERVICE HOURS

Estimated to 1-volt end-point per nominal 1.5-volt unit. Based on intermittent use of 3 to 4 hours per day at room temperature. (For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.		Weight		Voltage	Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.		30	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	4	1.5	—	—	—	—	—	—	—	870	—	—	—	—
—	741[1]	2	13	1.5	—	—	—	—	—	—	—	460	—	—	270	—
—	743	2	1	1.5	—	—	—	—	—	—	—	300	—	225	175	—
—	742	1	6	1.5	—	—	—	—	—	—	—	170	—	120	90	—
8F[2]	—	2	10	1.5	—	—	1100	600	450	—	—	400	—	320	230	190
4F	—	1	4	1.5	—	—	600	340	230	—	—	160	—	110	95	60
—	A-2300	11	—	2.5	—	—	—	—	2000	1715	1500	1333	1250	1200	1000	854
20F2	—	13	12	3.0	—	—	—	—	1100	—	—	850	—	775	600	500
2F2H	—	1	6	3.0	600	—	340	130	95	—	—	60	—	42	30	—
2F2BP[3]	—	1	5	3.0	600	—	340	130	95	—	—	60	—	42	30	—
F2BP	—	—	12	3.0	340	—	130	45	30	—	—	—	—	—	—	—
G3[4]	—	1	5	4.5	370	200	150	50	35	—	—	—	—	—	—	—
—	746	1	4	4.5	—	225	—	—	—	—	—	—	—	—	—	—
—	718[5]	2	13	6.0	—	415	—	—	—	—	—	—	—	—	—	—
F4PI	—	1	6	6.0	340	150	130	45	30	—	—	—	—	—	—	—

[1] Same life figures apply to 745, wt. 2 lb. 13 oz.
[2] Same life figures apply to 8FL, wt. 2 lb. 15 oz.
[3] Same life figures apply to 2F4, volts 6, wt. 2 lb. 11 oz.
[4] Same life figures apply to G5, volts 7½, wt. 2 lb. 2 oz.
[5] Same life figures apply to 747, wt. 2 lb. 13 oz.

If batteries of another make are to be used, locate ones of similar size and weight on these tables and comparable performance may be expected.

● POWER FOR PORTABLES

Dry-cell batteries are the only practical source of supply for equipment which must be transported on foot. From certain considerations they may also be the best source of voltage for a receiver whose filaments may be operated from a storage battery, since no problem of noise filtering is involved.

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.

Tables 7-III and 7-IV give service life of representative types of batteries for 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.

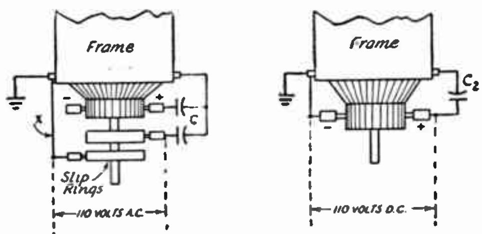


Fig. 7-41 — Connections used for eliminating interference from gas-driven generator plants. C should be 1 μfd., 300 volts, paper, while C_2 may be 1 μfd. with a voltage rating of twice the d.c. output voltage delivered by the generator. X indicates an added connection between the slip ring on the grounded side of the line and the generator frame.

Keying and Break-In

If the proper keying of a transmitter entailed only the ability to turn on and off the output, keying would be a simple matter. Unfortunately, perfect keying is as difficult to obtain as perfect voice quality, and so is not a matter to be dismissed lightly. The keying of a transmitter can be considered satisfactory if the power output is reduced to zero with the key open, or "up," and reaches full output when the key is closed, or "down." The keying system should accomplish this without producing objectionable transients or "clicks," which cause interference with other amateur stations and with local broadcast reception. Furthermore, the keying process should cause no "chirp," which means that the transmitter output frequency should not be affected by the keying process.

Keying Principles and Characteristics

Back-Wave

When the transmitter output is not reduced to zero under key-up conditions, the signal is said to have a **back-wave.** If the amount of back-wave is appreciable, the keying will be difficult to read. A pronounced back-wave may result when the amplifier feeding the antenna is keyed, as a result of the excitation energy feeding through an incompletely-neutralized stage. Magnetic coupling between antenna coils and one of the driver stages on the operating frequency is also a cause of back-wave. Direct radiation from a driver stage ahead of the keyed stage will result in a back-wave, but this type is generally heard only within a few miles of the transmitter, unless the driver stage is fairly high-powered.

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 that completely cut off the plate voltage when the key is open. It will occur in grid-block keying systems if the blocking voltage is not great enough, or in power-supply primary keying systems if only the final-stage power-supply primary is keyed. A vacuum-tube keyer will give a back-wave if the "open" key resistance is too low.

Key Clicks

If a transmitter is keyed in such a manner that the power output rises instantly to its full value or drops immediately to zero, the resultant short rise and decay times produce signals (at the times of closing and opening the key) extending from the signal frequency to several hundred kilocycles on either side. These signals are called "key clicks," and they will cause interference to other amateurs and other services. Consequently, keying systems must be used that increase the rise and decay times of the keyed characters, since this results in less click energy removed from the signal frequency.

The simple process of making and breaking any circuit with current flowing through it will produce a brief burst of r.f. energy. This effect can be noticed in a radio receiver when an electric light or other appliance in the house is turned on or off. It is, therefore, not only necessary to delay the rise and decay times of the keyed transmitter to prevent interference with other services, but it may be necessary to filter the r.f. energy generated at the key contacts if this energy is found to interfere with broadcast reception in the amateur's house or vicinity. This interference is also called "key clicks."

Getting back to the discussion of rise and decay times, tests have shown that practically all operators prefer to copy a signal that 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. On the other hand, 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. If it is too "soft" the dots and dashes will tend to run together and the characters will be difficult to copy. The keying should be adjusted so that for all normal hand speeds (15 to 35 w.p.m.) the readability will be satisfactory without causing unnecessary interference to the reception of other signals near the transmitter frequency.

Chirps

Keying should have no effect upon the frequency of the transmitter. In many cases where sufficient pains have not been taken, keying will cause a frequency change, or "chirp," at the instant of opening or closing the key. The resultant signal is unpleasant and, in cases of extreme chirp, difficult to copy. Multistage transmitters keyed in a stage following the oscillator are generally free from chirp, unless the keying causes line-voltage changes which in turn affect the oscillator frequency. When the oscillator is keyed, as is done for "break-in" operation, particular care must be taken to insure that the signal does not have keying chirps.

Break-In Operation

In code transmission, there are intervals between dots and dashes, and slightly longer intervals between letters and words, when no power is being radiated by the transmitter. If the receiver can be made to operate at normal sensitivity during these intervals, it is possible for the receiving operator to signal the transmitting operator, by holding his key down. This is useful during the handling of messages, since the receiving operator can immediately signal the transmitting operator if he misses part of the message. It is also useful in reducing the time necessary for calling in answer to a "CQ." The ability to hear signals during the short "key-up" intervals is called **break-in operation.**

Selecting the Stage To Key

It is highly advantageous from an operating standpoint to design the c.w. transmitter for break-in operation. In most cases this requires that the oscillator be keyed, since a continuously-running oscillator will create interference in the receiver and prevent break-in on or near one's own frequency. On the other hand, it is easier to avoid a chirpy signal by keying a stage or two following the oscillator. Since the effect of a chirp is multiplied with frequency, it is quite difficult to obtain chirpless oscillator keying in the 14- and 28-Mc. bands. In any case, however, the stages following the keyed stage (or stages) must be provided with sufficient fixed bias to limit the plate currents to safe values when the key is up and the tubes are receiving no excitation voltage. 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 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 following the keyed stage so that they draw a key-up current of about 5 per cent of the normal key-down value.*

The power broken by the key is an important consideration, both from the standpoint of safety to the operator and that of sparking at the key contacts. Keying of the oscillator or a low-power stage is favorable on both counts. The use of a keying relay is recommended when a high-power circuit is keyed.

Keying Circuits

Only general circuits can be shown for keying, since the final decision on where and how to key rests with the amateur and depends upon the power level and type of operation.

● PLATE-CIRCUIT KEYING

Any stage of the transmitter can be keyed by opening and closing the plate-power circuit. Fig. 8-1 shows how the key can be connected to key the plate circuit (A) or the screen circuit (B). The circuit of Fig. 8-1A shows the key in the negative power lead, although it could be placed in the positive lead, at the point marked "X." Either system is recommended only for low-voltage circuits, of the order of 300 or less, unless a relay is used, because of the danger of accidental electrical shock.

Fig. 8-1B shows the key in the screen lead of an electron-coupled oscillator, and can be considered a variation of 8-1A that has the desirable advantage of breaking less current at a lower voltage.

Both of the circuits shown in Fig. 8-1 respond well to the use of key-click filters, and

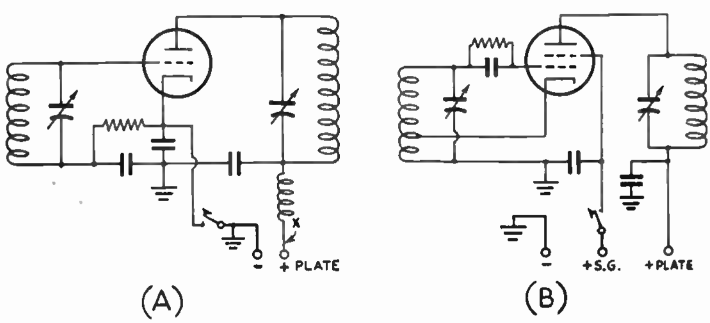


Fig. 8-1 — Plate-circuit keying is shown at A, and screen-grid keying is shown at B. Oscillator circuits are shown in both cases, but the same keying methods can be used with amplifier circuits. Notice the similarity between A and Fig. 8-5 — the only difference is in the way the grid return is connected.

are particularly suitable for use with crystal- and self-controlled oscillators, which are generally operated at low voltage and low power.

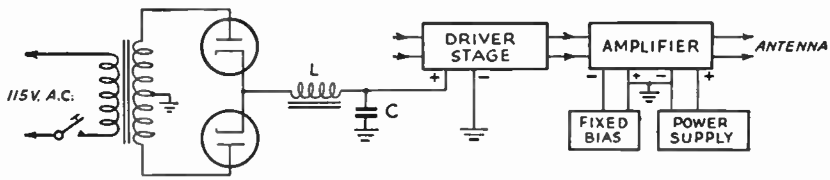
In any transmitter where a driver stage requires the same supply voltage as the screen of the driven stage, the positive lead to the driver stage and to the screen grid of the amplifier can be keyed simultaneously, with excellent results. Usually no fixed bias will be required on the grid of the amplifier, since the key-up plate current will have a low value.

Generally an oscillator will operate at a very low plate voltage, but some refuse to. In the case of the latter, an improvement in keying can sometimes be obtained by using a high value of resistance across the key that will permit the oscillator to draw some plate current (without oscillating). No one value of resistance can be recommended, since every case will be different, but several different values of resistance should be tried, increasing in value until the oscillator stops.

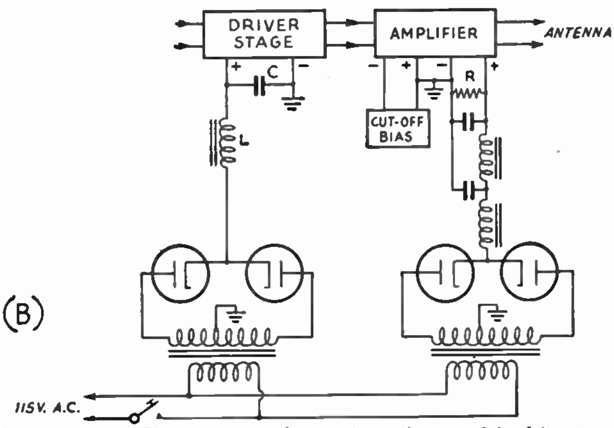
PRIMARY KEYING

A popular method of keying high-powered amplifiers is shown in Fig. 8-2. In its simplest form, as shown in 8-2A, it consists of keying the primary of the plate transformer supplying power to one or more of the driver stages. It has the advantage that the filter, *LC*, acts as a keying filter and prevents clicks. However, too much filter cannot be used or the keying will be too soft, and a single section is all that can normally be used. Since this will introduce some a.c. modulation on the keyed stages, it is essential that the amplifier driven by the keyed stage have sufficient excitation to operate as a Class C amplifier, which tends to eliminate the modulation existing in the excitation voltage. Primary keying of the final plate power supply alone is not recommended, since it is practically impossible to comply with FCC regulations about "adequately-filtered power supply" and still avoid keying that is too soft.

Primary keying of the driver power supply requires that the following amplifier stage (or stages) be biased to prevent excessive current under key-up conditions. If this bias exceeds the cut-off value for the tube (or tubes) a slightly more elaborate version of primary keying can be used, as shown in Fig. 8-2B. The primaries of both driver and final-amplifier plate supplies are keyed, and the system has the advantage that the final-amplifier plate voltage remains substantially constant under key-up or key-down conditions, and thus no clicks can be introduced by the sudden changes in final-amplifier plate voltage as the excitation is applied or removed. The final-



(A)



(B)

Fig. 8-2 — Primary-keying circuits. The circuit at A shows primary keying of the driver-stage (or stages) power supply, followed by an amplifier biased to or close to cut-off. The circuit of B uses primary keying of both driver and final supplies, and has the advantage that the key-up and key-down voltages on the final amplifier remain substantially constant, thus eliminating the chance of clicks being introduced by the final-amplifier plate-supply regulation.

In either case, *L* and *C* should be as small as possible, consistent with sufficient filtering and rectifier-tube limits. *R* in B need be only about 1000 ohms per volt. If a plate voltmeter is used, the bleed through it is sufficient, since the only function is to remove any long-standing charge from the power supply. A heavy bleed current will reduce the effectiveness of the keying system. See text for other bleeder suggestions.

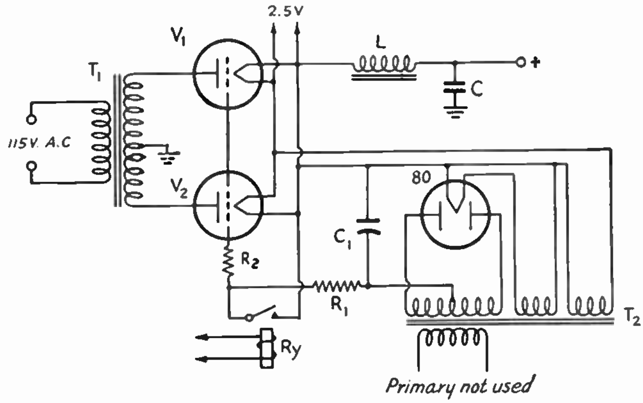


Fig. 8-3 — Grid-controlled rectifier keying. Circuit is similar to Fig. 8-2, and the values of *L* and *C* are the same. A well-insulated keying relay, *Ry*, is used to control the bias on the rectifiers V_1 and V_2. The bias voltage is obtained from a small receiver power-supply transformer T_2, the 80 rectifier, and filter condenser C_1. T_2 does not need to be insulated for the full plate-supply voltage (obtained from T_1) because it is excited from the filament transformer for V_1 and V_2. It should be well insulated to ground, however. R_1 limits the short-circuit on the bias supply and can be approximately 50,000 ohms in value.

amplifier plate supply will remain charged for several minutes after the last transmission, however, and extreme caution must be exercised. As a safety measure, the final-amplifier power supply can be discharged by a relay that shorts the supply through a 1000-ohm resistor, or the bias can be removed and the final-amplifier tube will discharge the power supply.

The keying system shown in Fig. 8-2B has been used to key an entire transmitter for break-in operation. The oscillator and multiplier/driver stages take their plate power from the supply with the small filter, while the final amplifier is powered from the heavily-filtered supply. It is essential, however, in a transmitter keyed for break-in in this manner, that the oscillator be free from chirp, and this point should be checked carefully before using the system on the air.

In using primary keying up to several hundred watts, direct keying in the primary circuit is satisfactory. For higher powers, however, a suitable keying relay should be used, because of the arcing at the contacts.

Fig. 8-3 shows **grid-controlled rectifier tubes** in the power supply. By applying suitable bias to the tubes when the key is up, no current flows through the tubes. When the key is closed, the bias is removed and the tubes conduct. The system can be used in the same way that primary keying was used in Fig. 8-2A and B. This system is used only in high-powered high-voltage supplies.

BLOCKED-GRID KEYING

An amplifier tube can be keyed by applying sufficient negative bias voltage to the control or suppressor grid to cut off plate-current flow when the key is up, and by removing this blocking bias when the key is down. When the bias is applied to the control grid, its value will be considerably higher than the nominal cut-off bias for the tube, since the r.f. excitation voltage must be overcome. The fundamental circuits are shown in Fig. 8-4A and B. The circuits can be applied to oscillator tubes as well as amplifiers. Suppressor-grid

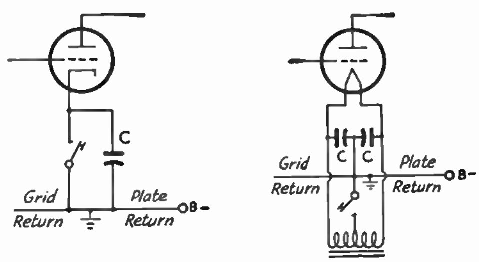


Fig. 8-5 — Cathode and center-tap keying. The condensers *C* are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

keying will not completely turn off a Tri-tet crystal oscillator or electron-coupled self-controlled oscillator, and is likely to cause serious chirps with the latter.

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 that controls the rise time on make, the rise time increasing as $R_2 \times C_1$ is made larger. $C_1 \times (R_1 + R_2)$ controls the decay time on break in the same manner. Since grid current flows through R_2 in Fig. 8-4A when the key is closed, operating bias is developed, and R_2 is the normal grid leak for the tube. Thus C_1 only is varied to obtain the proper rise time.

With blocked-grid keying only a small current is broken compared with other systems, and sparking at the key is slight.

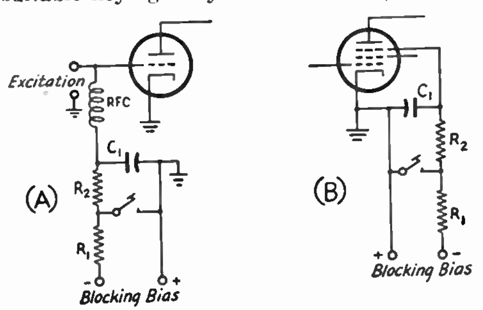


Fig. 8-4 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 is the normal value of grid leak for the tube.

CATHODE KEYING

Keying the cathode circuit of a tube simultaneously opens the grid and plate circuits of the tube. This is shown in Fig. 8-5A. The condenser *C* serves as a short path for the r.f. energy, since the keying leads are often long. When a filament-type tube is keyed in this manner, the key is connected in the filament-transformer center-tap lead, as in Fig. 8-5B, and the system is called center-tap keying. The condensers *C* serve the same purpose as in cathode keying.

Cathode (or center-tap) keying results in less sparking at the key contacts than does plate-supply keying, for the same plate power. When used with an oscillator it does not respond as readily to key-click filtering as does plate-circuit keying, but it is an excellent method for amplifier keying. If plate voltages above 300 are used, it is highly advisable to use a keying relay, to avoid accidental electrical shock at the key.

KEYING RELAYS

A keying relay can be substituted for a key in any of the keying circuits shown in this chapter. Most keying relays operate from 6.3 or 115 volts a.c., and they should be selected for their speed of operation and adequate insulation for the job to be done. Adequate current-handling capability is also a factor. A typical circuit is shown in Fig. 8-6.

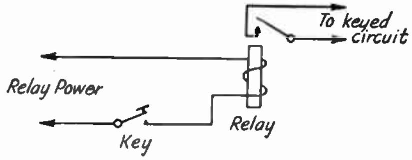


Fig. 8-6 — A keying relay can always be substituted for the key, to provide better isolation from the keyed circuit. An r.f. filter is generally required at the key, and the keying filter is connected in the keyed circuit at the relay contacts.

The relay-coil current that is broken by the key will cause clicks in the receiver, and an r.f. filter (see later in this chapter) is often necessary across the key. The normal keying filter connects at the relay armature contacts in the usual manner. Vibration effects of the keying relay upon the oscillator circuit should be avoided.

Key-Click Reduction

As pointed out earlier, interference caused by the key breaking current and the fast rise and decay times of the keyed characters is called "key clicks." The elimination of the interference depends upon its type.

R.F. FILTERS

Key clicks caused by the spark (often very minute) at the key contacts can be minimized by isolating the key from the rest of the wiring by a small r.f. filter. Such a filter usually consists of an r.f. choke in each key lead, placed right at the key terminals and by-passed on the line side by a small condenser. Such a circuit is shown in Fig. 8-7. Suitable values are best found by experiment, although 2.5-mh. r.f. chokes and a 0.001-μfd. condenser represent good starting points. The chokes must be capable of carrying the current that is broken, and the condenser must have a voltage rating equal at least to the voltage across the key under key-up conditions. Sometimes a small condenser directly across the key terminals is also necessary to remove the last trace of click. This type of r.f. filter is required in nearly every keying installation, in addition to the various circuits to be described in the following few paragraphs.

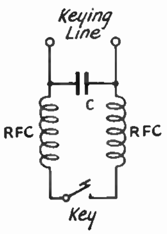


Fig. 8-7 — R.f. filter used for eliminating the radiation effects of sparking at the key contacts. Suitable values for best results with individual transmitters must be determined by experiment. Values for *RFC* range from 2.5 to 80 millihenrys and for *C* from 0.001 to 0.1 μfd.

Keying Filters

A filter used to give a desired shape to the keyed character, to eliminate clicks on the amateur bands and adjacent frequencies, is called a **keying filter** or **lag circuit.** In its simplest form it consists of a condenser and an inductance, connected as in Fig. 8-8. This type of keying filter is suitable for use in the circuits shown in Figs. 8-1 and 8-5. The optimum values of capacitance and inductance must be found by experiment but are not very critical. If a high-voltage low-current circuit is being keyed, a small condenser and a large inductance will be required, while a low-voltage high-current circuit needs a large condenser and small inductance to reduce the clicks properly. 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 set of conditions, increasing

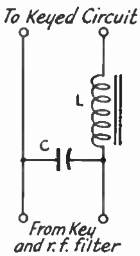


Fig. 8-8 — Lag circuit used for shaping the keying character to eliminate unnecessary sidebands. Actual values for any given circuit must be determined by experiment, and may range from 1 to 30 henrys for *L* and from 0.05 to 0.5 μfd. for *C*, depending on the keyed current and voltage.

the inductance will reduce the clicks on "make" and increasing the capacitance will reduce the clicks on "break."

Primary keying is adjusted by changing the filter values (*L* or *C* in Fig. 8-2). Since it is unlikely that a variety of chokes will be available to the operator, capacitance changes are usually all that can be made. If the keying is found too "soft," the value of *C* must be reduced.

Blocked-grid keying is adjusted by changing the values of resistors and condensers in the circuit, as outlined under the description of the circuit. The values required for individual installations will vary with the amount of blocking voltage and the value of grid leak.

Tube Keying

A tube keyer is a convenient device for keying the transmitter, because it allows the keying characteristic to be adjusted easily and also removes all dangerous voltages from the key itself. The current broken by the key is negligible and usually no r.f. filter is required at the key. A tube keyer uses a tube (or tubes in parallel) to control the current in the plate or cathode circuit of the stage being keyed. The keyer tube turns off the current flow when a high negative voltage is applied to the grid of the keyer tube. The keying characteristic is shaped through the time constants of the grid circuit of the keyer tube, in exactly the same way that it is controlled in blocked-grid keying. When a tube keyer is used to replace the key in a plate or cathode circuit, the power output of the stage may be reduced somewhat because the voltage drop across the keyer lowers the plate voltage or adds cathode bias, but this is of little importance and can be minimized by using more keyer tubes in parallel.

A Vacuum-Tube Keyer

A tube-keyer unit is shown in Figs. 8-9 and 8-10. T_1, the 80 rectifier, and C_1 and R_1 form the power-supply section that furnishes the blocking voltage for the keyer tubes. S_1 and S_2 and their associated resistors and condensers are included to allow the operator to select the keying characteristic he wants. A simplified version could omit the switches and extra components, since once the values have been selected the components can be soldered permanently in place. The rule for adjusting the keying characteristic is the same as for blocked-grid keying. However, large values of resistors and small values of condensers can be used, since there is no value of grid leak determined by the tube that dictates a starting point.

As many 45s may be added in parallel as desired. The voltage drop through a single tube varies from about 90 volts at 50 ma. to 50 volts at 20 ma. Tubes added in parallel will reduce the drop in proportion to the number of tubes used.

When connecting the output terminals of the keyer to the circuit to be keyed, the grounded output terminal of the keyer must be con-

Fig. 8-9 — A vacuum-tube keyer, built up on a 7 × 9 × 2-inch chassis with space for four or less keyer tubes and the power-supply rectifier. The resistors and condensers that produce the lag are underneath, controlled by the knobs at the right. The jack is for the key, while terminals at the left are for the keyed circuit.

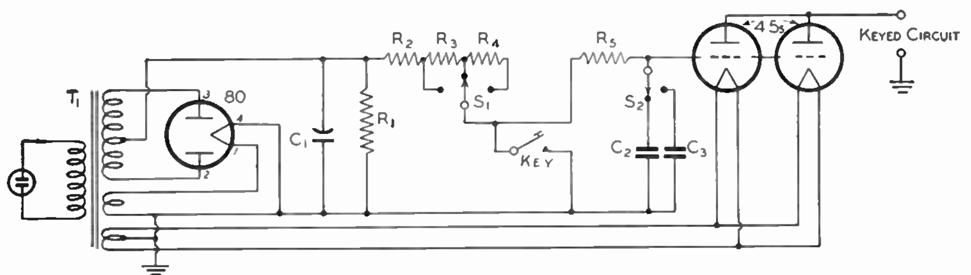


Fig. 8-10 — Wiring diagram of a practical vacuum-tube keyer similar to the one in Fig. 8-9.

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.0033-μfd. mica.
C_3 — 0.0047-μfd. mica.
R_1 — 0.22 megohm, 1 watt.
R_2 — 50,000 ohms, 10 watts.
R_3, R_4 — 4.7 megohms, 1 watt.
R_5 — 0.47 megohm, 1 watt.
S_1, S_2 — 3-position 1-circuit rotary switch.
T_1 — 325–0–325 volts, 5 volts and 2.5 volts (Thordarson T-13R01).

nected to the transmitter ground. Thus the keyer can be used only in negative-lead or cathode keying.

When the key or keying lead has poor insulation, the resistance may become low enough (particularly in humid weather) to reduce the blocking voltage and allow the keyer tube to pass some current. This may cause a slight back-wave, but can be cured by better insulation or reduced values of R_2, R_3, R_4 and R_5.

Checking Transmitter Keying

One of the best ways to check your transmitter keying is to enlist the aid of a near-by amateur and trade stations with him for a short time. Not only will you be able to check your own key clicks and chirps, if they exist, but if you have any complaint about the other fellow's signal this is a convenient way to let him know!

● A SIGNAL MONITOR

Lacking a conveniently-local amateur, your next best bet is to check your signals with a signal monitor. This consists of a completely-shielded battery-operated simple receiver. The complete shielding reduces the signal from the transmitter to the point where it is possible to listen without having the signal "block" the monitor. The monitor can be used to listen to a harmonic of the transmitter, in the case of high-powered transmitters, or the monitor can be used at some distance from the transmitter and remote keying leads run out for the test. A typical signal monitor is shown in Figs. 8-11, 8-12, 8-13 and 8-14.

The monitor is a two-tube regenerative receiver, as can be seen from the circuit diagram, Fig. 8-12. The 1T4 detector has a medium-C tank circuit and a low value of grid leak (R_1) to reduce blocking effects. Capacity control of regeneration is obtained through C_5. The 1S4 audio amplifier gives reasonable headphone volume with the 45-volt plate supply. By-pass condensers across the headphone jack attenuate signals picked up by the 'phone cords. The jack, J_1, is insulated from the panel by fiber washers to avoid shorting the plate-supply battery.

The monitor is built in a 5 × 9 × 6-inch "utility cabinet." The chassis is a small piece of sheet aluminum supported by the two variable condensers. The variable condensers fasten to the front panel (cover) of the cabinet by their single-hole mountings, and the tapped mounting brackets of the condensers then serve as two brackets to support the chassis. The side walls of the cabinet bearing on the chassis contribute to the rigidity of the assembly.

A shield can is mounted over the antenna post, and is only removed when the signal isn't strong enough to be heard otherwise. Normally the shield can will be left in place. The can is one of the small aluminum cans 35-mm. film is packed in. The cover of the can is fastened to the panel with two screws, and a National XS-7 steatite bushing serves as a feed-through and antenna terminal.

The coils are wound on Silver Type 125 coil forms, which are low-loss tube bases. The winding data are given in Fig. 8-12. For any one range, both coils are wound in the same direction, and the grid and plate leads are taken off at the outside ends of the coils. It isn't necessary to wind coils for every amateur band, since one's listening should be done on the band in use or a higher-frequency one. Running one's hands over the 'phone cords or touching the cabinet at any point should result in no change in frequency. If any is noted, it indicates that the shielding and filtering are not adequate, and the grounding of C_6 and C_7 and the bonding of the cabinet should be thoroughly checked.

● SIGNAL CHECKING WITH A RECEIVER

If keying the transmitter does not affect the line voltage, the station communications receiver

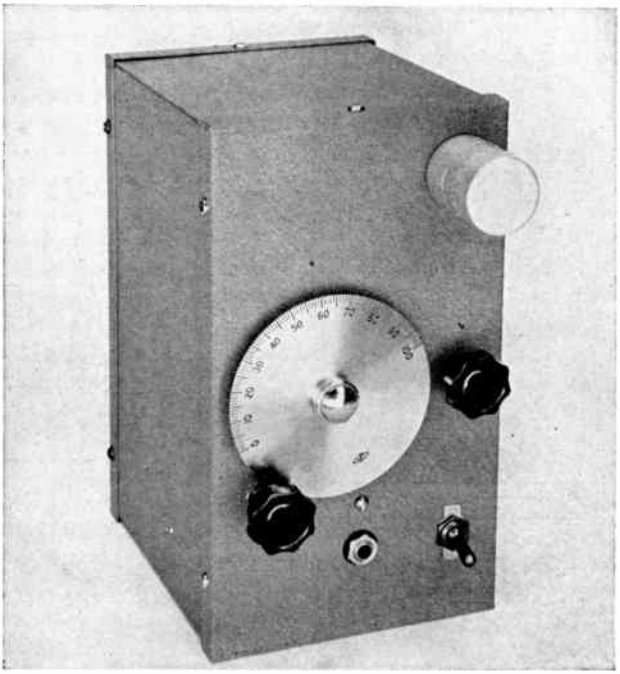


Fig. 8-11 — **A two-tube battery-operated monitor. The extra knob is a regeneration control, and the little aluminum box in the upper corner is a shield can for the antenna terminal.**

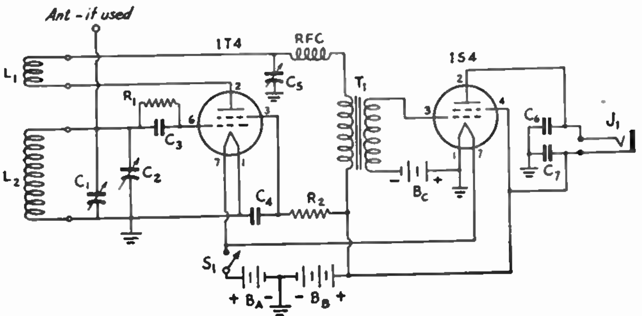


Fig. 8-12 — Wiring diagram of the battery-powered monitor.

C_1 — 10-$\mu\mu$fd. midget variable (Bud LC-1648).
C_2 — 3–30-$\mu\mu$fd. mica trimmer.
C_3 — 47-$\mu\mu$fd. mica.
C_4, C_6, C_7 — 0.001-μfd. mica.
C_5 — 75-$\mu\mu$fd. midget variable (Hammarlund HF-75 or Bud LC-1645).
R_1 — 0.1 megohm, ½ watt.
R_2 — 10,000 ohms, ½ watt.
L_1 — 7 Mc.: 5 turns. 14 Mc.: 3 turns. 28 Mc.: 2 turns.
L_2 — 7 Mc.: 14½ turns. 14 Mc.: 6½ turns. 28 Mc.: 2⅝ turns. L_1 and L_2 are close-wound with No. 24 d.c.c. on 1⅜-inch diameter tube-base forms. Final adjustment of tuning range made by spacing top turns of L_2 and/or setting C_2.
B_A — 1½-volt dry cell (Eveready No. 6 or equiv.).
B_B — 45-volt "B" battery, small size (Burgess XX30).
B_C — 4½ volts. Three "penlite" cells connected in series and wrapped with friction tape.
J_1 — Open-circuit jack.
RFC — 2.5-mh. r.f. choke (National R-100S).
T_1 — Interstage audio transformer, 3:1 ratio (Thordarson T13A34).

can be used to check keying. The antenna should be disconnected from the receiver and the antenna posts shorted to ground. This method is satisfactory only when the line voltage is not affected by keying, since any changes in line voltage will probably affect the receiver frequency. Receivers with good shielding will be more satisfactory than those that allow signals to leak in through the receiver wiring.

Key Clicks

When checking for key clicks, the b.f.o. and a.v.c. of the monitoring receiver should be turned off. If the monitor of Fig. 8-11 is used, the regeneration control should be backed off until the detector is out of oscillation. The keying should be adjusted so that a slight click is heard as the key is closed but practically none heard as the key is opened. When the keying constants have been adjusted to meet this condition, the clicks will be about optimum for all normal amateur work. The receiver gain should be reduced during these tests, since false clicks can be generated in the receiver if the receiver is overloaded.

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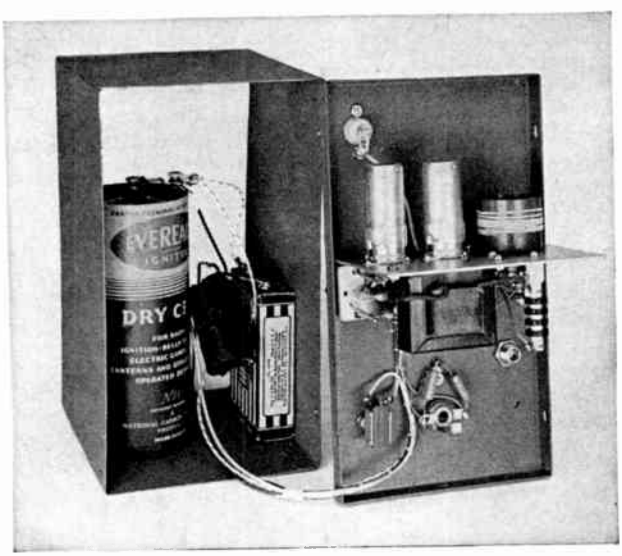


Fig. 8-13 — A view of the monitor disassembled, giving an idea of the arrangement of parts and batteries. The tube shields cover the audio amplifier (left) and the oscillating detector (center).

No clicks should be heard off the signal frequency. Checks should be made with no r.f. power but with the key breaking its normal current, to insure that local clicks are not generated by sparking at the key. To do a good job of checking clicks, those caused by sparking at the key must be eliminated independently of those generated by the keyed r.f. carrier.

Chirps

Keying chirps may be checked by tuning in the signal or one of its harmonics on the highest frequency range of the receiver or monitor and listening to the beat note in the normal manner. The gain should be sufficient to give moderate signal strength, but it should be low enough to avoid overloading. Adjust the tuning to give a low-frequency beat note and key the transmitter at several different speeds. Any chirp introduced by the keying will show up. The signal should be tuned in on either side of zero beat and at various beat frequencies for a complete check. Listening to a harmonic magnifies the effect of any chirp by the order of the harmonic and makes it easier to detect.

Oscillator Keying

The keying of an amplifier is relatively straightforward and requires no special treatment, but considerable care may be necessary with oscillator keying. Any oscillator, either crystal- or self-controlled, should oscillate at low voltages (in the order of two or three volts)

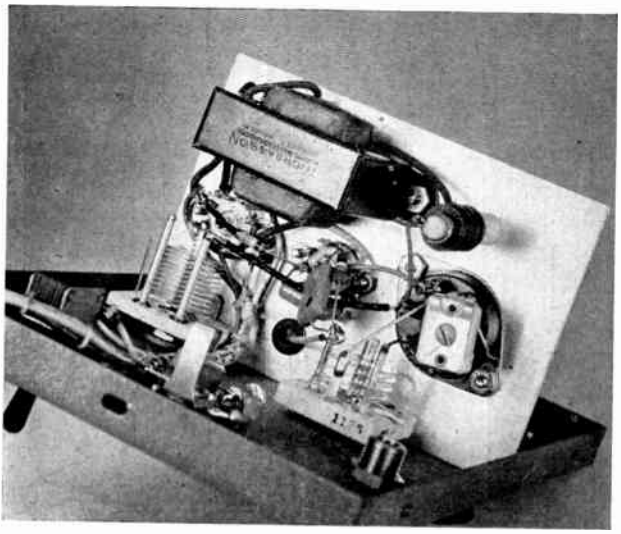


Fig. 8-14 — A close-up view of the underside of the monitor chassis, showing the main tuning condenser (right), the padding condenser mounted on the coil socket, and the feed-back-control condenser.

and have negligible change in frequency with plate voltage, if it is to key without chirps or clicks. A crystal oscillator will oscillate at low voltages if a regenerative type such as the Tri-tet or grid-plate is used and if an r.f. choke is connected in series with the grid-leak resistor, to reduce loading on the crystal. Crystal oscillators of this type are generally free from chirp unless the crystal is a poor one or if there is too much air gap in the crystal holder.

Self-controlled oscillators are more difficult to operate without chirp, but the important requirements are a high C/L ratio in the tank circuit, low plate (and screen) currents, and careful adjustment of the feed-back. A self-controlled oscillator intended to be keyed should be designed for best keying rather than maximum output.

Stages Following Keying

When a keying filter is being adjusted, the stages following the keyed stage should be made inoperative by removing the plate voltage. This allows the keying to be checked without masking by effects caused in the later stages. The following stages should then be connected in, one at a time, and the keying checked after each addition. An increase in click intensity (for the same carrier strength in the receiver) 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 will show up. The output condensers on the filters of the power supplies feeding these later stages can often be increased to good advantage in reducing clicks introduced by these stages.

Low-frequency parasitic oscillations in later stages can cause key clicks removed from the signal frequency by 50 or 100 kc. These clicks are often difficult to track down, but they cause considerable interference and cannot be tolerated. They are most common in beam-tetrode stages, and often can only be eliminated by neutralizing the tetrode stage. Since the parasitic oscillations are of a transient nature, and exist only during the make and break periods of keying, they are much harder to find than parasitics that are not transient.

MONITORING OF KEYING

Most operators find a keying monitor helpful in developing and maintaining a good "fist," especially if a "bug" or semiautomatic key is used. The most popular type of monitor is an audio oscillator that is keyed simultaneously

Fig. 8-15 — A top view of the "Monitone." The shaft of the screwdriver-adjusted potentiometer controlling tone and volume is located between the 6J5 and 6SL7 tubes. The right-hand switch controls the a.c., and the center switch cuts the tone oscillator in and out.

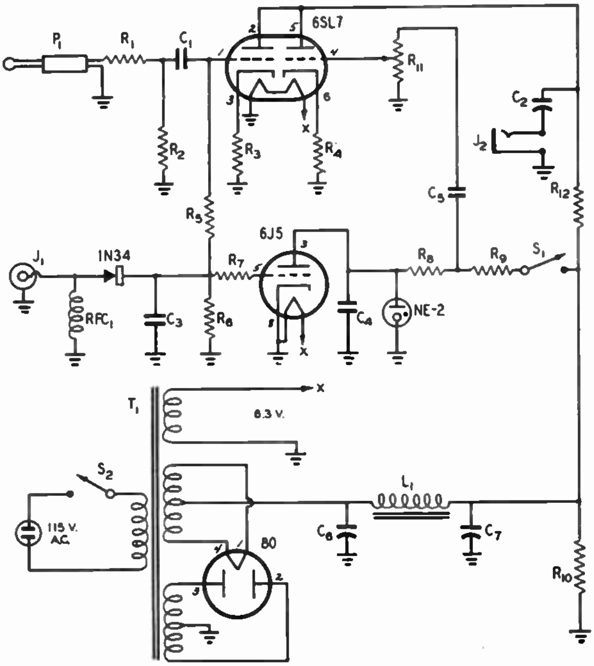


Fig. 8-16 — Circuit diagram of the "Monitone."

C_1 — 0.0047-μfd. 400-volt paper or mica.
C_2 — 0.1-μfd. 600-volt paper.
C_3 — 100-μμfd. mica.
C_4 — 0.001-μfd. paper or mica.
C_5 — 220-μμfd. paper or mica.
C_6, C_7 — 8-μfd. 450-volt electrolytic.
R_1 — 6800 ohms.
R_2 — 1000 ohms.
R_3, R_4 — 1200 ohms.
R_5 — 0.56 megohm.
R_6 — 1 megohm.
R_7 — 68,000 ohms.
R_8 — 4.7 megohms.
R_9 — 2.2 megohms.
R_{10} — 25,000 ohms, 5 watts.
R_{11} — 1-megohm volume control.
R_{12} — 22,000 ohms, 1 watt.
All resistors ½-watt unless otherwise specified.
L_1 — 40-ma. filter choke (Thordarson 13C26).
J_1 — Coaxial-connector jack.
J_2 — Open-circuit jack.
P_1 — 'Phone plug.
RFC_1 — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle.
T_1 — Small b.c. replacement transformer (Stancor P-6297).

with the transmitter. A unit that will key automatically with the transmitter (and also blank the receiver output at the same time) is the "Monitone" shown in Figs. 8-15 and 8-17. It requires no direct connection to the transmitter or key. When the key is up, signals from the receiver pass through the Monitone to the headphones. When the key is down, the receiver output is blanked and a sidetone appears in the headphones. The sidetone and blanking are keyed by the r.f. output of the transmitter, regardless of frequency.

The wiring diagram of the Monitone is given in Fig. 8-16. The 6SL7 acts as a dual amplifier, for the receiver output and for the sidetone oscillator (consisting of the neon bulb — NE-2, C_4 and $R_8 + R_9$). When r.f. from the transmitter is fed in at J_1 it is rectified by the 1N34 and a negative voltage appears at the grid of the 6J5 and Pin 1 of the 6SL7. This negative voltage cuts off the 6J5 and the one half of the 6SL7. The neon-bulb oscillator goes into action and the resultant tone is amplified by the other half of the 6SL7. S_1 is opened for 'phone operation, when the sidetone is not required and only receiver blanking is desirable.

The physical arrangement of parts can be seen in Figs. 8-15 and 8-17. The placement of parts is not critical, although the r.f. network (RFC_1, 1N34 and C_3) should be separated from the 6SL7.

Installation of the device consists of plugging the input lead into the headphone jack on the station receiver, the headphones into the jack on the Monitone, and the power plug into the a.c. outlet.

The final step is to couple the right amount of r.f. into the Monitone. A short piece of wire can be connected to the coaxial fitting on the back of the Monitone if the operating table is near the transmitter. If they are widely separated a piece of RG-59/U or ordinary shielded wire can be run from the coaxial fitting to a point near the final amplifier or feeders. (Caution — high voltage!) The length of the pick-up wire, either directly from the Monitone or extending beyond the shielding of the coaxial

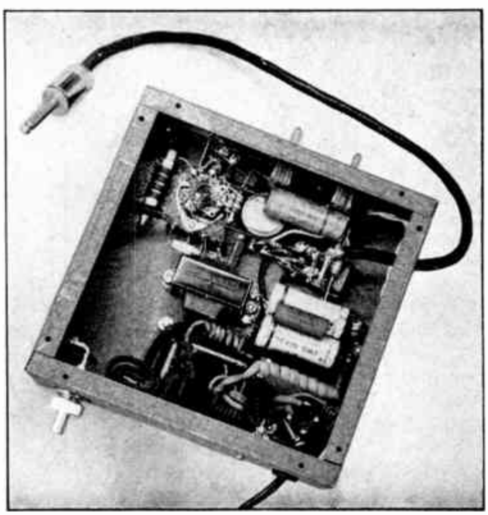


Fig. 8-17 — Bottom view of the "Monitone," showing the small neon lamp supported on its own leads and located directly above the center of the 6J5 socket. The r.f. feeder terminates on the tie-point that carries the 1N34.

line, will depend on the transmitter power being used. Only a foot or two will be needed.

Close the key and move the pick-up nearer or farther from the transmitter or feeder until the neon bulb in the Monitone glows. Find a point where a little less coupling will extinguish the neon — in other words adjust for the loosest coupling that will cause vigorous and sustained oscillation of the neon circuit. If only the final is keyed, care must be taken not to put the pick-up wire in the r.f. field of the driver stages — otherwise the oscillator will run continuously whenever the transmitter is switched on.

Break-In Operation

Break-in operation requires a separate receiving antenna, since none of the available antenna change-over relays is fast enough to follow keying. The receiving antenna should be installed as far as possible from the transmitting antenna. It should be mounted at right angles to the transmitting antenna and fed with low pick-up lead-in material such as coaxial cable or 300-ohm Twin-Lead, to minimize pick-up.

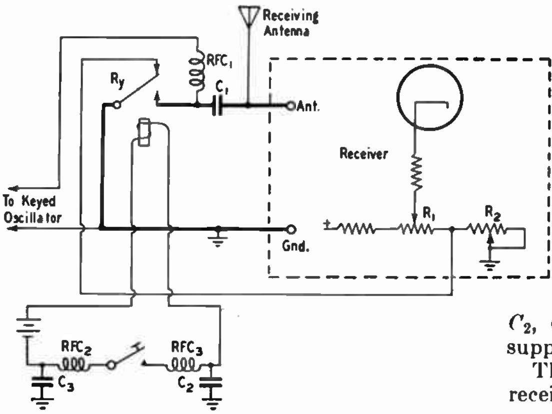


Fig. 8-18 — **Wiring diagram for smooth break-in operation. The leads shown as heavy lines should be kept as short as possible, for minimum pick-up of the transmitter signal.**

C_1, C_2, C_3 — 0.001 μfd.
R_1 — Receiver manual gain control.
R_2 — 5000- or 10,000-ohm wire-wound potentiometer.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke.
Ry — S.p.d.t. keying relay.

If a low-powered transmitter is used, it is often quite satisfactory to use no special equipment for break-in operation other than the separate receiving antenna, since the transmitter will not block the receiver too seriously. Even if the transmitter keys without clicks, some clicks will be heard when the receiver is tuned to the transmitter frequency because of overload in the receiver. An output limiter, as described in Chapter Five, will wash out these clicks and permit good break-in operation even on your transmitter frequency.

When powers above 25 or 50 watts are used, special treatment is required for quiet break-in on the transmitter frequency. A means should be provided for shorting the input of the receiver when the code characters are sent, and a means for reducing the gain of the receiver at the same time is often necessary. The system shown in Fig. 8-18 permits quiet break-in operation for higher-powered stations. It requires a simple operation on the receiver but otherwise is perfectly straightforward. R_1 is the regular receiver r.f. and i.f. gain control. The ground lead is lifted on this control and run to a rheostat, R_2, that goes to ground. A wire from the junction runs outside the receiver to the keying relay, R_y. When the key is up, the ground side of R_1 is connected to ground through the relay arm, and the receiver is in its normal operating condition. When the key is closed, the relay closes, which breaks the ground connection from R_1 and applies additional bias to the tubes in the receiver. This bias is controlled by R_2. When the relay closes, it also closes the circuit to the transmitter oscillator. C_2, C_3, RFC_2 and RFC_3 compose a filter to suppress the clicks caused by the relay current.

The keying relay should be mounted on the receiver as close to the antenna terminals as possible, and the leads shown heavy in the diagram should be kept short, since long leads will allow too much signal to get through into the receiver. A good high-speed keying relay should be used. If a two-wire line is used from the receiving antenna, another r.f. choke, RFC_4, will be required. The revised portion of the schematic is shown in Fig. 8-19.

● A DE LUXE BREAK-IN SYSTEM

In many instances it is quite difficult to key an oscillator without clicks *and* chirps. Most oscillators will key without apparent chirp

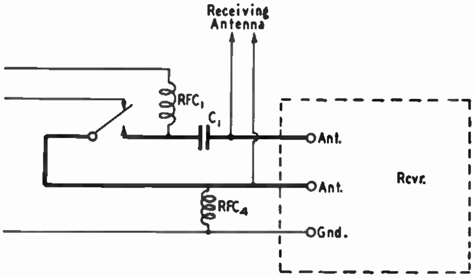


Fig. 8-19 — **Necessary circuit revision of Fig. 8-18 if a two-wire lead from the receiving antenna is used.** ***RFC***$_4$ **is a 2.5-mh. r.f. choke — other values are the same as in Fig. 8-18.**

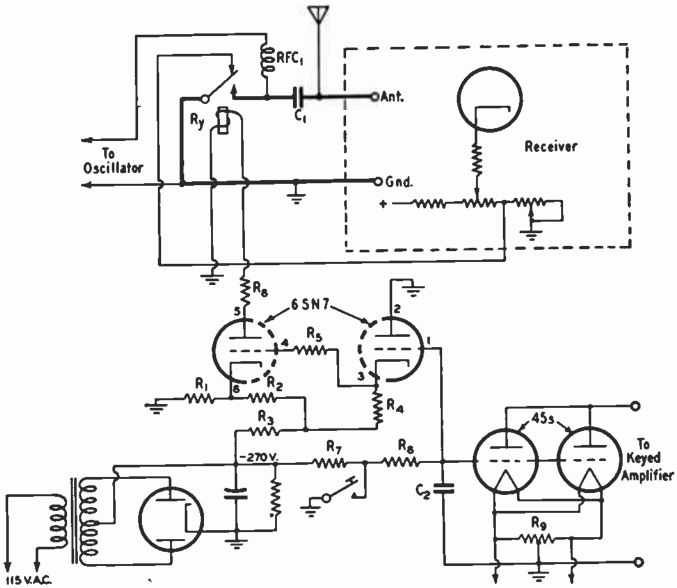


Fig. 8-20 — A de luxe break-in system that holds the oscillator circuit closed (and the receiver input shorted) during a string of fast dots but opens between letters or words.

C_1 — 0.001-μfd. mica.
C_2 — 0.0047-μfd. mica.
R_1 — 20,000 ohms, 10 watts, wire-wound.
R_2 — 1800 ohms.
R_3 — 1500 ohms.
R_4, R_5 — 1.0 megohm.
R_6 — 4700 ohms.
R_7 — 6.8 megohm.
R_8 — 0.47 megohm.
R_9 — 50-ohm center-tapped resistor, 2 watts.
All resistors 1-watt composition unless otherwise noted.
RFC_1 — 2.5-mh. r.f. choke.
Ry — High-speed relay, 1400-ohm 18-volt coil (Stevens-Arnold Type 172 Millisec relay).

if the rise and decay times are made very short, but this introduces key clicks that cannot be avoided. The system shown in Fig. 8-20 avoids this trouble by turning on the oscillator quickly, keying an amplifier with a vacuum-tube keyer, and turning off the oscillator after the amplifier keying is finished. The oscillator is turned on and off without lag, but the resultant clicks are not passed through the transmitter. Actually, with keying speeds faster than about 15 w.p.m., the oscillator will stay turned on for a letter or even a word, but it turns off between words and allows the transmitting station to hear the "break" signal of the other station. It requires one tube more than the ordinary vacuum-tube keyer and a special high-speed relay.

As can be seen from Fig. 8-20, the circuit is a combination of the break-in system of Fig. 8-18 and the tube keyer of Fig. 8-10, with a 6SN7 tube and a few resistors added. Normally the left-hand portion of the 6SN7 is biased to a low value of plate current by the drop through R_2 (part of the bleeder $R_1R_2R_3$) and the relay is open. When the key is closed and C_2 starts to discharge, the right-hand portion of the 6SN7 draws current and this in turn puts a less-negative voltage on the grid of the left-hand portion. The tube draws current and the relay closes. The relay will stay closed until the negative voltage across C_2 is close to the supply voltage, and consequently a string of dots or dashes (which doesn't give C_2 a chance to charge to full negative) will keep the relay closed. In adjusting the system, R_2 controls the amount of idling current through the relay and R_6 determines the voltage across the relay. R_7, R_8 and C_2 are the normal resistors and condenser for the tube keyer. When adjusted properly, the relay will close without delay on the first dot and open quickly during the spaces between words or slower letters. When idling, the voltage across the relay should be one or two volts — with the key down it should be 18 volts.

The oscillator should be designed to key as fast as possible, which means that series resistances and shunt capacitances should be held to a minimum. Negative plate-lead keying is slightly faster than cathode keying and should be used in the oscillator. The keyer tubes are connected in the cathode circuit of an amplifier stage, far enough removed in the circuit to avoid reaction on the oscillator.

● ELECTRONIC KEYS

Electronic keys, as contrasted with mechanical automatic keys, use vacuum tubes (and possibly relays) to form automatic dashes as well as automatic dots. Full descriptions of such devices can be found in the following *QST* articles:

Beecher, "Electronic Keying," April, 1940.
Grammer, "Inexpensive Electronic Key," May, 1940.
Savage, "Improved Switching Arrangement for Simplified Electronic Key," March, 1942.
Gardner, "New Electronic-Key Circuits," March, 1944.
Wiley, "Simplifying the Electronic Key," July, 1944.
"Electronic Bug Movement," Feb., 1945.
Snyder, "Versatile Electronic Key," March, 1945; correction, page 82, May, 1945.
Beecher, "Better Electronic Keyer," August, 1945.
DeHart, "De luxe Electronic Key," Sept., 1946; correction, page 27, Jan., 1947.
Gotisar, "The Dash Master," Aug., 1948.
Bartlett, "Further Advances in Electronic-Keyer Design," October, 1948; correction, page 10, Jan., 1949.

Radiotelephony

To transmit intelligible speech by radio it is necessary to **modulate** the normally-constant output of the radio-frequency section of a transmitter. **Modulation,** defined in the most simple terms, is the process of varying the transmitter output in a desired fashion. In the case of radiotelephony, it means varying the radio-frequency output in a way that follows the spoken word.

The unmodulated r.f. output of the transmitter is called the **carrier.** In itself, the carrier conveys no information to the receiving operator — other than that the transmitting station is "on the air." It is only when the carrier is modulated that it becomes possible to transmit a message.

METHODS OF MODULATION

The carrier as generated by the transmitter is a simple form of alternating current — practically a sine wave. As such, it has three "dimensions" that can be varied — its amplitude, its frequency, and its phase. Modulation can be applied successfully to any of the three.

In **amplitude modulation (AM)** the amplitude of the carrier is made to vary upward and downward, following similar variations in audio-frequency currents generated by a microphone. In this type of modulation the frequency and phase of the carrier are unaffected by the modulation. Amplitude modulation is today the most widely-used system in amateur stations.

In **frequency modulation (FM)** the frequency of the carrier is made to vary above and below the unmodulated carrier frequency, the frequency variations being made to follow the a.f. currents. The power output of the transmitter does not change in frequency modulation. The *phase* of the carrier does change, however, since frequency and phase are intimately related.

In **phase modulation (PM)** the phase of the carrier is advanced and retarded by the modulating audio-frequency current. The transmitter power does not vary with modulation, but the carrier frequency changes.

These definitions are quite broad, and detailed explanations of the three systems are given later in this chapter.

SIDEBANDS

No matter what the method of modulation, the process of modulating a carrier sets up new groups of radio frequencies both above and below the frequency of the carrier itself. These new frequencies that accompany the modulation are called **side frequencies,** and the frequency bands occupied by a group of them when the modulating signal is complex (as it is with voice modulation) are called **sidebands.** Sidebands always appear on *both* sides of the carrier; the band higher than the carrier frequency is called the **upper sideband** and the band lower than the carrier frequency is called the **lower sideband.** The modulation (that is, the intelligence) in the signal is carried in the sidebands, not in the carrier itself.

The result of this is that a modulated signal occupies a group or band of frequencies **(channel)** rather than the single frequency of the carrier alone. Just how much of a frequency band (that is, how wide a channel) is occupied depends upon the method of modulation and the frequency characteristics of the modulating signal itself.

A normal voice contains frequencies or tones ranging from perhaps a hundred cycles at the low end to several thousand cycles at the high end. Vowel sounds (*a, e, i, o, u*) are in general fairly low in frequency and contain most of the voice power. Consonants usually are characterized by higher frequencies, and the hissing sound of the letter "S" is particularly high up in the audio-frequency range. The timbre of a voice, or the thing that makes it possible for us to distinguish the voices of different individuals, results principally from overtones or harmonics. All these things add up to the fact that a fairly wide range of audio frequencies is needed for the accurate reproduction of a *particular* voice.

On the other hand, the frequency range required for good *intelligibility* is not nearly so wide as that needed for accurate reproduction or "fidelity." For the latter, an audio system that is "flat" — that is, has the same amplification at all frequencies — over the range up to about 10,000 cycles is required. But a system that "cuts off" above 2500 cycles — that is, has comparatively little output above that figure — will transmit everything that is necessary for *understandable* speech. The speech may sound a little less like the speaker's actual voice, but it will be thoroughly intelligible to the receiving operator.

This distinction between intelligibility and "quality" is extremely important. The *minimum* channel occupied by a 'phone transmitter, *no matter what the system of modulation,* is equal to *twice the highest audio frequency present in the modulation.* If audio frequencies up to 10,000 cycles are contained in the modulation, the channel will be at least twice 10,000 or 20,000 cycles (20 kc.) wide. But if there are no frequencies above 2500 cycles in the modulation, the channel will be only 5000 cycles (5 kc.) wide. In amateur bands where there is a great deal of congestion it is in everybody's interest that each transmitter should occupy no more than the minimum channel needed for transmitting *intelligible* speech. Taking up a wider frequency channel than that simply creates unnecessary interference.

Also, transmitting a wide range of audio frequencies in a congested band actually accomplishes nothing, insofar as "fidelity" is concerned; the receiving operator has to use so much receiver selectivity — in order to "copy" the signal at all — that the higher-frequency sidebands are rejected by the receiver. Those sidebands do, however, continue to interfere with stations operating on near-by carrier frequencies.

We have said that the *minimum* channel is equal to twice the highest audio frequency in the modulation. The actual channel occupied may be several times the minimum necessary channel-width. This depends on the system of modulation used, for one thing. For another, it depends on whether the system is operated properly or whether it is misadjusted. Improper operation of any sort invariably increases the channel-width. Since the amount of frequency space available for amateur operation is limited, no operator of an amateur 'phone station can avoid the obligation to confine his transmissions to the least possible space.

Amplitude Modulation

In amplitude modulation, as we have already stated, the amplitude or strength of the carrier is varied up and down from the unmodulated value. The several methods of making the carrier strength vary are discussed in a later section; for the moment let us look only at the end result that is the object of all the various amplitude-modulation systems.

In Fig. 9-1, the drawing at A shows the unmodulated r.f. carrier, assumed to be a sine wave of the desired radio frequency. The graph can be taken to represent either voltage or current, and each cycle has just the same height as the preceding and following ones.

In B, the carrier wave is assumed to be modulated by a signal having the shape shown in the small drawing above. The frequency of the modulating signal is much lower than the carrier frequency, so quite a large number of carrier cycles can occur during each cycle of the modulating signal. This is a necessary condition for good modulation, and always is the case in radiotelephony because the audio frequencies used are very low compared with the radio frequency of the carrier. (Actually, there would be very many times more r.f. cycles in each modulation cycle than are shown in the drawing; so many that it is impossible to make the drawing to actual scale.) When the modulating signal is "positive" (above its axis) the carrier amplitude is increased *above* its unmodulated amplitude; when the modulating signal is "negative" the carrier amplitude is *decreased.* Thus the carrier grows larger and smaller with the polarity and amplitude of the modulating signal.

The drawing at C shows what happens with a stronger modulating signal. In this case the strength of the modulation is such that on the "up" modulation the carrier amplitude is doubled at the instant the modulating signal reaches its positive peak. On the negative peak of the modulating signal the carrier amplitude just reaches zero; in other words, the carrier is "all used up."

Percentage of Modulation

When a modulated wave is detected in a receiver the sound that comes out of the loud-

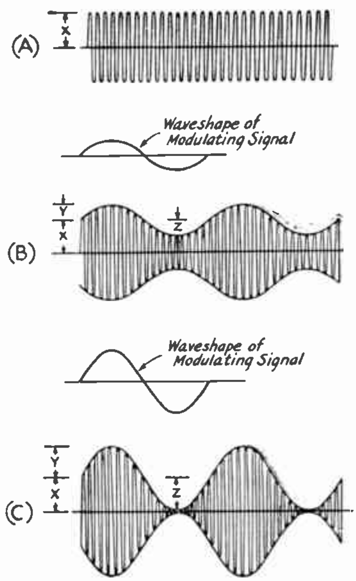


Fig. 9-1 — Graphical representation of (A) carrier unmodulated, (B) modulated 50%, (C) modulated 100%.

speaker or headset is caused by the modulation, not by the carrier. In other words, in detecting the signal the receiver eliminates the carrier and takes from it the modulating signal. The stronger the modulation, therefore, the greater is the useful receiver output. Obviously, it is desirable to make the modulation as strong or "heavy" as possible. A wave modulated as in Fig. 9-1C would produce considerably more useful signal than the one shown at B.

The "depth" of the modulation is expressed as a percentage of the unmodulated carrier amplitude. In either B or C, Fig. 9-1, X represents the unmodulated carrier amplitude, Y is the maximum *increase* in amplitude on the modulation up-peak, and Z is the maximum *decrease* in amplitude on the modulation down-peak. Assuming that Y and Z are equal, then the *percentage of modulation* can be found by dividing either Y or Z by X and multiplying the result by 100. In the wave shown in Fig. 9-1C, Y and Z are both equal to X, so the wave is modulated 100 per cent. In case the modulation is not symmetrical (Y and Z not equal), the larger of the two should be used for calculating the percentage of modulation.

The outline of the modulated wave is called the **modulation envelope.** It is shown by the thin line outlining the patterns in Figs. 9-1 and 9-2.

Power in Modulated Wave

The amplitude values shown in Fig. 9-1 correspond to current or voltage, so the drawings may be taken to represent instantaneous values of either. Now power varies as the square of either the current or voltage (so long as the resistance in the circuit is unchanged), so at the peak of the modulation up-swing the instantaneous power in the wave of Fig. 9-1C is four times the unmodulated carrier power (because the current and voltage are doubled). 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 value of unmodulated carrier power; that is, the power output of the transmitter increases 50 per cent with 100-per-cent modulation.

The complex waveform of speech does not contain as much power as there is in a pure tone or sine wave of the same peak amplitude. 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 regardless of the modulating waveform. Therefore, the peak output-power capacity of the transmitter must be the same for any type of modulating signal.

Overmodulation

If the carrier is modulated more than 100 per cent, a condition such as is shown in Fig. 9-2 occurs. Not only does the peak amplitude exceed twice the carrier amplitude, but there actually may be a considerable period during which the output is entirely cut off. Therefore the modulated wave is distorted, and the modulation contains harmonics of the audio modulating frequency.

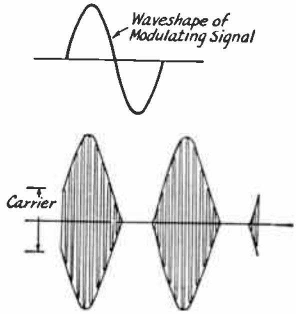


Fig. 9-2 — An overmodulated r.f. carrier wave.

The sharp "break" when the carrier is suddenly cut off on the modulation down-swing produces a type of distortion that contains a large number of harmonics. For example, it is easily possible for harmonics up to the fifth to be produced by a relatively small amount of overmodulation. If the modulating frequency is 2000 cycles, this means that the actual modulated wave will have sidebands not only at 2000 cycles, but also at 4000, 6000, 8000 and 10,000 cycles each side of the carrier frequency. The signal thus occupies five times the needed channel-width. It is obviously of first importance to prevent the modulation from exceeding 100 per cent, and thus prevent the generation of spurious sidebands — commonly called "splatter."

Carrier Requirements

For satisfactory amplitude modulation, the carrier *frequency* should be entirely unaffected by the application of modulation. If modulating the amplitude of the carrier also causes a change in the carrier frequency, the frequency will wobble back and forth with the modulation. This causes distortion and widens the channel taken by the signal. Thus unnecessary interference is caused to other transmissions. In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage that is isolated from the frequency-controlling oscillator by a **buffer amplifier.** Amplitude modulation applied directly to an oscillator always is accompanied by frequency modulation. Under existing regulations amplitude modulation of an oscillator is permitted only on frequencies above 144 Mc. Below that frequency the regulations require that an amplitude-modulated transmitter be completely free from frequency modulation.

Plate Power Supply

The d.c. power supply for the plate or plates of the modulated amplifier must be well filtered; if it is not, the plate-supply ripple will modulate the carrier and cause annoying hum. To be substantially hum-free, the ripple voltage should not be more than about 1 per cent of the d.c. output voltage.

In amplitude modulation the plate current varies at an audio-frequency rate; in other words, an alternating current is superimposed on the d.c. plate current. The output filter condenser in the plate supply must have low reactance, at the lowest audio frequency in the modulation, if the transmitter is to modulate equally well at all audio frequencies. The condenser capacitance required depends on the ratio of d.c. plate current to plate voltage in the modulated amplifier. The requirements will be met satisfactorily if the capacitance of the output condenser is at least equal to

$$C = 25 \frac{I}{E}$$

where C = Capacitance of output condenser in μfd.
I = D.c. plate current of modulated amplifier in milliamperes
E = Plate voltage of modulated amplifier

Example: A modulated amplifier operates at 1250 volts and 275 ma. The capacitance of the output condenser in the plate-supply filter should be at least

$$C = 25 \frac{I}{E} = 25 \times \frac{275}{1250} = 25 \times 0.22 = 5.5 \mu\text{fd.}$$

Linearity

Up to the limit of 100-per-cent modulation, the amplitude of the r.f. output should be directly proportional to the amplitude of the modulating signal. Fig. 9-3 is a graph of an ideal **modulation characteristic,** or curve, showing the relationship between r.f. output amplitude and modulating-signal amplitude. The modulation swings the amplitude back and forth along the curve *A* as the modulating signal alternately swings positive and negative. Assuming that the negative peak of the modulating signal is just sufficient to reduce the carrier amplitude to zero (modulating signal equal to -1 in the drawing), the same modulating signal peak in the *positive* direction ($+1$) should cause the r.f. amplitude to reach twice its unmodulated-carrier value. The ideal modulation characteristic is a straight line, as shown by curve *A*. Such a modulation characteristic is perfectly **linear.**

A **nonlinear** characteristic is shown by curve *B*. The r.f. amplitude does not reach twice the unmodulated carrier amplitude when the modulating signal reaches its positive peak. A modulation characteristic of this type gives a modulation envelope that is "flattened" on the up-peak; in other words, the modulation envelope is not an exact reproduction of the modulating signal. It is therefore distorted and harmonics are generated, causing the transmitted signal to occupy a wider channel than is necessary. A nonlinear modulation characteristic can easily result when a transmitter is not properly designed or is misadjusted.

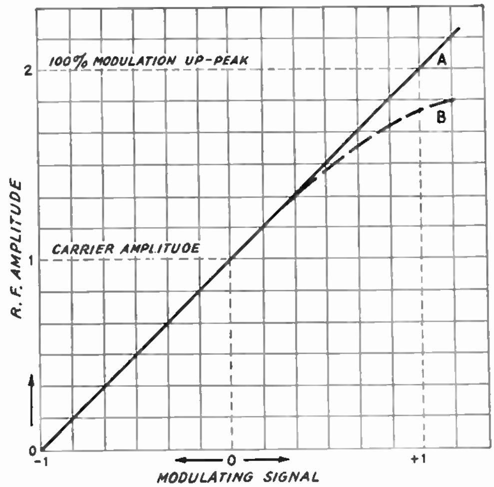


Fig. 9-3 — The modulation characteristic shows the relationship between the instantaneous amplitude of the r.f. output and the instantaneous amplitude of the modulating signal. The ideal characteristic is a straight line, as shown by curve *A*.

The **modulation capability** of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from nonlinearity. The maximum capability is 100 per cent on the down-peak but can be higher on the up-peak. The modulation capability should be as high as possible, so that the most effective signal can be transmitted for a given carrier power.

Types of Amplitude Modulation

The most widely-used 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.**

PLATE MODULATION

The most popular system of amplitude modulation is plate modulation. It is the simplest to apply, gives the highest efficiency in the modulated amplifier, and is the easiest to adjust for proper operation.

Fig. 9-4 shows 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 by the modulator 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.

As stated earlier, the average power output of the modulated stage must increase during modulation. The modulator must be capable of supplying to the modulated r.f. stage sine-wave 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 = D.c. plate voltage
I_p = D.c. plate current (ma.)

E_b and I_p are measured without modulation.

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. This will be the case when the amplifier operates under Class C conditions. The linearity then depends upon having sufficient grid excitation and proper bias, and upon the adjustment of circuit constants to the proper values.

Adjustment of Plate-Modulated Amplifiers

The general operating conditions for Class C operation have been described in Chapter Six. 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 of about 120 degrees at carrier plate voltage, and the grid excitation should be great enough so that the amplifier's plate efficiency will stay constant when the plate voltage is varied over the range from zero to twice the unmodulated value. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value, and then 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 must be transformed to the proper value for the modulator by using the correct output-transformer turns ratio. This point is considered in detail later in this chapter in the section on Class B modulator design.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause nonlinearity. The amplifier also must be completely free from parasitic oscillations.

Although the *effective* value of power input increases with modulation, as described above, the *average* d.c. plate power input to a plate-modulated amplifier does not change. This is because each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current on the next half-cycle of the modulating signal. The d.c. plate current to a properly-modulated amplifier is always constant, with or without modulation. On the other hand, an r.f. ammeter connected in the antenna or transmission line will show an increase in r.f. current with modulation.

Screen-Grid Amplifiers

Screen-grid tubes of the pentode or beam-tetrode type can be used as Class C plate-modulated amplifiers by applying the modulation to both the plate and screen grid. The usual method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 9-5. The dropping resistor, R, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier

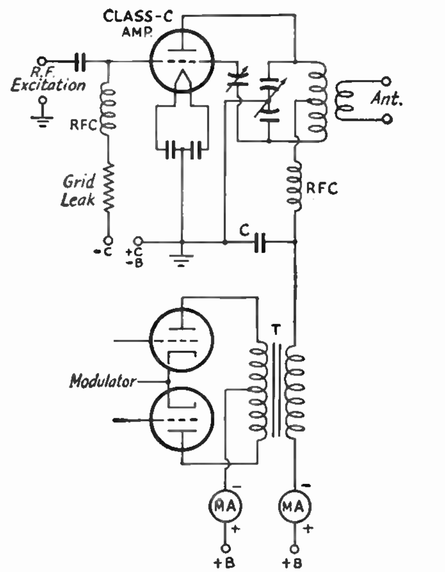


Fig. 9-4 — Plate modulation of a Class C r.f. amplifier. The r.f. plate by-pass condenser, C, in the amplifier stage should have reasonably high reactance at audio frequencies. (See section on Class B modulators.)

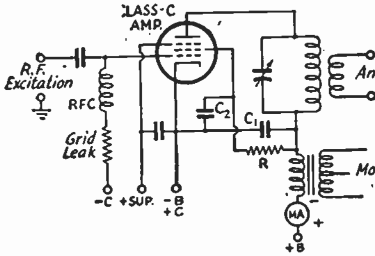


Fig. 9-5 — Plate and screen modulation of a Class C r.f. amplifier using a pentode tube. The plate r.f. by-pass condenser, C_1, should have reasonably high reactance at all audio frequencies. (See section on Class B modulators.) The screen by-pass, C_2, should be 0.002 μfd. or less in the usual case.

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 gives the power input to be used as the basis for determining the audio power required from the modulator.

Modulation of the screen along with the plate is necessary because both elements affect the plate current in a power-type screen-grid tube, and a linear modulation characteristic cannot be obtained by modulating the plate alone. However, at least some types of beam tetrodes (the 4-250A and 4-125A, for example) can be modulated satisfactorily by applying the modulating power to the plate circuit alone, *provided* the screen is "floating" at audio frequencies — that is, is not grounded for a.f. but is connected to its d.c. supply through an audio impedance. The circuit is shown in Fig. 9-6. The choke coil L_1 is the audio impedance in the screen circuit; its inductance should be large enough to have a reactance (at the lowest desired audio frequency) that is not less than the impedance of the screen. The latter can be taken to be approximately equal to the d.c. screen voltage divided by the d.c. screen current.

Choke Coupling

Fig. 9-7 shows the circuit of the choke-coupled system of plate modulation. The d.c. plate power for both the modulator tube and modulated amplifier is furnished from a common source through the modulation choke, *L*. This choke must have high impedance, compared to the modulating impedance of the Class C amplifier, 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 *lower* d.c. plate voltage than the modulator. The extent of the voltage difference is 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.

This type of modulation seldom is used except in very low-power portable sets, because a Class A modulator is required. The output of a Class A modulator is very low compared to the power obtainable from a pair of tubes of the same size operated Class B, so only a small amount of r.f. power can be modulated.

GRID-BIAS MODULATION

Fig. 9-8 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 varies the grid bias, and thereby the power output of the r.f. stage. The r.f. stage is operated as a Class C amplifier.

In this system the plate voltage is constant, and the increase in power output with modulation is obtained by making both 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 up-swing, be twice their carrier values. Thus at the modulation peak the power input is doubled, and since the plate efficiency also is doubled at the same instant the peak output power will be four times the carrier power. The maximum efficiency obtainable in practicable circuits is of the order of 70 to 80 per cent, so the carrier efficiency ordinarily cannot exceed about 35

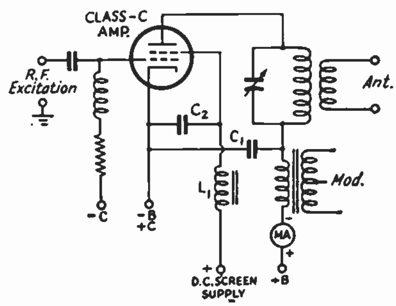


Fig. 9-6 — Plate modulation of a beam tetrode, using an audio impedance in the screen circuit. The value of L_1 is discussed in the text. See Fig. 9-5 for data on by-pass capacitors C_1 and C_2.

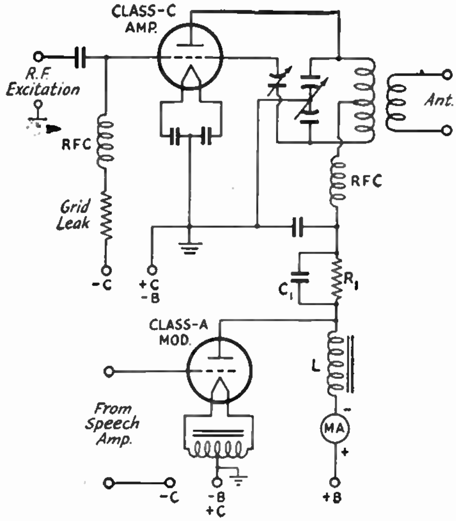


Fig. 9-7 — Choke-coupled plate modulation.

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.

Modulator Power

The increase in average carrier power with modulation is secured by varying the plate efficiency and d.c. plate input of the amplifier, so the modulator need supply only such power losses as may be occasioned by connecting it in the grid circuit. Since these are quite small, a modulator capable of only a few watts output will suffice.

The load on the modulator varies over the a.f. cycle as the rectified grid current of the modulated amplifier changes, so the modulator must have good voltage regulation. The purpose of the resistor *R* across the primary of the modulation transformer in Fig. 9-8 is to "swamp" such load changes by dissipating most of the audio power in the resistor. Generally, this resistor should be approximately equal to the load resistance required by the particular type of modulator tube used. The turns ratio of transformer *T* should be about 1-to-1 in most practical cases.

Grid-Bias Source

The change in instantaneous 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 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 should be chosen (see Chapter Seven). Grid-leak bias for a grid-modulated amplifier is unsatisfactory, and its use should never be attempted.

Driver Regulation

The load on the r.f. driving stage varies with modulation, and a linear modulation characteristic cannot 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 (which is less than the power required for ordinary Class C operation), and dissipating the extra power in a constant load such as a resistor. The variations caused by changes in load with modulation are thereby reduced because the variable load is only a fraction of the total load.

Operating Conditions

The d.c. plate input to the modulated amplifier, assuming a round figure of ⅓ (33 per cent) for the plate efficiency, should not exceed 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage. On the modulation up-peaks the d.c. plate current doubles instantaneously but the d.c. plate voltage does not change. The problem, therefore, is to choose a set of operating conditions that will give normal Class C efficiency when the plate current is *twice* the carrier value.

Example: Two tubes having plate dissipation ratings of 55 watts each are to be used with grid-bias modulation. With plate modulation, the ratings are 1250 volts and 250 ma. for the two tubes, so the plate-voltage/plate-current ratio is

$$\frac{E\ (\text{volts})}{I\ (\text{ma.})} = \frac{1250}{250} = 5$$

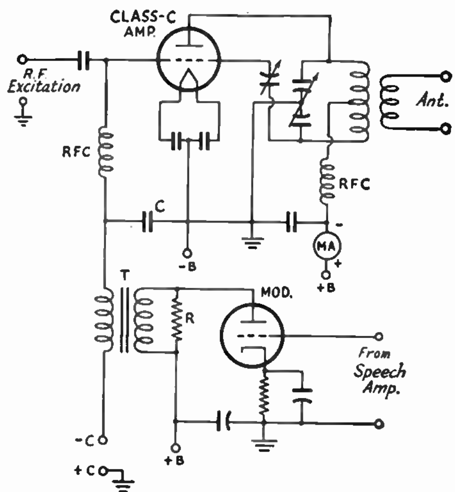


Fig. 9-8 — Grid-bias modulation of a Class C amplifier. The r.f. grid by-pass condenser, C, should have high reactance at audio frequencies (0.005 μfd. or less).

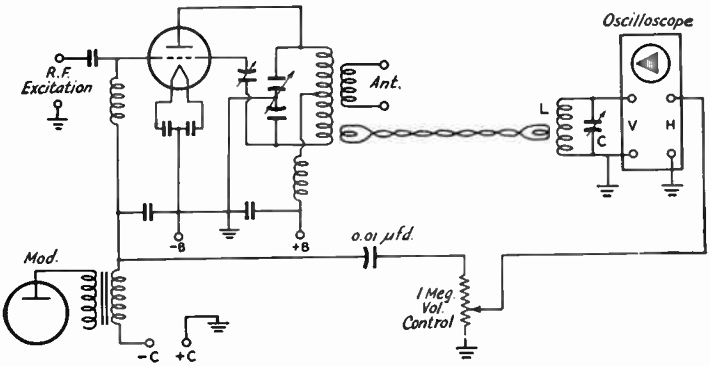


Fig. 9-9 — Adjustment set-up for grid-bias modulation. *L* and *C* should tune to the operating frequency, and may be coupled to the transmitter tank circuit through a twisted pair or other low-impedance line, using single-turn links at each end. The 0.01-µfd. blocking condenser that couples the audio voltage to the horizontal plates of the oscilloscope should have a voltage rating equal to about three times the grid bias.

With grid-bias modulation the maximum power input, at 33% efficiency, is

$P = 1.5 \times (2 \times 55) = 1.5 \times 110 = 165$ watts

The maximum recommended plate voltage for these tubes is 1500 volts. Using this figure, the plate current for the two tubes will be

$$I = \frac{P}{E} = \frac{165}{1500} = 0.11 \text{ amp.} = 110 \text{ ma.}$$

The plate-voltage/plate-current ratio at *twice* carrier plate current is

$$\frac{1500}{220} = 6.8$$

This is quite satisfactory. In this case it would be possible to use a lower plate voltage without having the plate-voltage/plate-current ratio drop to too low a value. At 1300 volts, for example, the ratio would be slightly over 5. However, at 1000 volts it would be only 3.

At 33% efficiency, the carrier output to be expected is 55 watts.

The tank-circuit L/C ratio should be chosen on the basis of *twice* the carrier plate current. In the example above, it would be based on a plate-voltage/plate-current ratio of 6.8. Note that if *carrier* conditions are used the ratio is 13.6, and a tank L/C ratio based on this figure would have a Q much too low for good coupling to the output circuit.

Since the amplifier operates in normal Class C fashion on the modulation up-peaks, the grid bias should be chosen for Class C operation at the plate voltage used. It may be higher if desired, but should never be lower. It is convenient to have an adjustable bias source for arriving at optimum operating conditions.

Adjustment

This type of modulated amplifier should be adjusted with the aid of an oscilloscope. The oscilloscope should be connected as shown in Fig. 9-9. A tone source for modulating the transmitter is a convenience, since a steady tone will give a steady pattern on the oscilloscope. A steady pattern is easier to study than one that flickers with voice modulation.

Having determined the permissible carrier plate current as previously described, apply r.f. excitation and plate voltage 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 (see later section in this chapter for use of the oscilloscope). If curvature occurs well below 100-per-cent modulation, the plate efficiency is too high. Increase the plate loading slightly 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 horizontal axis to twice the carrier amplitude. 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.

Suppressor Modulation

The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 9-10. The operating principles are the same as for grid-bias modulation. However, the r.f. excitation and modulating signals are applied to separate grids; this makes the system somewhat simpler to operate because 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 charac-

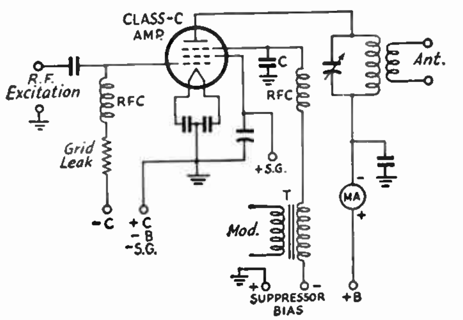


Fig. 9-10 — Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, *C*, should be the same as the grid by-pass condenser in grid-bias modulation (Fig. 9-8).

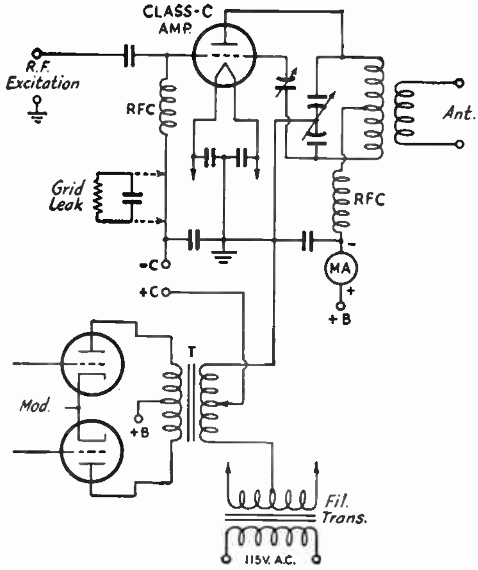


Fig. 9-11 — Circuit arrangement for cathode modulation of a Class C r.f. amplifier.

teristics, linear modulation up to practically 100 per cent can be obtained with negligible distortion.

The method of adjustment of this system is essentially the same as that described in the preceding paragraph.

CATHODE MODULATION

Circuit

The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 9-11. This type of modulation is a combination 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.

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 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. 9-12. 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).

Example: Assume that the r.f. tube to be used has a 100% plate-modulation rating of 250 watts input and will give a carrier power output of 190 watts at that input. Cathode modulation with 40% plate modulation is to be used. From Fig. 9-12, the carrier efficiency will be 56% with 40% plate modulation, the permissible d.c. input will be 65% of the plate-modulation rating, and the r.f. output will be 48% of the plate-modulation rating. That is,

Power input = 250 × 0.65 = 162.5 watts
Power output = 190 × 0.48 = 91.2 watts

The required audio power, from the chart, is equal to 20% of the d.c. input to the modulated amplifier. Therefore

Audio power = 162.5 × 0.2 = 32.5 watts

The modulator should supply a small amount of extra power to take care of losses in the grid circuit. These should not exceed four or five watts.

Modulating Impedance

The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m = Percentage of plate modulation (expressed as a decimal)
E_b = D.c. plate voltage on modulated amplifier
I_b = D.c. plate current of modulated amplifier

Example: Assume that the modulated amplifier in the example above is to operate at a plate

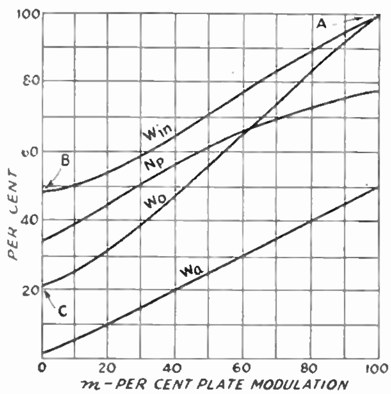


Fig. 9-12 — Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class C telephony tube ratings.

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.

potential of 1250 volts. Then the d.c. plate current is

$$I = \frac{P}{E} = \frac{162.5}{1250} = 0.13 \text{ amp. (130 ma.)}$$

The modulating impedance is

$$m\frac{E_b}{I_b} = 0.4\frac{1250}{0.13} = 3846 \text{ ohms}$$

The modulating impedance is the load into which the modulator must work, just as in the case of pure plate modulation. This load must be matched to the load required by the modulator tubes by proper choice of the turns ratio of the modulation transformer.

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 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.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter. That is, when directly-heated tubes are modulated their filaments must be supplied from a separate transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio-frequency modulation.

Adjustment of Cathode-Modulated Amplifiers

In most respects, the adjustment procedure is similar to that for grid-bias modulation. The critical adjustments are 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 connected in the same way as for grid-bias modulation. 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 as also will too-high excitation. The cathode current will be practically constant with or without modulation when the proper operating conditions have been established.

Speech Equipment

In designing speech equipment it is necessary to "work from both ends." That is, we must know, simultaneously, (1) the amount of audio power the modulation system must furnish and (2) the output voltage developed by the microphone when it is spoken into from normal distance (a few inches) with ordinary loudness. It then becomes possible to choose the number and type of amplifier stages needed to generate the required audio power without overloading or distortion anywhere along the line.

The starting point is the microphone.

● MICROPHONES

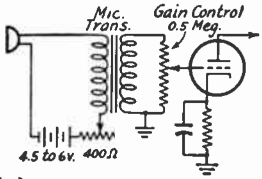
In this age, no one needs an introduction to the microphone. However, there are several different types of them, considerably different in characteristics. Before considering the various types, it is necessary to define a few terms used in connection with microphones.

The level of a microphone is its electrical output for a given sound intensity. 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. It decreases approximately as the square of the distance. Because of these variables, only approximate values based on averages of "normal" speaking voices can be given. The values in the following paragraphs are based on close talking; that is, with the microphone about an inch from the speaker's lips.

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 intelligible speech can be obtained if the output of the microphone does not vary more than a few decibels at any frequency within a range of about 200 to 2500 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.

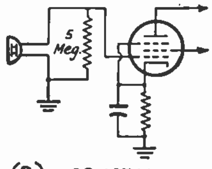
Carbon Microphones

The **carbon 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 be-

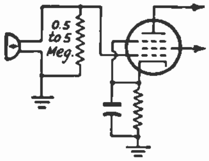


(A) S.B CARBON

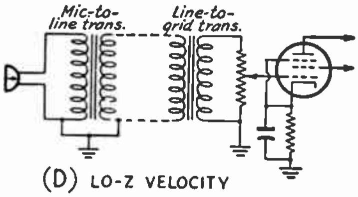
Fig. 9-13 — Speech input circuits used with various types of microphones.



(B) CRYSTAL



(C) HI-Z VELOCITY



(D) LO-Z VELOCITY

ing one connection and the metal backplate the other. Fig. 9-13A shows connections for carbon microphones. A rheostat is included for adjusting the button current to the value as specified with the microphone. 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.

Good-quality 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 can be assumed to be available at the grid of the amplifier tube. The usual button current is 50 to 100 ma.

Crystal Microphones

The **crystal microphone** makes use of the piezoelectric properties of Rochelle salts crystals. This type of microphone requires no battery or transformer and can be connected directly to the grid of an amplifier tube. It is the most popular type of microphone among amateurs, for these reasons as well as the fact that it has good frequency response and is available in inexpensive models.

The "communications-type" crystal microphone uses a diaphragm mechanically coupled to a crystal. This type of construction gives good sensitivity and adequate frequency response for speech. In higher-fidelity types the sound acts directly on a pair of crystals cemented together, with plated electrodes. The level with the latter construction is considerably less. The input circuit for either model of crystal microphone is shown in Fig. 9-13B.

Although the level of crystal microphones varies with different models, an output of 0.03 volt or so 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 are attenuated as the value of load resistance is lowered. A grid-resistor value of at least 1 megohm should be used for reasonably flat response, 5 megohms being a customary figure.

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 air particles set in motion by the sound.

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. 9-13C). 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. 9-13D.

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 secondary of the coupling transformer.

The **dynamic microphone** somewhat resembles a dynamic loudspeaker. 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 the generated voltage is proportional to the frequency of the sound waves and the amplitude is 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.

THE SPEECH AMPLIFIER

In common terminology, the audio-frequency amplifier stage that actually causes the r.f. carrier output to be varied is called the **modulator,** and all the amplifier stages preceding it comprise the **speech amplifier.** Depending on what sort of modulator is used, the speech amplifier may be called upon to deliver a power output ranging from practically zero (only voltage required) to 20 or 30 watts.

Before starting the design of a speech amplifier, therefore, it is necessary to have selected a suitable modulator for the transmitter. This selection must be based on the power required to modulate the transmitter 100 per cent, and this power in turn depends on the type of modulation system selected, as already described. With the modulator picked out, its **driving-power** requirements (audio power required to excite the modulator to full output) can be determined from the tube tables in Chapter Twenty-Five. Generally speaking, it is advisable to choose a tube or tubes for the last stage of the speech amplifier that will be capable of developing at least 50 per cent more power than the rated driving power of the modulator. This will provide a factor of safety so that losses in coupling transformers, etc., will not upset the calculations. A "skimpy" driver, or one designed without a safety factor, usually cannot excite the modulator to full output without being itself overloaded. The inevitable result is speech distortion, generation of unnecessary sidebands, and a "broad" transmitter.

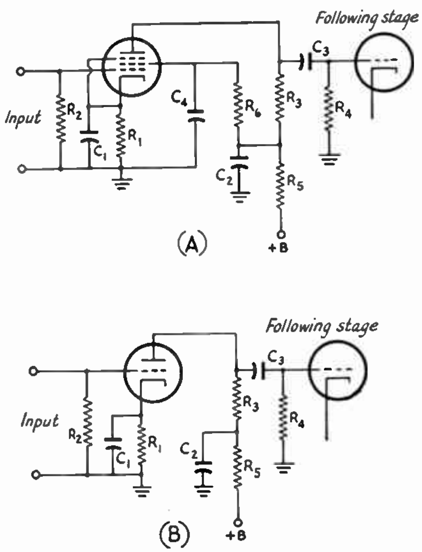


Fig. 9-14 — Resistance-coupled voltage-amplifier circuits. A, pentode; B, triode. Designations are as follows:

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 Table 9-1. Values in the decoupling circuit, C_2R_5, are not critical. R_5 may be about 10% of R_3; an 8- or 10-μfd. electrolytic condenser is usually large enough at C_2.

Voltage Amplifiers

If the last stage in the speech amplifier is a Class AB_2 or Class B amplifier, the stage ahead of it must be capable of sufficient power output to drive it. However, if the last stage is a Class AB_1 or Class A amplifier the preceding stage can be simply a voltage amplifier.

From there on back to the microphone, all stages are voltage amplifiers. These are always operated Class A, not only to simplify the design by avoiding driving power, but because just as much *voltage* can be secured from a Class A amplifier as from any other type.

The important characteristics of a voltage amplifier are its **voltage gain,** maximum undistorted **output voltage,** and its **frequency response.** The voltage gain is the voltage-amplification ratio of the stage. The output voltage is the maximum a.f. voltage that can be secured from the stage without distortion; we cannot figure on any greater output voltage than this, no matter what the gain of the stage, without running into the overload region. The amplifier frequency response should be adequate for voice reproduction; this requirement is easily satisfied.

The voltage gain and maximum undistorted output voltage depend on the operating conditions of the amplifier. Data on the popular types of tubes used in speech amplifiers are given in Table 9-I, for resistance-coupled amplification. The output voltage is in terms of *peak* voltage rather than r.m.s.; this method of rating is preferable because it makes the rating independent of the waveform. Exceeding the peak value causes the amplifier to distort, so it is more useful to consider only peak values in working with amplifiers.

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, because with transformers a sufficiently high load impedance cannot be obtained without considerable frequency distortion. Typical circuits are given in Fig. 9-14 and design data in Table 9-I.

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. With transformer coupling, tubes should be operated under the Class A conditions given in the tube tables in Chapter Twenty-Five.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 9-15. The circuit at A combines 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; 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.

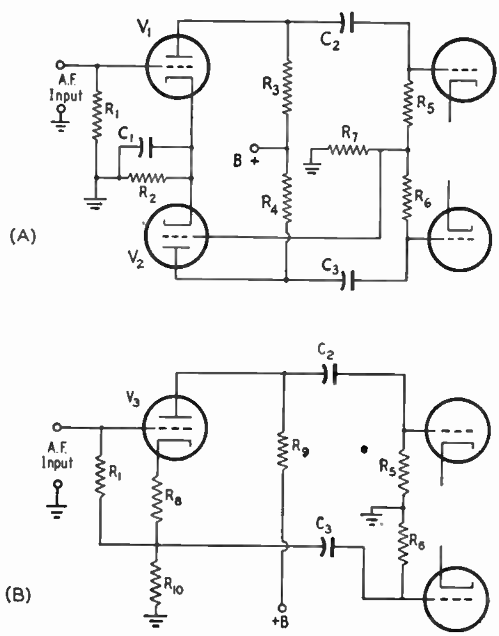


Fig. 9-16 — Self-balancing phase-inverter circuits. V_1 and V_2 may be a double triode such as the 6SN7GT or 6SL7GT. V_3 may be any of the triodes listed in Table 9-I, or one section of a double triode.

R_1 — Grid resistor (1 megohm or less).
R_2 — Cathode resistor; use one-half value given in Table 9-I for tube and operating conditions chosen.
R_3, R_4 — Plate resistor; select from Table 9-I.
R_5, R_6 — Following-stage grid resistor (0.22 to 0.47 megohm).
R_7 — 0.22 megohm.
R_8 — Cathode resistor; select from Table 9-I.
R_9, R_{10} — Each one-half of plate load resistor given in Table 9-I.
C_1 — 10-μfd. electrolytic.
C_2, C_3 — 0.01- to 0.1-μfd. paper.

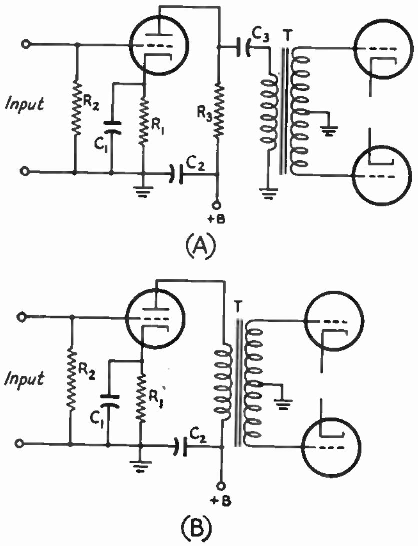


Fig. 9-15 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B for transformer coupling. Designations correspond to those in Fig. 9-14. In A, values can be taken from Table 9-I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given in the tube tables for the particular type of tube used.

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) to a following Class AB_2 or Class B stage.

Phase Inversion

Push-pull output may be secured with resistance coupling by using "phase-inverter" circuits as shown in Fig. 9-16.

The circuit shown in Fig. 9-16A is known as the "self-balancing" type. The amplified voltage from V_1 appears across R_5 and R_7 in series. The drop across R_7 is applied to the grid of V_2, and the amplified voltage from V_2 appears across R_6 and R_7 in series. This voltage is 180 degrees out of phase with the voltage from V_1,

TABLE 9-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.

	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)[1]	Voltage Gain[2]
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
6J7, 7C7	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	200
		2.0	2.95	2300	0.04	4.0	0.0025	100	230
6AU6, 6SH7	0.1	0.1	0.2	500	0.13	18.0	0.019	76	109
		0.22	0.24	600	0.11	16.4	0.011	103	145
		0.47	0.26	700	0.11	15.3	0.006	129	168
	0.22	0.22	0.42	1000	0.1	12.4	0.009	92	164
		0.47	0.5	1000	0.098	12.0	0.007	108	230
		1.0	0.55	1100	0.09	11.0	0.003	122	262
	0.47	0.47	1.0	1800	0.075	8.0	0.0045	94	248
		1.0	1.1	1900	0.065	7.6	0.0028	105	318
		2.2	1.2	2100	0.06	7.3	0.0018	122	371
6AQ6, 6AT6, 6Q7, 6SL7GT (one triode)	0.1	0.1	—	1500	—	4.4	0.027	40	34
		0.22	—	1800	—	3.6	0.014	54	38
		0.47	—	2100	—	3.0	0.0065	63	41
	0.22	0.22	—	2600	—	2.5	0.013	51	42
		0.47	—	3200	—	1.9	0.0065	65	46
		1.0	—	3700	—	1.6	0.0035	77	48
	0 47	0.47	—	5200	—	1.2	0.006	61	48
		1.0	—	6300	—	1.0	0.0035	74	50
		2 2	—	7200	—	0.9	0.002	85	51
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
6S37[3] (one triode)	0 1	0 1	—	750	—	—	0.033	35	29
		0.25	—	930	—	—	0.014	50	34
		0 5	—	1040	—	—	0.007	54	36
	0 25	0.25	—	1400	—	—	0.012	45	39
		0.5	—	1680	—	—	0.006	55	42
		1.0	—	1840	—	—	0.003	64	45
	0.5	0.5	—	2330	—	—	0.006	50	45
		1.0	—	2980	—	—	0.003	62	48
		2.0	—	3280	—	—	0.002	72	49
6J5, 7A4, 7N7, 6S7GT (one triode)	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
6C4	0 047	0 047	—	870	—	4.1	0.065	38	12
		0 1	—	1200	—	3.0	0.034	52	12
		0 22	—	1500	—	2.4	0.016	68	12
	0 1	0 1	—	1900	—	1.9	0.032	44	12
		0.22	—	3000	—	1.3	0.016	68	12
		0.47	—	4000	—	1.1	0.007	80	12
	0 22	0.22	—	5300	—	0.9	0.015	57	12
		0.47	—	800	—	0.52	0.007	82	12
		1.0	—	11000	—	0.46	0.0035	92	12

[1] Voltage across next-stage grid resistor at grid-current point.
[2] At 5 volts r.m.s. output.
[3] Cathode-resistor values are for phase-inverter service.

thus giving push-pull output. The part that appears across R_7 therefore opposes the voltage from V_1 across R_7, thus reducing the signal applied to the grid of V_2. The negative feed-back so obtained tends to regulate automatically the voltage applied to the phase-inverter tube so that the output voltages from both tubes are substantially equal — as they must be for distortionless reproduction. The self-balancing circuit also has the advantage of compensating for variations in the characteristics of the two tubes. The gain is slightly less than twice the gain of a single-tube amplifier using the same operating conditions.

The single-tube circuit shown in Fig. 9-16B also is inherently balanced. In this case the plate load resistor is divided into two equal parts, R_9 and R_{10}, one being connected to the plate in the normal way and the other between cathode and ground. Since the voltages at the plate and cathode are 180 degrees out of phase, the grids of the following tubes are fed equal a.f. voltages in push-pull. The grid return of V_3 is made to the junction of R_8 and R_{10} so normal bias will be applied to the grid. This circuit is highly degenerative because of the way R_{10} is connected. The voltage gain is less than 2 even when a high-μ triode is used at V_3.

Gain Control

A means for varying the over-all gain of the amplifier is a practical necessity. Without it, there would be no way to keep the final output down to the proper level for modulating the transmitter except to talk at just the right intensity. The common method of gain control is to adjust the value of a.c. voltage applied to the grid of one of the amplifiers by means of a voltage divider or potentiometer.

The gain-control potentiometer should be near the input end of the amplifier, at a point where the a.c. voltage level is so low that there is no danger of overloading in the stages ahead of the gain control. With carbon microphones the gain control may be placed directly across the microphone-transformer secondary. With other types of microphones, however, the gain control usually will affect the frequency response of the microphone when connected directly across it. The control therefore is usually placed in the grid circuit of the second stage.

DESIGNING THE SPEECH AMPLIFIER

The steps in designing a speech amplifier are as follows:

1) Determine the power needed to modulate the transmitter and select the modulator. In the case of plate modulation, this will nearly always be a Class B amplifier. Select a suitable tube type and determine from the tube tables in Chapter Twenty-Five the driving power required.

2) As a safety factor, multiply the required driver power by at least 1.5.

3) Select a tube, or pair of tubes, that will deliver the power determined in the second step. This is the last speech-amplifier stage. Receiver-type power tubes can be used (beam tubes such as the 6L6 may be needed in some cases) so the receiving-tube tables in Chapter Twenty-Five may be consulted. If the speech amplifier is to drive a Class B modulator, use a Class A or AB_1 amplifier if it will give enough power output.

4) If the last speech-amplifier stage *has* to operate Class AB_2, use a medium-μ triode (such as the 6J5 or corresponding types) to drive it. In the extreme case of driving 6L6s to maximum output, two triodes should be used in push-pull in the driver. In either case transformer coupling will have to be used, and transformer manufacturers' catalogs should be consulted for a suitable type.

5) If the last speech-amplifier stage operates Class A or AB_1, it may be driven by a voltage amplifier. If the last stage is push-pull, the driver may be a single tube coupled through a transformer with a balanced secondary, or may be a dual-triode phase inverter. Determine the signal voltage required for full output from the last stage. If the last stage is a single-tube Class A amplifier, the peak signal is equal to the grid-bias voltage; if push-pull Class A, the peak signal voltage is equal to twice the grid bias; if Class AB_1, twice the bias voltage when fixed bias is used; if cathode bias is used, twice the bias figured from the cathode resistance and the no-signal plate current.

6) From Table 9-I, select a tube capable of giving the required output voltage and note its rated voltage gain. A double-triode phase inverter (Fig. 9-16A) will have approximately twice the output voltage and twice the gain of one triode operating as an ordinary amplifier. If the driver is to be transformer-coupled to the last stage, select a medium-μ triode and calculate the gain and output voltage as previously described.

7) Divide the voltage required to drive the last stage by the gain of the preceding stage. This gives the peak voltage required at the grid of the next-to-the-last stage.

8) Find the output voltage, under ordinary conditions, of the microphone to be used. This information should be obtained from the manufacturer's catalog. If not available, the figures given in the section on microphones in this chapter will serve.

9) Divide the voltage found in (7) by the output voltage of the microphone. The result is the over-all gain required from the microphone to the grid of the next-to-the-last stage. To be on the safe side, double or triple this figure.

10) From Table 9-I, select a combination of tubes whose gains, when multiplied together, give approximately the figure arrived at in (9). These amplifiers will be used in cascade. In

general, if high gain is required it is advisable to use a pentode for the first speech-amplifier stage, but it is *not* advisable to use a second pentode because of the possibility of feed-back and self-oscillation. In most cases a triode will give enough gain, as a second stage, to make up the total gain required. If not, a third stage, also a triode, may be used.

SPEECH-AMPLIFIER CONSTRUCTION

Once a suitable circuit has been selected for a speech amplifier, the construction problem resolves itself into avoiding two difficulties — excessive hum, and unwanted feed-back. For reasonably humless operation, the hum voltage should not exceed about 1 per cent of the maximum audio output voltage — that is, the hum should be about 40 db. below the output level. Unwanted feed-back, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feed-back can be minimized by isolating each stage with "decoupling" resistors and condensers, by avoiding layouts that bring the first and last stages near each other, and by shielding of "hot" points in the circuit, such as grid leads in low-level stages.

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 that 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 any audio transformers that operate at fairly-high power levels; this will minimize magnetic coupling to the grid circuit and thus reduce hum or audio-frequency feed-back. It is always a safe plan, although not an absolutely necessary one, to build the speech amplifier and its power supply as separate units.

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 usually 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. If the center-tap is grounded, the heater leads to each tube should be twisted together to reduce the magnetic field from the heater current. With either type of connection, it is advisable to lay heater leads in the corner formed by a fold in the chassis, bringing them out from the corner to the tube socket by the shortest possible path.

In a high-gain amplifier it is sometimes helpful if the first tube has its grid connection brought out to a top cap rather than to a base pin; in the latter type the grid lead is exposed to the heater leads inside the tube and hence may pick up more hum. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

When metal tubes are used, always ground the shell connection to the chassis. Glass tubes used in the low-level stages of high-gain amplifiers must be shielded; tube shields are obtainable for that purpose. It is a good plan to enclose the entire amplifier in a metal box, or at least provide it with a cane-metal cover, to avoid feed-back difficulties caused by the r.f. field of the transmitter; r.f. picked up on exposed wiring leads or tube elements causes overloading, distortion, and frequently oscillation.

When using paper condensers as by-passes, be sure that the terminal marked "outside foil" is connected to ground. This utilizes the outside foil of the condenser as a shield around the "hot" foil. When paper condensers are used as coupling condensers between stages, always connect the outside-foil terminal to the side of the circuit having the lowest impedance to ground. Usually, this will be the plate side rather than the following-grid side.

INCREASING THE EFFECTIVENESS OF THE 'PHONE TRANSMITTER

The design principles outlined so far in this section are perfectly straightforward and apply to amplifiers designed for any purpose. However, the effectiveness of an amateur 'phone transmitter can be increased to a remarkable extent by taking advantage of speech characteristics and of the requirements in *voice* communication.

Measures that may be taken to make the modulation more effective include band compression (filtering), volume compression, and speech clipping.

Compressing the Frequency Band

Most of the intelligibility in speech is contained in the medium band of frequencies; that is, between about 500 and 2500 cycles. On the other hand, the major portion of speech power is normally concentrated below 500 cycles. It is these low frequencies that modulate the transmitter most heavily. If they are eliminated, the frequencies that carry most of the actual communication can be increased in amplitude without exceeding 100-per-cent modulation, and the effectiveness of the transmitter is correspondingly increased.

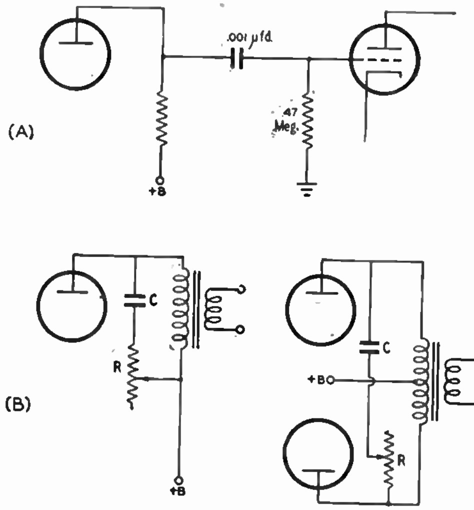


Fig. 9-17 — A, use of a small coupling condenser to reduce low-frequency response; B, tone-control circuits for reducing high-frequency response. Values for *C* and *R* are discussed in the text; 0.01 μfd. and 25,000 ohms are typical.

One simple way to reduce low-frequency response is to use small values of coupling capacitance between resistance-coupled stages, as shown in Fig. 9-17A. A time constant of 0.0005 second for the coupling condenser and following-stage grid resistor will have little effect on the amplification at 500 cycles, but will practically halve it at 100 cycles. In two cascaded stages the gain will be down about 5 db. at 200 cycles and 10 db. at 100 cycles. When the grid resistor is ½ megohm a coupling condenser of 0.001 μfd. will give the required time constant.

The high-frequency response can be reduced by using "tone control" methods, utilizing a condenser in series with a variable resistor connected across an audio impedance at some point in the speech amplifier. The best spot for the tone control is across the primary of the output transformer of the speech amplifier, as in Fig. 9-17B. The condenser should have a reactance at 1000 cycles about equal to the load resistance required by the amplifier tube or tubes, while the variable resistor in series may have a value equal to four or five times the load resistance. The control can be adjusted while listening to the amplifier, the object being to cut the high-frequency response as much as possible without unduly sacrificing intelligibility.

Compressing the audio-frequency band not only puts more modulation power in the optimum frequency band but also reduces hum, because the low-frequency response is reduced, and helps reduce the width of the channel occupied by the transmission, because of the reduction in the amplitude of the high audio frequencies.

Volume Compression

It is obviously desirable to modulate the transmitter as completely as possible — without, of course, overmodulating and setting up spurious sidebands. However, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use automatic gain control that follows the *average* (not instantaneous) variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier.

A practical circuit for this purpose is shown in Fig. 9-18. 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 power 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 an adjustable positive bias on the rectifier cathodes. This prevents the limiting action from beginning until a desired microphone input level is reached. R_2, R_3, C_2, C_3 and C_4 filter the audio frequencies from the rectified output. The output of the rectifier may be connected to the suppressor grid of a 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. If a transformer having a center-tapped secondary is not available, a half-wave rectifier may be used instead of the full-wave circuit shown, but it will be harder to get satisfactory filtering.

The over-all gain of the system must be high enough so that full output can be secured at a moderately low voice level.

Speech Clipping and Filtering

Earlier in this chapter it was pointed out that with sine-wave 100-per-cent modulation the average power increases to 150 per cent of the unmodulated carrier power, but that in speech waveforms the average power content is considerably less than in a sine wave, when

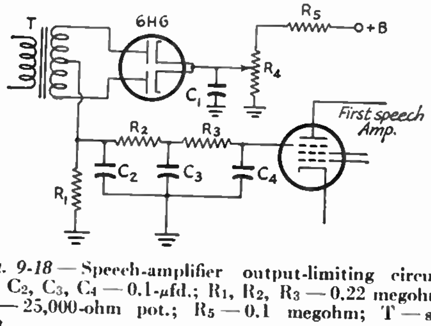


Fig. 9-18 — Speech-amplifier output-limiting circuit. C_1, C_2, C_3, C_4 — 0.1-μfd.; R_1, R_2, R_3 — 0.22 megohm; R_4 — 25,000-ohm pot.; R_5 — 0.1 megohm; T — see text.

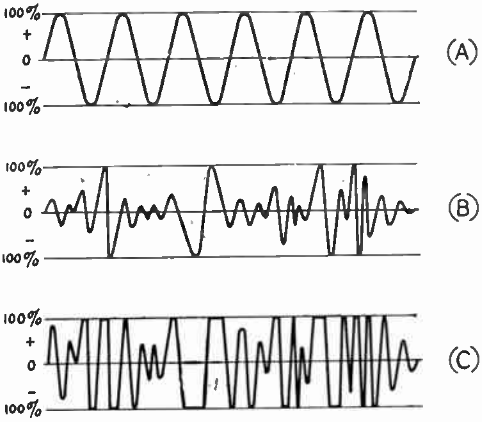


Fig. 9-19 — The normal speech wave (B) has high peaks but low average energy content. When the peaks are clipped the signal may be increased to a considerably higher power level without causing overmodulation (C).

both waveforms have the same *peak* amplitude. Nevertheless, it is the peak conditions that count in modulation. This is shown in the drawings of Fig. 9-19. The upper drawing, A, represents a sine wave having a maximum amplitude that just modulates a given transmitter 100 per cent. The same maximum amplitude will modulate the same transmitter 100 per cent regardless of the waveform of the modulating signal. The speech wave at B, therefore, also represents 100-per-cent modulation.

Now suppose that the amplitude of the wave shown at B is increased so that its power is comparable with — or even higher than — the power in a sine wave, but that everything above the 100-per-cent modulation mark is cut off. We then have a wave such as is shown at C, which is the wave at B increased in amplitude but with its peaks "clipped." This signal will not modulate the transmitter more than 100 per cent, but the voice power will be several times as great. The wave is not exactly like the one at B, so the result will not sound exactly like the original. However, such clipping can be used to secure a worth-while increase in modulation power without sacrificing *intelligibility*. The clipping can be done in the speech amplifier, and once the system is properly adjusted *it will be impossible to overmodulate the transmitter* no matter how much gain is used ahead of the clipper — because the clipper will hold the maximum output amplitude to the same value no matter what the amplitude of the signal applied to it.

But by itself the clipper is not enough. Although the clipping takes place in the audio system, the signal applied to the modulated r.f. amplifier has practically the same waveshape that the modulation envelope *would have had* if the signal were unclipped and the transmitter were badly overmodulated. In other words, clipping generates the same high-order harmonics that overmodulation does. It is therefore necessary to prevent the higher audio frequencies from reaching the modulator. In other words, the frequencies above those needed for intelligible speech must be filtered out, *after* clipping and *before* modulation. The filter required for this purpose should have relatively little attenuation at frequencies below about 2500 cycles, but very great attenuation for all frequencies above 3000 cycles.

It is possible to use as much as 25 db. of clipping before intelligibility is lost; that is, if the original peak amplitude is 10 volts, the signal can be clipped to such an extent that the resulting maximum amplitude is less than one volt. If the original 10-volt signal represented the amplitude that caused 100-per-cent modulation on peaks, the clipped and filtered signal can then be amplified up to the same 10-volt peak level for modulating the transmitter, with a very considerable increase in modulation power.

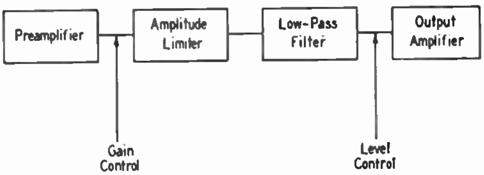


Fig. 9-20 — Block diagram of speech-clipping and filtering amplifier.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 db. or so) there is almost no perceptible change in "quality" but the voice power is four or five times as great as in ordinary modulation.

Before drastic clipping can be used, the speech signal must be amplified up to 10 times more than is necessary for normal modulation. Also, the hum and noise must be much lower than the tolerable level in ordinary amplification, because the noise in the output of the amplifier increases in proportion to the gain.

The clipper-filter system is shown in block form in Fig. 9-20. The limiter is a peak-limiting rectifier of the same general type that is used in receiver noise limiters. It must clip both positive and negative peaks. The gain control sets the amplitude at which clipping starts. Following the low-pass filter for eliminating the harmonic distortion frequencies is a second gain control, the "level" control. This control is set initially so that the amplitude-limited output of the clipper-filter modulates the transmitter 100 per cent. Thereafter it need not be touched. The clipper-filter system is consequently an automatic "overmodulation-preventer," and is a worth-while addition to the transmitter on that account even though deep clipping is seldom used.

Practical circuits are illustrated in a speech amplifier described later in this chapter.

Speech Amplifier with Push-Pull Triode Output

The speech amplifier shown in Fig. 9-21 is a general-purpose unit of straightforward design. Using a pair of power triodes in the output stage, it is capable of an actual undistorted output power of about 8 watts. It can therefore be used to drive a Class B modulator of moderate power output. It is also suitable for use as a grid-bias modulator for high-power transmitters. The gain of the amplifier is ample for the ordinary communications-type crystal microphone.

As shown in the circuit diagram, Fig. 9-22, the amplifier has a pentode first stage using a 6SJ7. A medium-μ triode, a 6J5, is used in the second stage. The gain control is in the grid circuit of this tube. The third stage uses a 6SL7GT in the self-balancing phase-inverter circuit, to obtain push-pull output for the grids of the output tubes. The final stage has two 6B4Gs in push-pull, operated Class AB_1. The power supply for the amplifier is included on the same chassis.

The circuits of individual stages are basically as described earlier in this chapter. R_4 and R_{10} are decoupling resistors in the 6SJ7 and 6J5 stages, respectively, to prevent unwanted feed-back. These resistors, in combination with C_3 and C_7, also provide some additional power-supply hum filtering for the first two stages where the signal level is low. Condenser C_5, which is shunted across the gain control, R_5, when S_1 is closed, serves to reduce the gain at frequencies above about 2500 cycles. This, as explained earlier in this chapter, is desirable because it reduces the width of the channel occupied by the transmitter. R_{17} and C_{12} are the cathode-bias resistor and by-pass condenser, respectively, for the output stage. C_{12} should not be omitted unless the output stage operating conditions are changed so that the amplifier operates purely Class A. When the plate current varies, as it does in Class AB_1 operation, the varying current through R_{17} will introduce considerable distortion unless the resistor is by-passed by a low-impedance condenser.

In the power-supply circuit, S_3 is used for shutting off the plate voltage while leaving the heater power on the tubes. A two-section condenser-input smoothing filter is used. The plate voltage for the output amplifier is taken from the first section; this makes the voltage available for the plates somewhat higher because it avoids the voltage drop through L_2. Hum at this point is inconsequential because of the high power level.

Parts Layout

The speech section occupies the left-hand side of the chassis and the power-supply section the right. Controls along the front chassis edge are the tone-control switch, S_1, gain control, R_5, microphone connector, "B" switch, S_3, and a.c. switch, S_2. The 6SJ7 is behind the microphone connector on the chassis, and the 6J5 is to its left, near the gain control. The 6SL7 phase inverter and 6B4G output tubes are located behind the 6J5.

At the right, the power transformer is at the rear of the chassis, the 5Y3GT rectifier in front, and the first filter choke, L_1, is to the left of the rectifier tube. The output transformer is at the rear center of the chassis.

The bottom view shows the cathode resistor, R_{17}, for the 6B4Gs at the lower right, together with its by-pass condenser, C_{12}. Just above is the second filter choke, L_2. The filter condensers, C_{13}, C_{14} and C_{15}, are the larger tubular units located to the left. The resistors and condensers associated with individual stages are grouped about the appropriate tube sockets. The terminals of the output transformer, T_1, project through a cut-out in the chassis, and secondary leads are brought out to a terminal strip.

A shielded lead should be used from the microphone connector to

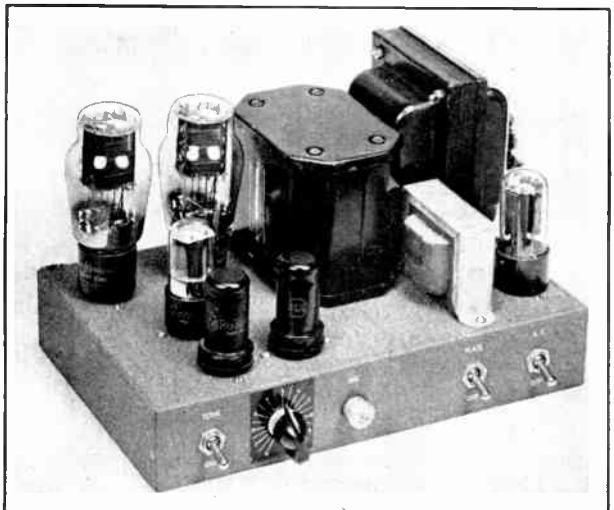


Fig. 9-21 — This amplifier uses 6B4Gs (equivalent to 6A3s) as output tubes and will deliver 8 watts of undistorted power. It is complete with power supply on a 7 × 11 × 2-inch chassis.

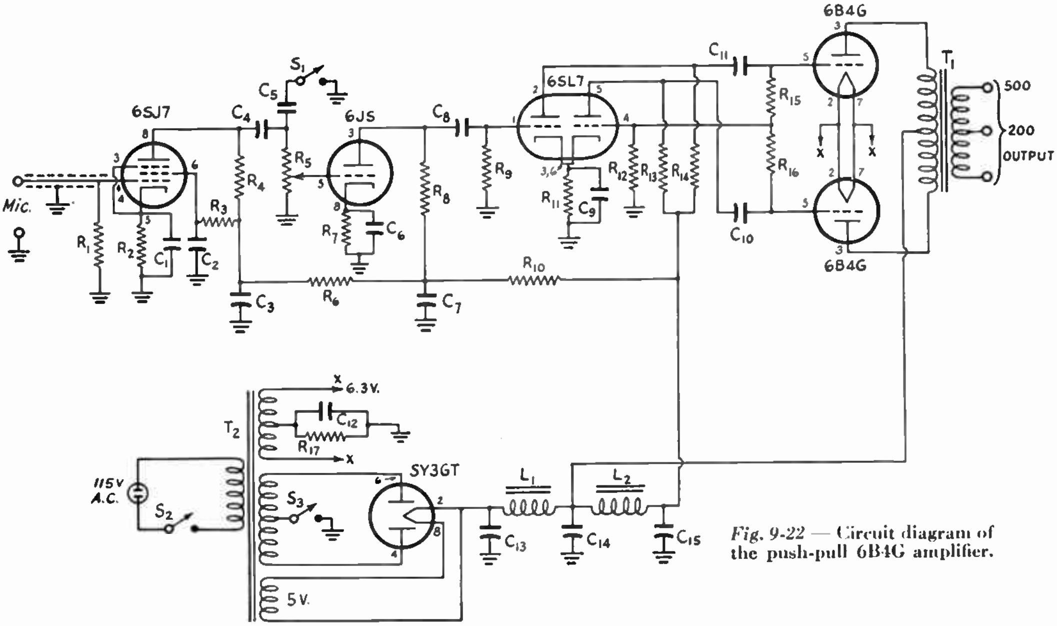


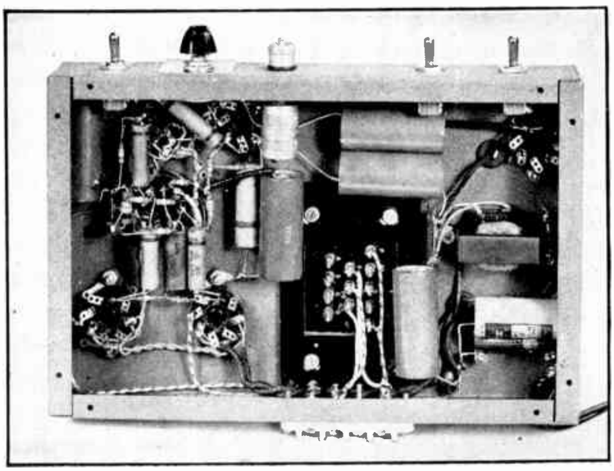
Fig. 9-22 — Circuit diagram of the push-pull 6B4G amplifier.

C_1, C_6, C_9 — 20-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_7, C_{13}, C_{14}, C_{15} — 10-μfd. 450-volt electrolytic.
C_4, C_8, C_{10}, C_{11} — 0.01-μfd. 600-volt paper.
C_5 — 0.001-μfd. 500-volt mica.
C_{12} — 50-μfd. 100-volt electrolytic.
R_1 — 1 megohm, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4, R_{12}, R_{13}, R_{14}, R_{15}, R_{16} — 0.22 megohm, ½ watt.
R_5 — 0.5-megohm volume control.
R_6 — 47,000 ohms, ½ watt.
R_8 — 82,000 ohms, ½ watt.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 10,000 ohms, 1 watt.
R_{11} — 1500 ohms, 1 watt.
R_{17} — 750 ohms, 10 watts.
L_1 — 8-hy. 160-ma. filter choke (UTC R-20).
L_2 — 10-hy. 35-ma. filter choke (UTC R-55).
S_1, S_2, S_3 — S.p.s.t. toggle.
T_1 — Output transformer, p.p. plates (5000 ohms) to line (UTC PA-16).
T_2 — 700 volts c.t., 110 ma.; 5 volts, 3 amp.; 6.3 volts, 4.5 amp. (Stancor P-4080).

the grid prong on the 6SJ7 socket, but there are otherwise no special constructional precautions to observe — other than those mentioned in the section on general considerations in speech-amplifier construction.

The output transformer shown in the photographs is designed for working into a 500- or 200-ohm line. This type of transformer may be used when the speech amplifier is located at some distance from the Class B modulator or other unit it is to drive. If desired, a Class B input transformer can be substituted at T_1. In that case, the leads to the modulator-tube grids should be shielded as a precaution against hum or r.f. pick-up. The transformer selected should be designed for working from a 5000-ohm plate-to-plate load to the grids of the modulator tubes selected.

Fig. 9-23 — Bottom view of the push-pull 6B4G amplifier. Output-transformer terminals are brought out to a connection strip on the rear edge of the chassis.



Speech-Amplifier Circuit with Volume Compression

Fig. 9–24 is the circuit diagram of a complete speech amplifier incorporating volume compression. It has sufficient gain for working from a crystal microphone and has a power output (6 watts or more, depending upon the efficiency of the output transformer) sufficient to drive a Class B modulator to an output of about 250 watts. The automatic gain-control circuit uses a separate amplifier and rectifier combined in one tube, a 6SQ7. 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.

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 Class AB_1, without grid current.

The amount of compression is controlled by 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.

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 decreases 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. Special care should be used in making this adjustment.

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 undesirable. The object of adjustment should be to obtain as much compression as possible without danger of overcompression.

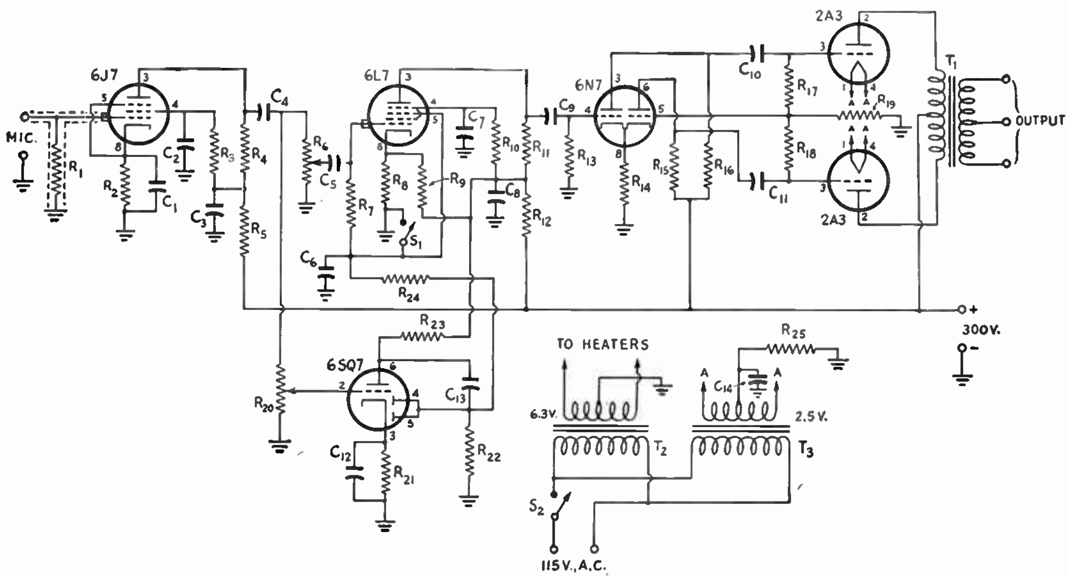


Fig. 9·24 — Circuit diagram of the Class-A 2A3 volume-compression speech amplifier.

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.
C_{14} — 50-μfd. 100-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2, R_8 — 1200 ohms, ½ watt.
R_3, R_7 — 2.2 megohms, ½ watt.
R_4, R_{13}, R_{22}, R_{24} — 0.47 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6, R_{20} — 0.5-megohm variable.
R_9 — 0.22 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} — 0.22 megohm, ½ watt.
R_{21} — 1700 ohms, ½ watt.
R_{25} — 750 ohms, 10 watts.
S_1, S_2 — S.p.s.t. switch.
T_1 — Output transformer to match p.p. 2A3s to Class B grids.
T_2 — Filament transformer, 6.3 volts, 2 amperes.
T_3 — Filament transformer, 2.5 volts, 5 amperes.

A Clipper-Filter Speech Amplifier

The amplifier shown in Fig. 9-25 has a usable output of about 4 watts (sine wave) and includes a clipper-filter for increasing the effectiveness of the modulator and for confining the channel-width to the frequencies needed for intelligible speech. The output stage uses a 6V6 with negative feed-back; this reduces the effective plate resistance of the tube to a low value. The unit therefore can be used to drive a Class B modulator that does not require more than 4 watts on the grids. It can also be used as a complete modulator unit for grid-bias modulation.

As shown in the circuit diagram, Fig. 9-26, the first tube is a 6SJ7. The second stage is one section of a 6SL7GT. With S_3 thrown to the left-hand position, the output of this stage is connected to the grid of a 6J5, which in turn drives the 6V6. Under these conditions the amplifier operates conventionally and has fairly wide frequency response. With S_3 thrown to the right, the output of the first 6SL7GT section is fed to the 6AL5 clipper, and the clipped output is then fed to the grid of the second section of the 6SL7GT. The output of this tube goes through a low-pass filter and thence through a second gain control, R_{15}, to the grid of the 6J5. Thus the clipper-filter feature can be used or not as desired.

The first two stages are resistance-coupled amplifiers following ordinary practice. In the last stage, use is made of the center-tap on the primary of the output transformer to obtain feed-back voltage that is applied to the grid of the 6V6 through the plate resistor, R_{18}, of the 6J5. If a different type of transformer is used, not having a center-tap, a voltage divider can be connected across the primary to obtain the feed-back voltage, as described in the section on negative feed-back in this chapter.

The amplifier has its own power supply, as shown in the diagram and photographs.

Considerations

The clipper circuit uses two diodes, one to clip positive and the other to clip negative peaks, in shunt with a load resistor, R_{11}. The diodes are biased so that they are nonconducting until the signal amplitude reaches about 2 volts. When conducting, the diode resistance is low compared to the resistance of R_{11}, and also compared to the series resistance R_{10}. Under these conditions, all of the voltage in excess of the 2-volt bias appears as a voltage drop in R_{10} (and in the plate resistance of the preceding stage), with the result that the voltage across R_{11} cannot exceed 2 volts.

For convenience, the bias for the diodes is taken from the cathode resistor of the 6V6 by a voltage-dividing arrangement. As shown in Fig. 9-26, the plate of one diode is connected to ground, R_{11} is returned to a point 2 volts above ground, and the cathode of the second diode is returned to a point 4 volts above ground. This makes the plate of each diode 2 volts negative with respect to its own cathode.

The filter shown in Fig. 9-27 is constructed of standard components, the chokes being 125-mh. units usually sold as r.f. chokes. The design of a filter using this value of inductance requires a fairly high capacitance and a low value of load resistance. The constants listed give a sharp cut-off between 2500 and 3000 cycles, with very large attenuation (averaging 45 db. below the response at 1000 cycles) at all frequencies above 3000 cycles. However, the low value of load or terminating resistor, 2000 ohms, greatly decreases the voltage amplification of the 6SL7GT section as compared to what could be obtained with a normal load. The over-all gain with R_{15} at maximum is about the same as with S_3 in the "normal-amplifier" position, despite the extra stage, when the input signal is below the clipping level. Once clipping begins, of course, the output voltage cannot rise above the clipping level no matter how high the amplitude of the input signal.

Construction

The amplifier is built on a 6 × 14 × 3-inch chassis. The input end of the speech amplifier is at the left end and the power supply is at the right. A shield is placed over the 6SL7GT to prevent hum pick-up and to protect the

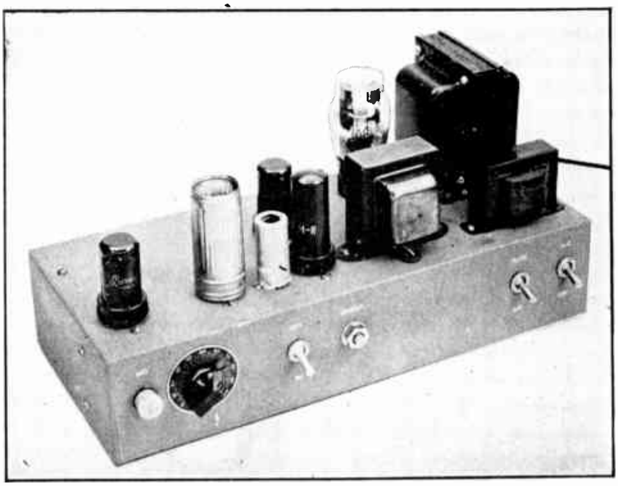


Fig. 9-25 — A 4-watt output amplifier with speech clipping and filtering. It uses a 6V6 output tube with negative feed-back, and has its power supply on the same chassis.

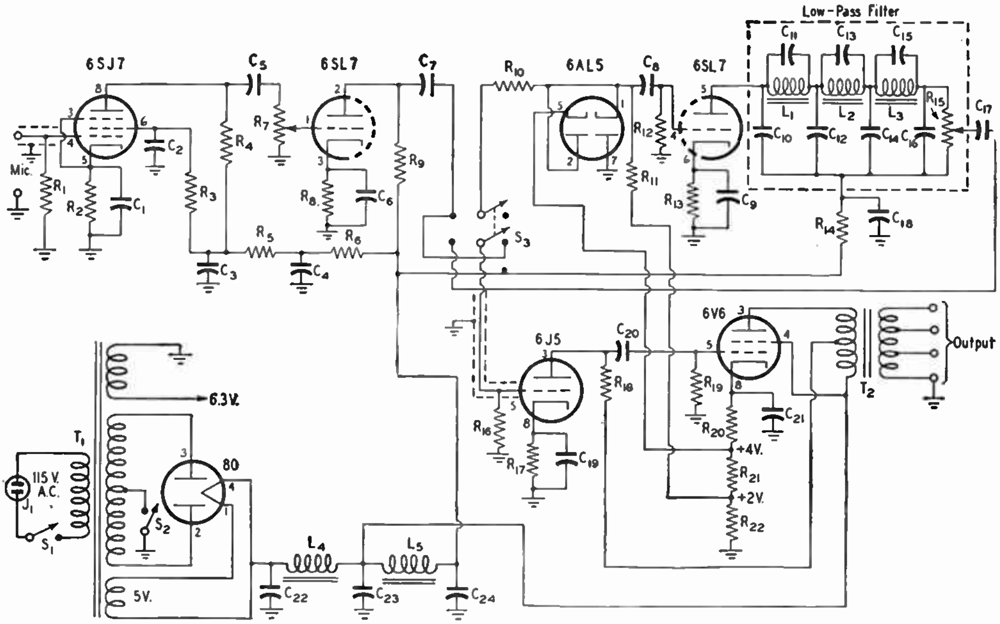


Fig. 9-26 — Circuit diagram of the clipper-filter speech amplifier.

C_1, C_6, C_9, C_{19} — 10-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_4, C_{18}, C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
C_5, C_7, C_8, C_{17}, C_{20} — 0.01-μfd. 600-volt paper.
C_{10}, C_{11}, C_{13} — 0.015-μfd. paper.
C_{12} — 0.03-μfd. paper.
C_{14} — 0.05-μfd. paper.
C_{15} — 0.003-μfd. mica.
C_{16} — 0.06-μfd. paper.
C_{21} — 50-μfd. 50-volt electrolytic.
C_{24} — 16-μfd. 450-volt electrolytic.
R_1 — 1 megohm, ½ watt.
R_2, R_{13} — 1000 ohms, ½ watt.
R_3 — 1.2 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_{10} — 47,000 ohms, ½ watt.
R_6 — 0.1 megohm, ½ watt.
R_7 — 2-megohm volume control.
R_8 — 3300 ohms, ½ watt.
R_9, R_{12}, R_{19} — 0.47 megohm, ½ watt.
R_{11} — 0.15 megohm, ½ watt.
R_{14} — 10,000 ohms, 1 watt.
R_{15} — 2000-ohm wire-wound volume control.
R_{16} — 0.33 megohm, ½ watt.
R_{17} — 1500 ohms, ½ watt.
R_{18} — 82,000 ohms, ½ watt.
R_{20} — 150 ohms, 10 watts.
R_{21}, R_{22} — 39 ohms, 2 watts.
L_1, L_2, L_3 — 125 mh.
L_4 — 10 henrys, 60 ma.
L_5 — 10 henrys, 35 ma.
J_1 — 115-v. a.c. connector.
S_1, S_2 — S.p.s.t. toggle.
S_3 — D.p.d.t. toggle.
T_1 — Power transformer, 350 volts each side c.t., 70 ma.; 5 volts, 2 amp.; 6.3 volts, 3 amp. (Stancor P-4078).
T_2 — Output transformer, 5000 ohms (total primary) to line or voice coil.

tube from r.f. fields from the transmitter. The 6AL5 is between the 6SL7GT and the 6V6. The 6J5 is just to the rear of the 6V6, and the output transformer, T_2, is to its right. Along the front edge of the chassis are the microphone connector; gain control, R_7; clipper-filter switch, S_3; the "output" control, R_{15}; and — at the far right — the "B" voltage and a.c. toggle switches.

The low-pass filter is built as a unit on a 2 × 5-inch mounting board, as shown in Fig. 9-27. The coils are kept well separated and are mounted so that their axes are all at right angles. This prevents magnetic coupling between them, and is essential to good filter performance. In other respects the placement of parts in the filter is not critical. If the proper values of capacitance are not at hand, they can be made up by connecting smaller units in parallel. For example, a 0.01-μfd. paper and 0.005-μfd. mica can be paralleled to make 0.015 μfd. The filter unit occupies the upper right-hand corner in the bottom-view photograph, Fig. 9-28.

Particular care should be taken to reduce hum. The 6SJ7 grid lead must be shielded, and the heater wiring in the vicinity of the first two tubes should be kept in the corners of the chassis except where it is necessary to bring the ungrounded wire out to the socket terminal. It is worth while to try reversing the heater connections on the 6SJ7 to reduce hum. Reducing the gain at the lower frequencies also will reduce the hum in the output, and this may be done by decreasing the capacitance of C_5 and C_7 to 0.002 μfd. instead of the 0.01 μfd. specified.

The output transformer, T_2, in this unit is a low-impedance output type, with 500- and 200-ohm line taps as well as taps for a 'speaker voice coil. If the unit is close to the Class B modulator a Class B driver transformer can be substituted, if desired, or a 1-to-1 transformer can be used for grid-bias modulation.

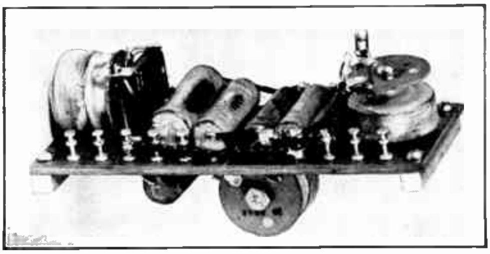


Fig. 9-27 — The low-pass filter is assembled as a unit on its own mounting board. Readily-available parts are used throughout.

Adjusting the Clipper-Filter Amplifier

The good effect of the low-pass filter in eliminating splatter can be entirely nullified if the amplifier stages following the filter can introduce appreciable distortion. That is a primary reason for the use of negative feedback in the output stage of the amplifier described. Amplifier stages following the unit must be operated well within their capabilities; in particular, the Class B output transformer (if a Class B modulator is to be driven) should be shunted by condensers to reduce the high-frequency response as described in the section on Class B modulators.

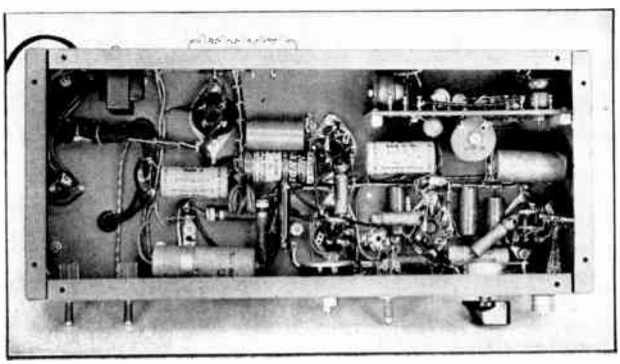
The setting of R_{15} is most important. It is most easily done with the aid of an oscilloscope (one having a linear sweep) and an audio oscillator, using the test set-up shown in the section on testing of speech equipment. Use a resistance load on the output transformer to reflect the proper load resistance (5000 ohms) at the plate of the 6V6. First set R_{15} at about ¼ the resistance from the ground end, switch in the clipper-filter, and apply a 500-cycle sine-wave signal to the microphone input. Increase the signal amplitude until clipping starts, as shown by flattening of both the negative and positive peaks of the wave. To check whether the clipping is taking place in the clipper or in the following amplifiers, throw S_3 to the "normal" or "out" position; the waveshape should return to normal. If it does not, return S_3 to the "in" position and reduce the setting of R_{15} until it does. Then reduce the amplifier gain by means of R_7 until the signal is just below the clipping level. At this point the signal should be a sine wave. Increase R_{15}, without touching R_7, until the wave starts to become distorted, and then back off R_{15} until distortion disappears.

Next, change the input-signal frequency to 2000 cycles, without changing the signal level. Slowly increase R_7 while observing the pattern. At this frequency it should be almost impossible to get anything except a sine wave through the filter, so if distortion appears it is the result of overloading in the amplifiers *following* the filter. Reduce the setting of R_{15} until the distortion disappears, even when R_7 is set at maximum and the maximum available signal from the audio oscillator is applied to the amplifier. The position of R_{15} should be marked at this point and the marked setting should never be exceeded.

To find the *operating* setting of R_{15}, leave the audio-oscillator signal amplitude at the value just under the clipping level and set up the complete transmitter for a modulation check, using the oscilloscope to give the trapezoidal pattern. With the Class C amplifier and modulator running, find the setting of R_{15} (keeping the audio signal just under the clipping level) that just gives 100-per-cent modulation. This setting should be below the maximum setting of R_{15} as previously determined; if it is not, the driver and modulator are not capable of modulating the transmitter 100 per cent and must be redesigned — or the Class C amplifier input must be lowered. Assuming a satisfactory setting is found, connect a microphone to the amplifier and set the amplifier gain control, R_7, so that the transmitter is modulated 100 per cent. Observe the pattern closely at different settings of R_7 to see if it is possible to overmodulate. If overmodulation does not occur at any setting of R_7, the transmitter is ready for operation and R_{15} may be locked in position; it need never be touched subsequently. If some overmodulation does occur, R_{15} should be backed off until it disappears and then locked.

In the absence of an oscilloscope the other methods of checking distortion described in the section on speech-amplifier testing may be used. The object is to prevent distortion in stages following the filter, so that when the clipping level is exceeded the following stages will be working within their capabilities.

Fig. 9-28 — Bottom view of the clipper-filter speech amplifier. Resistors and condensers are grouped around the sockets to which they connect.



6L6 Modulators for Low-Power Transmitters

Plate modulation for transmitters operating at final-stage plate power inputs up to 75 or 80 watts can be provided at relatively small cost by using Class AB 6L6s as modulators. The combined speech amplifier and modulator shown in Fig. 9-29 uses the 6L6s as Class AB_2 amplifiers and has an output (from the transformer secondary) of about 40 watts. The first stage is a 6SJ7 high-gain pentode amplifier, and is resistance coupled to one section of a 6SN7GT triode amplifier. The other section of the 6SN7GT is used as a single-tube phase inverter to obtain push-pull output. The grids of the push-pull 6L6s are driven by a 6SN7GT, with the two sections in push-pull, through transformer T_1. The gain control, R_6, is in the grid circuit of the first 6SN7GT section, and is shunted by condenser C_5 to reduce the high-frequency response. Condenser C_{11}, across the secondary of T_1, serves a similar purpose. The over-all circuit constants have been chosen so that the maximum response is in the most effective speech-frequency band. The response is down about 10 db. at 100 and 3000 cycles, as compared with the range 300–1500 cycles. The gain is more than sufficient for typical crystal microphones.

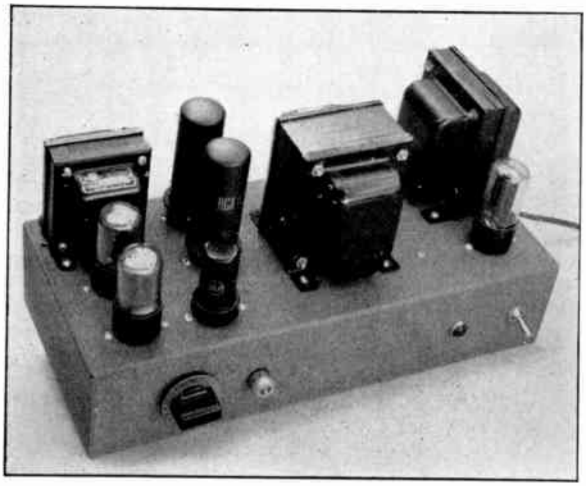


Fig. 9-29 — A 40-watt modulator of inexpensive construction. The second tube from the left, in the foreground, is the 6SJ7 first amplifier. The microphone connector is immediately below it on the chassis wall. Along the left edge, from the front, are the first and second 6SN7GTs and the driver transformer for the 6L6s. The output transformer is to the right of the 6L6s. The power transformer and rectifier are at the far right.

A power supply for the speech-amplifier stages and for the 6L6 heaters is included in the unit, but the power for the 6L6 plates and screens must be obtained from a separate supply. Fixed bias for the 6L6 grids is obtained from the built-in supply by taking the drop across R_{19}. This resistor, a potentiometer, should be adjusted so the voltage drop across it is 22.5 volts when the speech-amplifier stages are operating normally.

In building the amplifier, the usual precautions as to placement of components and wiring to avoid hum and feed-back should be observed. The microphone connector, J_1, should be located close to the 6SJ7 socket so the lead to the grid can be short. This lead also should be shielded.

The power supply for the 6L6s must have good voltage regulation, since the total current varies from approximately 95 ma. with no signal to 220 ma. at full output. A heavy-duty choke-input plate supply should be used; general design data will be found in Chapter Seven.

20-Watt Modulator

Fig. 9-32 is the circuit of a speech amplifier and modulator that has an output of approximately 20 watts. This circuit also uses 6L6s as output tubes, but the amplifier operates Class AB_1 and thus requires no driving power. Because of this, fewer voltage-amplifier stages are needed than in the case of the 40-watt amplifier. Push-pull input for the grids of the 6L6s is secured by using a single-plate-to-push-pull audio transformer between the 6J5 and the 6L6s. In this case it is economical to use a single power supply for the entire amplifier, so the low-voltage supply circuit shown in the 40-watt amplifier circuit may be omitted.

This amplifier can be used to plate-modulate

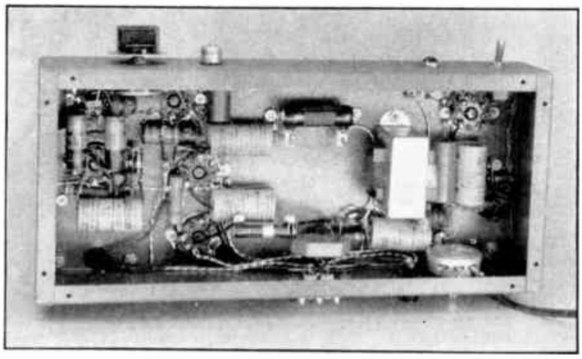


Fig. 9-30 — Underneath the chassis of the 40-watt modulator. The power-supply choke is mounted below chassis at the right. The bias-setting resistor, R_{19}, is on the rear chassis wall, at the lower right in this photograph. Other components are grouped near the tube socket with which they are associated.

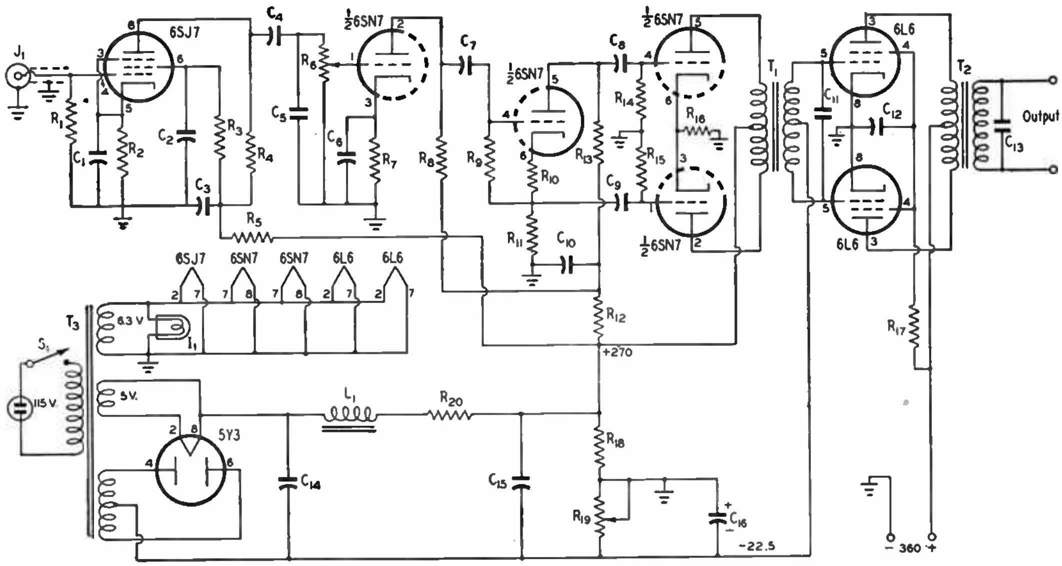


Fig. 9-31 — Circuit diagram of the 40-watt modulator.

C_1, C_6 — 25-μfd. 25-volt electrolytic.
C_2, C_4, C_7, C_8, C_9 — 0.1-μfd. 400-volt paper.
C_3, C_{10}, C_{12}, C_{14}, C_{15} — 8-μfd. 450-volt electrolytic.
C_5 — 470-$\mu\mu$fd. mica.
C_{11} — 0.1-μfd. 600-volt paper.
C_{13} — 0.01-μfd. 1200-volt mica.
C_{16} — 50-μfd. 50-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 0.5-megohm potentiometer.
R_8, R_{13} — 56,000 ohms, ½ watt.
R_9, R_{14}, R_{15} — 0.47 megohm, ½ watt.
R_{10} — 18,000 ohms, ½ watt.
R_{11} — 39,000 ohms, ½ watt.
R_{12} — 10,000 ohms, 1 watt.
R_{16} — 470 ohms, 1 watt.
R_{17} — 7500 ohms, 10 watts.
R_{18} — 7000 ohms, 25 watts.
R_{19} — 1000-ohm wire-wound potentiometer, 4 watts.
R_{20} — 1200 ohms, 1 watt.
L_1 — Smoothing choke; 12 henrys, 80 ma. (Thordarson T20C53).
I_1 — 6.3-volt pilot lamp.
J_1 — Microphone-cable connector (Amphenol).
T_1 — Class AB_2 driver transformer, p.p. plates to p.p. grids (Stancor A-4416).
T_2 — Modulation transformer, 3800 ohms to desired load (unit shown is Stancor A-3893).
T_3 — Power transformer: 350 volts each side center-tap, 70 ma.; 5 volts, 3 amp.; 6.3 volts, 3 amp. (Stancor P-4078).

an input of 40 watts to the r.f. amplifier. It is necessary, of course, to choose the proper output-transformer turns ratio to couple the modulator and modulated amplifier. The output stage is designed to work into a plate-to-plate load of 9000 ohms.

For the maximum power output of 20 watts, the plate supply for the amplifier must deliver 145 ma. at 360 volts. A condenser-input supply of ordinary design (Chapter Seven) may be used. The total plate current is approximately 120 ma. with no signal and 145 ma. at full output. If no more than 12 or 13 watts is needed, R_9 and R_{10} may be omitted and all tubes fed directly from a "B" supply giving approximately 175 ma. at 270 volts.

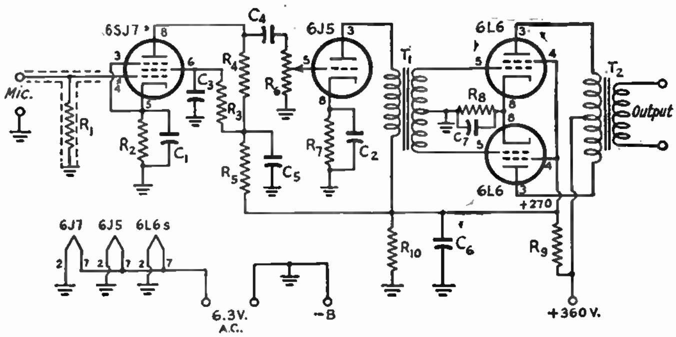


Fig. 9-32 — Circuit diagram of a low-cost modulator capable of power outputs up to 20 watts.

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.
C_7 — 50-μfd. 50-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1 watt.
R_8 — 250 ohms, 10 watts.
R_9 — 2000 ohms, 10 watts.
R_{10} — 20,000 ohms, 25 watts.
T_1 — Interstage audio transformer, single plate to p.p. grids, ratio 3:1.
T_2 — Output transformer, type depending on requirements.

An 807 Modulator and Speech Amplifier

The combined speech amplifier and modulator unit shown in Fig. 9-33 is simple and inexpensive in design and, with the exception of the plate supply for the modulator tubes, is contained on a chassis measuring 3 × 8 × 17 inches. With a 750-volt plate supply, the push-pull Class AB_2 807s are capable of a tube output of 120 watts, or enough to plate-modulate a Class C stage with 200 watts input, allowing for moderate losses in the modulation transformer.

As shown in Fig. 9-34, the first tube in the speech amplifier is a 6J7 (a 6SJ7 may be substituted). A 6SN7GT is used in the second stage, one section serving as a voltage amplifier and the other as a phase inverter of the self-balancing type. The gain control for the amplifier is in the grid circuit of the first half of the tube. The third tube, also a 6SN7GT, is a push-pull amplifier, transformer-coupled to the grids of the 807s.

A power supply for the three tubes preceding the 807s is built on the same chassis. Voltage for the 807 screens is taken from this same supply. The negative return of the supply goes to the chassis through the adjustable arm of potentiometer R_{17}, which is connected in series with the bleeder resistor, R_{16}. The voltage developed in the section of R_{17} below the adjustable arm is negative with respect to chassis, and is used to provide fixed bias for the 807s. C_{11} is connected across this section of R_{17} to by-pass any a.f. current that might flow through the resistor. A separate filament transformer is provided for the 807 heaters, since the total heater power required by all the tubes in the amplifier is somewhat in excess of the rating of the 6.3-volt winding on the ordinary small power transformer.

Resistors R_{14} and R_{15} and condenser C_8 are placed in the 807 screen circuit to suppress the r.f. parasitic oscillations that sometimes occur with these tubes. Their use is principally a precautionary measure, and they may not be required in some installations.

The frequency response of this unit is maximum in the range from about 200 to 2500 cycles, for greatest voice effectiveness and minimum width of the r.f. channel. Frequencies above 2500 cycles are attenuated by condensers C_{12} and C_{13}, the former across the secondary of the driver transformer and the latter across the secondary of the output transformer. The capacitance values given are about optimum for the types of transformers specified and should be close to optimum for other transformers of similar ratings. The voltage rating of C_{13} should be at least equal to the d.c. voltage on the modulated r.f. amplifier.

The photographs show the general layout of components. The 6J7 and 6SN7GT phase inverter are in line at the left-hand front edge of the chassis. The 6SN7GT driver and 5Y3GT rectifier are to the rear of the phase inverter.

The bottom view shows the by-pass condensers and resistors grouped around the sockets to which they connect. The bias-control potentiometer, R_{17}, is mounted on the rear edge of the chassis. A jack shield (National JS-1) covers the microphone jack, and the first-stage grid resistor, R_1, is mounted inside this shield. The lead to the 6J7 grid cap must be shielded and the shield grounded.

The No. 1 terminals of the driver transformer specified should be connected to the grids of the 807s. If a different transformer is used, it should have a primary-to-secondary ratio (total) of about 1-to-1 to couple the

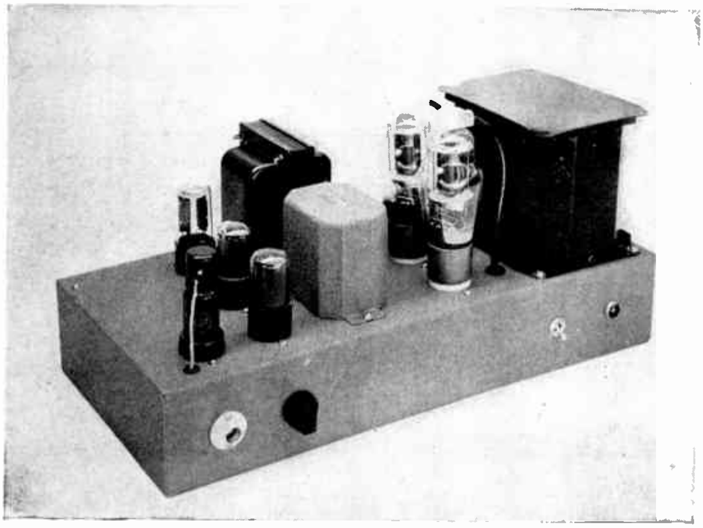


Fig. 9-33 — A speech amplifier and 807 modulator for plate modulation of transmitters up to 200 watts input. The microphone jack and the gain control are at the left end of the chassis. The audio components and tubes occupy the front section, and the power supply for the driver tubes is laid out along the rear edge. The driver transformer is in the center foreground, with the power-supply transformer directly behind it. The large transformer at the right is the modulation transformer.

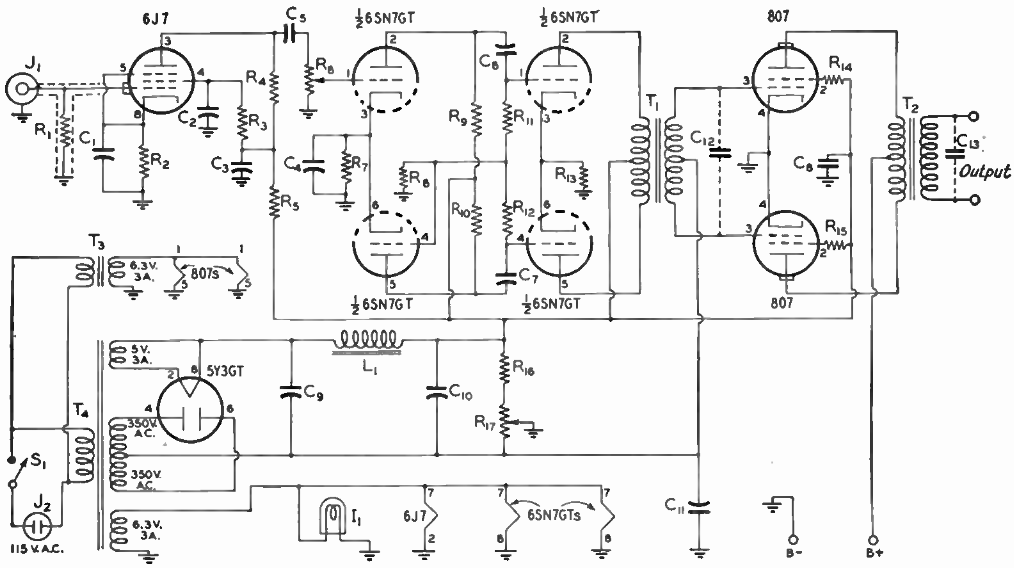
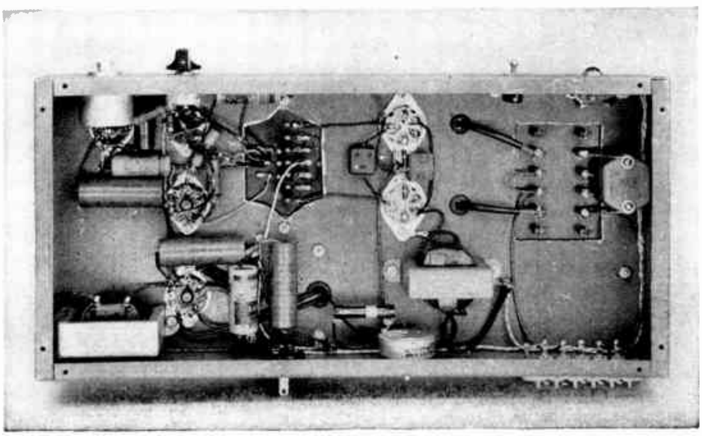


Fig. 9-34 — Circuit diagram of the push-pull 807 speech amplifier-modulator.

C_1 — 10-µfd. 50-volt electrolytic.
C_2 — 0.1-µfd. 400-volt paper.
C_3, C_9, C_{10} — 8-µfd. 450-volt electrolytic.
C_4, C_{11} — 50-µfd. 50-volt electrolytic.
C_5, C_6, C_7 — 0.01-µfd. 400-volt paper.
C_8 — 0.0068-µfd. mica.
C_{12} — 0.001-µfd. mica (see text).
C_{13} — 0.02-µfd. mica (see text).
R_1 — 1 megohm.
R_2, R_7 — 1500 ohms.
R_3 — 1.5 megohms.
R_4, R_8, R_{11}, R_{12} — 0.22 megohm.
R_5 — 47,000 ohms.
R_6 — 1-megohm volume control.
R_9, R_{10} — 0.1 megohm.
R_{13} — 470 ohms.
R_{14}, R_{15} — 100 ohms.
R_{16} — 15,000 ohms, 10 watts.
R_{17} — 1000-ohm wire-wound potentiometer.
(All resistors ½ watt unless otherwise noted.)
L_1 — Smoothing choke, 30 hy., 75 ma., 340-ohm d.c. resistance (Utah 4002).
I_1 — 6.3-volt a.c. pilot-lamp-and-socket assembly.
J_1 — Microphone-cable jack.
J_2 — Panel-mounting a.c. plug (Amphenol 61-M1).
S_1 — S.p.s.t. switch.
T_1 — Push-pull plates to push-pull grids (UTC S-9).
T_2 — Output transformer, type depending on requirements. A multitap transformer (UTC VM-3) is shown in photos.
T_3 — Filament transformer, 6.3 volts, 3 amp. (Thordarson T-21F10).
T_4 — Power transformer, 350 volts a.c. each side of center-tap, 70-ma. rating. Filament windings: 5 v., 3 amp.; 6.3 v., 3 amp. (Stancor P-4078).

6SN7GT and 807 grids properly. The output-transformer turns ratio will depend on the type of operation selected and the modulating impedance of the Class C amplifier. Operated at ICAS ratings, the 807s will deliver a tube output of 120 watts into a plate-to-plate load of 6950 ohms. This requires a plate supply capable of delivering 240 ma. at 750 volts. At CCS ratings the tubes will deliver 80 watts into a 6400-ohm load and require a 600-volt 200-ma. plate supply. The bias should be set, using R_{17}, to give −32 volts between the negative plate-supply terminal and chassis for ICAS operation, and to −30 volts for CCS operation.

Fig. 9-35 — Below-chassis view of the 807 modulator. The shielded microphone jack is in the upper left-hand corner. The filter choke is mounted in the lower left-hand corner and the 807 filament transformer is to the rear and slightly to the right of the 807 tube sockets. The condenser for attenuating the high audio frequencies, shown at the right-hand end of the chassis, is supported by No. 12 wire leads which connect to the output terminals of the modulation transformer.



Class-B Modulators and Drivers

CLASS-B MODULATORS

Plate modulation of all but low-power transmitters requires so much audio power that the Class B amplifier is the only practical type to use. (Included in the Class B category are high-power modulators of the Class AB_2 type; whether the operation is in one class or the other is principally a matter of degree.)

Class B modulator circuits are practically identical no matter what the power output of the modulator. The diagrams of Fig. 9-36 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 cathodes should be connected to ground.

Modulator Tubes

Class B audio ratings of various types of transmitting tubes are given in the tube tables of Chapter Twenty-Five. Choose a pair of tubes that is capable of delivering sine-wave audio power equal to somewhat more than half the d.c. input to the modulated Class C amplifier. It is sometimes convenient to use tubes that will operate at the same plate voltage as that applied to the Class C stage, because one power supply of adequate current capacity may then suffice for both stages.

In estimating the output of the modulator, remember that the figures given in the tables are for the *tube* output only, and do not include output-transformer losses. To be adequate for modulating the transmitter, the modulator should have a theoretical power capability about 25 per cent greater than the actual power needed for modulation.

Matching to Load

In giving Class B ratings on power tubes, manufacturers specify the plate-to-plate load impedance into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance of the Class C r.f. stage, so a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

$$N = \sqrt{\frac{Z_p}{Z_m}}$$

where N = Turns ratio, primary to secondary

Z_m = Modulating impedance of Class C r.f. amplifier

Z_p = Plate-to-plate load impedance for Class B tubes

Example: The modulated r.f. amplifier is to operate at 1250 volts and 250 ma. The power input is

$$P = EI = 1250 \times 0.25 = 312 \text{ watts}$$

so the modulating power required is $312/2 = 156$ watts. Increasing this by 25% to allow for losses and a reasonable operating margin gives $156 \times 1.25 = 195$ watts. The modulating impedance of the Class C stage is

$$Z_m = \frac{E}{I} = \frac{1250}{0.25} = 5000 \text{ ohms.}$$

From the tube tables a pair of Class B tubes is selected that will give 200 watts output when working into a 6900-ohm load, plate-to-plate. The primary-to-secondary turns ratio of the modulation transformer therefore should be

$$N = \sqrt{\frac{Z_p}{Z_m}} = \sqrt{\frac{6900}{5000}} = \sqrt{1.38} = 1.175{:}1.$$

Commercial Class B output transformers usually are rated to work between specified primary and secondary impedances and frequently are designed for specific Class B tubes. In such a case, it will be unnecessary to calculate the turns ratio when the recommended tube combination is used. Many transformers are provided with primary and secondary taps, so that various turns ratios can be

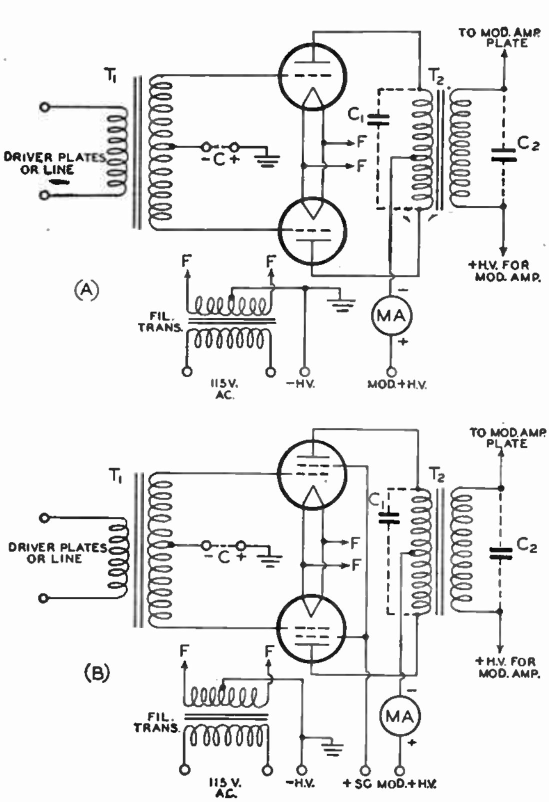


Fig. 9-36 — Class B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

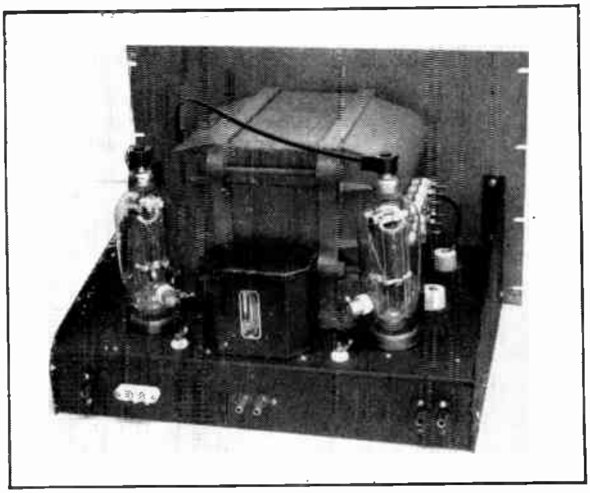


Fig. 9-37 — A typical chassis layout for a Class B modulator. Beyond adequate insulation for the voltages used, and sufficient ventilation for the modulator tubes, no particular constructional precautions are necessary. If the size of the components makes it necessary to use more than one chassis, the driver transformer may be included with the speech amplifier. In such case it is advisable to shield the "hot" audio leads to the modulator grids if they have to run any considerable distance.

obtained to meet the requirements of various tube combinations.

It may be that the exact turns ratio required by a particular tube combination cannot be secured, even with a tapped modulation transformer. *Small* departures from the proper turns ratio will have no serious effect if the modulator is operating well within its capabilities; if the actual turns ratio is within 10 per cent of the ideal value the system will operate satisfactorily. Where the discrepancy is larger, it is always possible to choose a new set of operating conditions for the Class C stage to give a modulating impedance that can be matched by the turns ratio of the available transformer. This may require operating the Class C amplifier at higher voltage and less plate current, if the modulating impedance must be increased, or at lower voltage and higher current if the modulating impedance must be decreased. However, this process cannot be carried too far without exceeding the ratings of the Class C tubes for either plate voltage or current, even though the power input is kept at the same figure. In such a case the only solution is to operate at reduced input and use less of the power available from the modulator.

Suppressing Audio Harmonics

Distortion in either the driver or Class B modulator itself will cause a.f. harmonics that may lie outside the frequency band needed for intelligible speech transmission. While it is almost impossible to avoid some distortion, it *is* possible to cut down the amplitude of the higher-frequency harmonics. The purpose of condensers C_1 and C_2 across the primary and secondary, respectively, of the Class B output transformer in Fig. 9-36 is to reduce the strength of harmonics and unnecessary high-frequency components existing in the modulation.

The condensers act with the leakage inductance of the transformer winding to form a rudimentary low-pass filter. The values of capacitance required will depend on the load resistance (modulating impedance of the Class C amplifier) and the leakage inductance of the particular transformer used. In general, capacitances between about 0.001 and 0.006 μfd. will be required; the larger values are necessary with the lower values of load resistance. A test set-up for measuring frequency response (described in a later section in this chapter) will quickly show the optimum values to use, if a small assortment of condensers is on hand for experimenting. The object is to find the combination of C_1 and C_2 that will give the most rapid reduction in response as the signal frequency is raised above about 2500 cycles.

The voltage rating of each condenser should at least be equal to the d.c. voltage at the transformer winding with which it is associated. In the case of C_2, part of the total capacitance required usually is supplied by the plate by-pass or blocking condenser of the modulated amplifier, so C_2 need only be large enough to make up the difference.

Grid Bias

Many modern transmitting tubes designed for Class B audio work can be operated without grid bias. Besides eliminating the need for a grid-bias supply, this reduces the variation in grid impedance over the audio-frequency cycle and thus gives the driver a more constant load into which to work. With these tubes, the grid return lead from the center-tap of the driver transformer secondary is simply connected to the filament center-tap or cathode.

When the tubes require bias, it should always be supplied from a *fixed* voltage source. Neither cathode bias nor grid-leak bias can be used with a Class B amplifier; with both types the bias changes with the amplitude of the signal voltage, whereas proper operation demands that the bias voltage be unvarying no matter what the strength of the signal. When only a small amount of bias is required it can be obtained conveniently from a few dry cells. When greater values of bias are required, a heavy-duty "B" battery may be used if the grid current does not exceed 40 or 50 milliamperes on voice peaks. Even though the batteries are charged by the grid current rather than discharged, a battery will deteriorate with time and its internal resistance will increase. When the increase in internal resistance becomes appreciable, the battery tends to act like a grid-leak resistor and the bias varies with the applied signal. Batteries should be checked with a voltmeter occasionally while the amplifier is operating. If the bias varies more than 10 per cent or so with voice excitation the battery should be replaced.

As an alternative to batteries, a regulated bias supply may be used. This type of supply is described in Chapter Seven.

Plate Supply

The plate supply for a Class B modulator should be sufficiently well filtered to prevent hum modulation of the r.f. stage. An additional requirement is that the output condenser of the supply should have low reactance, at 100 cycles or less, compared to the load into which each tube is working. (This load is one-fourth of 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.

To keep distortion at a minimum, the voltage regulation of the plate supply should be as good as it can be made. If the d.c. output voltage of the supply varies with the amount of current taken, it should be kept in mind that the voltage at *maximum* current determines the amount of power that can be taken from the modulator without distortion. A supply whose voltage drops from 1500 at no load to 1250 at the full modulator plate current is a 1250-volt supply, so far as the modulator is concerned, and any estimate of the power output available should be based on the lower figure.

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 possible to the recommended value for the tube.

Overexcitation

When a Class B amplifier is overdriven in an attempt to secure more than the rated power, distortion increases rapidly. The high-frequency harmonics which result from the distortion modulate the transmitter, producing spurious sidebands which 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. It *will* happen if the modulator is incapable of delivering the power required to modulate the transmitter fully, or if the Class C amplifier is not adjusted to give the proper modulating impedance.

As previously stated, the tubes used in the Class B modulator should be capable of somewhat more than the power output nominally required. 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. Even though means may be incorporated in the speech amplifier to attenuate frequencies above those necessary for intelligible speech, it is still possible for high-frequency sidebands to be radiated if distortion occurs in the modulator, or if the transmitter is overmodulated.

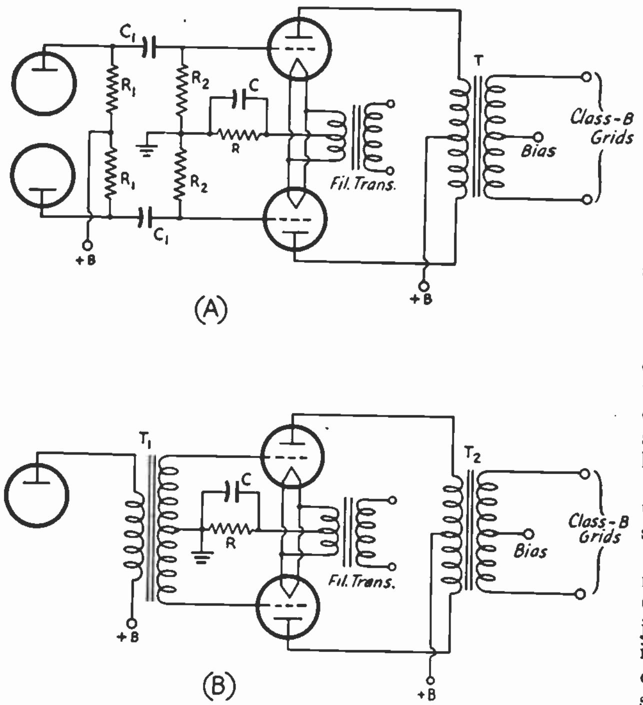


Fig. 9-38 — Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling condenser and grid resistor, respectively; values also may be taken from Table 9-I. In both circuits the output transformer, T, T_2, should have the proper turns ratio to couple between the driver tubes and the Class B grids. T_1 in B is usually a 2:1 transformer, secondary to primary. R, the cathode resistor, should be calculated for the particular tubes used. The value of C, the cathode by-pass, is determined as described in the text.

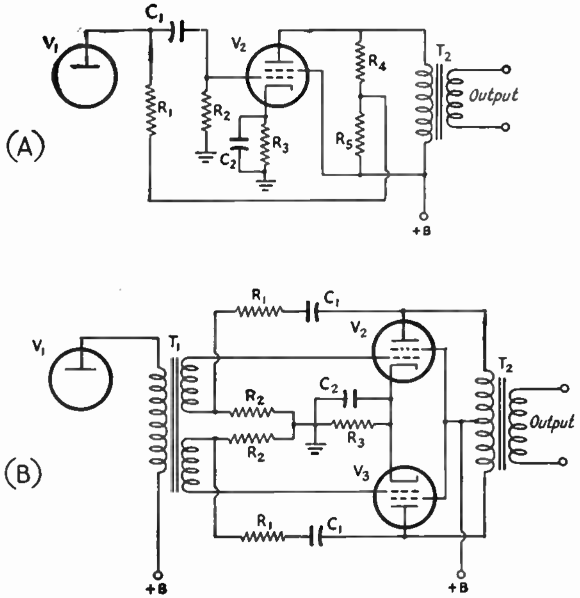


Fig. 9-39 — Negative feed-back circuits for drivers for Class B modulators. A — Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μfd.; C_2, 50 μfd.

B — Push-pull beam-tetrode driver. If V_1 is a 6J5 and V_2 and V_3 6L6s, the following values are suggested: R_1, 0.1 megohm; R_2, 22,000 ohms; R_3, 250 ohms; C_1, 0.1 μfd.; C_2, 100 μfd.

Such high-frequency harmonics can be reduced by connecting condensers across both the primary and secondary of the output transformer as previously described.

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 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.

DRIVERS FOR CLASS-B MODULATORS

Class B amplifiers are driven into the grid-current region, so power is consumed in the grid circuit. The preceding stage or 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, because the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source that will maintain the waveform of the signal without distortion even though the load varies. That is, the driver stage must have good **regulation.** To this end, it should be capable of delivering somewhat more power than is consumed by the Class B grids, as previously described in the discussion on speech amplifiers. It is also desirable to use 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.

The driver transformer, T or T_2 in Fig. 9-38, may couple directly between the driver tube and the modulator grids or may be designed to work into a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the output of the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator; the second transformer not only introduces additional losses but also impairs the voltage regulation of the driver stage.

Driver Tubes

The variation in grid resistance of a Class B amplifier over the audio-frequency cycle poses a special problem in the driver stage. To avoid distortion, the driver output *voltage* (not power) must stay constant (for a fixed signal voltage on its grid) regardless of the variations in load resistance.

The fundamental requirement for good voltage regulation in any electrical generator is that the internal resistance must be low. In a vacuum-tube amplifier, this means that the tubes must have a low value of plate resistance. The best tubes in this respect are low-μ triodes (the 6A3 is an example) and the worst are tetrodes and pentodes as represented by the 6V6 and 6L6. This does not mean that tetrodes (or pentodes) cannot be used, but it does mean that they should not be used without taking measures to reduce the effective plate resistance (see next section).

In selecting a driver stage always choose Class A or AB_1 operation in preference to Class AB_2. This not only simplifies the speech-amplifier design but also makes it easier to apply negative feed-back to tetrodes for reduction of plate resistance. It is possible to obtain a tube power output of approximately 25 watts (from 6L6s) without going beyond Class AB_1 operation; this is ample driving power for the popular Class B modulator tubes, even when a kilowatt transmitter is to be modulated.

The rated tube output (as shown by the

tube tables) should be reduced by about 20 per cent to allow for losses in the Class B input transformer. If two transformers are used, tube-to-line and line-to-grids, allow about 35 per cent for transformer losses. Another 25 per cent should be allowed, if possible, as a safety factor and to improve the voltage regulation.

Fig. 9-38 shows representative circuits for a push-pull triode driver using cathode bias. If the amplifier operates Class A, the cathode resistor need not be by-passed, because the a.f. currents from each tube flowing in the cathode resistor are out of phase and cancel each other. However, in Class AB operation this is not true; considerable distortion will be generated at high signal levels if the cathode resistor is not by-passed. The by-pass capacitance required can be calculated by a simple rule: the cathode resistance in ohms multiplied by the by-pass capacitance in microfarads should equal at least 25,000. The voltage rating of the condenser should be equal to the maximum bias voltage. This can be found from the maximum-signal plate current and the cathode resistance.

Example: A pair of 6A3s is to be used in Class AB_1, self-biased. From the tube tables, the cathode resistance should be 780 ohms and the maximum-signal plate current 120 ma. From Ohm's Law,

$$E = RI = 780 \times 0.12 = 93.6 \text{ volts}$$

From the rule mentioned previously, the by-pass capacitance required is

$$C = 25{,}000/R = 25{,}000/780 = 32\ \mu\text{fd.}$$

A 40- or 50-μfd. 100-volt electrolytic condenser would be satisfactory.

Negative Feed-Back

Whenever tetrodes or pentodes are used as drivers for Class B modulators, negative feed-back should be used in the driver stage. This will reduce the distortion caused by the variable load resistance represented by the Class B grids. It also reduces the distortion inherent in the driver stage itself, when properly applied. The effect of feed-back is to reduce the apparent plate resistance of the driver, and this in turn helps to maintain the a.f. output voltage at a more constant level (for a constant signal on the grid) when the load resistance varies. It is readily possible to reduce the plate resistance to a value comparable to or lower than that of low-μ triodes such as the 6A3.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-39. Fig. 9-39A shows resistance coupling between the preceding stage and a single tetrode, such as the 6V6, that operates at the same plate voltage as the preceding stage. Part of the a.f. voltage across the primary of the output transformer is fed back to the grid of the tetrode, V_2, through the plate resistor of the preceding tube, V_1. The amount of voltage so fed back is determined by the voltage divider, R_4R_5. The total resistance of R_4 and R_5 in series should be large compared to the rated load resistance of V_2. Instead of the voltage divider, a tap on the transformer primary can be used to supply the feed-back voltage, if such a tap is available.

The amount of feed-back voltage that appears *at the grid* of tube V_2 is determined by R_1, R_2 and the plate resistance of V_1, as well as by the relationship between R_4 and R_5. Calculation of the feed-back voltage, although not mathematically difficult, is not ordinarily practicable because the plate resistance of V_1 is seldom known at the particular operating conditions used. Circuit values for a typical tube combination are given in detail in Fig. 9-39.

The push-pull circuit in Fig. 9-39B requires an audio transformer with a split secondary. The feed-back voltage is obtained from the plate of each output tube by means of the voltage divider, R_1R_2. The blocking condenser, C_1, prevents the d.c. plate voltage from being applied to R_1R_2; the reactance of this condenser should be low, compared with the sum of R_1 and R_2, at the lowest audio frequency to be amplified. Also, the sum of R_1 and R_2 should be high compared with the rated load resistance for V_2 and V_3.

In this circuit the feed-back voltage that is developed across R_2 also appears at the grid of V_2 (or V_3) because there is no appreciable current flow (in the usual audio range) through the transformer secondary and grid-cathode circuit of the tube, provided the tubes are not driven to grid current. If the grid-cathode impedance of the tubes is relatively low, as it is when grid current flows, the feed-back voltage decreases because of the voltage drop through the transformer secondary. The circuit should not be used with tubes that are operated Class

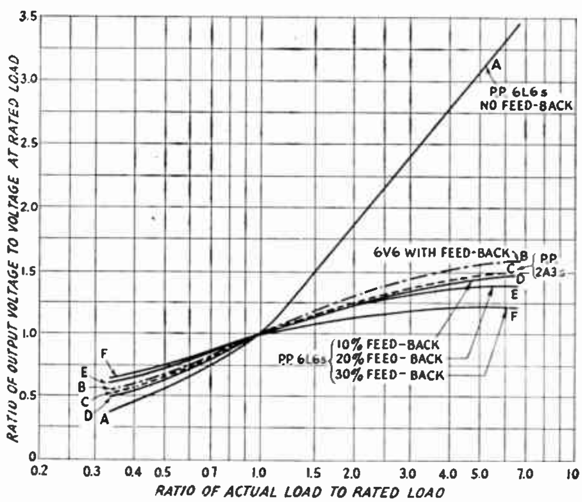


Fig. 9-40 — Output voltage regulation of two types of beam-tetrode drivers with negative feed-back. For comparison, the regulation with a pair of 2A3s (no feed-back) also is shown.

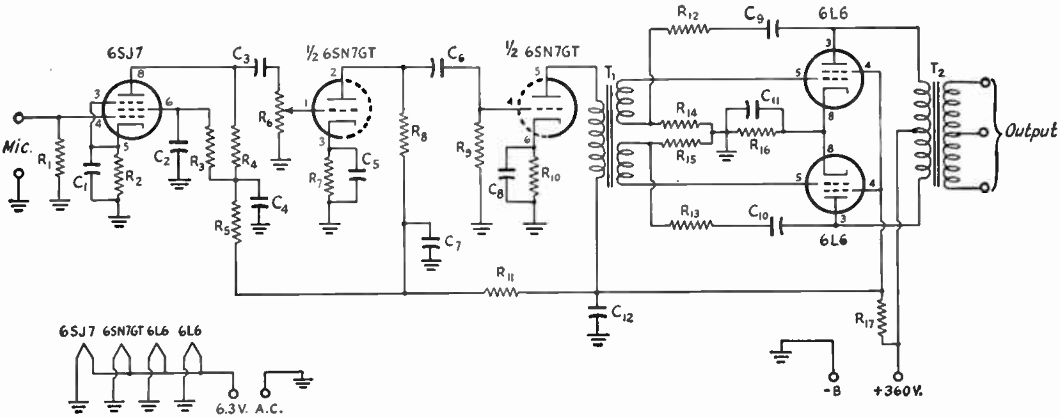


Fig. 9-41 — Circuit diagram of speech amplifier using 6L6s with negative feed-back, suitable for driving Class B modulators up to 500 watts output.

C_1, C_5, C_8 — 20-μfd. 25-volt electrolytic.
C_2, C_9, C_{10} — 0.1-μfd. 400-volt paper.
C_3, C_6 — 0.01-μfd. 600-volt paper.
C_4, C_7, C_{12} — 10-μfd. 450-volt electrolytic.
C_{11} — 100-μfd. 50-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_8 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 10,000 ohms, ½ watt.
R_{12}, R_{13} — 0.1 megohm, 1 watt.
R_{14}, R_{15} — 22,000 ohms, ½ watt.
R_{16} — 250 ohms, 10 watts.
R_{17} — 2000 ohms, 10 watts.
T_1 — Interstage audio, 2:1 secondary (total) to primary, with split secondary winding.
T_2 — Class B input transformer to suit modulator tubes.

AB_2. The per cent feed-back is

$$n = \frac{R_2}{R_1 + R_2} \times 100$$

where n is the feed-back percentage, and R_1 and R_2 are connected as shown in the diagram. The higher the feed-back percentage, the lower the effective plate resistance. However, if the percentage is made too high the preceding tube, V_1, may not be able to develop enough voltage, through T_1, to drive the push-pull stage to maximum output without itself generating harmonic distortion. Distortion in V_1 is not compensated for by the feed-back circuit. If V_2 and V_3 are 6L6s operated self-biased in Class AB_1 with a load resistance of 9000 ohms, V_1 is a 6J5, and T_1 has a turns ratio of 2-to-1, total secondary to primary, it is possible to use over 30-per-cent feed-back without going beyond the output-voltage capabilities of the 6J5. Actually, it is unnecessary to use more than about 20-per-cent feed-back. This value reduces the effective plate resistance to the point where the output voltage regulation is better than that of 6A3s or 2A3s without feed-back.

Instead of the voltage-divider arrangement shown in Fig. 9-39B for obtaining feed-back voltage, a separate winding on the output transformer can be used, provided it has the proper number of turns to give the desired feed-back percentage. Special transformers are available for this purpose.

The improvement in constancy of output voltage resulting from the use of negative feed-back is shown graphically in Fig. 9-40. In order to compare the various types of tubes, the variation in output voltage is shown as a percentage of the output voltage when the tubes are working into the rated load. The load resistance also is expressed as a percentage of the rated load resistance for the particular tube, or pair of tubes, used.

● SPEECH-AMPLIFIER CIRCUIT WITH NEGATIVE FEED-BACK

A circuit for a speech-amplifier suitable for driving a Class B modulator is given in Fig. 9-41. In this amplifier the 6L6s are operated Class AB_1 and will deliver up to 20 watts to the grids of the Class B amplifier. The feed-back circuit requires no adjustment, but does require an interstage transformer with two separate secondary windings (split secondary).

This amplifier may be constructed along the same lines as in Fig. 9-29, observing the same precautions with respect to shielding the 6SJ7 grid circuit. Although the power output is the same as from the amplifier of Fig. 9-32, an additional voltage-amplifier stage is incorporated in the circuit. This is necessary because the voltage fed back from the plates to the grids of the 6L6s opposes the voltage from the preceding stage, so the latter must be increased in order to maintain the same power output from the 6L6s. In turn, this necessitates more over-all voltage gain than is required to drive Class AB_1 p.p. 6L6s without feed-back.

The output transformer, T_2, should be selected to work between a 9000-ohm plate-to-plate load and the grids of whatever Class B tubes will be used. The power-supply requirements for this amplifier are essentially the same as for the amplifier of Fig. 9-32.

Checking 'Phone-Transmitter Operation

SPEECH EQUIPMENT

Every 'phone transmitter requires checking before it is initially put on the air. An adequate job can be done with equipment that is neither elaborate nor expensive. A simple set-up is shown in Fig. 9-42. The only equipment that is not likely to be already at hand is the audio oscillator (the construction of a very simple one is described in Chapter Sixteen). The voltmeter — one that operates at audio frequencies is necessary — is available in any multirange volt-ohm-milliammeter that has a rectifier-type a.c. range. The headset is included for aural checking of the amplifier performance.

A two-step attenuator for the output of the audio oscillator is recommended so that a wide range of output voltages can be smoothly controlled. Also, R_3 should have relatively low resistance — 500 ohms or less; operating at low impedance will minimize stray-hum pick-up, which might cause false results when the amplifier gain is high.

As a preliminary check, cover the microphone input terminals with a metal shield (with the audio oscillator and attenuator disconnected) and, while listening in the headset, note the hum level with the amplifier gain control in the off position. The hum should be very low under these conditions. Then increase the gain-control setting to maximum and observe the hum; it will no doubt increase. Then connect the audio oscillator and attenuator and, starting from minimum signal, increase the setting of R_1 until the voltmeter indicates full power output. (The voltage should equal $\sqrt{PR}$, where P is the expected power output in watts and R is the load resistance — R_4 in the diagram.) Listen carefully to the tone while increasing R_1 to see if there is any change in its character. When it begins to sound like a musical octave instead of a single tone, distortion is beginning. Assuming that the output is substantially without audible distortion at full output, substitute the microphone for the audio oscillator and speak into it in a normal tone while watching the voltmeter. Reduce the gain-control setting until the meter "kicks" nearly up to the full-power reading on voice peaks. Note the hum level, as read on the voltmeter, at this point; the hum level should not exceed one or two per cent of the voltage at full output.

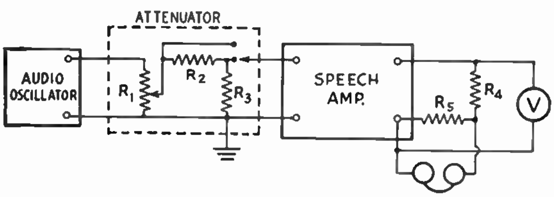


Fig. 9-42 — Simple test set-up for checking a speech amplifier. The audio-oscillator frequency range should be from about 100 to 5000 cycles. It is not necessary that it be continuously variable; a number of "spot" frequencies will be satisfactory. Suitable resistor values are: R_1, 50,000-ohm potentiometer; R_2, 4700 ohms; R_3, 470 ohms; R_4, rated load resistance for amplifier output stage; R_5, determine by trial for comfortable headphone level (25 to 100 ohms, ordinarily). V is a high-resistance a.c. voltmeter, multirange rectifier type.

If the hum level is too high, the amplifier stage that is causing the trouble can be located by temporarily short-circuiting the grid of each tube, in turn, to ground. When shorting a particular grid makes a marked decrease in hum, the hum presumably is coming from a *preceding* stage, although it is possible that it is getting its start in that particular grid circuit. If shorting a grid does *not* decrease the hum, the hum is originating either in the plate circuit of that tube or the grid circuit of the next. Aside from wiring errors or a defective tube, objectionable hum usually originates in the first stage of the amplifier.

If distortion occurs below the point at which the expected power output is secured, the stage in which it is occurring can be located by working from the last stage toward the front end of the amplifier, applying a signal to each grid in turn from the audio oscillator and adjusting the signal voltage for maximum output. In the case of push-pull stages, the signal may be applied to the primary of the interstage transformer — *after* disconnecting it from the plate-voltage source. Assuming that normal design principles have been followed and that all stages are theoretically working within their capabilities, the probable causes of distortion are wiring errors (such as accidental short-circuit of a cathode resistor), defective components, or use of wrong values of resistance in cathode and plate circuits.

An oscilloscope having amplifiers and a linear sweep circuit is a useful instrument in testing audio amplifiers because it provides a ready check on waveform and thus shows distortion instantly. It may be connected across the output circuit as shown in Fig. 9-43, and also may be moved from stage to stage to check the waveform at the grid as well as at the plate. When connected to circuits that are not at ground potential for d.c., a condenser (about 0.1 µfd.) should be connected in series with the "hot" oscilloscope lead. The hot lead preferably should be shielded so that it will not pick up stray hum and introduce it into the amplifier.

CLASS-B MODULATORS

Once the speech amplifier is in satisfactory working condition, a Class B modulator can be checked by similar means. A circuit is shown in Fig. 9-44. The resistance of R_1 should be equal to the modulating impedance of the

Class C amplifier to be modulated, and the resistor should have a power rating equal to the rated power output of the modulator. Calculate the voltage to be expected across R_1 at full output; if it exceeds the range of the meter the meter may be connected across say half or one-fourth of R_1 and the readings multiplied by 2 or 4, respectively. Only a few ohms will be needed at R_2, in the average case, to give a good signal in the headphones. As a safety precaution, ground the output terminal to which the headphones are connected and use a resistor at R_2 that has ample current-carrying capacity.

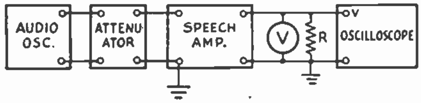


Fig. 9-43 — Test set-up using an oscilloscope for checking waveform.

Hum will seldom be a problem in the modulator. Distortion may be checked as described previously; the oscilloscope is excellent for this purpose. If a variable-frequency audio oscillator is used, a check on the frequency response of the over-all system can be obtained by varying its frequency (check its output voltage at each frequency change) and observing the variation in the modulator output voltage. The high-frequency response of the system can be attenuated by trying condensers of various values across the primary and secondary of the output transformer, as pointed out in the discussion on Class B modulators. The response above 3000 cycles should be small compared to the response in the 200- to 2500-cycle region so that the channel occupied by the transmitter will not be excessive. A simple way to check this is to apply a sine-wave signal of about 1500 cycles and increase its amplitude until distortion becomes noticeable; when this occurs the tone no longer sounds pure but sounds like a musical octave. The condenser values should then be adjusted until the test tone sounds pure again at the same signal amplitude.

THE MODULATED AMPLIFIER

Proper adjustment of a 'phone transmitter is aided immeasurably by the oscilloscope; it will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary. They do, however, give a different type of pattern than is obtained without them.

When using the tube without a sweep circuit, radio-frequency voltage from the modulated amplifier is applied directly to the vertical deflection plates of the tube, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the amplitude of the horizontal signal varies, the r.f. output of the transmitter also varies, and this produces a wedge-shaped pattern or **trapezoid** on the screen. If the oscilloscope has a horizontal sweep, the r.f. voltage is applied to the vertical plates as before (never through an amplifier) and the sweep produces a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a **wave-envelope** modulation pattern.

Oscilloscope connections for both types of patterns are shown in Fig. 9-45. The connections for the wave-envelope pattern are somewhat simpler than those for the trapezoidal figure. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a twisted-pair line and pick-up coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate r.f. harmonics, and the tuning control provides a means for adjustment of the pattern height.

To get a wave-envelope pattern the position of the pick-up coil should be varied until a carrier pattern, Fig. 9-46B, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, 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. This is illustrated by Fig. 9-46D, where the point X represents the sweep line (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.

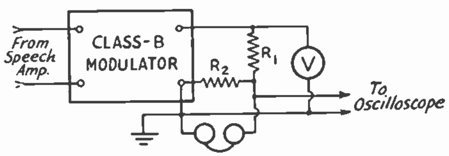


Fig. 9-44 — Set-up for checking a Class B modulator.

Connections for the trapezoidal pattern are shown in Fig. 9-45B. The vertical plates are coupled to the transmitter tank circuit through a pick-up loop; alternatively, the tuned input circuit to the oscilloscope may be used. The horizontal plates are coupled to the output of the modulator through a voltage divider, R_1R_2. R_2 should be a potentiometer so the audio voltage can be adjusted to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control. 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 to $1.4E$ 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,

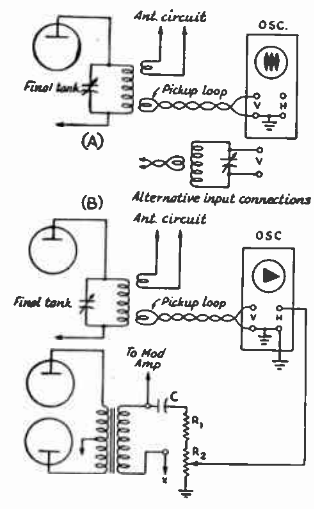


Fig. 9-45 — Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation.

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 in the circuit. With plate modulation, this is twice the d.c. voltage applied to the plate of the modulated amplifier.

Trapezoidal patterns for various conditions of modulation are shown in Fig. 9-46 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 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, X, 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.

Modulation Monitoring

It is always desirable to modulate as fully as possible, but 100-per-cent modulation should not be exceeded — particularly in the downward direction — because harmonic distortion will be introduced and the channel-width increased. This causes unnecessary interference to other stations. The oscilloscope is the best instrument for continuously checking the modulation. However, simpler indicators may be used for the purpose, once calibrated.

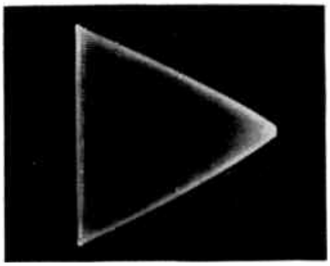
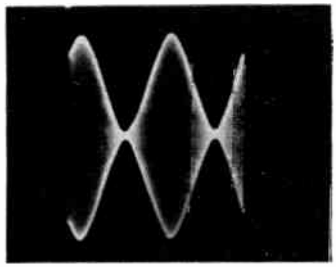
A convenient indicator, when a Class B modulator is used, is the plate milliammeter in the Class B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity that give 100-per-cent modulation on voice peaks, and simultaneously observe the maximum Class B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice 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 that represents 100-per-cent modulation.

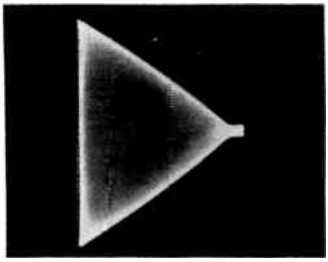
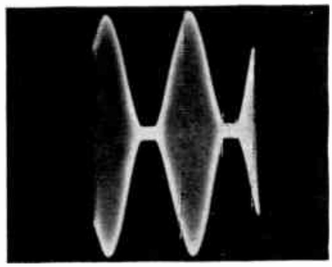
The plate milliammeter of the modulated r.f. stage also is of some value as an indicator of overmodulation. The average plate current stays constant if the amplifier is linear, so the reading will be the same whether or not the transmitter is modulated. When the amplifier is overmodulated, especially 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 under some operating conditions the average plate current will remain constant even though the amplifier is considerably overmodulated. Therefore an indicator of this type is not wholly reliable unless it has been checked previously against an oscilloscope.

Linearity

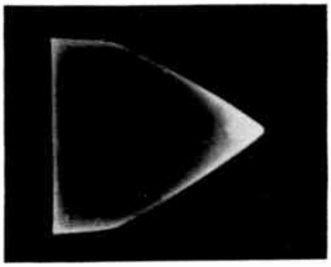
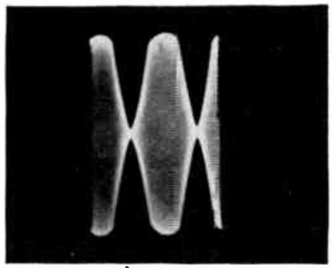
The linearity of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave-envelope pattern, and less auxiliary equipment is required. The connections

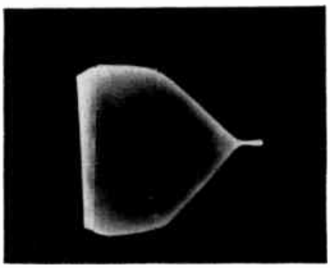
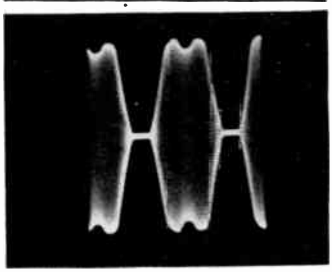
Properly-operated 'phone transmitter modulated 100 per cent.

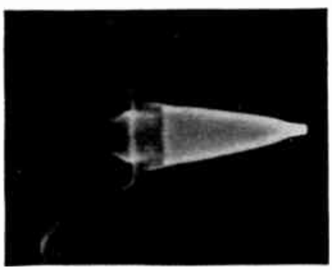
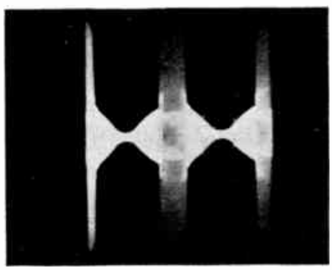
Overmodulation of a transmitter having high modulation capability. Distortion occurs only on the down-peaks.

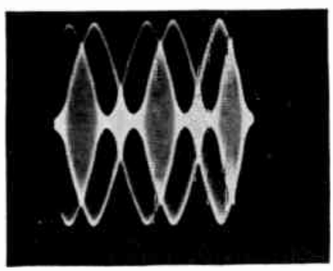
Nonlinearity in modulated r.f. stage, frequently caused by insufficient excitation of a plate-modulated amplifier or overexcitation of a grid-bias modulated amplifier. The amplifier modulates linearly in the downward direction but the up-peaks are flattened.

Overmodulation and nonlinear operation (insufficient modulation capability). These patterns are similar to those directly above, but with the modulation carried beyond 100 per cent in the downward direction.

Overmodulation and parasitic oscillations in the modulated amplifier. The trapezoidal pattern also shows phase distortion caused by incorrect coupling between the oscilloscope and audio system.

Left — Phase distortion caused by incorrect coupling between audio system and oscilloscope. *Right* — Multiple pattern caused by incorrect setting of oscilloscope time-base control. In both cases the wave is modulated 100 per cent.

PHOTOGRAPHS OF TYPICAL OSCILLOSCOPE PATTERNS

These photographs show various conditions of modulation as displayed by the wedge or trapezoidal patterns in the left-hand column and the wave-envelope patterns in the right-hand column.

(Photographs reproduced through courtesy of the Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

are the same as for measuring modulation percentage (Fig. 9-45B). 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. Nonlinearity will be shown by curvature of the sides. Curvature near the point, causing it to approach the axis more slowly 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. 9-46F). If a small vertical line remains, the amplifier should be reneutralized; 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. The usual reason is that the amplifier has been adjusted to have too-high carrier efficiency without modulation.

Fig. 9-47 shows typical patterns of both the trapezoid and wave-envelope 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. 9-46 and 9-47 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 may be 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 section. 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, and patterns in which the sides of the trapezoid are elliptical instead of straight, 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. 9-45B.

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

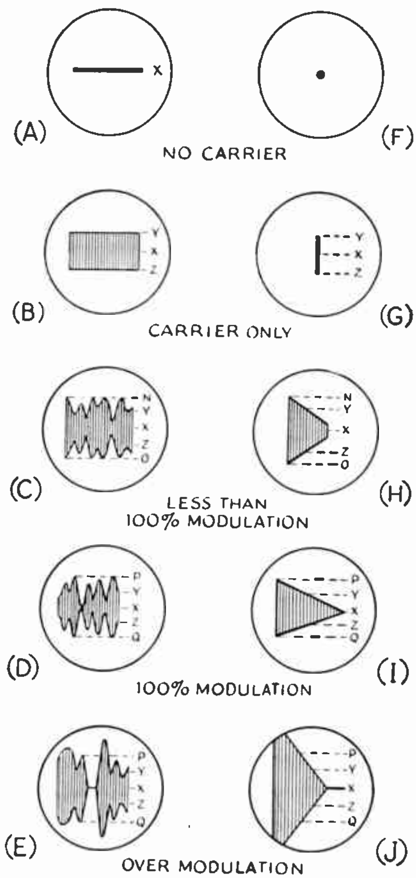


Fig. 9-46 — Wave-envelope and trapezoidal patterns representing different conditions of modulation.

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 capacitance 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 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 capacitance 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. 9-47B; the pattern with an upward kick will look like Fig. 9-47A, with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and Hum on Carrier

Noise and hum may be detected by listening to the signal on a receiver, provided the receiver is far enough away 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

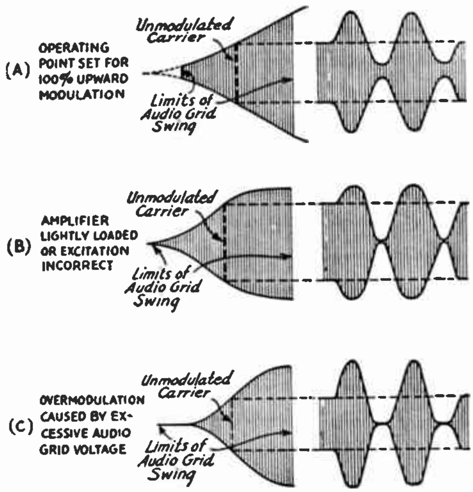


Fig. 9-47 — Oscilloscope patterns representing proper and improper adjustments for grid-bias or cathode modulation. Trapezoidal pattern at left; wave-envelope pattern at right. The pattern obtained with a correctly-adjusted amplifier is shown at A. The other drawings indicate nonlinear modulation from typical causes.

of the r.f. stages have insufficient smoothing. 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 that 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 has the same appearance as ordinary modulation on the carrier. 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 is needed for checking spurious sidebands outside the normal communication channel.

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. 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 3 to 4 kilocycles from the carrier should be of negligible strength in a properly-modulated 'phone transmitter. The causes are overmodulation or nonlinear operation.

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 is necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Direct coupling or unsymmetrical coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible because 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 r.f. power in a dummy antenna, 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.

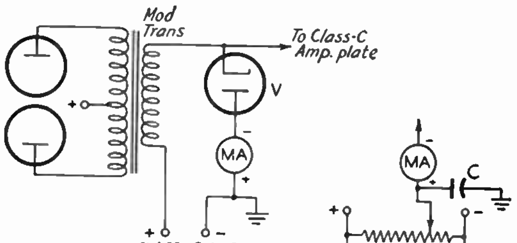


Fig. 9-48 — A negative-peak overmodulation indicator. Milliammeter *MA* may be any low-range instrument (up to 0-50 ma. or so). The inverse peak-voltage rating of the rectifier, *V*, must be at least equal to the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit can be used to indicate modulation in excess of any desired value below 100 per cent.

Overmodulation Indicators

The most positive method of *preventing* overmodulation is the clipper-filter system described earlier, when properly set up and adjusted. In the absence of such a system — or even with it, just to be safe — some form of overmodulation indicator should be in constant use when the transmitter is on the air.

The best device for this purpose is the cathode-ray oscilloscope. The trapezoidal and wave-envelope patterns are equally useful. A 60-cycle sinusoidal sweep will be quite satisfactory for the wave-envelope pattern. Either pattern should be watched particularly for the bright spots at the axis that accompany overmodulation in the downward direction. The speaking-voice intensity should be kept below the level that shows 100-per-cent modulation on the 'scope.

Overmodulation on negative peaks is more likely to result in spurious sidebands than overmodulation in the upward direction because of the sharp break that occurs when the carrier is suddenly cut off and on. The milliammeter in the negative-peak indicator of Fig. 9-48 will show a reading on each overmodulation peak that carries the instantaneous voltage on the plate of the Class C modulated amplifier "below zero" — that is, negative. The rectifier, *V*, cannot conduct so long as the negative half-cycle of audio output voltage is less than the d.c. voltage applied to the r.f. tube.

The inverse peak voltage rating of the rectifier tube must be at least twice the d.c. voltage applied to the plate of the modulated Class C amplifier. The filament transformer likewise must have insulation rated to withstand twice the d.c. plate voltage. Either mercury-vapor or high-vacuum rectifiers can be used. The 15-volt breakdown voltage of the former will introduce a slight error, since the plate voltage must go at least 15 volts negative before the rectifier will ionize, but the error is inconsequential at plate voltages above a few hundred volts.

The effectiveness of the monitor is improved if it indicates at somewhat less than 100-per-cent modulation, as it will then warn of the danger of overmodulation before it actually occurs. It can be adjusted to indicate at any desired modulation percentage by making the meter return to a point on the power-supply bleeder as shown in the alternative diagram. The by-pass condenser, *C*, insures that the full audio voltage appears across the indicator circuit. The modulation percentage at which the system indicates is determined by the ratio of the d.c. voltage between the meter tap and the positive terminal to the total d.c. voltage.

Frequency and Phase Modulation

The primary advantage of frequency or phase modulation over amplitude modulation comes from the fact that noise or "static," whether natural or set up by electrical machines, is fundamentally an amplitude effect. An AM detector responds to noise just as readily as to the desired modulation on a signal. However, if the receiving system responds only to frequency or phase changes and is insensitive to amplitude variations, it will give normal reception of an FM or PM signal but will not receive noise.

This statement, although an oversimplification, conveys the basic idea. In practice it is only partially accurate; the improvement that can be realized by using FM or PM instead of AM depends on the strength of the received signal, the character of the noise, and the way the noise is distributed over the receiver passband. In general, the wider the channel occupied by the signal the better the noise suppression — if the signal strength is above a certain **threshold** value. The wider the channel occupied by the signal, the stronger the signal required to reach the threshold. The noise suppression in the receiver is most effective when the noise is evenly distributed over the receiver passband and least so when the noise appears on one side or the other of the incoming carrier. (The noise itself usually is properly distributed, but misalignment in receiver circuits will cause uneven response over the passband.) The noise suppression also is most marked when the noise is of the "impulse" type, having a high peak amplitude but short duration.

In amateur work, FM and PM have been used not so much because of the possibility of an improved signal-to-noise ratio but because of more-or-less incidental advantages. For example, in the ultrahigh and superhigh frequency ranges some tubes do not lend themselves well to amplitude modulation, but can easily be frequency-modulated. On the lower frequencies FM and PM are often used because they cause less interference than AM in unshielded broadcast receivers in the vicinity.

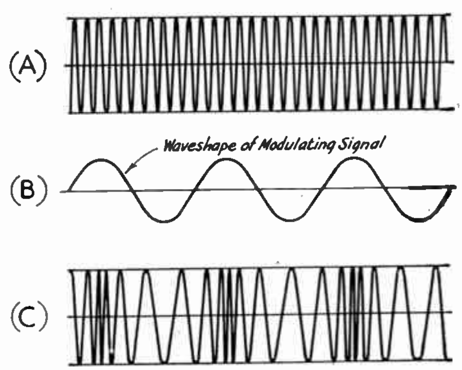


Fig. 9-49 — **Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.**

Frequency Modulation

The fundamental principle of frequency modulation is easy to understand. Suppose we have an oscillator operating at a frequency of, say, 3900 kc. Further suppose that we vary the oscillator tuning control back and forth so that at one extreme the frequency is 3905 kc. and at the other, 3895 kc.; that is, plus and minus 5 kc. on either side of the carrier frequency. Imagine that the tuning is varied back and forth in that fashion at 1000 times per second. Then we are *frequency modulating* the oscillator at an audio frequency of 1000 cycles.

The **frequency deviation** is the maximum change in frequency from the carrier frequency; in this example it is 5 kc. So long as the tuning control is varied between the same two extremes, the frequency deviation is the same no matter how rapidly the control is varied; i.e., no matter what the modulating frequency. In other words, we can make the deviation any reasonable figure we want, whether it is a few hundred cycles or tens of kilocycles, and it is not affected by the modulating frequency.

Fig. 9-49 is a representation of frequency modulation. In the unmodulated carrier each cycle occupies the same time as the preceding one. When a modulating signal is applied, the carrier frequency is increased during one half-cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the drawing by the fact that the r.f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency is proportional to the instantaneous amplitude of the modulating signal — or, to use the analogy above, to the position of the oscillator tuning control — so the frequency deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative. That is, the frequency deviation follows the changes in the amplitude of the modulating signal.

Phase and Frequency

Phase modulation is a little more difficult. To understand the difference between FM and PM it is necessary to appreciate that the frequency of an alternating current is determined by the *rate at which its phase changes.* A current in which the phase changes rapidly has a higher frequency than one in which the phase changes slowly. For example, if the phase moves through 360 degrees in one second the frequency is one cycle per second, but if the phase moves through 1080 degrees in one second (3 $\times$ 360 degrees) there are three complete cycles in one second.

If the phase moves along at a constant speed the frequency also is constant. But if the rate of phase change is speeded up or slowed down there is an accompanyingshift in the frequency. If the speed is increased the frequency becomes higher; if the speed is decreased the frequency becomes lower.

Now suppose we have a transmitter operating at a fixed frequency; a frequency that is unaffected by tuning an amplifier that is a few stages removed from the frequency-controlling oscillator. We cannot change the *frequency*, but we can *shift the phase* of the r.f. current by adjusting the tuning control of the amplifier. We might, for example, shift the phase of the current in that circuit 10 degrees by detuning the tank circuit from resonance. Once the detuning is finished the phase shift is permanent, but there is still just exactly the same number of cycles per second as before — so the frequency is exactly the same as it was in the first place. *But during the time that the phase shift is taking place there is a change in frequency.* If the phase is advanced (moved forward) the frequency increases; if it is retarded (slowed down) the frequency decreases.

This "instantaneous" frequency change would never be noticed in tuning an amplifier tank circuit, because the frequency deviation depends on the *speed* with which the phase is shifted. Any manual adjustment would be too slow to make an observable frequency change. But when the phase is shifted back and forth at an audio-frequency rate the frequency deviation *is* observable, and it is directly proportional to the rate at which the phase is shifted. The rate of phase shift is naturally proportional to the total number of degrees through which the phase is shifted; it is also proportional to the amplitude of the modulating signal (a large signal will shift the phase more, in the same time, than a small signal), and to the frequency of the modulating signal because the phase shift is more rapid the greater the number of times it is shifted per second.

To summarize, then, in FM the carrier frequency deviation is proportional to the amplitude of the modulating signal but not to its frequency. In PM the deviation is proportional to *both* the amplitude and frequency of the modulating signal. Fig. 9-49 is just as representative of PM as it is of FM, because it is impossible to tell the two apart when there is only one modulating frequency.

Modulation Depth

In FM or PM there is no condition that corresponds exactly to overmodulation in AM. "Percentage of modulation" has to be defined a little differently for these systems. Practically, "100-per-cent modulation" is reached when the transmitted signal occupies a channel just equal to the bandwidth for which the *receiver* is designed. If the channel occupied is wider than the receiver can accept, the receiver distorts the signal and the end effect is much the same as overmodulation in AM. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25-per-cent modulation. Until the maximum width is set for the channel, percentage of modulation has no meaning.

In amateur work no specifications have been set up for channel-width except in the case of "narrow-band" FM or PM (frequently abbreviated NFM), where the channel-width is defined as being the same as that of a properly-modulated AM signal. That is, the channel-width for NFM does not exceed twice the highest audio frequency in the modulating signal. NFM transmissions based on an upper audio limit of 3000 cycles therefore should occupy a channel no wider than 6 kc.

FM and PM Sidebands

From the descriptions given above of the fundamentals of frequency and phase modulation, it might be concluded that the channel occupied by the transmission would be no greater than the frequency deviation on each side of the carrier. However, if we applied the same line of reasoning to amplitude modulation we should reach the conclusion that an AM signal takes up no more space than the carrier alone, since only the *amplitude* of the carrier varies. Both conclusions would be wrong; the fact is that both FM and PM set up sidebands, just as AM does. In the case of FM and PM, single-tone modulation sets up a whole series of pairs of sidebands that are harmonically related to the modulating frequency, whereas in AM there is only one pair of sidebands.

The number of "extra" sidebands that occur in FM and PM depends on the relationship between the modulating frequency and the carrier frequency deviation. The ratio between the frequency deviation, in cycles per second, and the modulating frequency, also in cycles per second, is called the **modulation index.** That is,

$$\textit{Modulation index} = \frac{\textit{Carrier frequency deviation}}{\textit{Modulating frequency}}$$

Example: The maximum frequency deviation in an FM transmitter is 3000 cycles either side of the carrier frequency. The modulation index when the modulating frequency is 1000 cycles is

$$\text{Modulation index} = \frac{3000}{1000} = 3$$

At the same deviation with 3000-cycle modulation the index would be 1; at 100 cycles it would be 30, and so on.

The modulation index is also equal to the phase shift (in radians). In PM the index is constant regardless of the modulating frequency; in FM it varies with the modulating frequency, as shown in the previous example. To identify any particular FM system, the limiting modulation index — that is, the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used — is called the **deviation ratio.**

Fig. 9-50 shows how the amplitudes of the carrier and the various sidebands vary with the modulation index. This is for single-tone modulation; the first sideband (actually a pair, one above and one below the carrier) is displaced from the carrier by an amount equal to the modulating frequency, the second is twice the modulating frequency away from the carrier, and so on. For example, if the modulating frequency is 2000 cycles and the carrier frequency is 29,500 kc., the first sideband pair is at 29,498 kc. and 29,502 kc., the second pair

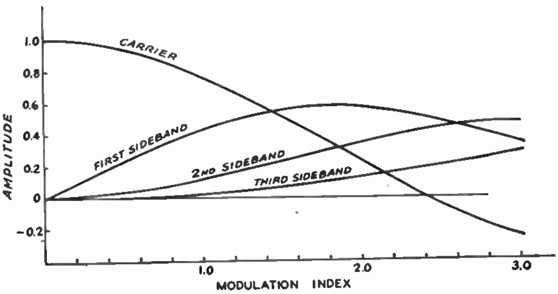


Fig. 9-50 — How the amplitude of the pairs of sidebands varies with the modulation index in an FM or PM signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

is at 29,496 kc. and 29,504 kc., the third at 29,494 kc. and 29,506 kc., etc. The amplitudes of these sidebands depend on the modulation index, not on the frequency deviation. In AM, regardless of the percentage of modulation (so long as it does not exceed 100 per cent) the sidebands would appear *only* at 29,498 and 29,502 kc. under the same conditions.

Note that, as shown by Fig. 9-50, the carrier strength varies with the modulation index. (In amplitude modulation the carrier strength is constant; only the sideband amplitude varies.) At a modulation index of approximately 2.4 the carrier disappears entirely and then becomes "negative" at a higher index. This simply means that its phase is reversed as compared to the phase without modulation. In FM and PM the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index. In AM the sideband power is supplied by the modulator in the case of plate modulation, and by changing the power input and efficiency in the case of grid-bias modulation.

Fig. 9-50 can be carried out to considerably-higher modulation indexes, in which case it will be found that more and more additional sidebands are set up and that the carrier goes through several "zeros" and reversals in phase.

Frequency Multiplication

In amplitude modulation it is customary amateur practice to apply the modulation to the final r.f. stage of the transmitter. If a lower-level stage is modulated, a special type of operation is necessary in the following r.f. stages to pass the modulation envelope without distortion. These "linear" amplifiers are rather difficult to adjust properly and must be operated at low plate efficiency. Consequently, the simplest and most economical transmitter design results when the final stage is modulated.

In frequency or phase modulation there is no change in the amplitude of the signal with modulation. Consequently, an FM or PM signal can be amplified by an ordinary Class C amplifier without distortion. The modulation can take place in a very low-level stage and the signal can then be amplified by either frequency multipliers or straight amplifiers. In fact, this is the usual practice. The audio power required for modulating an FM or PM transmitter is negligible.

If the modulated signal is passed through one or more frequency multipliers, the modulation index is multiplied by the same factor that the carrier frequency is multiplied. For example, suppose that the controlling oscillator in the transmitter is on 3.5 Mc. and the final output is on 28 Mc. The total frequency multiplication is 8 times, and any FM or PM applied to the oscillator will likewise be multiplied by 8 in the 28-Mc. output. If the frequency deviation is 500 cycles at 3.5 Mc., it will be 4000 cycles at 28 Mc.

Frequency multiplication offers a means for obtaining practically any desired amount of frequency deviation, whether or not the modulator itself is capable of giving that much deviation without distortion. Also, if the same oscillator is modulated in a transmitter that operates on several bands, the frequency deviation is different on each band. The amount of frequency multiplication after modulation always must be taken into account in determining whether or not the final frequency deviation is the desired value on a given band.

● NARROW-BAND FM OR PM

Where FM or PM is used in crowded 'phone bands (particularly below 29 Mc.) it is of utmost importance that the transmissions should occupy no greater channel-width than would be occupied by an AM signal. It is evident from Fig. 9-50 that this requirement can be met only by using a relatively small modulation index. It must be realized that the higher-order sidebands always are present, even at very small indexes. It is therefore necessary to set an arbitrary level above which the extra sidebands should not go. If the modulation index (with single-tone modulation) does not exceed about 0.6 the most important extra sideband, the second, will be at least 20 db. below the unmodulated carrier level, and this should represent an effective channel-width about equivalent to that of an AM signal. In the case of speech, a somewhat higher modulation index can be used. This is because the energy distribution in a complex wave is such that the modulation index for any one frequency component is reduced, as compared to the index with a sine wave having the same peak amplitude as the voice wave.

The chief advantage of narrow-band FM or PM for frequencies below 30 Mc. is that it eliminates or reduces certain types of interference to broadcast reception. Also, the modulating equipment is relatively simple and inexpensive. However, assuming the same unmodulated carrier power in all cases, narrow-band FM or PM is not as effective as AM. As shown by Fig. 9-50, at an index of 0.6 the amplitude of the first sideband is about 25 per cent of the unmodulated-carrier amplitude; this compares with a sideband amplitude of 50 per cent in the case of a 100-per-cent modulated AM transmitter. In other words, so far as effectiveness is concerned, a narrow-band FM or PM transmitter is about equivalent to a 100-per-cent modulated AM transmitter operating at one-fourth the power input. This assumes that the receiving system is equally efficient in all cases. This very often is not true on the low-frequency bands, since communications receivers are designed primarily for AM reception.

Methods of Frequency and Phase Modulation

FREQUENCY MODULATION

The simplest and most satisfactory device for amateur FM is the reactance modulator. This is 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 capacitance. Fig. 9-51 is a representative circuit. The control-grid circuit of the 6L7 tube is connected across the small capacitance, C_1, which is in series with the resistor, R_1, across the oscillator tank circuit. Any type of oscillator circuit may be used. The resistance of R_1 is made large compared to the reactance of C_1, so the r.f. current through R_1C_1 will be practically in phase 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. The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage, 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. The frequency increases in proportion to the amplitude of the lagging plate current of the modulator. The value of plate current is determined by the voltage on the No. 3 grid of the 6L7; hence the oscillator frequency will vary when an audio signal voltage is applied to the No. 3 grid.

If, on the other hand, C_1 and R_1 are interchanged 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.

A circuit using a receiving-type r.f. pentode of the high-transconductance type, such as the 6SG7, is shown in Fig. 9-52. In this case, both r.f. and audio are applied to the control grid. The audio voltage, introduced through a radio-frequency choke, *RFC*, varies the transconductance of the tube and thereby varies the r.f. plate current. The capacitance C_8 corresponds to C_1 in Fig. 9-51; it represents the input capacitance of the tube. (It is possible, also, to omit C_1 from Fig. 9-51 and depend upon the input capacitance of the 6L7 instead; the only disadvantage is that there is then no control over the modulator sensitivity. Likewise, a 3–30-$\mu\mu$fd. trimmer condenser can be connected at C_8 in Fig. 9-52 to permit controlling the sensitivity.) In Fig. 9-52 the r.f. circuit is

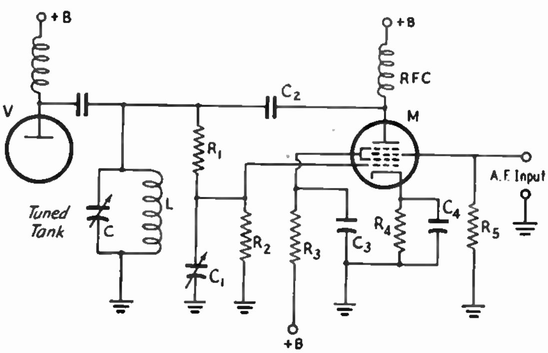


Fig. 9-51 — **Reactance-modulator circuit using a 6L7 tube.**

C — R.f. tank capacitance. C_1 — 3-30 $\mu\mu$fd. C_2 — 220 $\mu\mu$fd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 10-μfd. electrolytic in parallel with 0.01-μfd. paper.
L — R.f. tank inductance.
R_1 — 47,000 ohms.
R_2, R_5 — 0.47 megohm.
R_3 — 33,000 ohms.
R_4 — 330 ohms.
RFC — 2.5 mh.

series-fed, which is advantageous if the r.f. tube and the modulator can be operated at the same plate voltage. The use of different plate voltages on the two tubes calls for the parallel-feed arrangement shown in Fig. 9-51.

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 6.5 Mc. and the output frequency is to be 52 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 8000 cycles at the output frequency.

A reactance modulator can be connected to a crystal oscillator as well as to the self-controlled type. However, the resulting signal is more phase-modulated than it is frequency-modulated, for the reason that the frequency deviation that can be secured by varying the tuning of a crystal oscillator is quite small.

Design Considerations

The sensitivity of the modulator (frequency change per unit change in grid voltage) depends on the transconductance of the modulator tube. It 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, it is desirable to use the highest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 9-52 is used in connection with a 7-Mc. oscillator, a linear deviation of 1500 cycles above and below the carrier frequency can be secured when the oscillator tank capacitance is approximately 200 μμfd. A peak a.f. input of two volts is required for full deviation.

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. Therefore 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.

Speech Amplification

The speech amplifier preceding the modulator follows ordinary design, 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 more than suffice for crystal microphones.

PHASE MODULATION

The same type of reactance-tube circuit that is used to vary the tuning of the oscillator tank in FM can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for PM. Hence the modulator circuits of Figs. 9-51 and 9-52 can be used for PM if the reactance tube works on an amplifier tank instead of directly on a self-controlled oscillator.

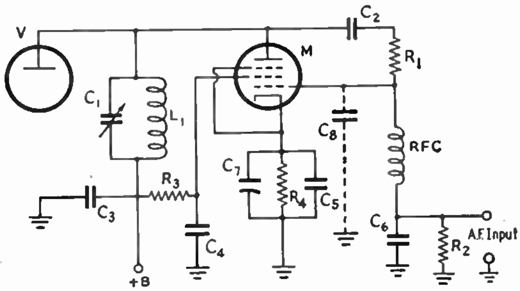


Fig. 9-52 — Reactance modulator using a high-transconductance pentode (6SG7, 6AG7, etc.).
C_1 — R.f. tank capacitance (see text).
C_2, C_3 — 0.001-μfd. mica.
C_4, C_5, C_6 — 0.0047-μfd. mica.
C_7 — 10-μfd. electrolytic.
C_8 — Tube input capacitance (see text).
R_1, R_2 — 0.47 megohm.
R_3 — Screen dropping resistor; select to give proper screen voltage on type of modulator tube used.
R_4 — Cathode bias resistor; select as in case of R_3.
L_1 — R.f. tank inductance.
RFC — 2.5-mh. r.f. choke.

The phase shift that occurs when a circuit is detuned from resonance depends on the amount of detuning and the Q of the circuit. The higher the Q, the smaller the amount of detuning needed to secure a given number of degrees of phase shift. If the Q is at least 10, the relationship between phase shift and detuning (in kilocycles either side of the resonant frequency) will be substantially linear over a range of about 25 degrees. From the standpoint of modulator sensitivity, the Q of the tuned circuit on which the modulator operates should be as high as possible. On the other hand, the effective Q of the circuit will not be very high if the amplifier is delivering power to a load, since the load resistance reduces the Q. There must therefore be a compromise between modulator sensitivity and r.f. power output from the modulated amplifier. An optimum figure for Q appears to be about 20; this allows reasonable loading of the modulated amplifier and the necessary tuning variation can be secured from a reactance modulator without difficulty.

It is advisable to modulate at a very low power level — preferably in a transmitter stage where receiving-type tubes are used. A practical phase-modulator unit is described later in this chapter.

Reactance-Modulator Unit for Narrow-Band FM

The FM speech-amplifier and modulator unit shown in Figs. 9-53 and 9-54 uses a pentode reactance modulator in a circuit which is basically that of Fig. 9-52. It differs only in the detail that the audio signal is applied to the control grid in parallel with the r.f. voltage from the oscillator, instead of the series-feed arrangement shown in Fig. 9-52. Because of the parallel feed, resistor R_4 is incorporated in the circuit to prevent r.f. from appearing in the plate circuit of the speech-amplifier tube.

The unit uses miniature tubes for the sake of making a compact assembly that can be mounted in any convenient spot near the VFO tuned circuit. In Fig. 9-53 it is shown mounted on the outside of the VFO case. When this type of mounting is used the unit should be placed so that the lead between the VFO tuned circuit and the modulator is as short as possible. If there is space available, it is preferable to mount the unit inside the VFO cabinet.

The chassis for the unit is 4 inches long by 2 inches wide, and has a mounting lip 2 inches deep. As shown in the photographs, it is formed from a piece of aluminum with the edges turned over to stiffen it. The various components are easily accommodated underneath. The r.f. leads should be kept short and separated as much as possible from the audio and power-supply wiring.

Filament and plate power can usually be taken from the VFO supply, since the total plate current is only a few milliamperes. Filament current required is 0.6 amp. The microphone input is carried through a shielded lead to the unit, thus the microphone connector can be placed in any convenient location on the VFO unit itself. Once the proper setting of the gain control is found it need not be touched again, so screwdriver adjustment is quite adequate.

The adjustment of reactance modulators is discussed in a later section in this chapter.

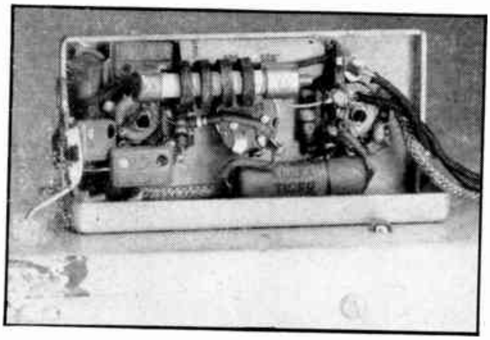


Fig. 9-54 — Underneath the modulator unit. The r.f. connection to the VFO goes through the feed-through bushing at the left.

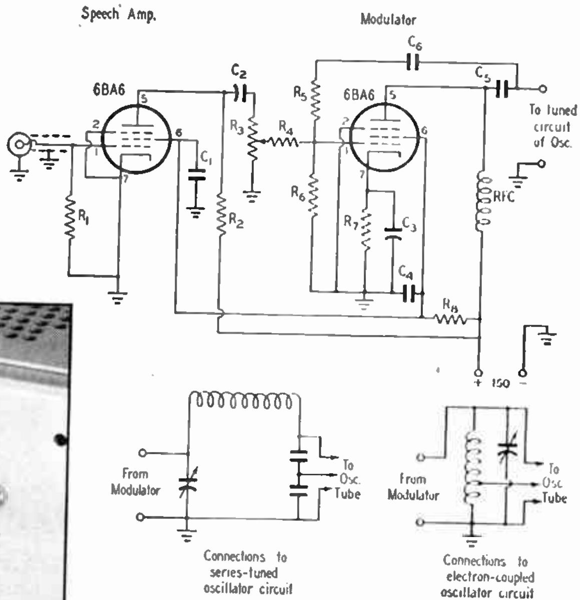


Fig. 9-55 — Circuit diagram of the narrow-band FM modulator unit.

C_1 — 680-μμfd. mica.
C_2, C_4 — 0.01-μfd. paper, 400 volts.
C_3 — 0.025-μfd. paper, 200 volts.
C_5, C_6 — 47-μμfd. mica.
R_1 — 1.2 megohms, ½ watt.
R_2, R_8 — 0.22 megohm, ½ watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 0.1 megohm, ½ watt.
R_5 — 10,000 ohms, ½ watt.
R_6 — 0.47 megohm, ½ watt.
R_7 — 390 ohms, ½ watt.
RFC — 2.5-mh. r.f. choke.

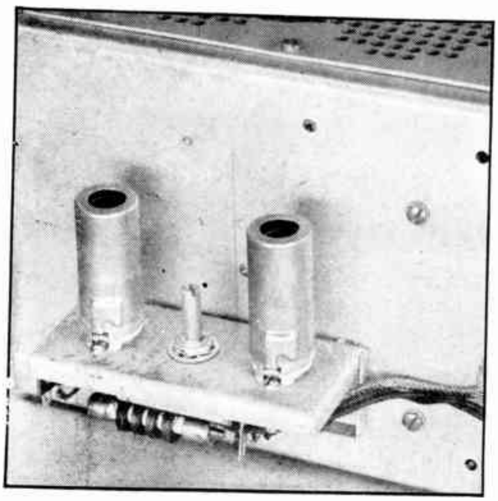


Fig. 9-53 — Miniature reactance modulator that can be used with any VFO. The shielded lead is for microphone input; the other two wires bring in filament and plate supply.

A Narrow-Band PM Exciter Unit

The unit shown in Figs. 9-56, 9-57 and 9-58 will deliver from 10 to 20 watts of phase-modulated output, depending upon the plate voltage used, and it can be used to replace a crystal oscillator of that power level in any existing transmitter. It can also be used with an existing VFO to obtain a phase-modulated signal. A low-pass filter is incorporated to limit the modulation frequencies to those below 3000 cycles, thus making it easier to comply with the regulations on narrow-band FM and PM.

As can be seen in the wiring diagram, Fig. 9-56, a 6J5 Pierce-type oscillator circuit is used to excite a 6SK7 r.f. amplifier. If VFO control is used, the 6J5 can be removed from its socket and the VFO output introduced at J_1. To accommodate various output levels from the VFO, a gain control, R_3, is included in the cathode circuit of the 6SK7 amplifier. The plate circuit of the 6SK7 amplifier is reactance-modulated (to give the PM signal) by a 6SG7 reactance modulator, and the output of the 6SK7 drives a 2E26 r.f. amplifier. With 500 volts on the plate of the 2E26, 15 watts output can be obtained. The plate voltage can be raised to 600 if more output is required. The microphone input at J_4 is amplified through a 6SJ7 and a 6J5, and the low-pass filter is connected between the 6J5 and the 6SG7 modulator tube. The degree of modulation is controlled by the setting of R_{15}, the audio gain control.

Construction

The unit is built on a 7 × 12 × 3-inch chassis, and the location of the components can be seen from Figs. 9-57 and 9-58. A shield can (Millen 80016) is used over L_1, to avoid regeneration, and a small shield extends up 1 inch around the 2E26 for the same reason. No special care is necessary in wiring the unit, except that C_5 or C_6 should be mounted across the 6SK7 socket, to shield the grid pin from the plate pin, and the audio-circuit wiring should be kept away from the r.f. circuits. The r.f. input and output, from J_1 and J_3, is most conveniently run in short pieces of RG-58/U

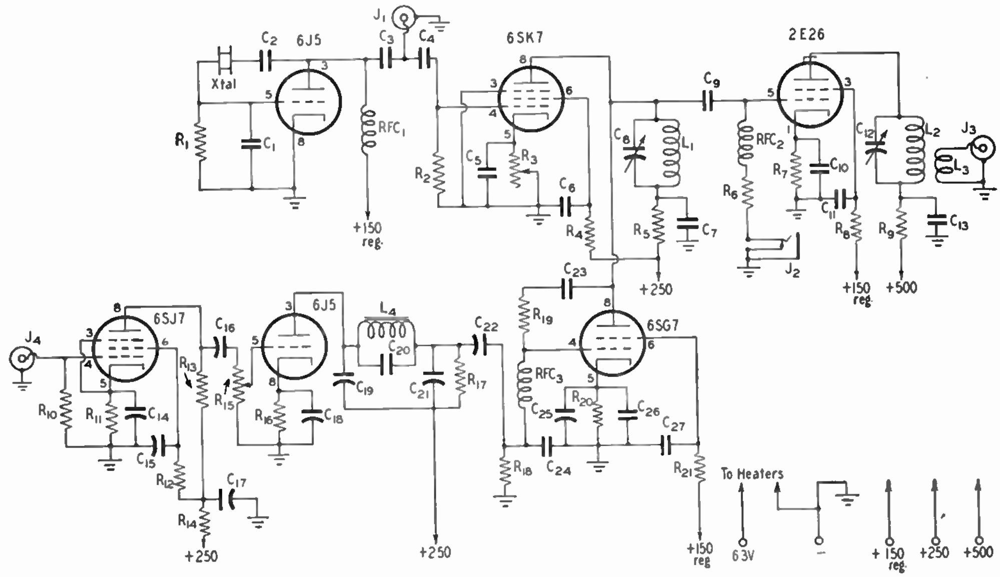


Fig. 9-56 — Wiring diagram of the narrow-band phase-modulation unit.

C_1 — 22-μμfd. mica.
C_2, C_5, C_6, C_7, C_{10}, C_{11}, C_{13}, C_{23}, C_{24}, C_{26}, C_{27} — 0.001-μfd. mica.
C_3, C_4, C_9 — 100-μμfd. mica.
C_8 — 50-μμfd. midget variable (Millen 21050).
C_{12} — 50-μμfd. variable (Millen 22050).
C_{14}, C_{18}, C_{25} — 10-μfd. electrolytic, 25 volts.
C_{15} — 0.1-μfd. paper, 200 volts.
C_{16} — 0.05-μfd. 400-volt paper.
C_{17} — 8-μfd. electrolytic, 150 volts.
C_{19}, C_{21}, C_{22} — 0.01-μfd. 400-volt paper.
C_{20} — 0.006-μfd. 200-volt paper.
R_1, R_2, R_4, R_{14} — 47,000 ohms.
R_3 — 5000-ohm potentiometer, wire-wound.
R_5, R_8, R_{21} — 470 ohms.
R_6 — 12,000 ohms.
R_7 — 330 ohms, 1 watt.
R_9 — 100 ohms.
R_{10}, R_{12} — 1.0 megohm.
R_{11} — 820 ohms.
R_{13} — 0.22 megohm.
R_{15} — 0.25-megohm volume control.
R_{16} — 1000 ohms.
R_{17} — 3900 ohms.
R_{18} — 0.1 megohm.
R_{19} — 22,000 ohms.
R_{20} — 220 ohms.
Resistors ½ watt unless otherwise specified.
L_1, L_2 — 3.5 Mc.: 40 turns No. 26 e., 1" d., close-wound.
L_3 — 9 turns No. 22 enam., close-wound next to cold end of L_2.
L_4 — 0.25 henry (Millen 34400-250).
J_1, J_3 — Cable connector (Jones S-201).
J_2 — Closed-circuit midget jack.
J_4 — Microphone-cable connector (Amphenol PC1M).
RFC_1, RFC_2 — 2.5 mh.

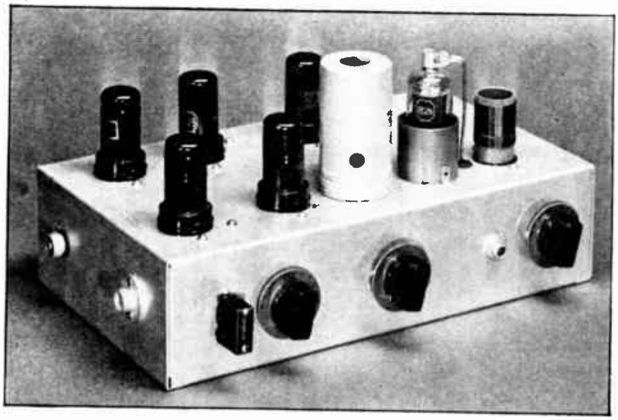


Fig. 9-57 — The narrow-band PM exciter is built in simple style. The r.f. stages are mounted along the front, and from left to right the tubes are 6J5, 6SK7 and 2E26. Note the shield can for the 6SK7 plate coil. The tubes at the rear, from left to right, are 6SJ7, 6J5 and 6SG7.

cable, but no other shielding should be necessary. The excitation control, R_3, was made to be adjusted by a screwdriver, since ordinarily there is little need for changing the setting once it has been established.

The power-supply requirements are 500 volts at 60 ma. (or whatever is required for input to the 2E26 plate), 250 volts at 25 ma., 150 volts regulated (by a VR-150) at about 12 ma., and 6.3 volts a.c. at 2.3 amperes. If the 150-volt supply is not regulated, C_{27} should be shunted with an 8-μfd. electrolytic condenser, to avoid audio degeneration.

Tuning

To set the unit in operation after the power supply has been connected, allow the heaters to warm up and apply plate power. If the crystal oscillator is being used, the circuit will work without adjustment and, with R_3 set at minimum resistance, a meter plugged in at J_2 should read 3 or 4 ma. when C_8 is tuned to resonate the 6SK7 plate circuit. With VFO input, the excitation may be more than this. The 2E26 plate circuit can be tuned using a milliammeter in the 500-volt lead or by reading grid current in the following stage. When the two r.f. circuits have been resonated, the gain of the 6SK7 stage should be reduced, by increasing the resistance at R_3, until the grid current is between 2.5 and 3 ma. By checking the signal in a receiver (at reduced gain and with no antenna connected, as described in a later section), the proper setting of R_{15} for the microphone in use can be found. The audio gain control, R_{15}, will normally be set near maximum for work on 3.9 Mc., but it will be necessary to reduce the setting for 14- and 29-Mc. operation. Working with a 3.9-Mc. crystal or VFO, best results on 75-meter 'phone will be obtained when the receiving operator uses his crystal filter for pure PM reception.

If desired, the 2E26 may be operated as a frequency doubler to 7 Mc., by substituting a coil at L_2 having half the number of turns in the same winding length. Still another alternative is to substitute such a coil for L_1 in the plate circuit of the 6SK7, and double again in the 2E26 to 14 Mc. In such case L_2 should have one-fourth the number of turns given under Fig. 9-56, in the same winding length. With this second alternative the modulation is applied at 7 Mc., which will reduce the phase shift, at the final output frequency, to one-half the value obtained when the modulation takes place at 3.5 Mc.

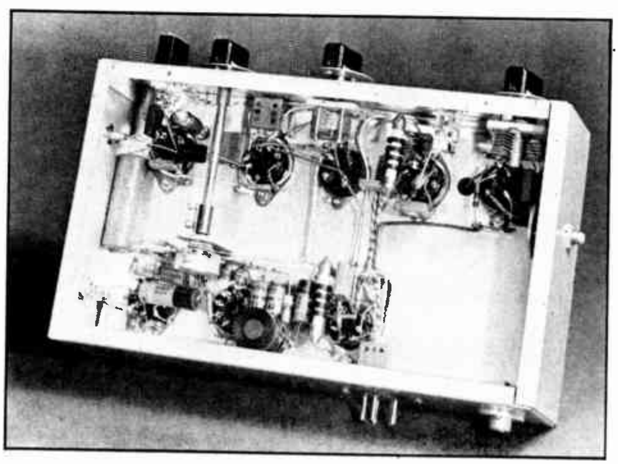


Fig. 9-58 — The r.f. and audio wiring of the PM exciter are kept separated as much as possible. To carry out this scheme, the audio gain control is mounted on a small bracket, and a long shaft is brought out to the panel.

Checking FM and PM Transmitters

Accurate checking of the operation of an FM or PM transmitter requires different methods than the corresponding checks on an AM set. This is because the common forms of measuring devices either indicate amplitude variations only (a d.c. milliammeter, for example), or because their indications are most easily interpreted in terms of amplitude. There is no simple instrument that indicates frequency deviation in a modulated signal directly.

However, there is one favorable feature in FM or PM checking. The modulation takes place at a very low level and the stages following the one that is modulated do not affect the linearity of modulation so long as they are properly tuned. Therefore the modulation may be checked *without putting the transmitter on the air*, or even on a dummy antenna. The power is simply cut off the amplifiers following the modulated stage. This not only avoids unnecessary interference to other stations during testing periods, but also keeps the signal at such a low level that it may be observed quite easily on the station receiver. A good receiver with a crystal filter is an essential part of the checking equipment of an FM or PM transmitter, particularly for narrow-band FM or PM.

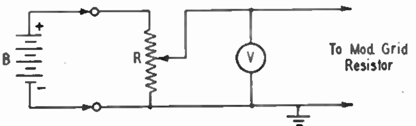


Fig. 9-59 — **D.c. method of checking frequency deviation of a reactance-tube modulated oscillator. A 500- or 1000-ohm potentiometer may be used at *R*.**

The quantities to be checked in an FM or PM transmitter are the linearity and frequency deviation. Because of the essential difference between FM and PM the methods of checking differ in detail.

Reactance-Tube FM

It was explained earlier that in FM the frequency deviation is the same at any audio modulation frequency if the audio signal amplitude does not vary. Since this is true at *any* audio frequency it is true at zero frequency. Consequently it is possible to calibrate a reactance modulator by applying an adjustable d.c. voltage to the modulator grid and noting the change in oscillator frequency as the voltage is varied. A suitable circuit for applying the adjustable voltage is shown in Fig. 9-59. The battery, *B*, should have a voltage of 3 to 6 volts (two or more dry cells in series). The arrows indicate clip connections so that the battery polarity can be reversed.

The oscillator frequency deviation should be measured by using a receiver in conjunction with an accurately-calibrated frequency meter, or by any means that will permit accurate measurement of frequency differences of a few hundred cycles. One simple method is to tune in the oscillator on the receiver (disconnect the receiving antenna, if necessary, to keep the signal strength well below the overload point) and then set the receiver b.f.o. to zero beat. Then increase the d.c. voltage applied to the modulator grid from zero in steps of about ½ volt and note the beat frequency at each change. Then reverse the battery terminals and repeat. The frequency of the beat note may be measured by comparison with a calibrated audio-frequency oscillator, or by comparison with a piano or other musical instrument (see Chapter Twenty-Four for frequencies of musical tones). Note that with the battery polarity positive with respect to ground the radio frequency will move in one direction when the voltage is increased, and in the other direction when the battery terminals are reversed. When a number of readings has been taken a curve may be plotted to show the relationship between grid voltage and frequency deviation.

A sample curve is shown in Fig. 9-60. The usable portion of the curve is the center part which is essentially a straight line. The bending at the ends indicates that the modulator is no longer linear; this departure from linearity will cause harmonic distortion and will broaden the channel occupied by the signal. In the example, the characteristic is linear 1.5 kc. on either side of the center or carrier frequency. This is the maximum deviation permissible at the frequency at which the measurement is made. At the final output frequency the deviation will be multiplied by the same number of times that the measurement frequency is multiplied. This must be kept in mind when the check is made at a frequency that differs from the output frequency.

A good modulation indicator is a "magic-eye" tube such as the 6E5. This should be connected across the grid resistor of the reactance modulator as shown in Fig. 9-61. Note its deflection (using the d.c. voltage method as in

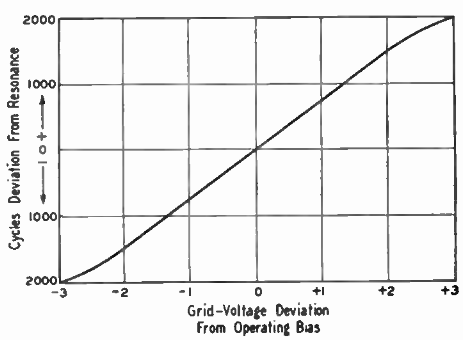


Fig. 9-60 — **A typical curve of frequency deviation *vs.* modulator grid voltage.**

Fig. 9-59) at the maximum deviation to be used. This deflection represents "100-per-cent modulation" and with speech input the gain should be kept at the point where it is just reached on voice peaks. If the transmitter is used on more than one band, the gain control should be marked at the proper setting for each band, because the signal amplitude that gives the correct deviation on one band will be either too great or too small on another. For narrow-band FM the proper deviation is approximately 2000 cycles (based on an upper a.f. limit of 3000 cycles and a deviation ratio of 0.7) at the final *output* frequency. If the output frequency is in the 29-Mc. band and the oscillator is on 7 Mc., the deviation at the *oscillator* frequency should not exceed 2000/4, or 500 cycles.

Checking with a Crystal-Filter Receiver

With PM the d.c. method of checking just described cannot be used, because the frequency deviation at zero frequency also is zero. For narrow-band PM it is necessary to check the actual channel-width occupied by the transmission. (The same method also can be used to check FM.) For this purpose it is necessary to have a crystal-filter receiver and an a.f. oscillator that generates a 3000-cycle sine wave.

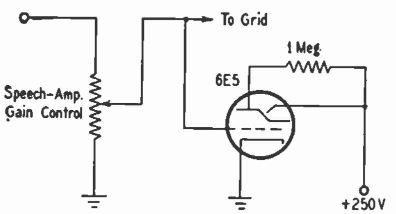


Fig. 9-61 — 6E5 modulation indicator for FM or PM modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.

Keeping the signal intensity in the receiver at a medium level, tune in the carrier at the *output* frequency. Do not use the a.v.c. Switch on the beat oscillator, and set the crystal filter at its sharpest position. Peak the signal on the crystal and adjust the b.f.o. for any convenient beat note. Then apply the 3000-cycle tone to the speech amplifier (use the connections shown in Fig. 9-42 to avoid overloading) and increase the audio gain until there is a small amount of modulation. Tuning the receiver on either side of the carrier frequency will show the presence of sidebands 3 kc. from the carrier on both sides. With low audio input, these two should be the *only* sidebands detectable.

Now increase the audio gain and tune the receiver over a range of about 10 kc. on both sides of the carrier. When the gain becomes high enough, a second set of sidebands spaced 6 kc. on either side of the carrier will be detected. The signal amplitude at which these sidebands become detectable is the maximum speech amplitude that should be used. If the 6E5 modulation indicator is incorporated in the modulator, its deflection with the 3000-cycle tone will be the "100-per-cent modulation" deflection for speech.

When this method of checking is used with a reactance-tube modulated FM (not PM) transmitter, the linearity of the system can be checked by observing the *carrier* as the a.f. gain is slowly increased. The beat-note frequency will stay constant so long as the modulator is linear, but nonlinearity will be accompanied by a shift in the average carrier frequency that will cause the beat note to change in frequency. If such a shift occurs at the same time that the 6-kc. sidebands appear, the extra sidebands may be caused by modulator distortion rather than by an excessive modulation index. This means that the modulator is not able to shift the frequency over a wide-enough range. The 6-kc. sidebands should appear *before* there is any shift in the carrier frequency.

R.F. Amplifiers

The r.f. stages in the transmitter that follow the modulated stage may be designed and adjusted as in ordinary operation. In fact, there are no special requirements to be met except that all tank circuits should be carefully tuned to resonance (to prevent unwanted r.f. phase shifts that might interact with the modulation and thereby introduce hum, noise and distortion). In neutralized stages, the neutralization should be as exact as possible, also to minimize unwanted phase shifts. With FM and PM, all r.f. stages in the transmitter can be operated at the manufacturer's maximum c.w.-telegraphy ratings, since the average power input does not vary with modulation as it does in AM 'phone operation.

The output of the transmitter should be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated-carrier value when the transmitter is modulated. If 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.

Amplitude modulation accompanying FM or PM is just as much to be avoided as frequency or phase modulation that accompanies AM. A mixture of AM with either of the other two systems results in the generation of spurious sidebands and consequent widening of the channel. If the presence of AM is indicated by variation of antenna current with modulation, the cause is almost certain to be nonlinearity in the modulator. In very wide-band FM the selectivity of the transmitter tank circuits may cause the amplitude to decrease at high deviations, but this is not likely to occur on amateur frequencies at which wide-band FM would be used.

Single-Sideband Transmission

The most recent development in amateur radiotelephony is the introduction of practical single-sideband suppressed-carrier (s.s.s.c.) transmission. This system has tremendous potentialities for increasing the effectiveness of 'phone transmission and for reducing interference. Because only one of the two sidebands normally produced in modulation is transmitted, the channel width is immediately cut in half. However, when only one sideband is transmitted the carrier — which is essential in double-sideband transmission — no longer is necessary; it can be supplied without difficulty at the receiver. With the carrier eliminated there is a great saving in power at the transmitter — or, from another viewpoint, a great increase in effective power output. Assuming that the same final-amplifier tube or tubes are used either for normal AM or for s.s.s.c., it is possible to demonstrate that use of s.s.s.c. gives an effective gain of at least 9 db. over AM — equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that wrecks so much communication in congested 'phone bands.

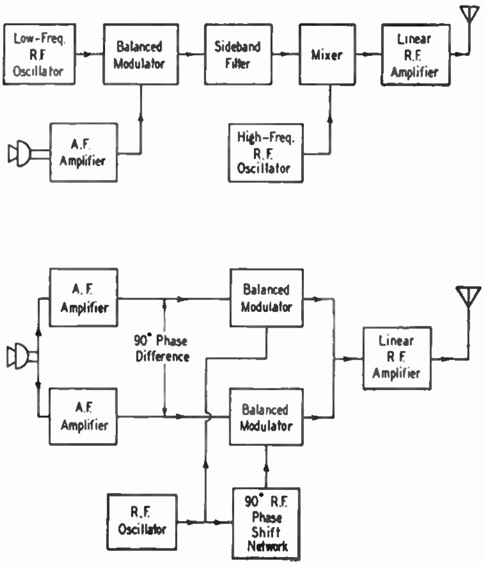


Fig. 9-62 — Two basic systems for generating single-sideband suppressed-carrier signals.

Two basic systems for generating s.s.s.c. signals are shown in Fig. 9-62. One involves the use of a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Filters having such characteristics can only be constructed for relatively low frequencies, so it is common practice to begin operations at about 10 kc. The oscillator output is combined with the audio output of a speech amplifier in a balanced modulator — one in which the carrier is "neutralized" out, leaving only the two sidebands. One of the sidebands is passed by the filter and the other rejected, so that a single-sideband signal is fed to the mixer. The s.s.s.c. signal is there mixed with the output of a high-frequency r.f. oscillator to produce the desired output frequency. For additional amplification a linear r.f. amplifier (Class A or Class B) must be used.

The second system is based on the vector relationships between the carrier and sidebands in a modulated signal. As shown in the diagram, the audio signal is split into two components that are identical except for a phase difference of 90 degrees. The output of the r.f. oscillator (which may be at the final output frequency, if desired) is likewise split into two separate components having a 90-degree phase difference. One r.f. and one audio component are combined in each of two separate modulators, and the outputs of the two modulators are then combined. In this process one sideband is balanced out and the other is accentuated.

Further details and recent developments in amateur equipment for single-sideband work will be found in the following *QST* references:

Goodman, "What Is Single-Sideband Telephony?" *QST*, January, 1948.

Villard, "Single-Sideband Operating Tests," *QST*, January, 1948.

Nichols, "A Single-Sideband Transmitter for Amateur Operation," *QST*, January, 1948.

Norgaard, "What About Single Sideband?" *QST*, May, 1948.

Norgaard, "A New Approach to Single Sideband," *QST*, June, 1948.

Dawley, "An S.S.S.C. Transmitter Adapter," *QST*, July, 1948.

Villard, "A Simple Single-Sideband Transmitter," *QST*, November, 1948.

Goodman, "The Basic 'Phone Exciter," *QST*, January, 1949.

Antennas and Transmission Lines

The radio-frequency power that is generated by a transmitter serves a useful purpose only when it is radiated out into space in the form of electromagnetic waves. It is the antenna's job to convert the power into radio waves as efficiently as possible, and to direct the waves where they will do the most good in communication. To do so, the antenna usually must be located well above the ground and kept as far as possible from buildings, trees, and other objects that might absorb energy. This raises a problem, because by some means or another the power that is generated inside the station, in the transmitter, must be conveyed to the antenna. The usual means is a **transmission line.**

There is thus a natural association between antennas and transmission lines — an association that has frequently led to the quite mistaken belief that an antenna fed by a particular type of transmission line is a better (or worse) radiator than exactly the same type of antenna fed by a different type of transmission line. The fact is that a transmission line can be used to carry power to any sort of device — not just an antenna — capable of receiving it. Nor does the antenna care by what means it gets the power; the amount it receives will be radiated just as well no matter by what system it was conveyed to the antenna.

While it would be dangerous to carry the comparison very far, there are nevertheless some similarities between transmission lines at radio frequencies and the lines used for carrying 60-cycle power from the generating station to the consumer. Some lines are best adapted to carrying power at high voltage and relatively low current; the reverse is true of others. Like a lot of other electrical devices, some antennas want a relatively large current at low voltage, while others want high voltage at low current. We connect a 115-volt lamp, for example, to a 115-volt power line, but if we have a 6-volt lamp and want to run it from the 115-volt line we have to use a transformer to reduce the voltage. Similarly, if an antenna wants high current at low voltage (low impedance) and the transmission line is of a type that is best adapted to carrying power at low current and high voltage (high impedance), we need a transforming device comparable to the transformer used with the 6-volt lamp.

Fig. 10-1 — The principal elements in the system connecting the transmitter and antenna.

The power company's generators usually do not generate the voltage that is wanted on the power line, so an appropriate transformer is connected between the generator and the line. Similarly, the voltage generated in the tank circuit of the final amplifier in the transmitter usually has to be transformed to a value that "fits" the transmission line used. At radio frequencies, it is more convenient to talk in terms of impedance rather than voltage, so we speak of "impedance transformations" rather than "voltage transformations." In general, we have a complete system like that shown in block form in Fig. 10-1. Perhaps this looks complicated, but the power-line analogy should help make it understandable. The equipment itself is not particularly complex, and seems even less so when the underlying necessity for it is appreciated.

Transmission Lines

At power-line frequencies — and even at rather high radio frequencies when we are dealing with tuned circuits that are physically rather small — it is habitual to think of current as flowing "around" the circuit. In a series circuit, for example, it is assumed that the current has the same value at every point in the circuit. Indeed, all the explanations of circuit action in Chapter Two are based on this assumption.

The assumption can be true only if electrical and magnetic effects take place instantaneously all around the circuit. The fact is, though, that the action is *not* instantaneous. The fastest that an electromagnetic field can travel is 300,000,000 meters, or 186,000 miles, per second. This is a tremendous speed, and is so great that in many circuits the action *appears* to be instantaneous. When that is so we can ignore the fact that it takes a certain amount of time for an electrical effect that occurs at one point to be felt at another a short distance away.

But there are other circuits in which time becomes an all-important factor. The transmission lines used to carry radio-frequency power are typical circuits in which time cannot be neglected.

CURRENT FLOW IN LONG LINES

Suppose we have a battery connected to a pair of parallel wires that extends to a very great distance, as in Fig. 10-2. At the moment

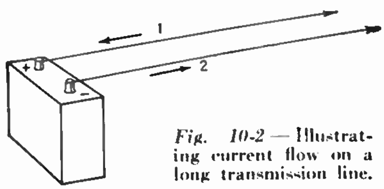


Fig. 10-2 — Illustrating current flow on a long transmission line.

the battery is connected to the wires, electrons in wire No. 1 near the positive terminal of the battery will be attracted to the battery, and the same number of electrons in wire No. 2, near the battery terminal, will be repelled outward along the wire. The directions are shown by the arrows. Thus a current flows in both wires at the instant the battery is connected. These currents do *not* flow throughout the entire length of both wires simultaneously. They start instantaneously *in both wires* at the battery terminals, but a definite time interval will elapse before they are evident at a distance from the battery.

The time interval may be very small. For example, one-millionth of a second (one microsecond) after the connection is made the currents in the wires will have traveled 300 meters, or nearly 1000 feet, from the battery terminals. Note that they flow in both wires simultaneously, even though there may be no connection between the two wires at the end (which is infinitely far away) to form what we ordinarily think of as a closed circuit.

The current is in the nature of a charging current, flowing to charge the capacitance between the two wires. But unlike an ordinary condenser, the conductors of this "linear" condenser have appreciable inductance. In fact, we may think of the line as being composed of a whole series of small inductances and capacitances connected as shown in Fig. 10-3, where each coil is the inductance of a very short section of one wire and each condenser is the capacitance between two such short sections.

Characteristic Impedance

An indefinitely-long chain of coils and condensers connected as in Fig. 10-3, where each L is the same as all others and all the Cs have the same value, has an interesting and important peculiarity. To an electrical impulse applied to one end, the combination (or transmission line) appears to have an impedance that is approximately equal to $\sqrt{L/C}$, where L and C are the inductance and capacitance per unit length. Furthermore, this impedance is purely resistive. The line will "look like" such an impedance only when it is infinitely long, but even a short line can be made to "think" it is infinitely long by means to be described a little later.

This inherent line impedance is called the **characteristic impedance** or **surge impedance** of the line. Its value is determined by the inductance and capacitance per unit length. These quantities in turn depend upon the size of the line conductors and the spacing between them. The closer the two conductors of the line and the greater their diameter, the higher the capacitance and the lower the inductance. A line with large conductors closely spaced will have low impedance, while one with small conductors widely spaced will have relatively high impedance.

The characteristic impedance of the line is a very important property. For one thing, it determines the amount of current that can flow when a voltage is applied to the line. When a line is infinitely long, the current is simply equal to E/Z_o, where E is the voltage applied to the line and Z_o is the characteristic impedance. This has nothing to do with the *resistance* of the conductors; in fact, in this simplified picture of a transmission line we

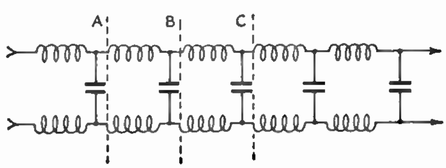


Fig. 10-3 — Equivalent of a transmission line in lumped circuit constants.

have tacitly assumed that the conductors do not have any resistance. The line is an impedance (like any circuit composed of L and C, without any R) that does not consume power. Actually, of course, the conductors do have resistance, so power cannot be transmitted along the line without some loss. But if the line is properly constructed and operated, this loss will be small compared with the amount of power carried to the load to do useful work.

R.F. on Lines

Bearing in mind that *time* must elapse before the currents initiated at the "input" end of the line — that is, the end to which the source of power is connected — can appear some distance away, consider now what happens when a radio-frequency voltage is applied to a transmission line. Suppose an r.f. generator is connected to a long line as shown in Fig. 10-4. To make the figures easy, assume that the frequency is 10 Mc., or 10,000,000 cycles per second. Then each cycle will occupy 0.1 microsecond, as shown by the drawing of the applied voltage. Suppose that the points *B* and *D* along the line are 30 meters away from *A* and *C*, respectively. If the current travels with the velocity of light, in 0.1 microsecond (one cycle) it will move 30 meters (300,000,000 meters divided by 10,000,000 cycles) along the line. This is a distance of one wavelength. Thus any currents observed at *B* and *D* occur just one cycle later in time than the currents at *A* and *C*. To put it another way, the currents initiated at *A* and *C* do not appear at *B* and *D*, one wavelength away, until the applied voltage has had time to go through a complete cycle.

Since the applied voltage is always changing, the currents at *A* and *C* are changing in proportion. The current a short distance away from *A* and *C* — for instance, at *X* and *Y* — is not the same as the current at *A* and *C* because the current at *X* and *Y* was caused by a value of voltage that occurred slightly earlier in the cycle. This is true all along the line; at any instant the current anywhere along the line from *A* to *B* and *C* to *D* is different from the current at every other point in that same distance. The series of drawings shows how the instantaneous currents might be distributed if we could snapshot them at intervals of one-quarter cycle. The current travels out from the input end of the line in waves.

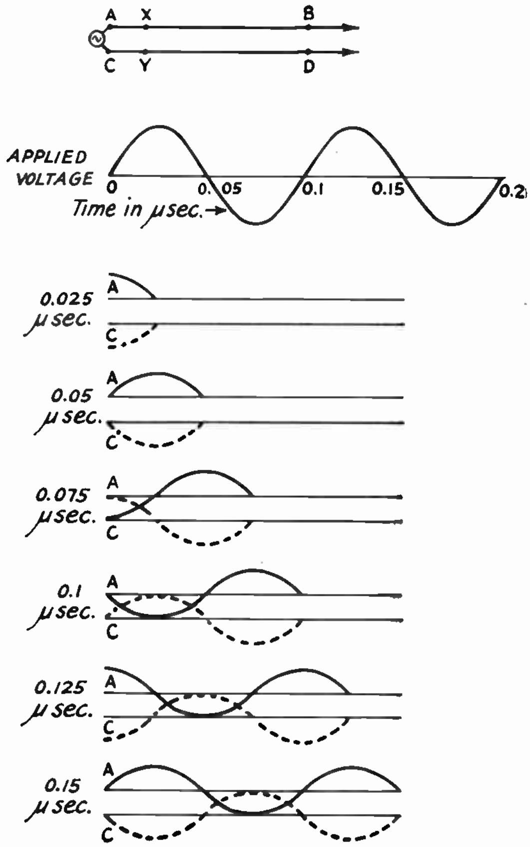


Fig. 10-4 — Progression of radio-frequency current flow in a transmission line.

At any selected point on the line the current goes through its complete range of a.c. values in the time of one cycle just as it does at the input end. Therefore (if there are no losses) an ammeter inserted in either conductor would read exactly the same current at any point along the line, because the ammeter averages the current over a whole cycle. The *phases* of the currents at any two separated points would be different, but an ammeter would not show this.

"Matched" Lines

In this picture of current traveling along a transmission line we have assumed that the line was infinitely long. Lines have a definite length, of course, and they are connected to or **terminated** in a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, the current traveling along the line to the load does not find conditions changed in the least when it meets the load; in fact, the load just looks like still more transmission line of the same characteristic impedance. Consequently, connecting such a load to a short transmission line allows the current to travel in exactly the same fashion as it would on an infinitely-long line.

In other words, a short line terminated in a purely-resistive load equal to the characteristic impedance of the line acts just as though it were infinitely long. Such a line is said to be **matched.** In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

STANDING WAVES

Now suppose that the line is terminated in a load that is not equal to the line's characteristic impedance. To take an extreme case, suppose that the output end of the line is short-circuited, as in Fig. 10-5.

With the infinitely-long line (or its matched counterpart) the impedance was the same at any point on the line and therefore the ratio of voltage to current was the same at any point on the line. However, the impedance at the end of the line in Fig. 10-5 is zero — or at least extremely small. A given amount of power in a very low impedance will result in a very large current and a very small voltage, as compared with the current-voltage ratio that exists in a few hundred ohms — which is a typical impedance value for some types of transmission lines. Something has to happen, therefore, when the power traveling along the transmission line meets the short-circuit at the end.

What happens is that the outgoing power, on meeting the short-circuit, simply reverses its direction of flow and goes back along the transmission line toward the input end. It has nowhere else to go. There is a very large current in the short-circuit, but substantially no voltage across the line at this point. We now have a voltage and current representing the power going outward toward the short-circuit, and a second voltage and current representing the reflected power traveling back toward the source.

Consider only the two current components

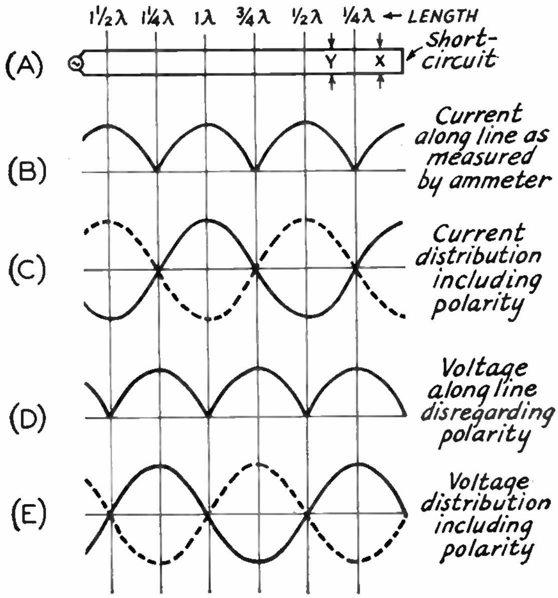


Fig. 10-5 — Standing waves of voltage and current along a short-circuited transmission line.

for the moment. The reflected current travels at the same speed as the outgoing current, so its instantaneous value will be different at every point along the line, in the distance represented by the time of one cycle. At some points along the line the outgoing and reflected currents will be in phase while at other points they will be completely out of phase. At the out-of-phase points the currents cancel each other (if the outgoing and reflected currents have the same value, as they will if all the power is reflected) and so at those points the resultant current is zero. At the in-phase points the two currents add numerically. At in-between points the two currents are neither completely in nor completely out of phase and so their sum is not equal to their numerical sum, but is something less.

The points at which the currents are in and out of phase depend only on the *time* required for them to travel and so depend only on the *distance* along the line from the point of reflection. The phase is completely reversed when the current travels for one-half cycle — that is, a distance of one-half wavelength — and is back in the in-phase condition when the current has traveled for one whole cycle, or one wavelength.

In the short-circuit at the end of the line the total current is high and the two current components are in phase. Therefore at a distance of *one-half* wavelength back along the line from the short-circuit the outgoing and reflected components will again be in phase and the current will have its maximum value. This is also true at any point that is a multiple of a half-wavelength from the short-circuited end of the line. The distance along the line is one-half wavelength because the current has to travel the distance twice in order to "meet itself coming back."

Since a total distance of one-half wavelength gives a complete reversal of phase, the outgoing and reflected currents will cancel at a point *one-quarter* wavelength, along the line, from the short-circuit. At this point, then, the current will be zero. It will also be zero at all points that are an *odd* multiple of one-quarter wavelength from the short-circuit.

If the current along the line is measured at successive points with an ammeter, it will be found to vary about as shown in Fig. 10-5B. The same result would be obtained by measuring the current in either wire, since the ammeter cannot measure phase. However, if the phase could be checked, it would be found that in each successive half-wavelength section of the line the currents at any given instant are flowing in opposite directions, as indicated by the solid line in Fig. 10-5C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first wire, as a result of the electron movement discussed in connection with Fig. 10-2. This is indicated by the broken curve in Fig. 10-5C. The variations in current intensity along the transmission line are referred to as **standing waves.** The point of maximum line current is called a **current loop** and the point of minimum line current a **current node.**

Voltage Relationships

Since the end of the line is short-circuited, the voltage at that point has to be zero. This can only be so if the voltage in the outgoing wave is met, at the end of the line, by an equal voltage of opposite polarity. In other words,

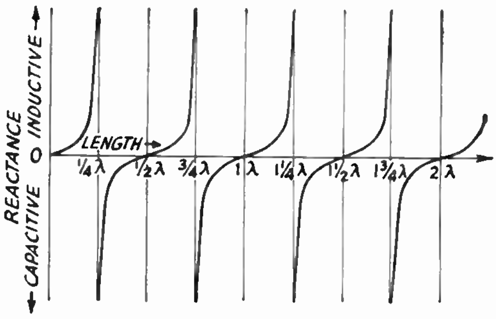


Fig. 10-6 — Input reactance *vs.* length of a short-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.

the phase of the voltage wave is *reversed* when reflection takes place from the short-circuit. This reversal is equivalent to an extra half-cycle or half-wavelength of travel. As a result, the outgoing and returning voltages are in phase a quarter wavelength from the end of the line, and again out of phase a half-wavelength from the end. The standing waves of voltage, shown at D in Fig. 10-5, are therefore displaced by one-quarter wavelength from the standing waves of current. The drawing at E shows the voltages on both wires when phase is taken into account. The polarity of the voltage on each wire reverses in each half-wavelength section of transmission line. A voltage maximum on the line is called a **voltage loop** and a voltage minimum is called a **voltage node.**

At either the odd or even multiples of a quarter wavelength the impedance is a pure resistance, because at these points the current and voltage in the transmission line are exactly in phase.

A detailed study of the outgoing and reflected components of voltage and current will show that at a point such as X in Fig. 10-5, lying anywhere in the section of line between the short-circuit and the first quarter-wavelength point, the current lags behind the voltage. This is exactly what happens in an inductance, so it can be said that a section of short-circuited transmission line less than a quarter wavelength long has inductive reactance. The value of reactance is determined by the ratio of voltage to current at the input end of such a line. It is evident from B and D in Fig. 10-5 that the reactance is low when the line is quite short, and highest when the line is nearly a quarter wavelength long. The line also has inductive reactance when its length is between one-half and three-quarter wavelength, between one and one-and-one-quarter wavelengths, and so on.

On the other hand, in the section of line between one-quarter and one-half wavelength from the short-circuit the current leads the voltage, so a short-circuited line having a length between these two limits "looks like" a capacitive reactance to the generator to which it is connected. The reactance is highest when the line is just over one-quarter wavelength long, and lowest when the line is just less than one-half wavelength long. Fig. 10-6 shows the general way in which the reactance varies with different line lengths.

Input Impedance

It is apparent, from examination of B and D in Fig. 10-5, that at points that are a multiple of a half-wavelength — i.e., ½, 1, 1½ wavelengths, etc. — from the short-circuited end of the line the current and voltage have the same values that they do at the short-circuit. In other words, if the line were an exact multiple of a half-wavelength long the generator or source of power would "look into" a short-circuit. On the other hand, at points that are an odd multiple of a quarter wavelength — i.e., ¼, ¾, 1¼, etc. — from the short-circuit the voltage is maximum and the current is zero. Since $Z = E/I$, the impedance at these points is theoretically infinite. (Actually it is very high, but not infinite. This is because the current does not actually go to zero when there are losses in the line. Losses are always present, but usually are small enough so that the impedance is of the order of tens or hundreds of thousands of ohms.)

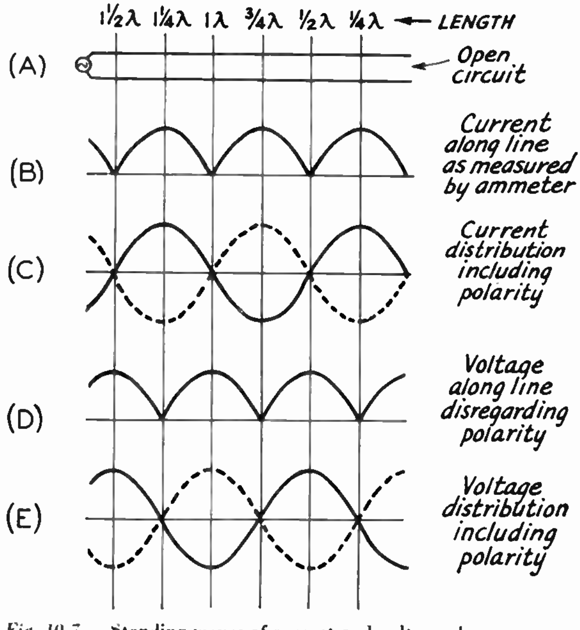


Fig. 10-7 — Standing waves of current and voltage along an open-circuited transmission line.

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source because it has nowhere else to go. In this case, the outgoing and reflected components of *current* must be equal and opposite in phase in order for the total current at the end of the line to be zero. The outgoing and reflected components of voltage are in phase, however, and add together. The result is that we again have standing waves, but the conditions are reversed. Fig. 10-7 shows the open-circuited line case. It may be compared directly with Fig. 10-5. The impedance looking into the line toward the open end is purely resistive at each multiple of one-quarter wavelength. It is very low at odd multiples of one-quarter wavelength, and very high at even multiples. In fact, an open-circuited line and short-circuited line behave just alike *if* the length of one differs by one-quarter wavelength from the length of the other.

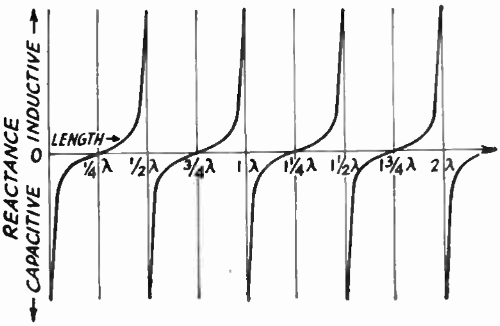


Fig. 10-8 — Input reactance *vs.* length of an open-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.

Fig. 10-8 shows how the reactance varies with line length for the open-circuited line. Comparing this with Fig. 10-6 shows that the reactance of any given length of line is of the opposite type to that obtained with a short-circuited line of the same length.

Lines Terminated in Resistive Load

An open- or short-circuited line does not deliver any power to a load, and for that reason is not, strictly speaking, a "transmission" line. However, the fact that a line of the proper length has an extremely high resistive input impedance at a given frequency or wavelength makes such lines useful as substitutes for the more common coil-and-condenser resonant circuits. With proper design, the effective Q of such a "linear" resonant circuit is much higher than is obtainable with coils and condensers. Linear circuits are particularly useful at v.h.f., and their application in that field is discussed in later chapters. In this chapter we are concerned with lines delivering power to a load such as an antenna.

Fig. 10-9 shows a line terminated in a resistive load. In such a case at least part of the outgoing power is absorbed in the load, and so is not available to be reflected back toward the source. Because only part of the power is reflected, the reflected components of voltage and current do not have the same magnitude as the outgoing components. Therefore there is no such thing as complete cancellation of either voltage or current at any point along the line. However, the *speed* at which the outgoing and reflected components travel is not affected by their amplitude, so the phase relationships are similar to those in open- or short-circuited lines.

It was pointed out earlier that if the load resistance, which we will call Z_r, is equal to the characteristic impedance, Z_o, of the line all the power is absorbed in the load. In such a case there is no reflected power and therefore no standing waves of current and voltage. This is a special case that represents the changeover point between "short-circuited" and "open-circuited" lines. If Z_r is less than Z_o, the current is largest at the load and the reflected component of voltage is out of phase with the outgoing component at the load. If Z_r is greater than Z_o, the voltage is largest at the load and the reflected component of current is out of phase with the outgoing component. Thus, if Z_r is less than Z_o the current will be minimum at a point one-quarter wavelength from the load and at every point an odd number of quarter wavelengths away, while the voltage will be maximum at these same points. The current will be maximum and the voltage minimum at points that are multiples of one-half wavelength from the load. If Z_r is greater than Z_o, the opposite is true. The two conditions are shown at B and C, respectively, in Fig. 10-9.

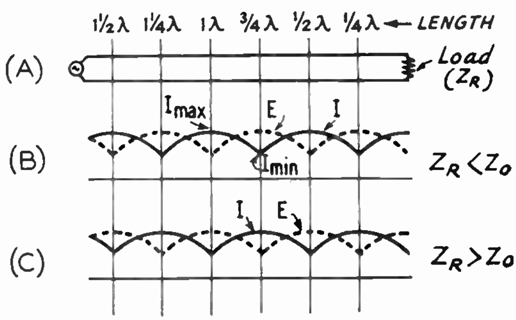


Fig. 10-9 — Standing waves on a transmission line terminated in a resistive load.

The impedance looking into the line toward the load will be purely resistive (if Z_r is a pure resistance) when the line length is a multiple of a quarter wavelength, just as in the open- and short-circuited cases. The input impedance is *equal* to Z_r when the line is an *even* multiple of a quarter wavelength long. If Z_r is less than Z_o the impedance is maximum when the line is an odd multiple of a quarter wavelength long. In such a case the impedance looking into the line is

$$Z_s = \frac{Z_o^2}{Z_r} \qquad \text{(10-A)}$$

where Z_s = Impedance looking into line (line length an odd multiple of one-quarter wavelength)
Z_r = Impedance of load (must be pure resistance)
Z_o = Characteristic impedance of line

Example: A quarter-wavelength line having a characteristic impedance of 500 ohms is terminated in a resistive load of 75 ohms. The impedance looking into the input or sending end of the line is

$$Z_s = \frac{Z_o^2}{Z_r} = \frac{(500)^2}{75} = \frac{250{,}000}{75} = 3333 \text{ ohms}$$

When Z_r is greater than Z_o, the input impedance reaches its minimum value when the line is an odd multiple of a quarter wavelength long. The value of input impedance in this case also is given by the equation above.

Example: A quarter-wave line is terminated in a resistive load of 1200 ohms. The characteristic impedance of the line is 600 ohms. Then the input impedance of the line is

$$Z_s = \frac{Z_o^2}{Z_r} = \frac{(600)^2}{1200} = \frac{360{,}000}{1200} = 300 \text{ ohms}$$

Impedance Transformation

If the formula in the preceding discussion is rearranged, we have

$$Z_o = \sqrt{Z_s Z_r} \qquad \text{(10-B)}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product. A quarter-wave line, in other words, has the characteristics of a transformer. This is a very useful attribute of transmission lines.

Example: A 600-ohm transmission line is to be used to feed an antenna that has a resistive impedance of 75 ohms. It is desired that the line operate without standing waves, and it must therefore be terminated in a resistive load equal to its characteristic impedance; i.e., 600 ohms. A quarter-wave line or "linear transformer" is to be used to match the 75-ohm load to the line impedance. To do this, the characteristic impedance of the quarter-wave transformer must be

$$Z_o = \sqrt{Z_s Z_r} = \sqrt{75 \times 600} = \sqrt{45{,}000}$$
$$= 212 \text{ ohms}$$

Reactance of Terminated Lines

We have seen that a short-circuited line less than one-quarter wavelength long exhibits inductive reactance. Also, a line of any length terminated in a resistive load equal to its characteristic impedance always looks like a pure resistance to the source of power. When the load is purely resistive and has any value between zero and Z_o, a line less than a quarter wave long will show inductive reactance, but the reactive effects decrease the closer the value of Z_r approaches Z_o.

On the other hand, an open-circuited line less than one-quarter wavelength long exhibits capacitive reactance. If the line is terminated in a resistive load larger than Z_o, it continues to show capacitive reactance but the reactive effects are less the closer Z_r approaches Z_o in value.

In general, then, a line terminated in a resistive impedance less than Z_o will show reactance variations with length similar to those of a short-circuited line as given in Fig. 10-6. A line terminated in a resistive impedance greater than Z_o will show reactance variations with length similar to those of an open-circuited line as given in Fig. 10-8. The magnitudes of the reactances will be smaller the closer Z_r approaches Z_o in value.

Loads That Are Not Pure Resistance

In most amateur applications of transmission lines the load is — or should be — a pure resistance. At least, every attempt is made to make it so. However, there are cases where the load has reactance as well as resistance, and recognizing the symptoms of reactance in the load is of value in indicating what steps should be taken to convert the load to a pure resistance.

The situation is easier to visualize if a line terminated in a pure reactance is considered first. For example, suppose the line is terminated in a capacitive reactance as shown in Fig. 10-10A. It does not matter to the line what physical form the reactance takes; the important thing is that in it the current leads the voltage. The reactance might be a condenser, for example — or it might simply be an additional section of transmission line that exhibits capacitive reactance at its input end, as indicated in Fig. 10-10B.

From Fig. 10-8, we can see that a section of open-circuited transmission line less than one-quarter wavelength long will have capacitive reactance. By proper choice of line length, any desired value of reactance can be obtained. Conversely, any "lumped" reactance, such as a condenser, connected to the end of the transmission line can be replaced by a section of open-circuited line of appropriate length. If the condenser capacitance is small and its reactance therefore is high, only a short length of line is required. If the condenser capacitance is large and its reactance consequently is low, the additional line section must be nearly a quarter wavelength long. In other words, connecting a condenser across the end of the transmission line is equivalent to lengthening the line, electrically. The amount of effective lengthening depends on the capacitance of the condenser.

Once the equivalent lengthening is determined, we can simply look upon the line as one having the new length and apply all that has been said previously. In the case just considered, this would mean that the point of maximum current, instead of appearing exactly a quarter wavelength from the end of the open-circuited line, would appear at something less than a quarter wavelength from the end. This is shown in Fig. 10-10C. The larger the capacitance of the terminating condenser the closer the current loop comes to the physical end of the line. All the other loops and nodes of both current and voltage would be changed accordingly.

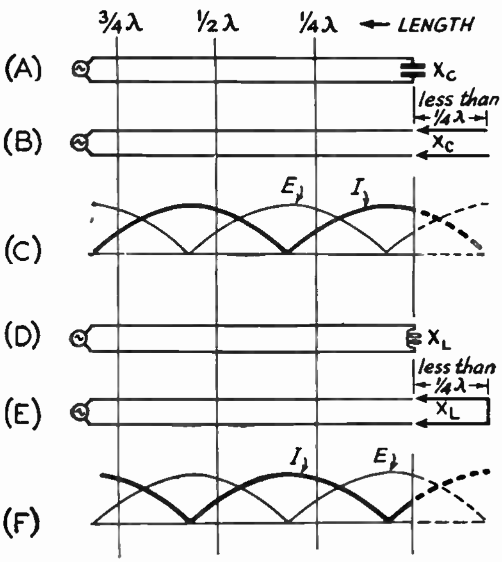


Fig. 10-10 — **Lines terminated in reactance. A reactive load is equivalent to a change in the length of the line.**

If the line is terminated in an inductance, we can substitute a short-circuited section of line less than one-quarter wavelength long for the lumped inductance. Thus, terminating a line in an inductance is equivalent to extending its length by something less than one-quarter wavelength and short-circuiting it. This is shown at D, E and F in Fig. 10-10. The larger the inductance, the greater the length of line, up to one-quarter wavelength, that must be added to obtain the electrical equivalent. When the equivalent section of line is substituted for the inductance, all that has been said about shorted lines applies, based on the new equivalent length.

When the load has both resistance and reactance the apparent length of the line is again increased. However, the amount of the apparent increase is affected by the resistance component of the load together with the reactive component. Inductive reactance will cause the first voltage maximum to appear less than one-quarter wavelength from the load, just as in Fig. 10-10F. Capacitive reactance will cause the first current maximum to appear less than one-quarter wavelength from the load, as in Fig. 10-10C. If the positions of either the voltage or current loops can be determined, it is always possible to tell whether the reactive component of the load is inductive or capacitive. When the load has both resistance and reactance, the voltage and current nodes do not reach zero because not all the outgoing power is reflected. The actual standing waves would be more like those shown in 10-9, but with the positions of the nodes and loops shifted as indicated in Fig. 10-10.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, as indicated in Fig. 10-11, is called the standing-wave ratio. It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is perfectly matched. (In that case the "maximum" and "minimum" current are the same, since the current does not vary along the line.) When the line is terminated in a purely-resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_r}{Z_o} \text{ or } \frac{Z_o}{Z_r} \qquad \textbf{(10-C)}$$

Where $S.W.R.$ = Standing-wave ratio

Z_r = Impedance of load (must be pure resistance)

Z_o = Characteristic impedance of line

Example: A line having a characteristic impedance of 300 ohms is terminated in a resistive load of 25 ohms. The s.w.r. is

$$S.W.R. = \frac{Z_o}{Z_r} = \frac{300}{25} = 12 \text{ to } 1$$

It is customary to put the larger of the two quantities, Z_r or Z_o, in the numerator of the fraction so that the s.w.r. will be expressed by a number larger than 1.

It is easier to measure the standing-wave ratio than some of the other quantities (such as the impedance of an antenna) that enter into transmission-line computations. Consequently, the s.w.r. is a convenient basis for work with lines. The higher the s.w.r., the greater the mismatch between line and load. Also, the higher the s.w.r. the more marked are the reactive effects when the line length is not an exact multiple of a quarter-wavelength. In practical lines, the loss in the line itself increases with the s.w.r.

Resonant and Nonresonant Lines

A transmission line terminated in a resistive load equal to its characteristic impedance is commonly called a **flat, nonresonant** or **untuned** line. The line is "flat" because there are no standing waves, hence a graph of the current along the line is a straight line. It is "nonresonant" because the input impedance of such a line is pure resistance and does not change when the line length is changed.

When there are standing waves the line is said to be **resonant** or **tuned.** In this case the input impedance depends critically on the length of the line and the characteristics of the load. The input impedance is a pure resistance only when the line length is such that a current or voltage loop appears at the input end. The previous discussion has shown that the positions of these loops depends upon the characteristics of the load. At all other lengths the input impedance consists of both reactance and resistance. Under these conditions the line acts something like a circuit that is not tuned to resonance; it is difficult to make it "take power" until something is done to "tune" it — that is, to eliminate the reactance. When this is done the input impedance of the line is purely resistive and its resistance may be matched to the transmitter for optimum power transfer.

It should be noted that if there are standing waves on the line the input impedance, even when the reactance is tuned out, is never equal to the characteristic impedance of the line. Depending on the length of the line, the characteristics of the load, and the s.w.r., the input resistance may be considerably higher or considerably lower than the line's characteristic impedance. This introduces an element of uncertainty in coupling to the transmitter. In one special case, when the load is a pure resistance and the line is exactly one-half wavelength long, the input impedance of the line is a pure resistance equal to the load impedance.

The reactive or resonance effects increase with the s.w.r., as previously pointed out. Generally speaking, a line is satisfactorily flat if the s.w.r. does not exceed about 1.5 to 1, but if the s.w.r. is much larger it becomes necessary to tune out the input reactance.

Radiation

Whenever a wire carries alternating current the electromagnetic fields travel away into space with the velocity of light. At power-line frequencies the field that "grows" when the current is increasing has plenty of time to return or "collapse" about the conductor when the current is decreasing, because the alternations are so slow. But at radio frequencies fields that travel only a relatively short distance do not have time to get back to the conductor before the next cycle commences. The consequence is that some of the electromagnetic energy is prevented from being restored to the conductor; in other words, energy is radiated into space in the form of electromagnetic waves.

The amount of energy radiated depends, among other things, on the length of the conductor in relation to the frequency or wavelength of the r.f. current. If the conductor is very short compared to the wavelength the energy radiated will be small. However, a transmission line used to feed power to an antenna is not short in this sense; in fact, it is almost always an appreciable fraction of a wavelength long and may have a length of several wavelengths.

The lines previously considered have consisted of two parallel conductors of the same diameter. Provided there is nothing in the system to destroy symmetry, at every point along the line the current in one conductor has the same intensity as the current in the other conductor at that point, but the currents flow in opposite directions. This was shown in Figs. 10-5C and 10-7C. This means that the fields set up about the two wires have the same intensity, but *opposite directions.* The consequence is that the total field set up about such a transmission line is zero; the two fields "cancel out." Hence no energy is radiated.

Actually, the fields do not completely cancel out because for them to do so the two conductors would have to occupy the same space, whereas they are slightly separated. However, the cancellation is substantially complete if the distance between the conductors is very small compared to the wavelength. Radiation will be negligible if the distance between the conductors is 0.01 wavelength or less, provided the currents in the two actually are balanced as described.

The amount of radiation also is proportional to the current flowing in the line. Because of the way in which the current varies along the line when there are standing waves, the effective current, for purposes of radiation, becomes greater as the s.w.r. is increased. For this reason the radiation is least when the line is flat. However, if the conductor spacing is small and the currents are balanced, the radiation from a line with even a high s.w.r. is inconsequential. A small unbalance in the line currents is far more serious.

There is no factual basis for the common belief that the presence of standing waves on a transmission line always means that the line is radiating a great deal of r.f. energy. Tuned lines are perhaps more subject to the stray coupling effects described later in this chapter, simply because they are frequently cut to resonant lengths while any random length can be used for a flat line. It is the stray coupling that gives rise to excessive line radiation, not the presence of the normal type of standing wave on the transmission line.

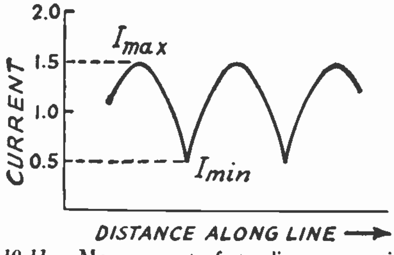


Fig. 10-11 — Measurement of standing-wave ratio. In this drawing, I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. $= I_{max}/I_{min} = 1.5/0.5 = 3$.

Practical Line Characteristics

The foregoing discussion of transmission lines has been based on a line consisting of two parallel conductors. Actually, the **parallel-conductor** line is but one of two general types. The other is the **coaxial** or **concentric** line. The coaxial line consists of a round conductor placed in the center of a circular tube. The inside surface of the tube and the outside surface of the smaller inner conductor form the two conducting surfaces of the line.

In the coaxial line the fields are entirely inside the tube, because the tube acts as a shield to prevent them from appearing outside. This reduces radiation to the vanishing point. So far as the electrical behavior of coaxial lines is concerned, all that has previously been said about the operation of parallel-conductor lines applies. There are, however, practical differences in their construction and use.

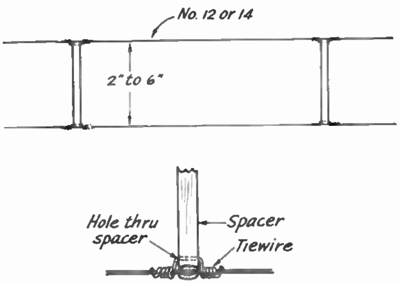


Fig. 10-12 — Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

Types of Construction

There are several constructional variations in both the basic types of transmission lines mentioned in the preceding section. Probably the most common type of transmission line used in amateur installations is a parallel-conductor line in which two wires (ordinarily No. 12 or No. 14) are supported a fixed distance apart by means of insulating rods called "spacers." The spacings used vary from two to six inches, the smaller spacings being necessary at frequencies of the order of 28 Mc. and higher so that radiation will be minimized. The construction is shown in Fig. 10-12. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 10-13. The characteristic impedance of such "open-wire" lines runs between about 400 and 600 ohms, depending on the wire size and spacing.

Parallel-conductor lines also are sometimes constructed of metal tubing of a diameter of ¼ to ½ inch. This reduces the characteristic impedance of the line. Such lines are mostly used as quarter-wave transformers, when different values of impedance are to be matched.

Two forms of "Twin-Lead" or "ribbon" transmission line are shown in Fig. 10-13. This is a parallel-conductor line with stranded conductors imbedded in low-loss insulating material (polyethylene). It has the advantages of light weight, compactness and neat appearance, together with close and uniform spacing. However, losses are higher in the solid dielectric than in air, and dirt or moisture on the line tends to change the characteristic impedance. Twin-Lead line is available in characteristic impedances of 75, 150 and 300 ohms.

The most common form of coaxial line consists of either a solid or stranded-wire inner conductor surrounded by polyethylene dielectric. Copper braid is woven over the dielectric to form the outer conductor, and a waterproof vinyl covering is placed on top of the braid. This cable is made in a number of different diameters. It is moderately flexible, and so is convenient to install. Some different types are shown in Fig. 10-13. This solid coaxial cable is commonly available in impedances approximating 50 and 70 ohms.

Air-insulated coaxial lines have lower losses than the solid-dielectric type, but are less used in amateur work because they are expensive and difficult to install as compared with the flexible cable. The common type of air-insulated coaxial line uses a solid-wire conductor inside a copper tube, with the wire held in the center of the tube by means of insulating "beads" at regular intervals.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_o = 276 \log \frac{b}{a} \qquad \textbf{(10-D)}$$

where Z_o = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

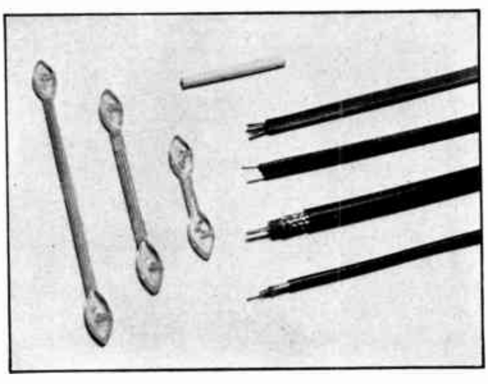


Fig. 10-13 — Typical manufactured transmission lines and spacers.

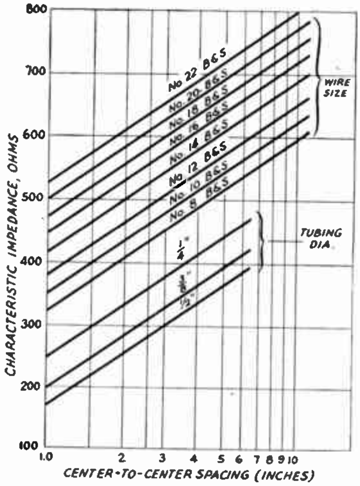


Fig. 10-14 — Chart showing the characteristic impedance of typical spaced-conductor parallel transmission lines. Tubing sizes given are for outside diameters.

It does not matter what units are used for a and b so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 10-14 for a number of common conductor sizes.

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_o = 138 \, log \frac{b}{a} \qquad \textbf{(10-E)}$$

where Z_o = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Again it does not matter what units are used for b and a, so long as they are the same. Curves for typical conductor sizes are given in Fig. 10-15.

The formula for coaxial lines is approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced. When the line is filled with a solid dielectric, the characteristic impedance as given by the chart should be multiplied by $1/\sqrt{K}$, where K is the dielectric constant of the material. In solid-dielectric parallel-conductor lines such as Twin-Lead the characteristic impedance cannot be calculated readily, because part of the electric field is in air as well as in the solid dielectric.

Electrical Length

In the discussion of line operation earlier in this chapter it was assumed that currents traveled along the conductors at the speed of light. Actually, the velocity is somewhat less, the reason being that electromagnetic fields travel more slowly in dielectric materials than they do in free space. In air the velocity is practically the same as in empty space, but a practical line always has to be supported in some fashion by solid insulating materials. The result is that the fields are slowed down; the currents travel a shorter distance in the time of one cycle than they do in space, and so the wavelength along the line is less than the wavelength would be in free space at the same frequency. (Wavelength is equal to velocity divided by frequency.)

Whenever reference is made to a line as being so many wavelengths (such as a "half-wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. Its actual physical length as measured by a tape always will be somewhat less. The physical length corresponding to an electrical wavelength is given by

$$Length\ in\ feet = \frac{984}{f} \cdot V \qquad \textbf{(10-F)}$$

where f = Frequency in megacycles
V = Velocity factor

The **velocity factor** is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 10-I.

Example: A 75-foot length of 300-ohm Twin-Lead is used to carry power to an antenna at a frequency of 7150 kc. From Table 10-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is

$$Length\ (\text{feet}) = \frac{984}{f} \cdot V = \frac{984}{7.15} \times 0.82$$
$$= 137.6 \times 0.82 = 112.8 \text{ ft.}$$

The line length is therefore $75/112.8 = 0.665$ wavelength.

Because a quarter-wavelength line is frequently used as a linear transformer, it is convenient to calculate the length of a quarter-wave line directly. The formula is

$$Length\ (\text{feet}) = \frac{246}{f} \cdot V \qquad \textbf{(10-G)}$$

where the symbols have the same meaning as above.

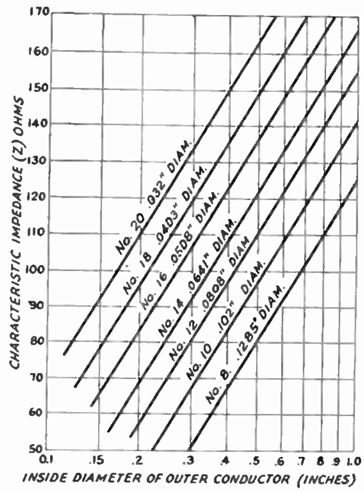


Fig. 10-15 — Chart showing characteristic impedance obtained with various air-insulated concentric lines.

Losses in Transmission Lines

There are three ways by which power may be lost in a transmission line: by radiation, by heating of the conductors (I^2R loss), and by heating of the dielectric, if any. Loss by radiation will occur if the line is unbalanced and, particularly with open-wire lines, may greatly exceed the heat losses. It can be reduced to a minimum by terminating the line in a balanced load and by symmetrical construction.

Heat losses in both the conductor and the dielectric increase with frequency. Conductor losses also are greater the lower the characteristic impedance of the line, because a higher current flows in a low-impedance line for a given power input. The converse is true of dielectric losses because these increase with the voltage, which is greater on high-impedance lines. The dielectric loss in air-insulated lines is negligible (the only loss is in the insulating spacers) and such lines operate at high efficiency when radiation losses are low. In solid-dielectric lines most of the loss is in the dielectric, the conductor losses being small.

It is convenient to express the loss in a transmission line in decibels per unit length, since the loss in db. is directly proportional to the line length. Losses in various types of lines operated without standing waves (that is, terminated in a resistive load equal to the characteristic impedance of the line) are given in Table 10-I. In these figures the radiation loss is assumed to be negligible.

When there are standing waves on the line the power loss increases as shown in Fig. 10-16. Whether or not the increase in loss is serious depends on what the original loss would have been if the line were perfectly matched. If the loss with perfect matching is very low, a large s.w.r. will not greatly affect the *efficiency* of the line — i.e., the ratio of the power delivered to the load to the power put into the line.

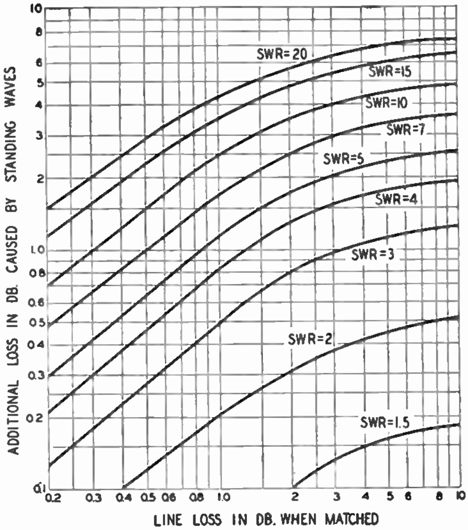


Fig. 10-16 — **Effect of standing-wave ratio on line loss. The ordinates give the *additional* loss in decibels for the total line loss, under perfectly-matched conditions, shown on the horizontal scale.**

Example: A 150-foot length of RG-11/U cable is operating at 7 Mc. with a 5-to-1 s.w.r. If perfectly matched, the loss from Table 10-I would be $1.5 \times 0.41 = 0.615$ db. From Fig. 10-16 the additional loss because of the s.w.r. is 0.73 db. The total loss is therefore $0.615 + 0.73 = 1.345$ db. The total power loss is just sufficient to make a detectable change in signal strength when observing conditions are ideal, but the additional loss caused by the s.w.r. is below the detectable (1 db.) level. With perfect matching the line efficiency is approximately 87 per cent. With the 5-to-1 s.w.r. the efficiency drops to about 73.5 per cent. With an open-wire line the loss caused by such an s.w.r. would be negligible, provided the line is well balanced to prevent radiation.

TABLE 10-I

Transmission-Line Velocity Factors and Attenuation

Type of Line	*Velocity Factor V*	** *Attenuation, db./100 ft.; Mc.*						*Capacitance per foot μμfd.*
		3.5	*7*	*14*	*28*	*50*	*144*	
Open-wire, 400 to 600 ohms	0.975*	0.03	0.05	0.07	0.1	0.13	0.25	
Parallel-tubing	0.95*	***						
Coaxial, air-insulated	0.85*	0.2	0.28	0.42	0.55	0.7	1.4	
RG-8/U (53 ohms)	0.66	0.28	0.42	0.64	1.0	1.4	2.6	29.5
RG-58/U (53 ohms)	0.66	0.53	0.8	1.2	1.9	2.7	5.1	28.5
RG-11/U (75 ohms)	0.66	0.27	0.41	0.61	0.92	1.3	2.4	20.5
RG-59/U (73 ohms)	0.66	0.56	0.82	1.4	1.8	2.5	4.6	21.0
Twin-Lead, 300 ohms	0.82	0.18	0.3	0.5	0.84	1.3	2.8	5.8
Twin-Lead, 150 ohms	0.77	0.2	0.35	0.6	1.0	1.6	3.5	10
Twin-Lead, 75 ohms	0.68	0.37	0.64	1.1	1.9	3.0	6.8	19
Transmitting Twin-Lead, 75 ohms	0.71	0.29	0.49	0.82	1.4	2.1	4.8	
Rubber-insulated twisted-pair or coaxial ****	0.56 to 0.65	0.96	1.6	2.5	4.2	6.2	13	

* Average figures for air-insulated lines taking into account effect of insulating spacers.

** For lines terminated in characteristic impedance.

*** Losses between open-wire line and air-insulated coaxial cable. Actual loss with both open-wire and parallel-tubing lines is higher than listed because of radiation, especially at higher frequencies.

**** Approximate figures for good-quality rubber insulation.

An appreciable s.w.r. on a solid-dielectric line may result in excessive loss of power at the higher frequencies. Such lines, whether of the parallel-conductor or coaxial type, should be operated as nearly flat as possible, particularly when the line length is more than 50 feet or so. As shown by Fig. 10-16, the increase in line loss is not too serious so long as the s.w.r. is below 2 to 1, but increases rapidly when the s.w.r. rises above 3 to 1. Tuned transmission lines such as are used with multiband antennas always should be air-insulated, in the interests of highest efficiency.

Unbalance in Parallel-Conductor Lines

When installing parallel-conductor lines care should be taken to avoid introducing electrical unbalance into the system. If for some reason the current in one conductor is higher than in the other, or if the currents in the two wires are not exactly out of phase with each other, the electromagnetic fields will not cancel completely and a considerable amount of power may be radiated by the line.

Maintaining good line balance requires, first of all, a balanced load at its end. For this reason the antenna should be fed, whenever possible, at a point where each conductor "sees" exactly the same thing. Usually this means that the antenna system should be fed at its electrical center. Even though the antenna appears to be symmetrical, physically, it can be unbalanced electrically if the part connected to one of the line conductors is inadvertently coupled to something (such as house wiring or a metal pole or roof) that is not duplicated on the other part of the antenna. Every effort should be made to keep the antenna as far as possible from other wiring or sizable metallic objects. The transmission line itself will cause some unbalance if it is not brought away from the antenna at right angles to it for a distance of at least a quarter wavelength.

In installing the line conductors take care to see that they are kept away from metal. The minimum separation between either conductor and all other wiring should be at least four or five times the conductor spacing. The shunt capacitance introduced by close proximity to metallic objects can drain off enough current (to ground) to unbalance the line currents, resulting in increased radiation. A shunt capacitance of this sort also constitutes a reactive load on the line, causing an impedance "bump" that will prevent making the line actually flat.

Coupling the Transmitter to the Line

In very general terms, the problem of coupling the transmission line and transmitter together is one of transforming the input impedance of the line into a value of impedance that will "load" the transmitter properly — that is, cause it to deliver the desired power output at as high efficiency as the transmitter design will permit. This is a question of impedance matching, and the impedance that must be matched is the value of resistance into which the tubes in the final stage of the transmitter should work. The value of this resistance is determined by the choice of tube operating conditions. The tubes are working into the proper resistance when the final tank circuit is tuned to resonance and the loading is such that the tubes are drawing rated plate current, as described in Chapter Six. The proper value of load resistance is thus reached automatically when the coupling is adjusted to bring the plate current up to the normal operating value. It is therefore not at all necessary to know what value of resistance is required. It is sufficient to note that, in general, it is in the neighborhood of a few thousand ohms, and is higher the higher the plate-voltage/plate-current ratio of the final stage.

The input impedance of the line can assume a wide range of values. As described earlier, it may be very much higher or very much lower than the impedance of the load at the end of the line, unless the line is matched to the load. Furthermore, it may or may not be a pure resistance, depending on the s.w.r., the line length, and the characteristics of the load.

Transforming Impedances

It was explained in Chapter Two that a resistive load tapped across part of a tuned circuit is equivalent to a higher value of resistance connected in parallel with the whole circuit. In other words, there is a transformer action in such an arrangement that enables us to change the value of a given resistance, such as R in Fig. 10-17A, into a new and higher value when the source of power looks into the terminals AB. Given reasonable values for L and C, the resistance looking into AB is determined practically wholly by the value of R and the position of the tap, so long as LC is tuned to resonance with the applied frequency. This is because the resonant impedance of LC alone (with R disconnected) is usually very high

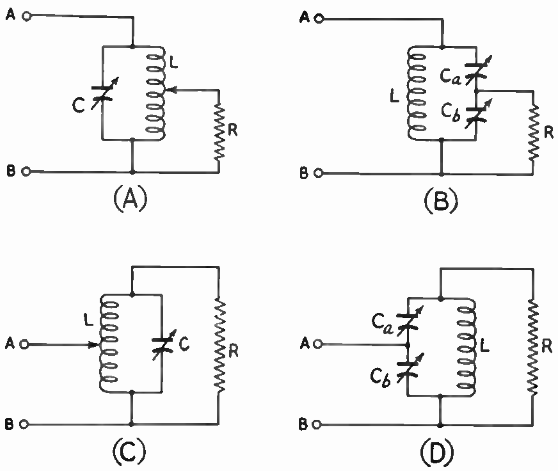


Fig. 10-17 — Using a resonant circuit for matching impedances.

compared with the resistance, R, of any practical load likely to be used, and also compared with any resistance that might be required between the terminals AB.

Fig. 10-17B shows a circuit that also provides a method for impedance transformation, using a capacitance voltage divider instead of tapping on the inductance. In this case, decreasing the capacitance of C_b (while increasing the capacitance of C_a correspondingly to maintain resonance) has the same effect as moving the tap toward the top of the coil in Fig. 10-17A. This type of circuit gives very smooth control. However, variable condensers of impracticable size would be necessary, to give as wide a range of impedance transformation as the circuit at A.

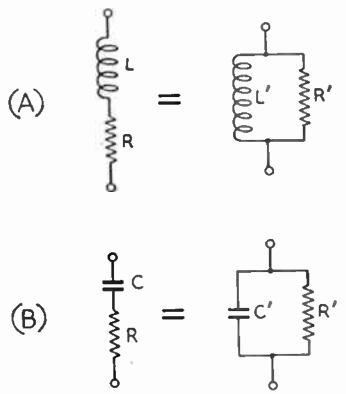


Fig. 10-18 — Series and parallel equivalents of a line whose input impedance has both reactive and resistive components.

When an r.f. amplifier is coupled to a transmission line the line impedance very seldom is larger than the load impedance required by the amplifier. However, should such a case arise the same circuits can be used by reversing the terminals. This is shown at C and D in Fig. 10-17. With R connected across the whole circuit, its resistance can be transformed to a lower value when the input terminals are tapped across part of the coil, as at A, or across C_b in Fig. 10-17B. The nearer the tap is to the bottom end of the coil, or the larger the capacitance of C_b compared with C_a, the smaller the resistance between terminals AB.

Complex Loads

In the foregoing it was assumed that the load, R, was a pure resistance. However, the input impedance of a line is more likely than not to have a reactive as well as a resistive component. This means, basically, that the current flowing into the line is not in phase with the voltage applied to the line. To represent such a condition by circuit symbols we can assume the input impedance of the line to consist either of a reactance (coil or condenser) in series with a resistance, or a reactance in parallel with a resistance. It does not matter which we choose, so long as the values assigned to the resistance and reactance are such that if the voltage were applied to the circuit instead of to the line, the current that flows would have exactly the same amplitude and phase angle as it actually does at the input terminals of the line.

These equivalent circuits are shown in Fig. 10-18. In practical work with lines it is not necessary to know the values of R, L or C. It is sufficient to know that they *symbolize* a condition that exists at the input end of the line — and then to know what to do about them. A few general points are worth noting: Given a fixed value of voltage, if the current at the input end of the line is high, then the impedance is relatively low; if the current is low, the impedance is relatively high. If the current is very nearly in phase with the voltage the reactance in the *series* equivalent circuit is small, but the reactance in the *parallel* equivalent circuit is large. On the other hand, if there is a considerable phase difference between current and voltage the reactance is large in the equivalent series circuit and is low in the equivalent parallel circuit. (In visualizing these reactances as coils and condensers it must be remembered that "large" and "small" are relative terms; for example, a "large" inductance at 28 Mc. would be a "small" inductance at 3.5 Mc. Also, the larger the capacitance of a condenser the smaller its reactance.)

Now suppose that a reactive line is to be connected to our impedance-transforming resonant circuit. Let us choose the parallel equivalent circuit, since it is somewhat easier to picture what happens. Fig. 10-19A shows a load with inductive reactance tapped across part of the resonant circuit (corresponding to Fig. 10-17A), and a load with capacitive reactance is shown in Fig. 10-19B. Imagine for the moment that the load has only reactance; the resistive component, R, is disconnected. Then, just as in the pure-resistance case previously discussed, a small reactance tapped across the coil L will appear as a larger reactance across the whole circuit, or between the input terminals AB. Thus, connecting a coil, L_1, across part of L is equivalent to connecting a larger coil across the whole circuit. Connecting a condenser, C_1, across part of L is equivalent to connecting a *smaller* condenser (larger reactance) across the whole circuit.

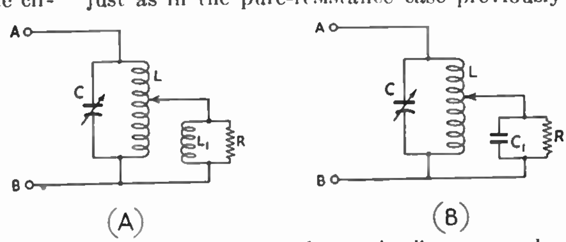


Fig. 10-19 — Circuit equivalent of a reactive line connected to a resonant circuit for impedance matching.

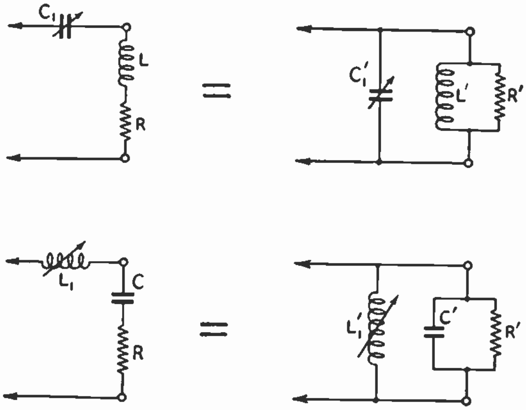


Fig. 10-20 — Methods for canceling the reactive component of the input impedance of a transmission line. In A the line input impedance is represented by L and R in series, or by L' and R' in parallel, and in B by C and R in series, or by C' and R' in parallel.

In either case this equivalent shunting reactance detunes the LC circuit from resonance, and C must be readjusted to bring it back. In the case of Fig. 10-19A, the capacitance of C must be increased because the "reflected" reactance in parallel with L decreases the total inductance (inductances in parallel) and so tunes the circuit to a higher frequency. The opposite is the case in Fig. 10-19B; the shunting reactance is capacitive and increases the total capacitance. Consequently the capacitance of C must be decreased to bring the circuit back to resonance.

The over-all effect, then, of coupling a reactive load to the circuit is to cause detuning as well as to cause the desired resistance loading. If the reflected reactance is large, corresponding to connecting a very large coil or a very small condenser across the whole LC circuit, it is readily possible to retune the circuit to resonance by adjusting C. The nearer the tap to the top end of L, the greater the change required in the tuning. But this simple method of compensating for the reactive component of the load is not always sufficient. In some cases the tap has to be moved so far up the coil, in order to obtain the right value of resistance loading, that the tuning condenser, C, no longer has sufficient range to compensate for the reflected reactance. When such a condition exists it is difficult, and sometimes impossible, to couple the desired amount of power to the transmission line.

Canceling Line Reactance

The remedy for this condition is to make the input end of the line look like a pure resistance *before* it is tapped on the impedance-transforming circuit. This can be done by "tuning out" the reactance of the line, by inserting a reactance of the same value but of the opposite kind. Again we have our choice between considering the line to be represented by reactance and resistance in series, or by reactance and resistance in parallel. The circuits are shown in Fig. 10-20. In A, a condenser, C_1, is used to cancel out the inductive reactance of the line, and in B an inductance, L_1, is used to cancel capacitive reactance. The same value of capacitance cannot be used for C_1 and C_1' under a given set of conditions because, as explained earlier, L and L' do not have the same values. For example, if L is small its parallel equivalent, L', is large, so a large capacitance would be required at C_1 and a small capacitance at C_1'. Because of limitations in practicable components (particularly in the capacitance range of variable condensers), there are conditions where the series circuit is the easiest to set up, from a practical standpoint. In others, the parallel circuit is easier to get working. For the large majority of cases either circuit will work equally well; from the standpoint of convenience, the parallel circuit is probably better.

To summarize, then, we have three general cases as shown in Fig. 10-21. If the line is purely resistive, or so nearly so that such reactance as is reflected across the LC circuit can be tuned out by readjusting C, the circuit at A may be used. Where the line shows more pronounced reactive effects, the line reactance can be tuned out, as indicated at B and C, so that the load tapped on L is purely resistive. It is easy to tell which should be used, inductance or capacitance, to compensate for the line reactance. If the line only (Fig. 10-21A)

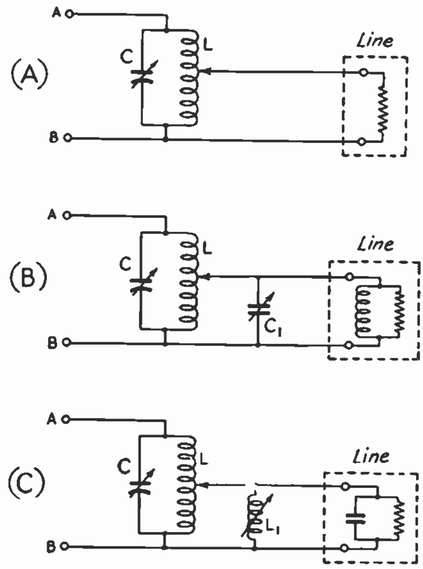


Fig. 10-21 — Methods of canceling line input reactance combined with impedance transformation.

is tapped across a very small portion of L, C will have to be readjusted slightly to bring the LC circuit back to resonance. If the capacitance of C has to be increased, a condenser, C_1, should be connected across the input terminals of the line. If the capacitance of C has to be decreased, an inductance, L_1, should be connected across the line. In either case the compensating reactance, C_1 or L_1, should be adjusted in value until the setting of C, for resonance with the applied frequency, is the same whether or not the line is tapped on L. When this condition is reached the loading may be adjusted by changing the tap position until the amplifier takes the desired plate current.

PRACTICAL COUPLING SYSTEMS

In practical work the two primary functions that a coupling system must perform — tuning out the line reactance, if any, and providing a method for control of loading on the transmitter — are not always enough. For one thing, it is desirable that the coupling system be such that the transmission line will operate only in the way it is intended that it should. For another, the coupling system should prevent transfer of any of the harmonic energy that always is present in the output of a transmitting amplifier. Both these points will be considered later in this section. For the moment, let us take a look at some of the simpler coupling systems.

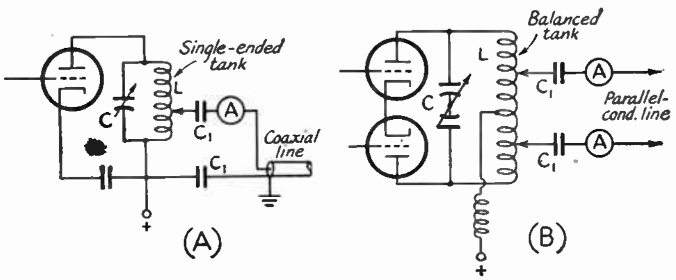


Fig. 10-22 — Simple methods of coupling to a transmission line. The blocking condensers, C_1, should be 0.001-μfd. (or larger) mica condensers having a voltage rating in excess of the maximum d.c. voltage applied to the final amplifier (including the voltage applied on modulation up-peaks). The coaxial line can be coupled to a balanced tank circuit by connecting the grounded shield to the center of the coil (through a blocking condenser) and tapping the inner conductor on one side of the center. The parallel-conductor line requires a balanced tank circuit.

The possibility of tapping the input end of the transmission line directly on the final-amplifier tank suggests itself from the discussion earlier. This method will work when the input impedance of the line is purely resistive, or nearly so. It can therefore be used with nonresonant or untuned lines, or with a resonant line when the line has the right length. As explained earlier, the input impedance of the line will be resistive when its length is a multiple of a quarter wavelength, provided the load at the output end of the line is a pure resistance. This will be so if the antenna itself is resonant, but will not be true if the antenna length is not correct for the operating frequency. The circuits are shown in Fig. 10-22. If the final amplifier is series-fed so that the tank circuit is "hot" with the plate voltage, it is necessary to connect a blocking condenser between the tank and the line. These circuits, although simple, are not recommended except perhaps in emergencies; there is little or no discrimination against harmonic frequencies.

Adjustment of this type of coupling is simple. First, resonate the amplifier tank circuit, with the line disconnected, by setting the tank condenser, C, to the minimum plate current point. Then tap the line across a turn or two of the tank coil, and readjust C for minimum plate current. The new minimum will be higher than with no load on the tank. Continue increasing the number of turns between the line taps, readjusting C each time, until the minimum plate current is the desired full-load value.

R.F. Ammeters

The r.f. ammeters shown in Fig. 10-22 and subsequent coupling circuits are useful accessories. The input impedance of the line is unaffected by any adjustments made in the coupling system (except for the effects of stray capacitance, as discussed later) so the greater the current flowing into the line the larger the amount of power delivered to the load. Measurement of r.f. current thus gives a check on the adjustment procedure and indicates when the largest power output is being obtained. Obviously, an adjustment that increases the input to the final stage of the transmitter without causing the line current to increase has simply increased the losses without increasing the output.

In the case of parallel-conductor lines two ammeters are shown, one in each conductor. This gives a check on line balance, since the two currents should be the same. It is not actually necessary to use two instruments; one ammeter can be switched from one side of the line to the other for comparative measurements. Also, it is to be understood that any current-indicating device (such as a flashlight lamp) that will work at r.f. may be used as a substitute for an actual ammeter.

The scale range required depends on the input impedance of the line and the power. The current to be expected can readily be calculated from Ohm's Law when the line is flat. In other cases the s.w.r. and the length of the line must be considered. The maximum current will occur when there is a current loop at the input end of the line, and if the load impedance and line impedance are known the input impedance at a current loop can be calculated from the formulas given earlier.

The ammeters are less useful when the input impedance of the line is high, because in that case the input current is quite small. It is to be noted that the value of current does not indicate, in any absolute sense, how well the system as a whole is working unless the actual value of the resistance component of the line input impedance is known. Current measurements taken on different lines, or on the same line if its length in wavelengths is changed, are not directly comparable.

Inductive Coupling

The circuits shown in Fig. 10-23, like those in Fig. 10-22, are useful only with lines having purely-resistive input impedance. The pick-up coil, which is inductively coupled to the tank coil, is in fact simply a substitute for the tapped portion of the tank coil in Fig. 10-22. The number of turns required in the pick-up coil depends upon the resistance represented by the input end of the line. For flat lines, the number is governed by the characteristic impedance of the line. For 50- or 70-ohm lines it may range from one or two turns, at frequencies of the order of 14 to 28 Mc., to several turns at 3.5 Mc. For higher-impedance lines it may take half as many turns as there are in the tank coil, to get adequate coupling. In both cases the coupling between the coils will have to be very tight. The link windings provided on commercial coils are not usually adequate for this type of coupling except for low-impedance lines at the higher frequencies. When the number of turns on the pick-up coil is fixed, the loading on the final amplifier can be varied by varying the coupling between the two coils. Inductive coupling of this type is somewhat better than direct coupling from the standpoint of harmonic transfer.

Pick-up coil coupling introduces some reactance into the tank circuit, because of the leakage reactance of the coupling coil. This must be compensated for by retuning the final tank circuit when the desired degree of coupling is reached. If very much retuning is required, or if the amplifier loads with loose coupling between the two coils, it is an excellent indication that the line is not actually flat.

When a "swinging-link" assembly is used to obtain this type of coupling, the loading on the final amplifier can be adjusted to the desired value by varying the coupling between the two coils. The tank condenser, C, should be readjusted to minimum plate current each time the coupling is changed. If the desired loading cannot be obtained there is no alternative but to use a different coupling system.

The pick-up coil may be wound directly over the final tank coil, in which case the correct number of turns may be determined by trial. The insulation between the coils must be adequate for the plate voltage used, if the amplifier is series-fed.

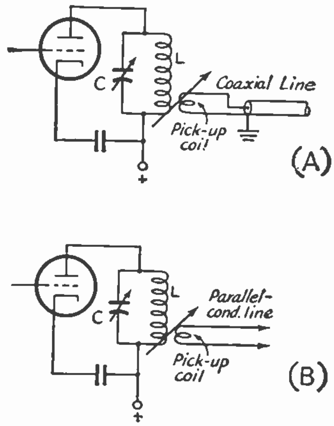


Fig. 10-23 — Using an untuned pick-up coil to couple to a transmission line. The method of adjustment is discussed in the text.

Series and Parallel Tuning

The circuits shown in Fig. 10-24 are useful with parallel-conductor lines operating at a relatively-high standing-wave ratio, particularly when the line length is such as to make the input impedance substantially a pure resistance. Assuming that the antenna is resonant, the optimum line lengths will be multiples of a quarter wavelength at the operating frequency. When the s.w.r. is high, the impedance at such points is considerably higher or considerably lower than the characteristic impedance of the line.

In these circuits the secondary, consisting of L_1, C_1 (and C_2, in the series circuit) and the input impedance of the line, is tuned to the operating frequency. As explained in Chapter Two, the degree of coupling between two resonant circuits is determined by their Qs, and it is necessary to keep the Qs fairly high (of the order of 10 or so). Assuming that the input impedance of the line is purely resistive, it can be inserted in series with the circuit (as in A) if its value is below about 100 ohms. The Q of the secondary circuit then can be brought to the proper value by making the reactance of L_1 of the order of 500 to 1000 ohms and setting the total capacitance of C_1 and C_2 to tune the circuit to resonance. With this type of tuning the current flowing into the line is rather large; in other words, the system is suitable for coupling into the line at a current loop.

On the other hand, if the line impedance is of the order of a few thousand ohms or more — which it will be at a voltage loop when the s.w.r. is high — the secondary circuit cannot be made to take power from the transmitter if the line resistance is inserted in series. The Q of the secondary circuit would be far too low to give adequate coupling. In such a case the parallel-tuned circuit at B may be used. As explained in Chapter Two, the Q of a circuit loaded by a parallel resistance will be equal to the resistance divided by the reactance of one of the tuned-circuit elements. Thus the reactance of L_1, in the parallel-tuned circuit, should not exceed a few hundred ohms at the operating frequency, to ensure a high-enough Q for good coupling.

If the input impedance of the line has reactance along with resistance, the reactance can be tuned out (within reasonable limits) by adjusting C_1 to compensate for it. This can easily be understood by reference to the section on canceling line reactance earlier in this chapter. The line will show reactance when its length is such that neither a voltage loop nor a current loop appears at the input end. There is a limit to how far the compensation can be carried, because at some line lengths the resistance component of the impedance has a value that is neither low enough to be inserted in series with the tuned secondary circuit, nor high enough to be placed in parallel with it. In such cases it is impossible to get adequate coupling between the final tank circuit and the line. In general, the value of the resistance component changes more slowly in the vicinity of a current loop than it does in the vicinity of a voltage loop. For this reason the series circuit is usable over a fair range of line lengths either longer or shorter than the length giving a current loop. There is less tolerance in the case of a voltage loop, so parallel tuning cannot be expected to give good results if the line length departs too far from the length giving a voltage loop at the input end.

Provided that both the final-tank circuit and the line-coupling circuit have Qs of the order of 10 or more, adequate power transfer can be obtained with fairly loose coupling between L and L_1. This is desirable, in that it aids in maintaining a balanced line and helps reduce the amount of harmonic energy transferred. The fact that the secondary circuit is tuned to the operating frequency is also greatly helpful in preventing harmonics from getting into the line.

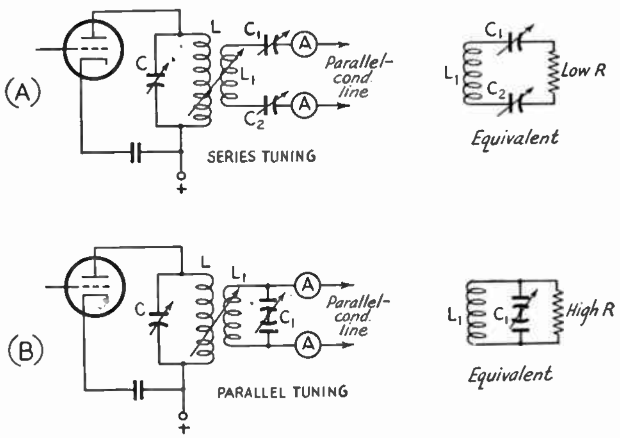


Fig. 10-24 — Series and parallel tuning. This method is particularly useful with resonant lines when the length is such as to bring either a current or voltage loop near the input end. Design data and methods of adjustment are given in the text.

Adjustment of Series Tuning

The tuning procedure with series tuning is as follows: With C_1 and C_2 at minimum capacitance, 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 that 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 that just brings the amplifier plate current to the rated value when C_1 and C_2 pass through resonance.

The values of inductance and capacitance to be used in the secondary circuit will depend largely on the frequency of operation. A coil of 3 or 4 turns (diameter 2 to 3 inches) will usually be adequate at 28 Mc., but 15 or 20 turns may be needed at 3.5 Mc. It is best to be able to adjust the number of turns on L_1 (by a tap), and there must of course be some means for varying the coupling between L and L_1. If the transmitter does not load properly, the secondary circuit may not be tuned to resonance; if it is found that maximum loading is secured with C_1 and C_2 at maximum capacitance, L_1 is too small and more turns must be used. If the maximum loading occurs with C_1 and C_2 at minimum, reduce the number of turns in L_1. In either case, make sure that C_1 and C_2 can be tuned *through* resonance so that the loading drops off at either higher or lower settings. Should it not be possible to get adequate loading even though the secondary circuit is resonated, increase L_1 and reduce C_1 and C_2 correspondingly, to raise the Q of the secondary circuit. If the practical limit of this process is reached and the transmitter output stage still does not load properly, the transmission-line length is not suitable for series tuning.

Two condensers are used in the series-tuned circuit in order to keep the line balanced to ground. This is because two identical condensers, both connected with either their stators or rotors to the line, will have the same capacitance to ground. A single condenser will slightly unbalance the circuit, since the frame has more capacitance to ground than the stator, but the unbalance is not serious unless the condenser is mounted near a large mass of metal, such as a chassis.

Adjustment of Parallel Tuning

Coupling and tuning adjustments with parallel tuning are carried out in much the same way as with series tuning. There is only one condenser to adjust, of course. Start with very loose coupling between L and L_1, resonate the secondary circuit by adjusting C_1 to make the final-amplifier plate current rise, then readjust C for minimum plate current. Increase the coupling in small steps, reresonating C_1 and C each time, until the desired loading is obtained.

Just as in the case of series tuning, it should be possible to tune *through* resonance with C_1. If the resonant point is at either maximum or minimum capacitance on C_1, change the number of turns on L_1 to bring the resonant point well on the condenser scale. In general, L_1 and C_1 will have about the same values as L and C, respectively, when the input impedance of the line is purely resistive. If the line shows reactance, the reactance can be tuned out, within limits, by adjustment of C_1 and, if necessary, by changing the number of turns on L_1 to achieve a combination that will permit the secondary circuit to resonate at the operating frequency.

If the input resistance of the line is very high, the secondary circuit will tune quite sharply. On the other hand, if the input resistance is relatively low the tuning will be broad and the resonance point will not be well marked. In such a case the number of turns in L_1 should be reduced and the capacitance of C_1 increased, to increase the Q of the circuit. This will permit power transfer with relatively loose coupling between L and L_1. Should it not be possible to load the transmitter properly with any combination of L_1 and C_1, the input resistance of the line is too low for parallel tuning.

In the parallel-tuned circuit C_1 is shown as a balanced or split-stator condenser. This type of condenser is used so that the system will be balanced to ground for stray capacitances. This is particularly desirable in the case of parallel tuning, because the voltage at the input end of the line is high, causing a relatively large current to flow through a small stray capacitance. An alternative scheme to maintain balance is to use two single-ended condensers in parallel, but with the frame of one connected to one side of the line and the frame of the other connected to the other side of the line. The same two condensers may be switched in series, as in Fig. 10-24A, when series tuning is to be used.

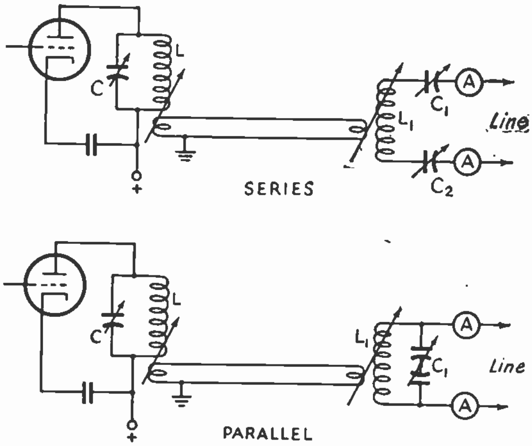


Fig. 10-25 — Link-coupled series and parallel tuning.

Link Coupling

The circuits shown in Fig. 10-24 require a means for varying the coupling between two sizable coils, a thing that is somewhat inconvenient constructionally. It is easier to use separate fixed mountings for the final tank and antenna coils and couple them by means of a link. As explained in Chapter Two, a short length of link line is equivalent to providing mutual inductance between two tuned circuits. Typical arrangements for series and parallel tuning are shown in Fig. 10-25. Although these drawings show variable coupling at both ends of the link circuit, a fixed link can be used at either end so long as a variable link is used at the other.

There is no essential difference between the tuning procedures with these circuits and those of Fig. 10-24. The only change is that the coupling is adjusted by means of a link instead of by varying the spacing between L and L_1.

"Universal" Antenna Couplers

An antenna-coupling system that is adaptable to a wide range of line input impedances can be constructed on the basis of the coupling principles described earlier. Combined with link coupling to the final tank circuit and provision for tuning out the input reactance of the line, such a system is suitable for working into either resonant or nonresonant lines, and introduces additional selectivity into the coupling system that helps discriminate against harmonics.

The circuit diagram is given in Fig. 10-26. The final tank is coupled to a second tuned circuit, L_1C_1, through a link. Taps are provided on L_1 so that the resistive component of the line impedance can be matched to the transmitter. To take care of cases where the input impedance of the line has a considerable reactive component, provision is made for switching in either a shunt capacitance or inductance, both of which are variable (see earlier discussion). The coupling should be variable at least at one end of the link circuit.

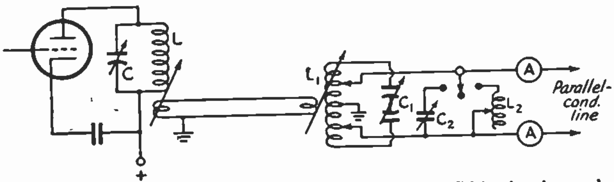


Fig. 10-26 — "Universal" antenna-coupling system. This circuit can be used with both resonant and nonresonant parallel-conductor lines.

In general, it is advisable to make the inductance of L_1 about the same as that of L, and to use for C_1 a condenser of the same capacitance as that used for C. The voltage rating for C_1 also should be the same as that of C. In other words, L_1C_1 may be a duplicate of LC for the operating frequency in use. The link coils can consist of two or three turns at each end. Provision should be made for tapping L_1 at frequent intervals — every turn, if possible. C_2 should have as large a maximum capacitance as is convenient — 250 to 500 $\mu\mu$fd. — but its voltage rating need not be high in the average case. For most installations where the power output does not exceed a few hundred watts a plate spacing of the order of 0.025 to 0.05 inch is sufficient. The inductance L_2 can consist of 20 or 25 turns approximately 2 inches in diameter and spaced 8 to 10 turns to the inch. The coil should be tapped every few turns.

The tuning procedure is as follows: First, disconnect the feeder taps on L_1 and use the loosest possible coupling, through the variable link coupling, to the final tank circuit. Tune C_1 until the plate current rises to a peak, indicating that L_1C_1 is resonated, and note the setting of C_1. Cut C_2 and L_2 out of the circuit and then connect the line taps across a turn or two at the center of L_1. Readjust C_1 to resonance, as indicated by a rise in plate current. It should be necessary to use closer coupling to get an observable change in plate current with the line connected. Note the new setting of C_1. If the capacitance is lower, switch in L_2 and find the tap that permits returning C_1 as nearly as possible to its original setting; if the capacitance is higher, switch in C_2 and adjust it to bring C_1 back to the original setting. Then increase the coupling, keeping C_1 at resonance as indicated by maximum plate current, and keeping C at resonance as indicated by minimum plate current. Continue until the minimum plate current reaches the desired load value. If C_1 flashes over as the coupling is increased, or if tuning C_1 back and forth a small amount either side of resonance makes it necessary to change the setting of C appreciably to maintain the final tank in resonance, the taps on L_1 are too close together. Move each tap one turn toward the ends of L_1, and again try increasing the coupling for rated load on the amplifier. When the proper loading is obtained, the tuning of L_1C_1 will be reasonably sharp, and changing the coupling will not necessitate more than "touching up" C to maintain resonance. If the taps on L_1 are too far apart the antenna tank circuit, L_1C_1, will be loaded heavily and its tuning will be broad. Under these conditions it may also be impossible to load the amplifier to rated plate current, even with the tightest available coupling. On the other hand, if the taps on L_1 are too close together the antenna tank will be too lightly loaded; its tuning will be critical and will affect the tuning of the plate tank circuit to a marked degree, and L_1 may overheat when the coupling is adjusted to make the amplifier take normal input.

When the reactive effects at the input end of the line are small, neither C_2 nor L_2 will be required. When this is the case, the setting of C_1 for resonance will not change much when the line is tapped on L_1. The greater the number of turns between the taps, the greater the detuning of the antenna tank by a given amount of reactance in the transmission-line input impedance.

This coupling system is equally effective with flat lines or those operating at a high s.w.r. If the line is actually flat, C_2 and L_2 will not be needed and the resonance setting of C_1 will not be affected by connecting the line. Regardless of the s.w.r., the positions of the line taps will depend on the resistive component of the line input impedance. If the resistance is low, the taps will be close together; if it is very high, the taps may have to be set right at the ends of L_1.

Coupling to Coaxial Lines

The principles of coupling to coaxial lines are just the same as for coupling to parallel-conductor lines. However, this type of line is unbalanced to ground, has inherently low impedance, and always should be operated with a low standing-wave ratio. The input impedance of a properly-operated coaxial transmission line therefore will be principally resistive, and of a value varying between perhaps 30 to 100 ohms, depending on the type of line and the s.w.r.

It is possible to couple such a line by means of a small coil inductively coupled to the final tank coil, as shown in Fig. 10-23A. The small amount of reactance introduced by the pick-up coil — and by the line, if the s.w.r. is slightly greater than 1 — can readily be tuned out by adjustment of the final tank condenser. However, additional selectivity is desirable for the purpose of reducing harmonic transfer from the final tank. Circuits are shown in Fig. 10-27. Except that it is adapted for single-ended rather than balanced operation, the circuit at A operates in much the same way as the circuit in Fig. 10-26. Also, because the load is known to be in the region of 100 ohms or less, it is possible to tap it across a capacity voltage divider (see earlier discussion) for impedance matching. This avoids the necessity for tapping L_1.

The circuit of Fig. 10-27B is similar in operation to that at A, but dispenses with the link circuit. For convenience, it uses a link coil on the final tank for inductive transfer of energy, the rest of the inductance in the antenna tank circuit being made up by L_1.

In the circuit at A, L_1 may be the same as L; in B, L_1 plus the pick-up coil should have about the same inductance as L. Except at perhaps 28 Mc., it is satisfactory, practically, to make L_1 the same as L in this circuit also, since the pick-up coil will not ordinarily have much inductance itself. In both circuits C_2 should have about the same capacitance as C, and C_1 should have approximately the value suggested in Fig. 10-27.

To adjust the circuit, set C_1 at maximum, loosen the coupling between L and the link or pick-up coil, and tune C_2 to resonance. This will be indicated, as usual, by a rise in the amplifier plate current. Adjust C to minimum plate current and increase the coupling in small steps, reresonating C_2 and C each time, until the amplifier plate current is normal. The loading on the antenna tank circuit is least when C_1 is at maximum capacitance, and increases when the capacitance of C_1 is decreased (with C_2 increased correspondingly to maintain resonance). The symptoms of under- and over-loading of the antenna tank are the same as described in connection with the universal antenna coupler. Adjust the loading by means of C_1, so that at normal plate input the antenna tank tuning is reasonably sharp and the setting of C is not greatly affected when C_2 is tuned a small amount either side of resonance.

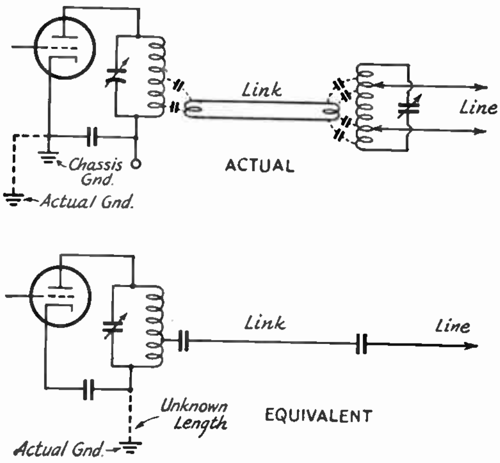


Fig. 10-28 — The stray capacitive coupling between coils in the upper circuit leads to the equivalent circuit shown below. The effect on the performance of the antenna system is discussed in the text.

Stray Coupling

In most of the circuits in Figs. 10-22 to 10-27, inclusive, a single-ended tank circuit has been indicated for the final amplifier. The amplifier itself has been shown only sketchily. The fact is that any type of antenna coupling circuit can be used with any type of amplifier — screen grid or neutralized triode, single-ended or push-pull. However, the actual arrangement, physically, of the circuit elements usually has an important bearing on the performance of the system. As it happens, a coupling system that is poorly designed, constructionally speaking, usually will do what it is supposed to do. But, equally important, it may do a lot of things it is *not* supposed to do.

Most of the unwanted effects that occur on transmission lines can be traced to stray capacitances in the system. Fig. 10-28 is an illustration. The upper drawing shows the ordinary link-coupled system as it might be used to couple into a parallel-conductor line. Inasmuch as a coil is a sizable metallic object, it will have capacitance to any other metallic objects in its vicinity, including other coils. Consequently there is capacitance between the final tank coil and its associated link coil, and between the antenna tank coil and its link. These capacitances

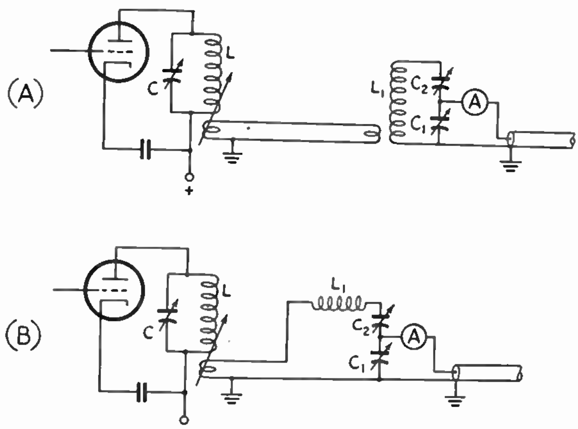


Fig. 10-27 — Coupling to coaxial lines. These circuits are used for harmonic suppression when working into a nonresonant coaxial line. Recommended capacitance values for C_1 are as follows: 28 Mc., 100 μμfd.; 14 Mc., 200 μμfd.; 7 Mc., 400 μμfd.; 3.5 Mc., 800 μμfd.

are small, but not negligible. In addition, the transmitter, particularly with metal-chassis construction, has appreciable capacitance to ground. Even if it did not, there is always a path from the transmitter to ground through the power wiring and the many stray capacitances associated with it.

There is a fundamental difference between the inductive coupling between coils and the capacitive coupling between them. Inductive coupling induces a voltage in the secondary coil that causes a current to flow, in common terminology, "around" the circuit. In Fig. 10-28, this means that the same current flows in both conductors of the link but, if the wires are parallel, the current flows in opposite directions in the two as it completes its travel around the loop. The same is true of the currents in the two conductors of the line. But with stray capacitive coupling the voltages at all points on the secondary coil are essentially in phase; for this type of coupling the secondary coil is just a mass of metal. Consequently, whatever current flows in the link (or in the line) flows in the *same* direction in both wires. Although both the link and line have two conductors and apparently form an ordinary go-and-return circuit, to the currents that flow as a result of capacitive coupling they simply look like a pair of conductors in parallel — in effect, that is, like a single conductor. The equivalent circuit is shown in the lower drawing in Fig. 10-28.

This single-wire circuit is an antenna system in itself, working in conjunction with a ground lead of unknown composition and length. It includes the regular antenna as well as the entire transmission line. If the various lengths happen to be just right, a fairly-large "parallel" current of this type can flow in it. This means that a considerable proportion of the total power output of the transmitter can be wasted in losses and radiation from a very undesirable sort of antenna system. Furthermore, despite the tuned tank circuits in the amplifier and antenna coupler, harmonic currents will flow in such an "antenna" even more readily than the fundamental current.

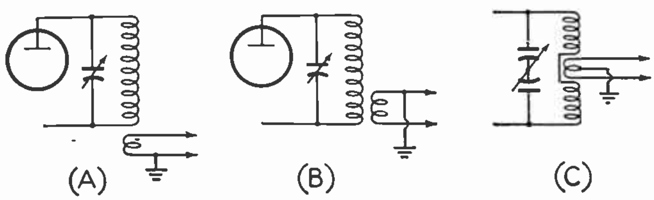


Fig. 10-29 — Methods of coupling and grounding link circuits to reduce energy transfer through stray capacitance.

There are other undesirable results, too. The fact that the power wiring becomes part of an "antenna" system means that the transmitter itself may perforce be at a considerable r.f. potential above ground. The chassis becomes "hot" with r.f., r.f. feed-back is prone to occur in speech equipment, and a considerable amount of r.f. power may be pumped into receiving and other equipment connected to the same a.c. power outlet. (A similar type of coupling in the input circuits of a receiver leads to stray pick-up of signals that may partially or completely mask the directive effects of the proper antenna.) On top of all this, it is impossible to tell much about the operation of the transmission line because the parallel current is more or less in phase with the regular line current in one wire and out of phase with it in the other. Thus the resultant currents in the two wires are unbalanced, and there is no way to separate the "parallel" and "line" currents in measurement.

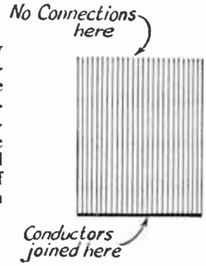


Fig. 10-30 — The Faraday screen. Stiff wire or small-diameter rod may be used for the conductors, separated by a distance about equal to the diameter of the wire or rod. The dimensions of the screen should be greater than the diameters of the coils to be shielded from each other.

These effects can only be eliminated if the stray capacitances are eliminated. However, they can be reduced by arranging the coils so the amount of energy coupled from the primary to the secondary is small, even though the capacitance itself still exists. This can be done by using a link coil that is physically small — that is, has few turns — and coupling it to the "cold" point on the tank coil. The cold point will be at the end of the coil that is grounded for r.f., either directly or through a by-pass condenser, in the case of single-ended tanks. In balanced tank circuits, the cold point is at the center. The coupling is further reduced if one side of the link circuit is grounded to the transmitter chassis as close as possible to the point where the tank itself is grounded. If the link is at the end of the tank coil the side farthest from the tank should be grounded, as indicated in Fig. 10-29A. If the link is wound *over* one end of the tank coil, ground the side toward the hot end of the tank, as indicated in Fig. 10-29B. With a balanced tank circuit the link should be at the center of the coil. In this case the best point to ground is the center of the link coil, but if this is impracticable good results will be secured by grounding either end of the coil. Ground directly to the chassis and keep the lead as short as possible.

This treatment of link circuits does not eliminate capacitive coupling. It simply makes it less troublesome, by making certain that the coupling occurs between parts of circuits that are not at high r.f. voltage. However, there are cases, particularly with balanced tank circuits, where the point on the tank coil that is cold for the fundamental frequency is hot at the even harmonics. This means that even though the transmitter and line behave properly on the fundamental frequency, harmonics still can be radiated at considerable intensity. The only way to be sure that these effects do not exist is to eliminate the stray capacitance entirely.

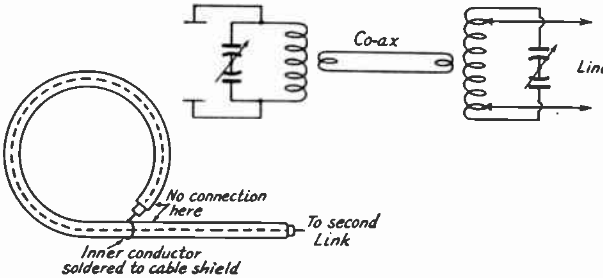


Fig. 10-32 — A shielded link coil constructed from coaxial cable. The smaller sizes of cable such as RG-59/U are most convenient, except when the coils have a diameter of 3 inches or more. For larger coils, RG-8/U or RG-11/U can be used.

Capacitive coupling between coils can be eliminated by means of a **Faraday** screen. This is a shield that prevents the electric field from one coil from reaching the other, but which has no effect on the magnetic field. As shown in Fig. 10-30, it consists of a group of parallel conductors, insulated from each other, and connected together at one end only. This forms an effective shield for the electric field, but since the conductors are open-circuited the voltages induced in them by the magnetic field cannot cause any current to flow. (Such current flow is essential to magnetic shielding with nonmagnetic materials, as explained in Chapter Two.)

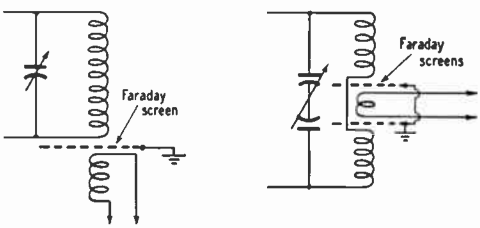


Fig. 10-31 — Installation of Faraday screens to eliminate capacitive coupling between coils.

The Faraday screen should be somewhat larger than the diameter of the coils with which it is used. It is simply mounted between the two coils that are to be shielded from each other, and then grounded to the chassis through a short lead, as indicated in Fig. 10-31. In the case of a balanced tank circuit with a swinging link, two shields must be used, one on each side of the link coil. In the case of fixed links wound over the tank coil, a satisfactory screen can be made by using several turns of the same type of coil, cutting them parallel to the axis to open-circuit the conductors, and then soldering them together at one end only. This shield can then be inserted between the tank coil and link, making sure that it is adequately insulated from both.

An alternative, and perhaps simpler, type of screening is shown in Fig. 10-32. In this case the inner conductor of a piece of coaxial cable is used to form a one-turn link. The outer conductor serves as an open-circuited shield around the turn, this shield being grounded to the chassis. The circuit to the link line is made by connecting the inner conductor to the outer conductor at the finish of the turn, as shown, and from there on the coaxial line is used to transfer the power to a second, and similar, link coil at the antenna tuner. This type of shielded link is simpler to make than the regular Faraday screen.

Aside from the adverse effects on the performance of the antenna system, stray capacitive coupling frequently is responsible for interference to near-by broadcast receivers. It is not difficult to appreciate that radiation taking place from transmission lines and power wiring is, in general, more likely to get into a broadcast receiver than radiation from an antenna that is intentionally kept away from other antennas — particularly when the receivers are connected to that same power wiring.

Antenna-Coupler Construction

The apparatus used to cancel line reactance and match the line resistance to the transmitter is commonly called an "antenna coupler" or "antenna tuner." (The name is really a misnomer, because the coupling and tuning equipment at the input end of the line does not have any effect on the antenna itself; if there is any antenna tuning to be done it must be done at the antenna, independently of the line.) The design principles and the important constructional points have been covered earlier in this chapter; in this section we show a few examples of typical construction.

Bearing in mind the precautions mentioned earlier as to maintaining balance in parallel-conductor transmission lines, it is usually good practice to install the coupling equipment close to the point where the line enters the station. This is a simple matter when the tuning equipment is link-coupled to the trans-

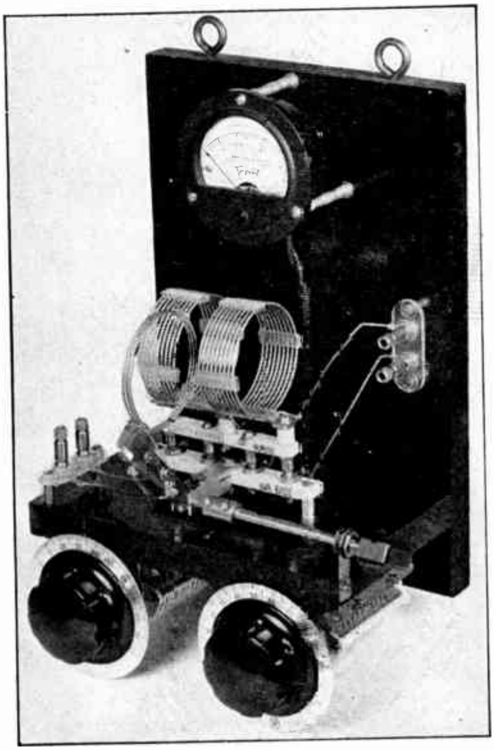


Fig. 10-33 — A wall-mounting antenna coupler for medium-power transmitters. This unit provides a choice of either series or parallel tuning for resonant feeders. Standard transmitting coils of the variable-link type are used.

mitter, since there are no particular restrictions on the length of the link that can be used. However, if the link line is fairly long it should be treated as a transmission line rather than merely as a means of providing mutual inductance between two separated coils. In such a case it is advisable to have variable coupling at both ends of the link. This permits matching the link line to the line tank circuit, and once the match is obtained the power output of the transmitter can be varied by changing the coupling at the transmitter tank. If the link line is not properly matched its current may be excessive, leading to unnecessary power loss.

The most desirable form of link line is coaxial cable. Properly handled, its losses are low; and since it is shielded it can be on or near metal objects with impunity.

SERIES-PARALLEL COUPLER FOR WALL MOUNTING

Fig. 10-33 shows a link-coupled coupler designed for series or parallel tuning of a resonant line. It is suitable for transmitters having a power output in the neighborhood of 250 watts. A higher-power version easily could be made using a similar layout, but substituting heavier coils and condensers with greater plate spacing.

As shown in Fig. 10-34, the change from series to parallel tuning is made by means of jumpers and extra pins on the coil plug bar. A separate coil is used for each band, and after determining which should be used, series or parallel tuning, on a particular band, the jumpers may be installed permanently or left off as required. The tuning condensers specified, together with a set of standard plug-in transmitting coils, should provide adequate coupling if the transmission-line length is such as to bring a voltage or current loop near the input end.

The unit is mounted on an 8 × 12 × ⅞-inch board for hanging on the wall in any convenient location near the entrance point of the feeders. The 2.5-ampere r.f. ammeter is mounted centrally by long wood screws through spacers at the top of the unit. A short length of twisted pair connects it to the thermocouple, secured in a horizontal position at the bottom of the backboard. The tuning condensers are mounted on the underside of a 4-inch shelf extending the width of the unit. Atop the shelf, the jack bar for the coil is supported on pillars by wood screws. An extension shaft to vary the degree of coupling is supported by a bushing fastened to a short strip of brass at the right of the shelf. A short length of 300-ohm ribbon (coaxial cable can be used instead) connects the input terminals to the movable link, while the output terminals are located at the middle right of the backboard. Two screw eyes at the top permit the unit to be hung from screws or nails in the wall.

RACK-MOUNTING SERIES-PARALLEL COUPLER

The rack-mounting coupling unit shown in Fig. 10-35 is suitable for power outputs of 25 to 50 watts, and provides either series or parallel tuning for resonant lines. Separate condensers are used for this purpose, and while

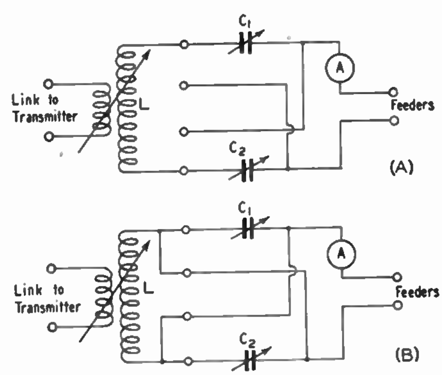


Fig. 10-34 — Circuit diagram of an antenna coupler for use with a medium-power transmitter. A — Series tuning. B — Parallel tuning.

C_1, C_2 — 100-μμfd. single section variable, 0.070-inch spacing (Cardwell MT-100-GS).

L — B & W BVL series.

A — 0–2.5 thermocouple r.f. ammeter.

three are required, this system has the advantage that no switching is necessary when changing from series to parallel tuning. It is also possible to cover a somewhat wider range of line input impedances with parallel tuning because the series condensers can be used to help cancel out inductive reactance that cannot be handled by the parallel circuit alone.

The coupler is mounted on a 5¼ × 19-inch panel. The parallel condenser, C_1, is in the center, with C_2 and C_3 on either side. The variable condensers are mounted on National GS-1 stand-off insulators which are fastened to the condenser tie-rods by means of machine screws with the heads cut off. Small ceramic shaft couplings are used to insulate the control knobs from the condenser shafts.

Clips with flexible leads attached are provided for the parallel condenser, C_1, so that the sections may be used either in series or parallel to form either a high-C or low-C tank circuit, as required. When the high-C tank is necessary the two stators are connected together by means of the clips, as indicated by the dotted lines in the circuit diagram, Fig. 10-36. When the two sections are connected in series for low-C operation the breakdown voltage is increased.

Two sets of variable condensers are suggested in the list of parts. The smaller receiving-type condensers with 0.03-inch air gap are satisfactory for transmitter power outputs up to 50 watts. The larger condensers, with 0.045-inch spacing, are required for transmitter outputs of the order of 100 watts.

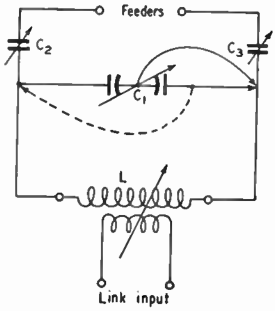


Fig. 10-36 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers that are operated at less than 1000 volts on the plate.

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_1 — 100 μμfd. per section, 0.045-inch spacing (National TMK-100-D) for high voltages; receiving type for low voltages (Hammarlund MCD-100).

C_2, C_3 — 250 μμfd., 0.026-inch spacing (National TMS-250) for high voltages; receiving type for low voltages (Hammarlund MC-250).

L — B & W JVL-series coils. Approximate dimensions for parallel tuning for each band are as follows:
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.

BANDSWITCHING UNIVERSAL COUPLER

The coupling unit shown in Figs. 10-37 and 10-39 is of the "universal" type discussed earlier. It is a bandswitching unit using commercially-available coils. Provision is made for switching either capacitance or inductance across the transmission line to compensate for its input reactance. Impedance matching is achieved by tapping the tank coils at the proper points.

In the circuit diagram, Fig. 10-38, only one set of coils is shown. For other bands the connections shown for L_1 and L_2 would be duplicated. Bandswitching is accomplished by a five-gang switch, S_1. Compensating reactances can be switched in or out of the circuit by S_2. The coupling links, L_2, are the shielded type using coaxial cable described earlier in this chapter (Fig. 10-32).

The coupler is wholly supported by a 7 × 19-inch relay-rack panel. The variable condensers are mounted from the panel by small stand-off insulators, and insulated couplings are used between the condenser shafts and the National Type AM dials. The tank condenser, C_1, is mounted at the right-hand end of the panel with the bandswitch, S_1, to its left. The four coils are grouped around the bandswitch, with the 28-Mc. coil placed so that the leads to it are the shortest. The coils are Millen 44000 series with the plug bases removed from the

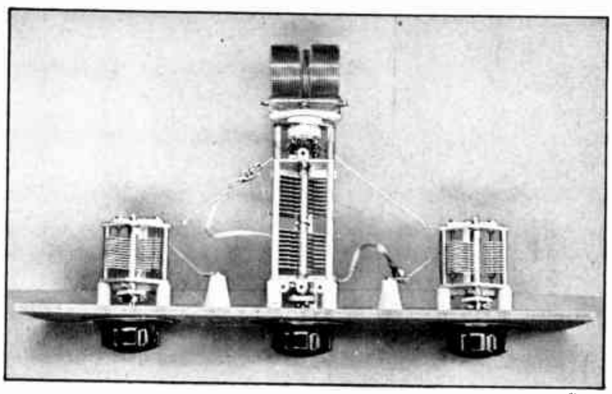


Fig. 10-35 — Rack-mounted coupler for low-power transmitters. This unit uses three variable condensers to provide either series or parallel tuning without condenser switching.

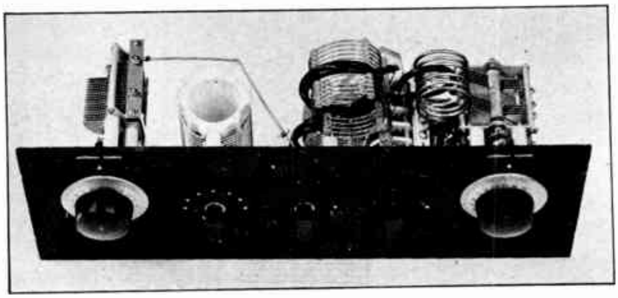


Fig. 10-37 — Bandswitching universal-type coupler for parallel-conductor lines. This unit can be used with transmitters having power outputs of the order of 100 watts.

3.5-, 7- and 14-Mc. coils. It is not practicable to remove the base from the 28-Mc. coil because it does not have the polystyrene supporting strip that is part of the lower-frequency coil

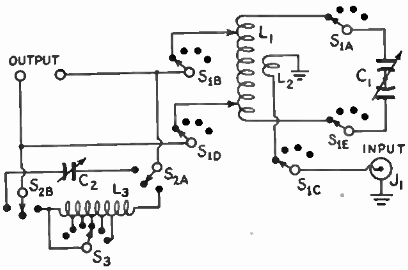


Fig. 10-38 — Circuit diagram of the bandswitching coupler. In this diagram the ground symbol indicates points that are connected together. Wiring to coils is shown for one band only, to avoid complicating the diagram; the wiring for other coils is identical.

C_1 — 100-μμfd.-per-section variable (Cardwell MR-100-BD).
C_2 — 335-μμfd. variable (Cardwell MR-335-BS).
L_1 — Millen 44000-series coils (see text).
L_2 — Shielded link; one turn for 28 and 14 Mc.; 2 turns for 7 and 3.5 Mc.
L_3 — 26 turns No. 12 on 2½-inch diameter form (National XR-10A), 7 turns per inch. Tapped 8, 14, 18, 22 and 24 turns from end to which arm of S_3 is connected.
J_1 — Coaxial-cable connector (Amphenol).
S_1 — 5-section 4-position ceramic wafer switch (Centralab 2546).
S_2 — 2-section 4-position ceramic wafer switch (Centralab 2543).
S_3 — 1-section 6-position ceramic wafer switch (Centralab 2501).

assemblies. The coils are partly supported by the wiring to the switch and partly by the polystyrene plate mounted on the back of the switch. The ends of the coil mounting strips are cemented into holes cut in the plate.

The compensating condenser, C_2, is mounted at the left-hand end of the panel. L_3 is mounted vertically to its right, with S_3 directly in front of it on the panel. S_2 is mounted centrally on the panel. The output terminals to the line are mounted above S_3. The link input terminal is a coaxial cable socket mounted on a small bracket in the lower right-hand corner.

The link coils, L_2, are supported by the wiring, and the coupling is changed by bending the link into or out of its associated tank coil. Since the links fit rather tightly in the tank coils, the pressure helps hold them in place once the proper coupling is determined. The link shields are all connected together and to the input connector; the inner conductors go to the switch contacts. The link coils are made from RG-59/U cable.

With the coils and condensers specified, this coupler can handle power outputs of the order of 100 to 150 watts. The method of adjustment is covered earlier in this chapter.

A WIDE-RANGE ANTENNA COUPLER

The photograph of Fig. 10-40 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 the various circuit combinations possible with this arrangement are given in Fig. 10-41.

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

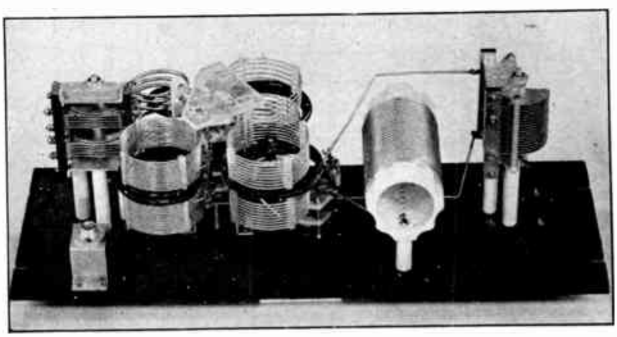


Fig. 10-39 — Rear view of the bandswitching coupler. Details of coil mountings are shown in this view.

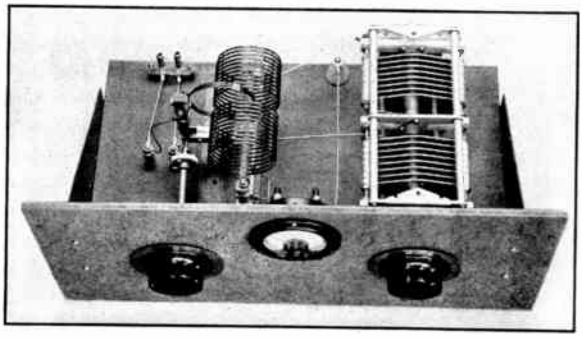


Fig. 10-40 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel 8¾ × 19 inches in size. The variable condenser is a split-stator unit with a capacitance of 200 $\mu\mu$fd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B & W TVL series. The r.f. ammeter has a 4-ampere scale.

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. 10-41.

The tuning condenser specified, together with a set of standard plug-in transmitting coils, should cover nearly 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 having an output of 500 watts with plate modulation and somewhat more on c.w. For higher-power 'phone, a tank condenser with larger plate spacing should be used.

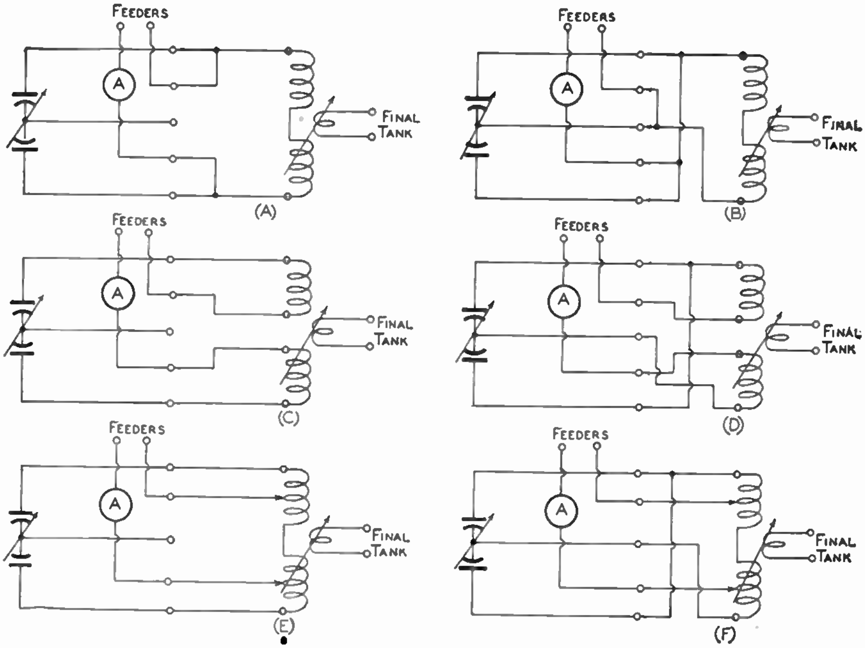


Fig. 10-41 — Circuit diagram of the wide-range rack-type antenna coupler. A — Parallel tuning, low *C*. B — Parallel tuning, high *C*. C — Series tuning, low *C*. D — Series tuning, high *C*. E — Parallel tank, low-impedance output, low *C*. F — Parallel tank, low-impedance output, high *C*. After the 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 only to plug in the right coil for each band, tune the condenser for resonance, and adjust the link loading.

Antennas

In selecting the type of antenna to use, the propagation characteristics of the frequency band or bands to be used should be given due consideration. These are outlined in Chapter Four. In general, antenna construction and location become more critical and important on the higher frequencies. On the lower frequencies (3.5 and 7 Mc.) the angle of radiation and plane of polarization may be of relatively little importance; at 28 Mc. and higher 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, directivity and gain.

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 in its direction of maximum radiation (broadside). The wave from an antenna in a slanting position contains both horizontal and vertical components.

Angle of Radiation

The wave angle (or vertical angle) 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 generally 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 that point.

Impedance

The impedance of the antenna at any point is the ratio of the voltage to the current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load represented by the antenna. It is a pure resistance only at current loops (maxima) and nodes (minima) on resonant antennas. The antenna impedance is high at the current node and low at the current loop.

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. When there are standing waves on an antenna, the parts of the wire carrying the higher current have the greater radiating effect. All resonant antennas have standing waves — only terminated types, like the terminated rhombic and terminated "V," have substantially uniform current along their lengths.

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 field is measured in the optimum direction of the antenna under test. In amateur work, the comparison antenna is generally a half-wave antenna at the same height and having the same polarization as the antenna under consideration. Power gain usually is expressed in decibels.

Front-to-Back Ratio

In unidirectional beams (antenna systems with maximum radiation in only one direction) the front-to-back ratio is the ratio of power radiated in the maximum direction to power radiated in the opposite direction. It is also a measure of the reduction in received signal when the beam direction is changed from that for maximum response to the opposite direction. Front-to-back ratio is usually expressed in decibels.

Ground Effects

The radiation pattern of any antenna that is many wavelengths distant from the ground and all other objects is called the **free-space pattern** of that antenna. The free-space pattern of an antenna is almost impossible to obtain in practice, except in the v.h.f. and u.h.f. ranges. Below 30 Mc., the location of the antenna with respect to ground plays an important part in determining the actual radiation pattern of the antenna.

When any antenna is near the ground the free-space pattern 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 effect of a perfectly-reflecting ground is such that the

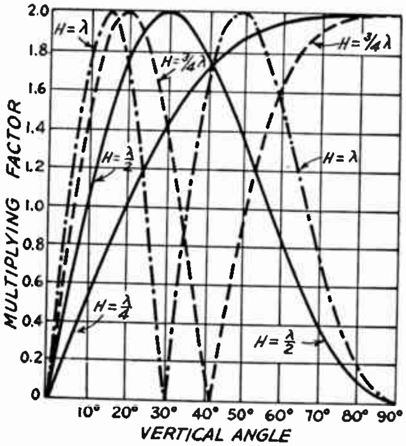


Fig. 10-42 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.

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. 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. 10-42 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.

Radiation Angle

The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies. It is advantageous, therefore, to erect the antenna at a height that 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 one-half wavelength at 14 Mc., and preferably three-quarters or one wavelength; at least one wavelength, and preferably higher, at 28 Mc. and the very-high frequencies. The physical height required for a given height in wavelengths 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 being preferable.

Imperfect Ground

Fig. 10-42 is based on ground having perfect conductivity, whereas the actual earth is not a perfect conductor. 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 over horizontal ground. 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 that 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. 10-43. 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 or quarter-wave antenna will radiate

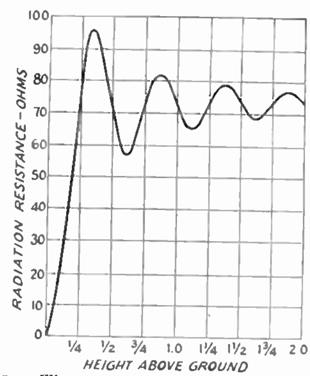


Fig. 10-43 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

equally well in all *horizontal* directions, so that it is substantially nondirectional, 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.

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.

The Half-Wave Antenna

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 variously known as a **half-wave dipole, half-wave doublet, or Hertz** antenna.

The length of a half-wavelength in space is:

$$\textit{Length (feet)} = \frac{492}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-H)}$$

The actual length of a half-wave antenna will not be exactly equal to the half-wave in space, but depends upon the thickness of the conductor in relation to the wavelength as shown in Fig. 10-44, where K is a factor that must be multiplied by the half-wavelength in free space to obtain the resonant antenna length. An additional shortening effect occurs with wire antennas supported by insulators at the ends because of the capacitance added to the system by the insulators **(end effect).** The following formula is sufficiently accurate for wire antennas at frequencies up to 30 Mc.:

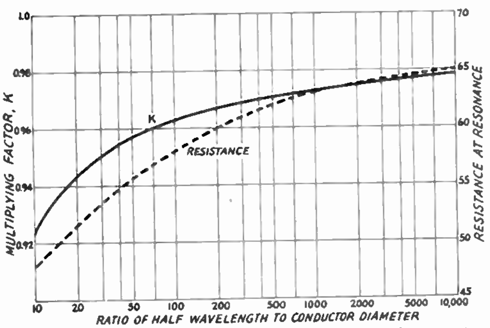


Fig. 10-44 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half-wavelength (Equation 10-H). The effect of conductor diameter on the impedance measured at the center also is shown.

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times 0.95}{\textit{Freq. (Mc.)}} = \frac{468}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-I)}$$

Example: A half-wave antenna for 7150 kc. (7.15 Mc.) is $\frac{468}{7.15} = 65.45$ feet, or 65 feet 5 inches.

Above 30 Mc. the following formulas should be used, particularly for antennas constructed from rod or tubing. K is taken from Fig. 10-44.

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times K}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-J)}$$

$$\textit{or length (inches)} = \frac{5905 \times K}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-K)}$$

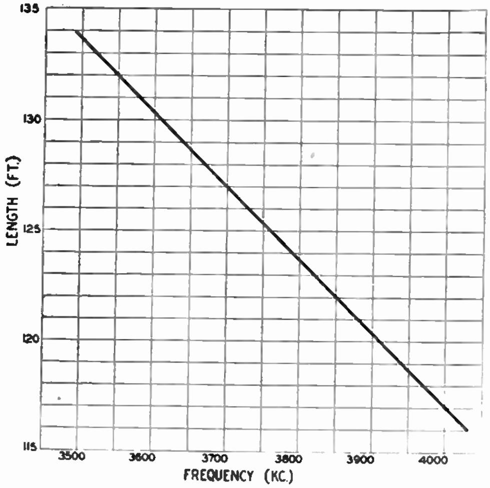
Example: Find the length of a half-wavelength antenna at 29 Mc., if the antenna is made of 2-inch diameter tubing. At 29 Mc., a half-wavelength in space is $\frac{492}{29} = 16.97$ feet, from Eq. 10-H. Ratio of half-wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2} = 101.8$. From Fig. 10-44, $K = 0.963$ for this ratio. The length of the antenna, from Eq. 10-J, is $\frac{492 \times 0.963}{29} = 16.34$ feet, or 16 feet 4 inches. The answer is obtained directly in inches by substitution in Eq. 10-K: $\frac{5905 \times 0.963}{29} = 196$ inches.

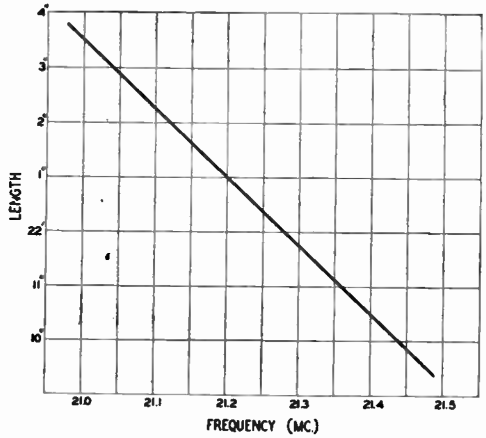
Current and Voltage Distribution

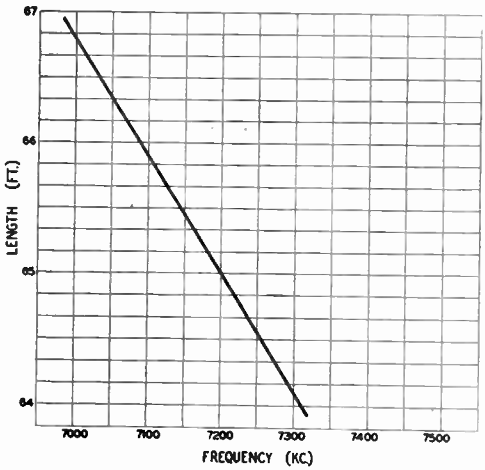
When power is fed to such an antenna, the current and voltage vary along its length. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage. The current does not actually reach zero at the current nodes, because of the end effect; similarly, the 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.** The radiation resistance is an *equivalent* resistance, a convenient conception to indicate the radiation properties of an antenna. The radiation resistance is the equivalent resistance that would dissipate the power the antenna radiates, with a current flowing in it equal to the antenna current at a current loop (maximum). The ohmic resistance of a half-wavelength antenna is ordinarily small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

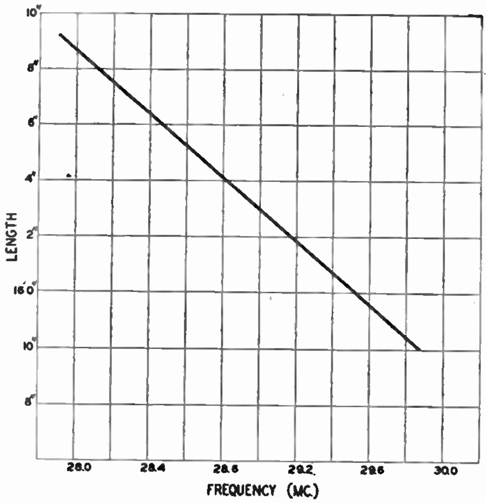
Impedance

The radiation resistance of an infinitely-thin 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 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









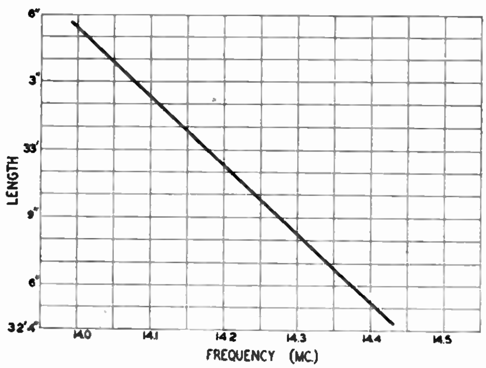


Fig. 10-45 — The above charts, based on Eq. 10-I, can be used to determine the length of a half-wave antenna of wire.

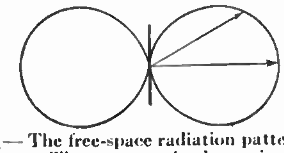


Fig. 10-46.— The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining 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.

is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends. The actual value at the ends will depend on a number of factors, such as the height, the physical construction, the insulators at the ends, 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 the wavelength, as shown in Fig. 10-44. If the diameter of the conductor is made large, the capacitance per unit length increases and the inductance per unit length decreases. Since the radiation resistance is affected relatively little, the decreased L/C ratio causes the Q of the antenna to decrease, so that the resonance curve becomes less sharp. Hence, the antenna is capable of working over a wide frequency range. This effect is greater as the diameter 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 perpendicular to the wire and zero along the direction of the wire, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 10-46, 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 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. The variation in radiation at various vertical angles from a half-wavelength horizontal antenna is indicated in Figs. 10-47 and 10-48.

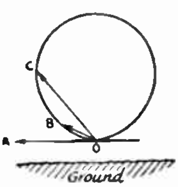


Fig. 10-47 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing of the free-space field pattern of a horizontal antenna.

FEEDING THE HALF-WAVE ANTENNA

Direct Feed

If possible, it is advisable to locate the antenna at least a half-wavelength from the transmitter and use a transmission line to carry the power from the transmitter to the antenna. However, in many cases this is impossible, particularly on the lower frequencies, and direct feed must be used. Three examples of direct feed are shown in Fig. 10-49. In the method shown at A, C_1 and C_2 should be about 150 μμfd. each for the 3.5-Mc. band, 75 μμfd. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil

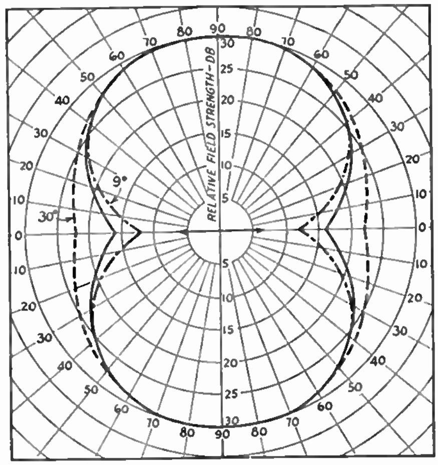


Fig. 10-48 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude 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.

connected between them should resonate to 3.5 Mc. with about 60 or 70 μμfd., for the 80-meter band, for 40 meters it should resonate with 30 or 35 μμfd., and so on. The circuit is adjusted by using loose coupling between the antenna coil and the transmitter tank coil and adjusting C_1 and C_2 until resonance is indicated by an increase in plate current. The coupling between the coils should then be increased until proper plate current is drawn. It may be necessary to reresonate the transmitter tank circuit as the coupling is increased, but the change should be small.

The circuits in Fig. 10-49B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the

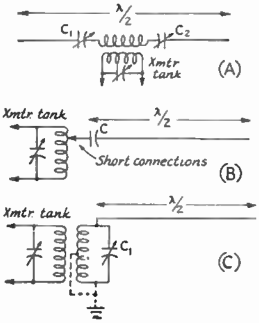


Fig. 10-49 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitive coupling; C, voltage feed, with inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

tap toward the "hot" or plate end of the tank coil — the condenser C may be of any convenient value that will stand the voltage, and it doesn't have to be variable. In the circuit at C, the antenna tuned circuit (C_1 and the antenna coil) should be similar to the transmitter tank circuit. The antenna tuned circuit is adjusted to resonance with the antenna connected but with loose coupling to the transmitter. Heavier loading of the tube is then obtained by tightening the coupling between the antenna coil and the transmitter tank coil.

Of the three systems, that at A is preferable because it is a symmetrical system and generally results in less r.f. power "floating" around the shack. The system of B is undesirable because it provides practically no protection against the radiation of harmonics, and it should only be used in emergencies.

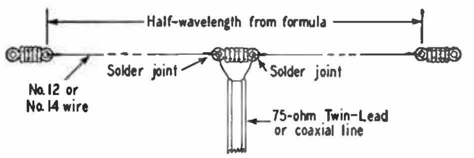


Fig. 10-50 — Construction of a half-wave doublet fed with 75-ohm line. The length of the antenna is calculated from Equation 10-I or Fig. 10-45.

Transmission-Line Feed for Half-Wave Antennas

Since the impedance at the center of a half-wavelength antenna is in the vicinity of 75 ohms, it offers a good match for 75-ohm two-wire transmission lines. Several types are available on the market, with different power-handling capabilities. They can be connected in the center of the antenna, across a small strain insulator to provide a convenient connection point. Coaxial line of 75 ohms impedance can also be used, but it is heavier and thus not as convenient. In either case, the transmission line should be run away at right angles to the antenna for at least one-quarter wavelength, if possible, to avoid current unbalance in the line caused by pick-up from the antenna. The antenna length is calculated from Equation 10-I, for a half-wavelength antenna. When No. 12 or No. 14 enameled wire is used for the antenna, as is generally the case, the length of the wire is the over-all length measured from the loop through the insulator at each end. This is illustrated in Fig. 10-50.

The use of 75-ohm line results in a "flat" line over most of any amateur band. However, by making the half-wave antenna in a special manner, called the **two-wire** or **folded doublet**, a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 10-51, with another version in Fig. 10-84B. The two differ only in the construction of the antenna

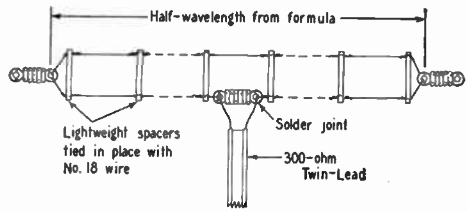


Fig. 10-51 — The construction of an open-wire folded doublet fed with 300-ohm line. The length of the antenna is calculated from Equation 10-I or Fig. 10-45.

proper. The open-wire line shown in Fig. 10-51 is made of No. 12 or No. 14 enameled wire, separated by lightweight spacers of Lucite or other material (it doesn't have to be a *low-loss* insulating material), and the spacing can be on the order of from 4 to 8 inches, depending upon what is convenient and what the operating frequency is. At 14 Mc., 4-inch separation is satisfactory, and 8-inch or even greater spacing can be used at 3.5 Mc.

If a half-wavelength antenna is fed at the center with other than 75-ohm line, or if a folded doublet is fed with other than 300-ohm line, standing waves will appear on the line and coupling to the transmitter may become awkward for some line lengths, as described earlier in this chapter. However, in many cases it is not convenient to feed the half-wave antenna with the correct line (as is the case where multiband operation of the same antenna is desired), and sometimes it is not convenient to feed the antenna at the center. Where multiband operation is desired (to be discussed later) or when the antenna must be

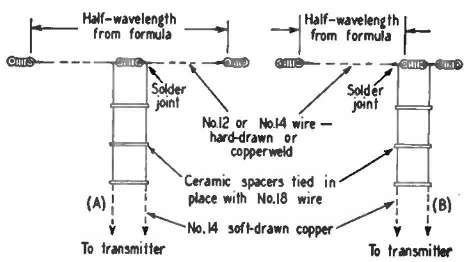


Fig. 10-52 — The antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 10-I or Fig. 10-45.

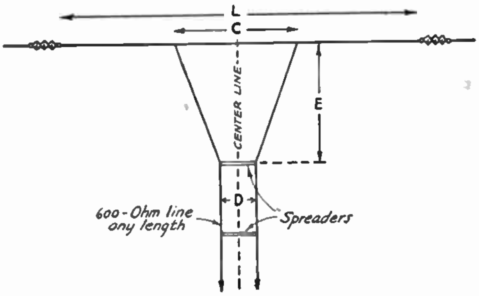


Fig. 10-53 — Delta-matched antenna system. The dimensions *C*, *D*, and *E* are found by formulas given in the text. It is important that the matching section, *E*, come straight away from the antenna without any bends.

fed at one end by a transmission line, an open-wire line of from 450 to 600 ohms impedance is generally used. The impedance at the end of a half-wavelength antenna is in the vicinity of several thousand ohms, and hence a standing-wave ratio of 4 or 5 is not unusual when the line is connected to the end of the antenna. It is advisable, therefore, to keep the losses in the line as low as possible. This requires the use of ceramic or Micalex feeder spacers, if any appreciable power is used. For low-power installations in dry climates, dry wood spacers that have been boiled in paraffin are satisfactory. Mechanical details of half-wavelength antennas fed with open-wire lines are given in Fig. 10-52. If the power level is low, below 100 watts or so, 300-ohm Twin-Lead can be used in place of the open line.

One method for offering a match to a 600-ohm open-wire line with a half-wavelength antenna is shown in Fig. 10-53. The system is called a **delta match.** The line is "fanned" as it approaches the antenna, to have a gradually-increasing impedance that equals the antenna impedance at the point of connection. The dimensions are fairly critical, but careful measurement before installing the antenna and matching section is generally all that is necessary. The length of the antenna, *L*, is calculated from Equation 10-I or Fig. 10-45. The length of section *C* is computed from:

$$C \text{ (feet)} = \frac{118}{Freq. \text{ (Mc.)}} \qquad \textbf{(10-L)}$$

The feeder clearance, *E*, is found from

$$E \text{ (feet)} = \frac{148}{Freq. \text{ (Mc.)}} \qquad \textbf{(10-M)}$$

Example: For a frequency of 7.1 Mc., the length

$L = \frac{468}{7.1} = 65.91$ feet, or 65 feet 11 inches.

$C = \frac{118}{7.1} = 16.62$ feet, or 16 feet 7 inches.

$E = \frac{148}{7.1} = 20.84$ feet, or 20 feet 10 inches.

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires 4¾-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or 3¾-inch spaced No. 16 wire.

Long-Wire Antennas

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. 10-54 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-

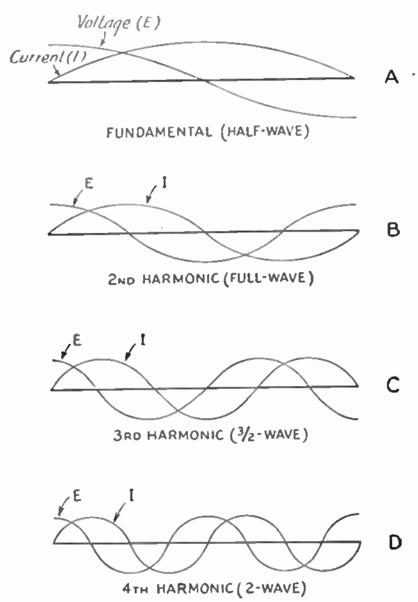


Fig. 10-54 — Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.

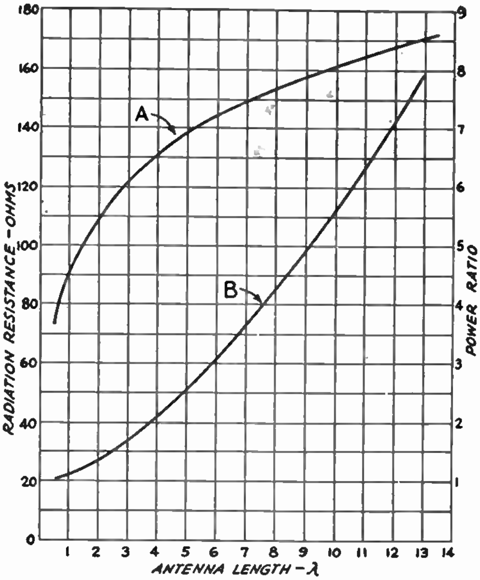


Fig. 10-55 — Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

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 multiband operation with one antenna.

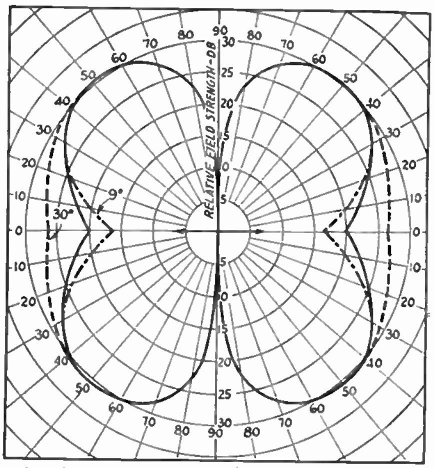


Fig. 10-56 — Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale: actual amplitudes will depend upon the height of the antenna.

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 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

$$\textit{Length}\ (\text{feet}) = \frac{492\,(N-0.05)}{\textit{Freq.}\ (\text{Mc.})} \qquad \textbf{(10-N)}$$

where N is the number of *half*-waves on the antenna.

Example: An antenna 4 half-waves long at 14.2 Mc. would be $\frac{492\ (4\ -\ 0.05)}{14.2} = \frac{492 \times 3.95}{14.2}$ = 136.7 feet, or 136 feet 8 inches.

It is apparent that an antenna cut as a half-wave for a given frequency will be slightly off

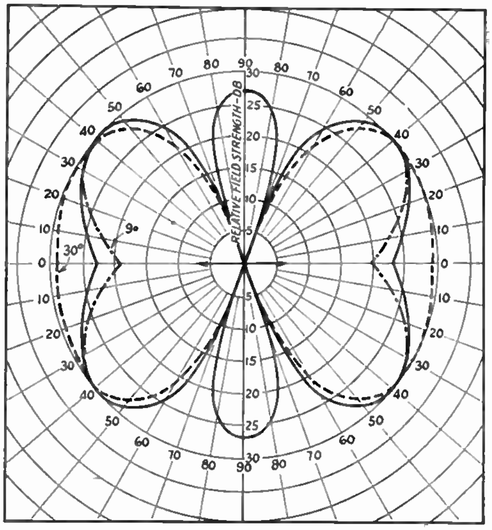
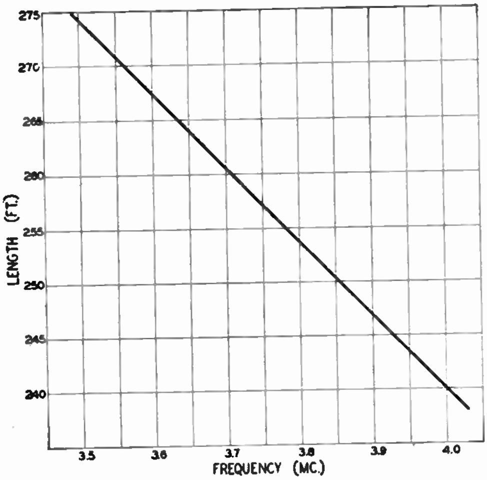


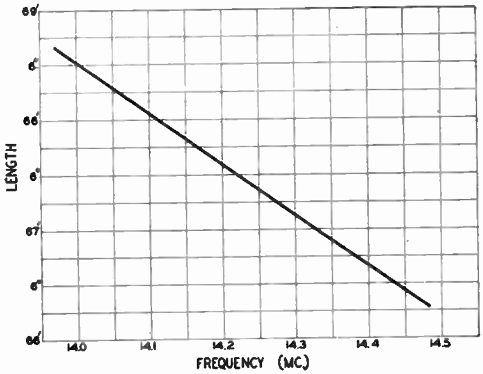
Fig. 10-57 — Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

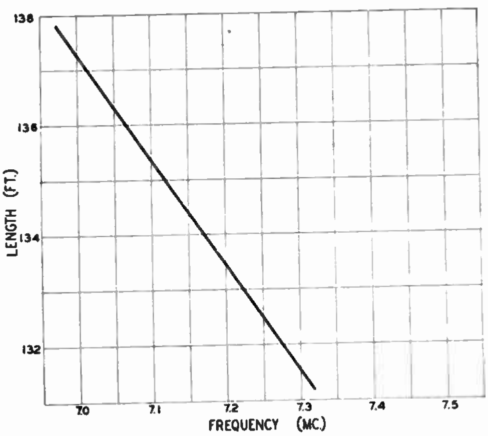
resonance at exactly twice that frequency (the second harmonic), because of the decreased influence of the end effects when the antenna is more than one-half wavelength long. The effect is not very important, except for a possible unbalance in the feeder system and consequent radiation from the feedline. If the antenna is fed in the exact center, no unbalance will occur at any frequency, but end-fed systems will show an unbalance in all but one frequency band, the band for which the antenna is cut.

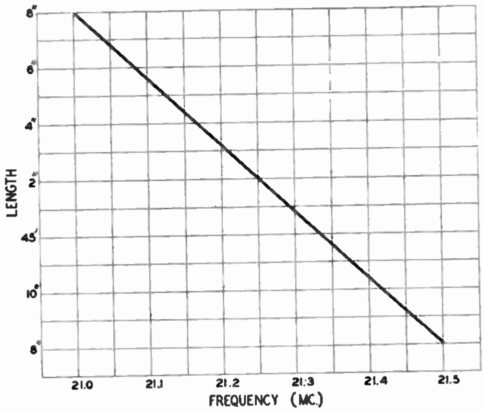
Impedance and Power Gain

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 di-









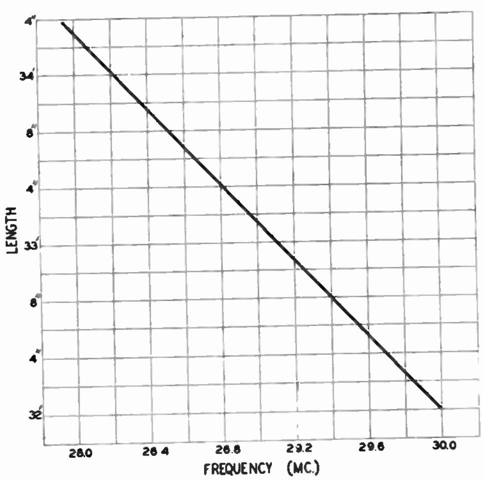


Fig. 10-58 — Charts for determining the over-all length of a one-wavelength antenna of wire, based on Eq. 10-N.

rection 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. 10-55 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. 10-56, 10-57 and 10-59, 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. 10-54. 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.

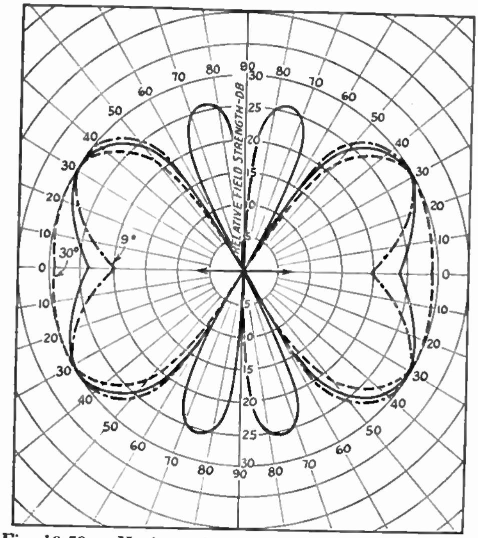


Fig. 10-59 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

No point on a long-wire antenna offers a reasonable impedance for a direct match to any of the common types of transmission lines. The most common practice is to feed the antenna at one end or at a current loop with a low-loss open-wire line and accept the resulting standing-wave ratio of 4 or 5. When a better match is required, "stubs" are generally used (described later in this chapter).

Multiband Antennas

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 nonresonant 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.

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 that 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. It is also wise not to attempt to use a half-wave antenna center-fed with coaxial cable on its harmonics. Higher-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

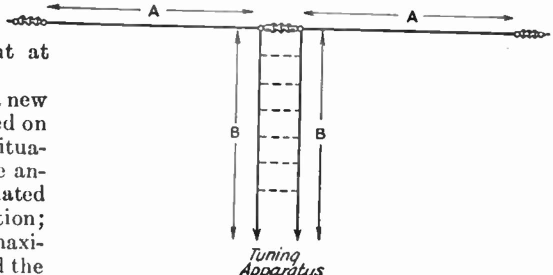


Fig. 10-60 — Practical arrangement of a shortened antenna. The total length, $A + B + B + A$, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table 10-III for lengths and tuning data.

TABLE 10–II
Multiband Resonant-Line Fed Antennas

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed: 120	60	4-Mc. 'phone	series
136	67	3.5-Mc. c.w. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
134	67	3.5-Mc. c.w. 7 Mc.	series parallel
67	33	7 Mc. 14 Mc. 28 Mc.	series parallel parallel
With center feed: 137	67	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel parallel
67.5	34	7 Mc. 14 Mc. 28 Mc.	parallel parallel 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 (3500–3600 kc.) that quadruples into the 14-Mc. band. Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

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

The most practical simple multiband antenna is one that is a half-wavelength long at the lowest frequency and is fed either at the center or one end with an open-wire line. Although the standing-wave ratio on the feedline will not approach 1.0 on any band, if the losses in the line are low the system will be efficient. From the standpoint of reduced feedline radiation, a center-fed system is superior to one that is end-fed, but the end-fed arrangement is often more convenient and should not be ignored as a possibility. The center-fed antenna will not have the same radiation pattern as an end-fed one of the same length, except on frequencies where the over-all length of the antenna is a half-wavelength or less. The end-fed antenna acts like a long-wire antenna on all bands (for which it is longer than a half-wavelength), but the center-fed one acts like two antennas of half that length fed in phase. For example, if a full-wavelength antenna is fed at one end, it will have a radiation pattern as shown in Fig. 10-56, but if it is fed in the center the pattern will be somewhat similar to Fig. 10-48, with the maximum radiation broadside to the wire. Either antenna is a good radiator, but if the radiation pattern is a factor, the point of feed must be considered.

Since multiband operation of an antenna does not permit matching of the feedline, some attention must be paid to the length of the feedline if convenient transmitter-coupling arrangements are to be obtained. Table 10-II gives some suggested antenna and feeder lengths for multiband operation. In general, the length of the feedline should be some integral multiple of a quarter wavelength at the lowest frequency.

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.

With center feed practically any convenient length of antenna can be used, if the feeder length is adjusted to accommodate at least

TABLE 10–III
Antenna and Feeder Lengths for Short Multiband Antennas, Center-Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	38	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel series series series or parallel
67.5	34	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
50	43	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	51	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	31	7 Mc. 14 Mc. 28 Mc.	parallel series parallel

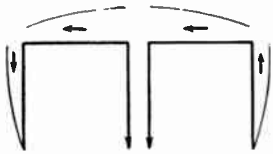


Fig. 10-61 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

one half-wave around the whole system.

A practical antenna of this type can be made as shown in Fig. 10-60. Table 10-III gives a few recommended lengths. However, the antenna can be made any convenient length, provided the total length of wire is a half-wavelength at the lowest frequency, or an integral multiple of a half-wavelength.

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. 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. The operation is illustrated in Fig. 10-61. Such an antenna will be a somewhat better radiator than a quarter-wavelength antenna on the lowest frequency, but is not so desirable for multiband 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. However, the fact that the radiation pattern is incapable of prediction does not detract from the general usefulness of the antenna.

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. 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 multiband work. The "V" antenna is shown in Fig. 10-62.

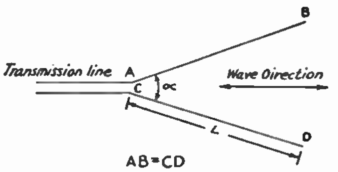


Fig. 10-62 — The basic "V" antenna, made by combining two long wires.

Fig. 10-63 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 ∝ 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. 10-63 is the vertical angle of maximum radiation. 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

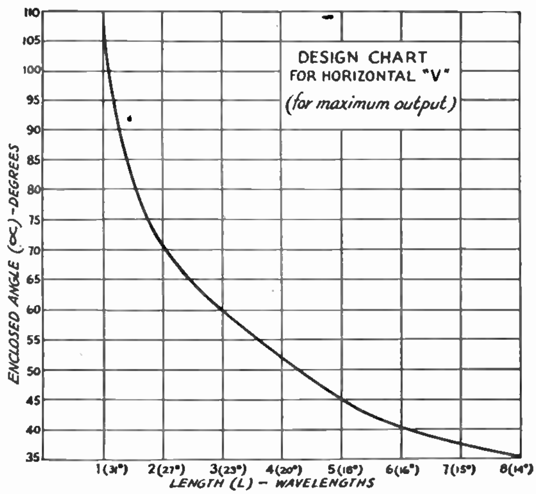


Fig. 10-63 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires. Values in parentheses represent approximate wave angle for height of one-half wavelength.

wires, but is not exactly twice the gain for a single long wire as given in Fig. 10-55. 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 eight-wavelength wire would be only approximately 9 db.

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. 10-62. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a nonresonant line. If the antenna wires are made multiples of a half-wave in length (use Equation 10-N for computing the length), the matching section will be closed at the free end. A stub can be connected across the resonant line to provide a match, as described later.

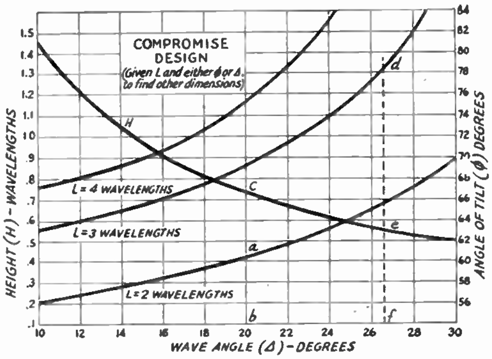


Fig. 10-65 — Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:

(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°.

THE RHOMBIC ANTENNA

The horizontal rhombic or "diamond" antenna is shown in Fig. 10-64. Like the "V," it requires a great deal of space for erection, but it is capable of giving excellent gain and directivity. It also can be used for multiband operation. In the terminated form shown in Fig. 10-64, it operates like a nonresonant 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. 10-64. 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. 10-65 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 two, three and four 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 from the 7- to 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

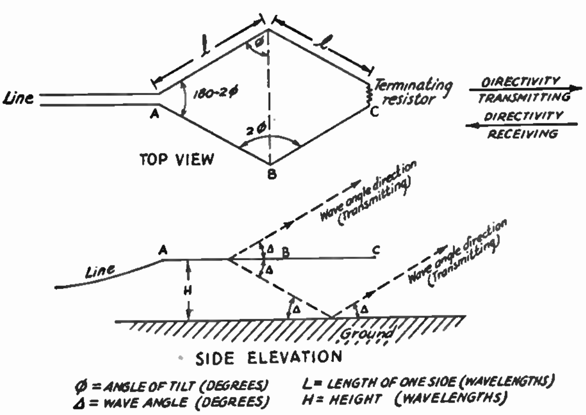


Fig. 10-64—The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

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 lightweight box or fiber 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 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 nonresonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to satisfactory 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.

Directive Arrays with Driven Elements

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 that 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 that the spacing and phasing has upon the radiation resistance of the elements, as well as upon their number.

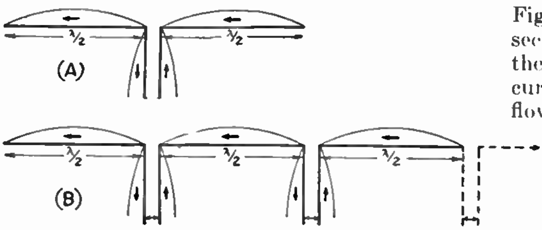


Fig. 10-66 — Collinear half-wave antennas in phase. The system at A is generally known as "two half-waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

Collinear Arrays

Simple forms of collinear arrays, with the current distribution, are shown in Fig. 10-66. 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. 10-66B. Note that quarter-wave phasing sections are used between elements; these give the reversal in phase necessary to make the currents in individual antenna elements all flow in the same direction at the same instant. Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder, 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, 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 10-IV. Although three-quarter 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.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mount-

TABLE 10-IV
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	4	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

ing 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.

Broadside Arrays

Parallel antenna elements with currents in phase may be combined as shown in Fig. 10-67 to form a **broadside** array, so named because

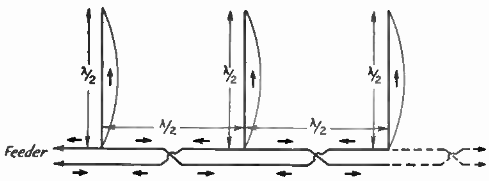


Fig. 10-67 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).

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. 10-68. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table 10-V 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 while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines or through quarter-wave matching sections and nonresonant lines. In Fig. 10-67, note the "crossing over" of the feeders, which is necessary to bring the elements into 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. 10-69. 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.

The arrays in Fig. 10-69 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 over-all performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the eight-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 between the second and third set of elements. 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. 10-69B, known as the "lazy-H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 10-70.

End-Fire Arrays

Fig. 10-71 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 ver-

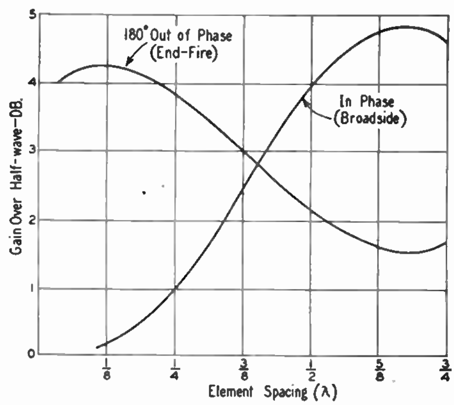


Fig. 10-68 — Gain vs. spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

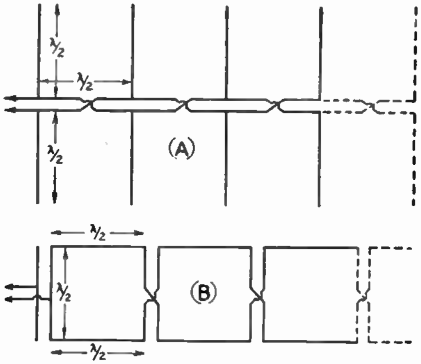


Fig. 10-69 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table 10-V) plus the gain of one set of collinear elements (Table 10-IV). 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.

tically 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. 10-68 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements to give a further increase in gain and directivity.

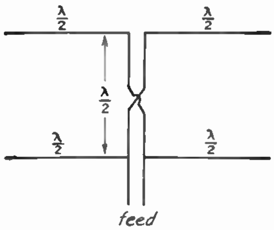


Fig. 10-70 — A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

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 or phasing stub.

Phasing

Figs. 10-69 and 10-71 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 10-71, 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 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. 10-69. 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 feedline connected to its center, as in Fig. 10-67.

Adjustment of Arrays

With arrays of the types just described, using half-wave spacing between elements, it

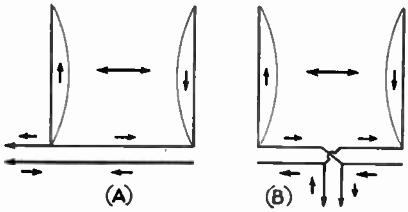


Fig. 10-71 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 10-68. Direction of maximum radiation is shown by the large arrows.

will usually suffice to make the length of each element that given by Equations 10-I or 10-J. The half-wave phasing lines between the parallel elements should be of open-wire construction, and their length can be calculated from:

$$\textit{Length of half-wave line (feet)} = \frac{480}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-O)}$$

Example: A half-wavelength phasing line for 28.8 Mc. would be $\frac{480}{28.8}$ = 16.66 feet = 16 feet 8 inches.

The spacing between elements can be made equal to the length of the phasing line. No special adjustments of line or element length or spacing are needed, provided the formulas are followed closely.

TABLE 10-V
Theoretical Gain *vs.* Number of Broadside Elements (Half-Wave Spacing)

No. of elements	*Gain*
2	4 db.
3	5.5
4	7
5	8
6	9

With collinear arrays of the type shown in Fig. 10-66B, the same formula may be used for the element length, while the length of the quarter-wave phasing section can be found from the following formula:

$$\text{Length of quarter-wave line (feet)} = \frac{240}{\text{Freq. (Mc.)}} \qquad \text{(10-P)}$$

Example: A quarter-wavelength phasing line for 14.25 Mc. would be $\frac{240}{14.25} = 16.84$ feet = 16 feet 10 inches.

If the array is fed in the 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 of 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.

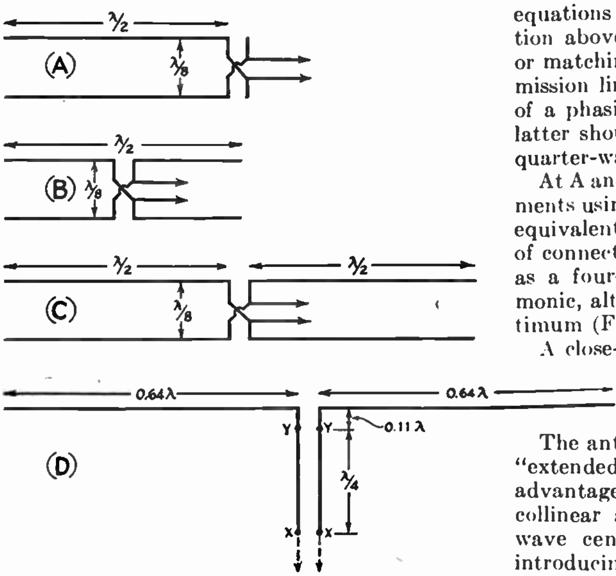


Fig. 10-72 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is 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 ⅛ 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.

The phasing sections can be made of 300-ohm Twin-Lead, if low power is used. However, the lengths of the phasing sections must be only 84 per cent of the length obtained in the two formulas above.

Example: The half-wavelength line for 28.8 Mc. would become 0.84 × 16.66 = 13.99 feet = 14 feet 0 inches

Using Twin-Lead for the phasing sections is most useful in arrays such as that of Fig. 10-66B, or any other system in which the element spacing is not controlled by the length of the phasing section.

Simple Arrays

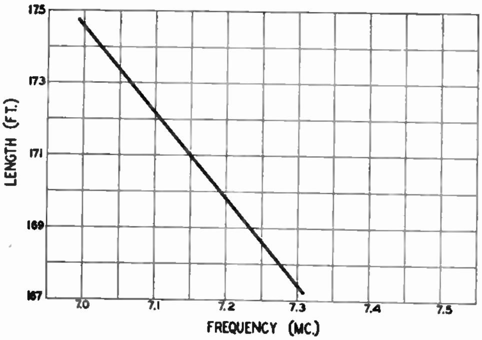
Several simple directive-antenna systems using driven elements have achieved rather wide use among amateurs. Four of these systems are shown in Fig. 10-72. Tuned feeders are assumed in all cases; however, a matching section readily can be substituted if a nonresonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths can be calculated from the equations for the antenna and from the equation above for the resonant transmission line or matching section. In cases where the transmission line proper connects to the midpoint of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

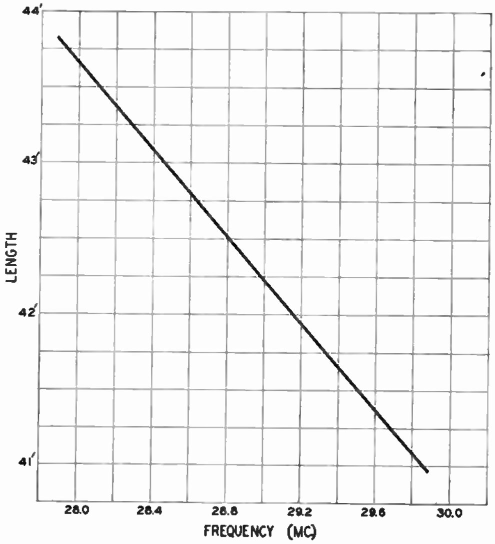
At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 10-68) for such operation.

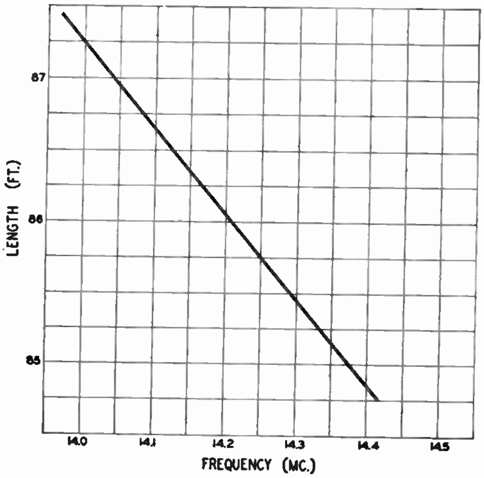
A close-spaced four-element array is shown at C. It will give about 2 db. more gain than the two-element array.

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 fairly 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 because of the effect of the collinear arrangement. The vertical pattern, however, will be the same as that of the antennas in A and B.







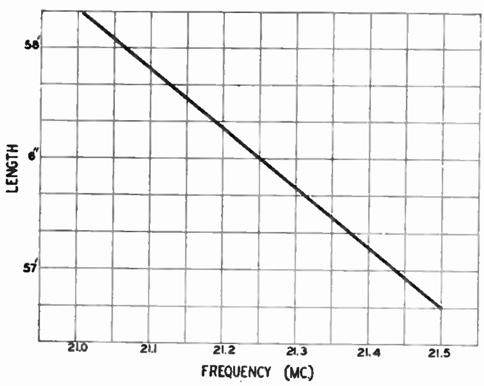


Fig. 10-73 — Charts for determining the over-all length of the flat-top portion of an extended double-Zepp antenna, as shown in Fig. 10-72D.

Matching the Antenna to the Line

Except in the several cases of half-wave antennas mentioned earlier, most antenna systems do not have center impedances that readily match open-wire lines or available solid-dielectric ones. However, any antenna can be matched to practically any line by any of the several means to be described. The matching is accomplished by first resonating the antenna to the proper frequency and then introducing either a matching transformer between the antenna and the line or by applying corrective stubs to the line.

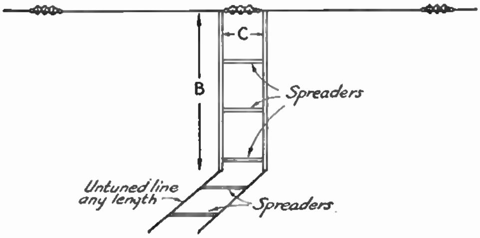


Fig. 10-74 — The "Q" antenna, using a quarter-wave impedance-matching section with close-spaced conductors.

An impedance mismatch of 10 or 20 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 1.5- or 2-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.

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.

"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, Equation 10-B. 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. 10-74. Important dimensions are the length of the antenna itself, 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.

The required characteristic impedance for the matching section is

$$Z_m = \sqrt{Z_1 Z_2} \qquad \text{(10-B)}$$

where Z_1 and Z_2 are the antenna and feedline impedances.

Example: To match a 600-ohm line to an antenna presenting a 72-ohm load, the quarter-wave matching section would require a characteristic impedance of $\sqrt{72 \times 600} = \sqrt{43{,}200} = 208$ ohms.

The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 10-14. 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 Equation 10-G. The length of the antenna can be calculated from Equations 10-I or 10-J.

This system has the advantage of the simplicity of adjustment of the 75-ohm feeder system and at the same time the superior insulation of an open-wire system.

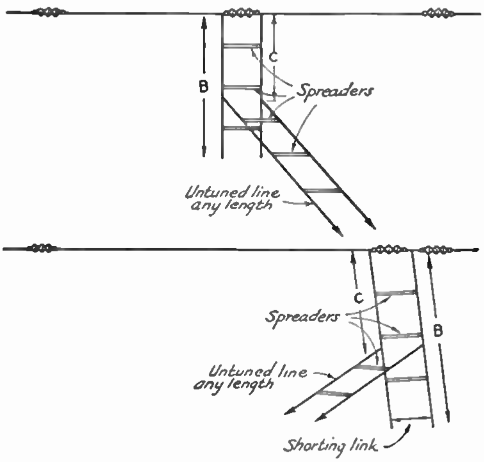


Fig. 10-75 — Antenna systems with quarter-wave open-wire linear impedance-matching transformers.

Linear Transformers

Fig. 10-75 shows two methods of coupling a nonresonant line to an antenna through a quarter-wave linear transformer or matching section. In the case of the center-fed antenna, the free end of the matching section, B, is open (high impedance) if the other end is connected to a low-impedance point (current loop) 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.

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. 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 Equations 10-1, 10-X and 10-P. 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

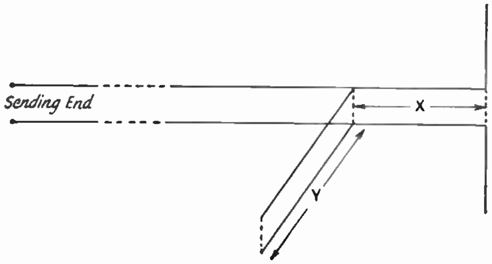


Fig. 10-76 — When antenna and transmission line differ in impedance, they may be matched by a short length of transmission line, Y, called a stub. Determination of the critical dimensions, X and Y, for proper matching depends on whether the stub is open or closed at the end.

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 the chapter on measurements. 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

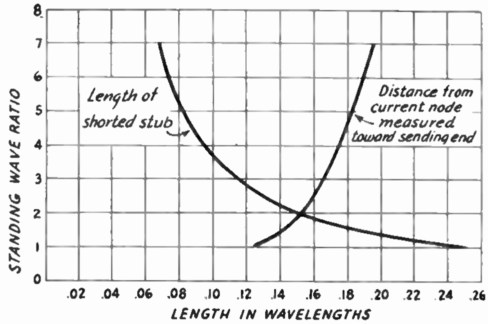


Fig. 10-77 — Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.

quarter wavelength will indicate whether or not standing waves are present.

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 waves, 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 a slight lengthening or shortening of the stub is necessary to tune out this reactance.

Matching Stubs

The operation of the quarter-wave matching transformer of Fig. 10-75 may be considered from another — and more general — viewpoint. Suppose that section C is looked upon simply as a continuation of the transmission line. Then the "free" end of the transformer becomes a "stub" line, shunting a section of the main transmission line. From this viewpoint, matching the line to the antenna becomes a matter of selecting the right type and length of stub and attaching it to the proper spot along the line.

Referring to Fig. 10-76, at any distance (X) from the antenna, the line will have an imped-

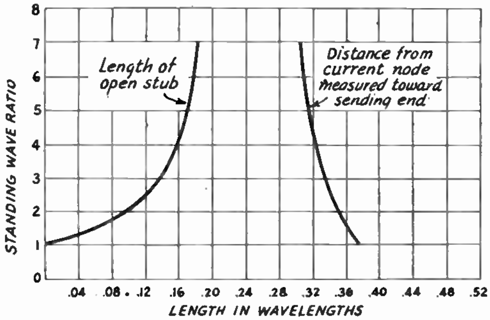


Fig. 10-78 — Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.

ance that 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 to 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 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 the characteristic impedance of the line by adjusting the distance X.

Distances X and Y may be determined experimentally, but since their values are interdependent the cut-and-try method is somewhat laborious. If the standing-wave ratio and the positions of the current loops and nodes can be measured, the length and position of the stub can be found from Figs. 10-77 and 10-78.

While it is relatively easy to locate the position of the current (or voltage) loops and nodes by examining the line with a neon bulb, r.f. galvanometer, or pick-up loop and crystal detector, other means are more direct for determining the standing-wave ratio. Several devices of this type are described in Chapter Sixteen, and the use of these also affords a simple method for determining the location of current loops (voltage nodes). With the meter or indicator in the line near the transmitter, points will be found on the transmission line where touching the line with a screwdriver will have a minimum effect on the meter indication. These points correspond to voltage nodes.

Once the standing-wave ratio is known, the length and position of the stub, in terms of wavelength, can be found directly from Figs. 10-77 and 10-78. The wavelength in feet for any frequency can be found from Equation 10-O.

Measuring Standing Waves

In adjusting a "Q-match" or linear transformer, or a delta or "T"-match to an antenna, one of the standing-wave indicators described in Chapter Sixteen should be used. If 300-ohm Twin-Lead is used, the simple "twin-lamp" indicator is the most convenient and the simplest to use. For lines of other impedance, or for coaxial line, the Micro-Match type or the bridge type should be used. In any event, the absolute value of standing-wave ratio is not as important as the proper adjustment for a minimum ratio, since ratios of 1.5-to-1 or less represent good amateur practice.

Where two-wire lines are used, the standing-wave-ratio indicator should give the same reading regardless of the polarity of the transmission line — any discrepancy indicates an unbalance in the line.

Directive Arrays with Parasitic Elements

Parasitic Excitation

The antenna arrays previously described 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, 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 reradiate 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. They are widely used to give additional gain and directivity to simple antennas.

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.

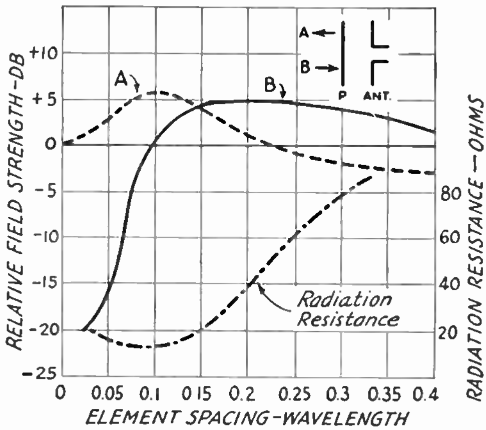


Fig. 10-79 — Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back 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.

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. 10-79, 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 formula for a half-wavelength antenna. 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.

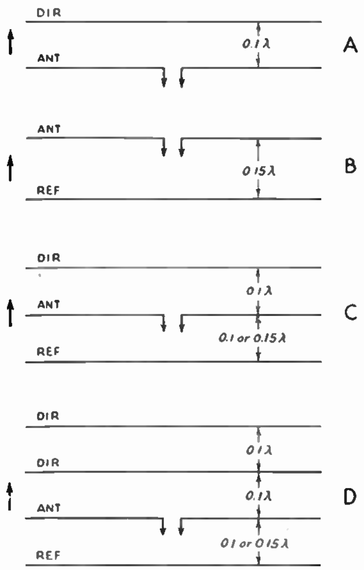


Fig. 10-80 — Half-wave antennas with parasitic elements. A, with director; B, with reflector; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 10-79, in the first two cases, and depends upon the spacing and length of the parasitic element. 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 the direction of maximum radiation.

Simple Systems: the Rotary Beam

Four practical combinations of antenna, reflector and director elements are shown in Fig. 10-80. 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. 10-79. 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 directivity 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.

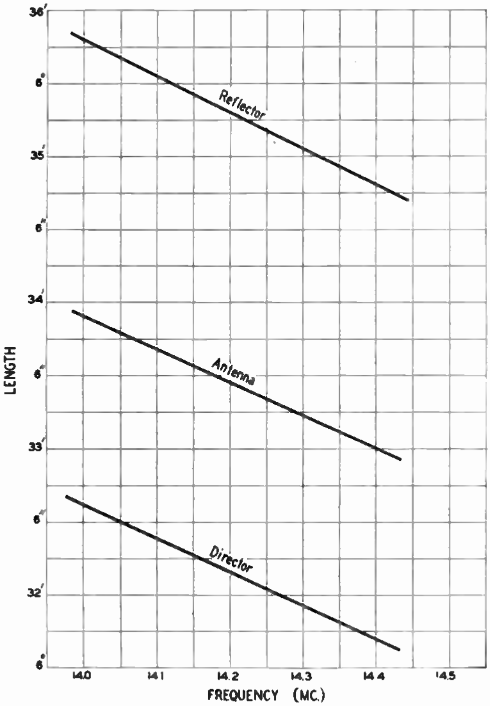
Arrays using more than one parasitic element, such as those shown at C and D in Fig. 10-80, will give more gain and directivity than is indicated for a single reflector or director by the curves of Fig. 10-79. The gain with a properly-adjusted three-element array (antenna, director and reflector) will 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.

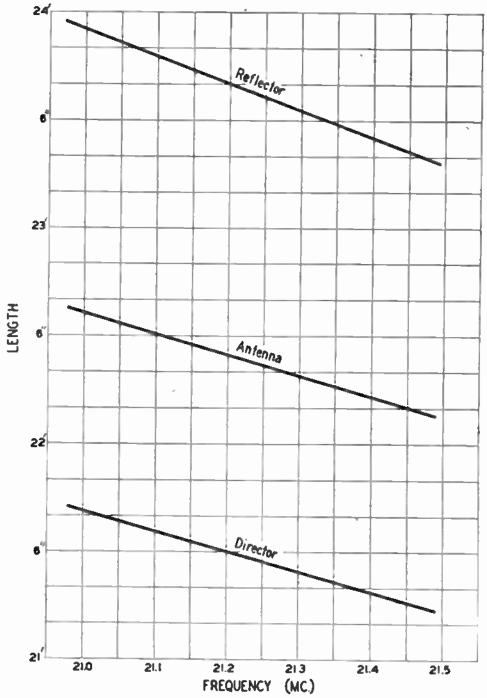
The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic resistance but also has lower *Q;* both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a single half-wave dipole. With 3- and 4-element arrays the radiation resistance of the driven element may be as low as 6 or 8 ohms, so that ohmic losses in the conductor can consume an appreciable fraction of the power. Low radiation resistance means that the antenna will work over only a small frequency range without retuning unless large-diameter conductors are used. In addition, the antenna elements should be rigid because if they are free to move with respect to each other, the array will tend to show troublesome 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. 10-82. 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. 10-82 preferably





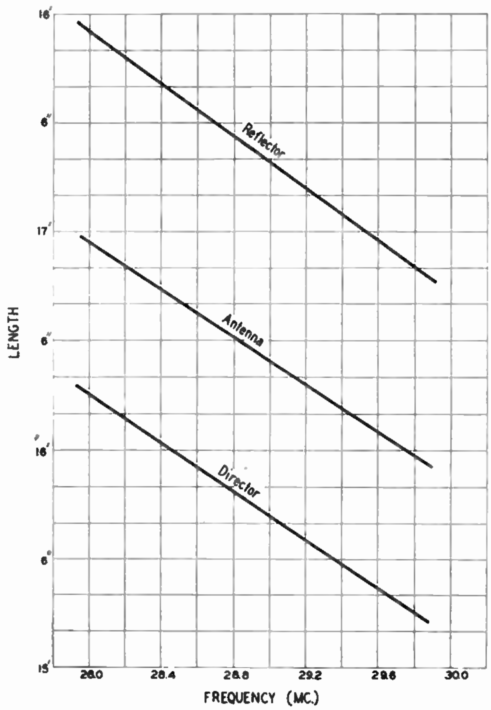


Fig. 10-81 — Director, antenna and reflector lengths for three-element beams, for element spacing of 0.1 to 0.2 wavelength. The greater spacing will result in slightly higher gain. The lengths indicated are for maximum gain — some improvement in front-to-back ratio may be obtained by adjustment of the reflector length.

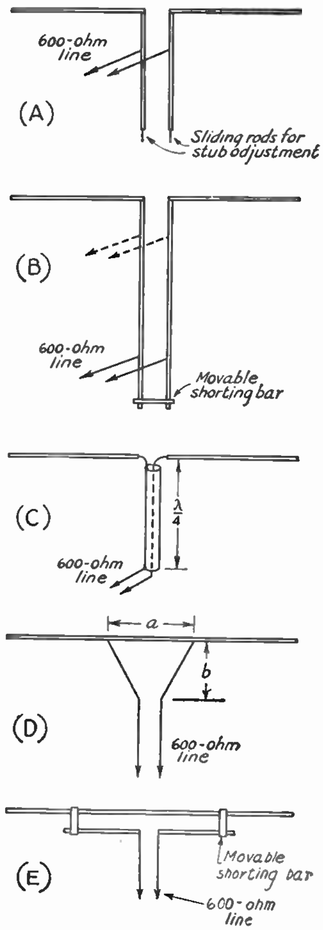


Fig. 10-82 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer; E, "T" matching transformer. Adjustment details are discussed in the text.

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 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 (RG-11/U), it will give a good 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 10-G.

The delta matching transformer shown at D 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*.

The system shown at E ("T"-match) resembles the delta match in principles of operation. It has the advantage that, with close spacing between the two parallel conductors, line radiation from the matching section is negligible whereas radiation from a delta may be considerable. It is adjusted by moving the shorting bars, keeping them equidistant from the center, until there are no standing waves on the line. The matching section may be made of the same type of conductor used for the driven element and spaced a few inches from it.

The "folded-dipole" type of antenna may be used as the driven element of a close-spaced parasitic array to secure an impedance step-up to the transmission line and also to broaden the resonance curve of the antenna. The folded dipole consists of two or more half-wave antennas connected together at the ends with the feeder connected to the center of only one of the antennas. The spacing between the parallel antennas should be small — of the order of the spacing used between wires of a transmission line. The current in the system divides in approximate proportion to the areas of the conductors, resulting in an impedance step-up at the input terminals. With two similar conductors (equal areas) the impedance step-up is 4-to-1; if there are three similar conductors (or if the one not connected to the transmission line has twice the diameter of the other) the step-up is 9-to-1; if the ratio of the areas is 3-to-1 the step-up is 16-to-1, and so on. Thus if a 3-conductor dipole (all conductors the same diameter) is used as the driven element of a four-element parasitic array the center impedance of approximately 8 ohms is multiplied by 9 and appears as approximately 72 ohms at the input terminals. Such a system therefore can be fed directly from a 70-ohm line with no additional means for matching.

Fig. 10-83 shows the impedance step-up obtained in a folded dipole when conductors of different sizes are used.

Sharpness of Resonance

Peak performance of a multielement parasitic array depends upon proper phasing or tuning of the elements, which can be exact for one frequency only. In the case of close-spaced arrays, which because of the low radiation resistance usually are quite sharp-tuning, the frequency range over which optimum results can be secured is only of the order of 1 or 2

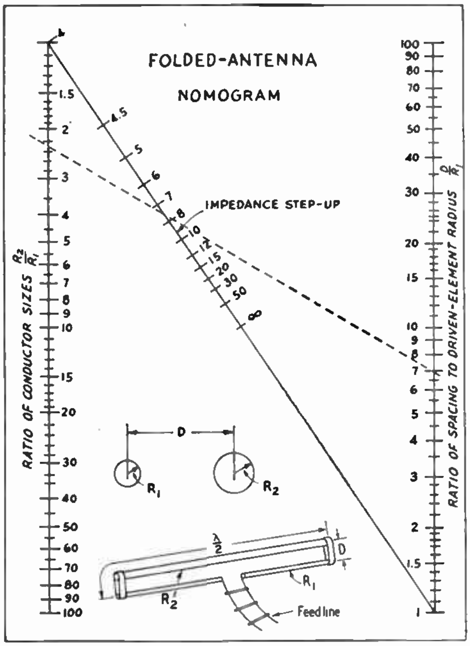


Fig. 10-83 — Nomogram for computing impedance step-up in a folded dipole with dissimilar conductors. The line at the left is the ratio of conductor diameters, and the line at the right is the ratio of conductor spacing (center-to-center) to the driven-element radius. The solid slanting line is the impedance step-up ratio. Laying a straightedge between any two known quantities will give the value of the third.

Example: Find the diameter of the large conductor when the driven-element diameter is 0.5 inch, line impedance 300 ohms, antenna impedance 40 ohms, and spacing 1.75 inches.

Impedance step-up required = 300/40 = 7.5

Spacing-to-element-radius ratio = 1.75/0.25 = 7

Laying a straightedge across the figure (dashed line), ratio of conductor diameters = 2.3

Diameter of large conductor = 2.3 × 0.5 = 1.15 inches

per cent of the resonant frequency, or up to about 500 kc. at 28 Mc. However, the antenna can be made to work satisfactorily over a wider frequency range by adjusting the director or directors to give maximum gain at the *highest* frequency to be covered, and by adjusting the reflector to give optimum gain at the *lowest* frequency. This sacrifices some gain at all frequencies, but maintains more uniform gain over a wider frequency range.

As mentioned in the preceding paragraphs, the use of large-diameter conductors will broaden the response curve of an array because the larger diameter lowers the Q. This causes the reactances of the elements to change rather slowly with frequency, with the result that the tuning stays near the optimum over a considerably-wider frequency range than is the case with wire conductors.

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 10-IV, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 10-68 or Table 10-V, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 10-79, 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.

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 that 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 at frequencies below the v.h.f. range, 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 loca-

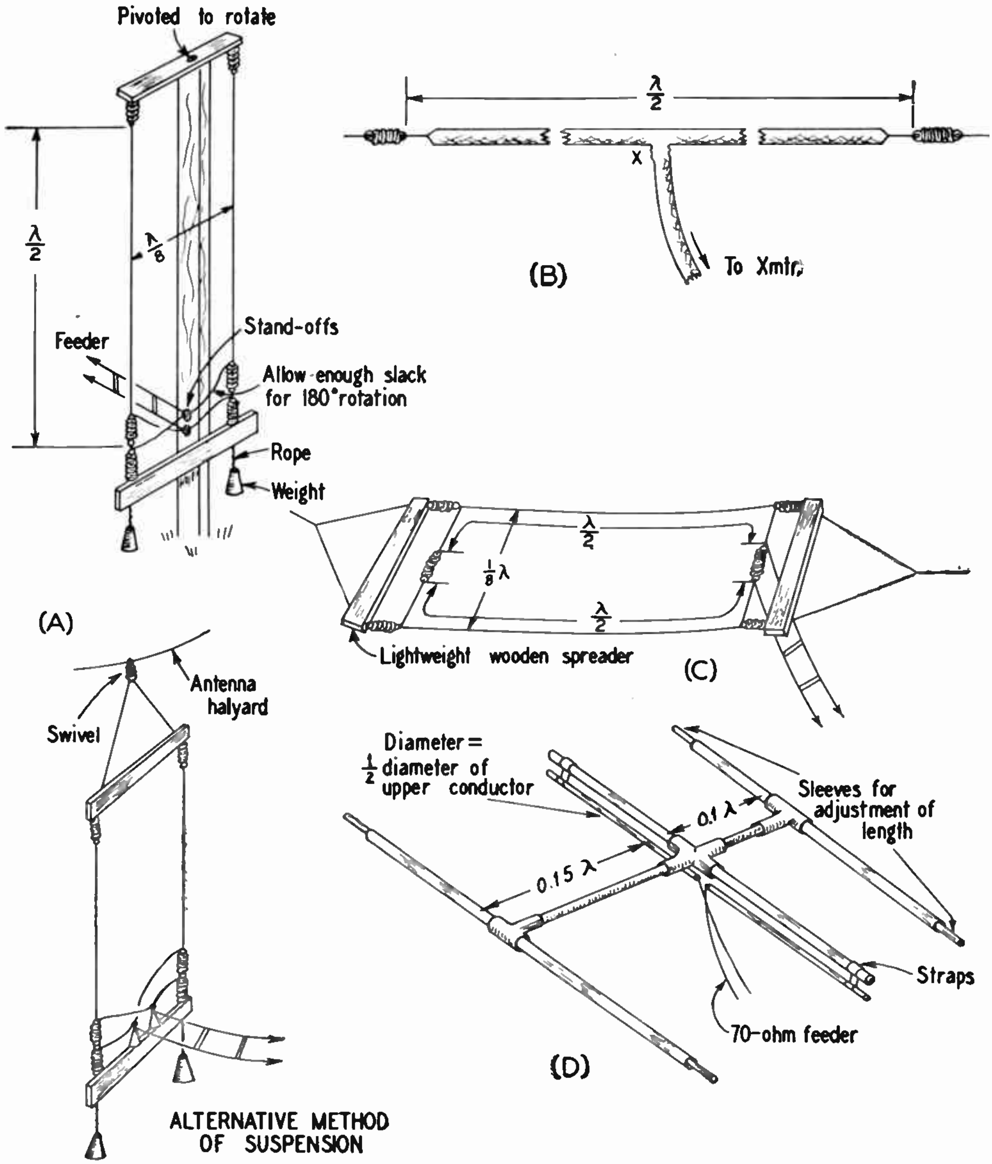
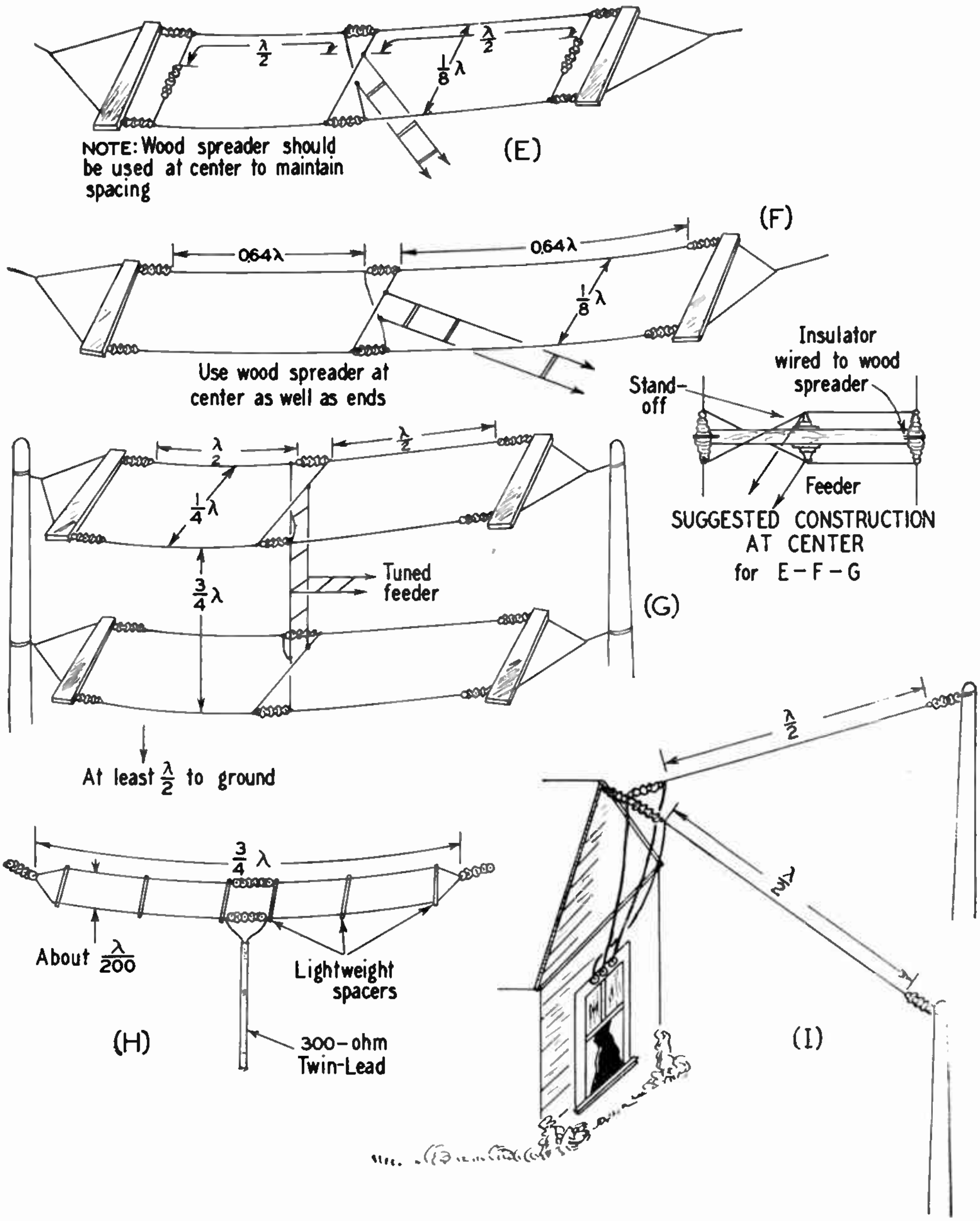


Fig. 10-84 — Some suggested antenna systems. A — Simple bidirectional rotatable end-fire array using ⅛-wave spacing between out-of-phase elements. It is suitable for either 14 or 28 Mc. and can be hand-rotated. It can also be suspended from the halyard holding another antenna, as suggested in the lower drawing. B — Folded dipole using 300-ohm Twin-Lead for both antenna and feeder. The junction X at the center is made by opening one conductor of the antenna section and soldering to the feeder leads. The joint may be made mechanically firm by heating the dielectric with a soldering iron, using extra bits of dielectric for a good bond. C — An end-fire array for use where space is limited. The ends of the two half-wave elements are folded to meet at an insulator in the center. The antenna may be made still shorter by increasing the spacing; spacings up to ¼ wavelength may be used. D — Pipe-assembly three-element beam ("plumber's delight") with folded-dipole driven element. Because all three elements are at the same r.f. potential at their centers it is possible to join them electrically as well as mechanically with no effect on the performance. Provision is made for adjusting the element lengths for optimum performance at a given frequency. E — An extension of the folding principle shown in C. The collinear in-phase elements give additional gain and directivity. F — End-



fire array with extended double-Zepps. This antenna should give a gain of about 7 db. in the direction perpendicular to the line of the antenna. G — An 8-element array combining broadside, end-fire and collinear elements. The gain of an antenna of this type is about 10 db. This antenna also can be used at half the frequency for which it is designed. H — A three-quarter wavelength folded antenna matches 500- or 600-ohm open-wire line, but 300-ohm Twin-Lead is quite satisfactory. Its pattern is quite similar to a half-wavelength antenna. Note that, unlike the half-wavelength folded dipole, the far side is *open* at the center. I — Using two half-wave antennas at right angles to change direction. With the three feeders indicated, either antenna alone can be fed as a Zepp and will radiate best perpendicular to its direction. By feeding the two together, leaving the third feeder wire idle, the optimum direction is the bisector of the angle between the wires. This system is most useful at high frequencies.

In these drawings, wavelength dimensions on conductors refer to lengths calculated for the conductor size as described in Equation 10-J.

The feeders to the various directive systems in A, C, E, F and G must be tuned if used as shown. For one-band operation, matching stubs may be attached to the feeders if a matched line is desired.

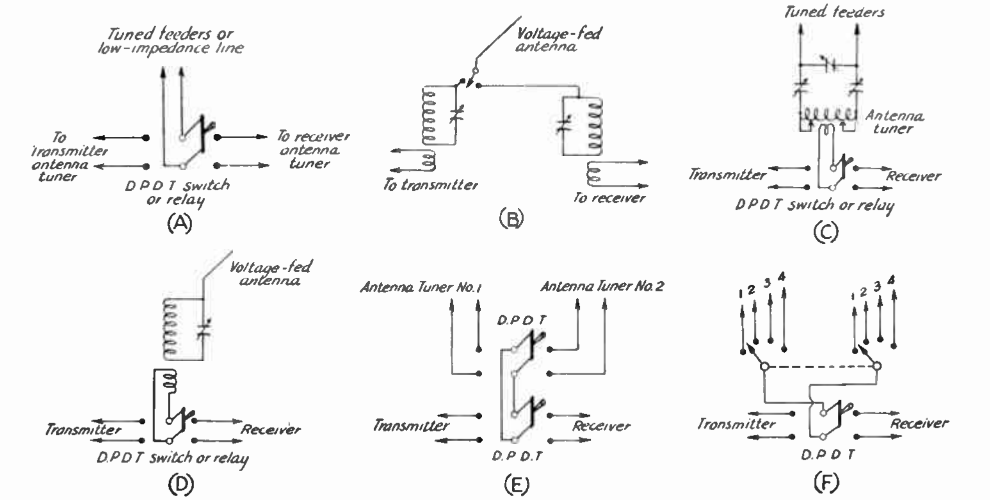


Fig. 10-85 — Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. 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.

tion, 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.

In selecting a directional receiving antenna it is preferable to choose a type that 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 feedline from the antenna should be balanced so that it will not pick up signals and destroy the directivity.

Antenna Switching

Switching of the antenna from receiver to transmitter is commonly done with a change-over relay, connected in the antenna leads or the coupling link from the antenna tuner. If the relay is one with a 115-volt a.c. coil, the switch or relay that controls the transmitter plate power will also control the antenna relay. If the convenience of a relay is not desired, porcelain knife switches can be used and thrown by hand.

Typical arrangements are shown in Fig. 10-85. If coaxial line is used, the use of a coaxial relay is recommended, although on the lower-frequency bands a regular switch or change-over relay will work almost as well.

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 and matching stubs 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.

At points of maximum voltage, insulation is most important, and Pyrex glass, Isolantite or steatite insulators with long leakage paths are recommended for the antenna. 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

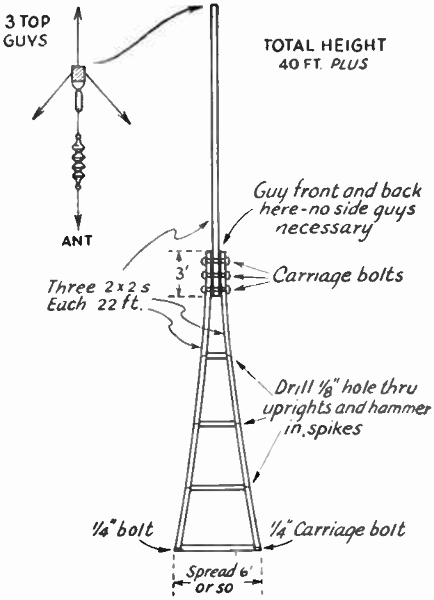


Fig. 10-86 — Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

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. 10-86 is satisfactory for heights up to 35 or 40 feet. 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.

SIMPLE 40-FOOT MAST

The mast shown in Fig. 10-87 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 halfway 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 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 meas-

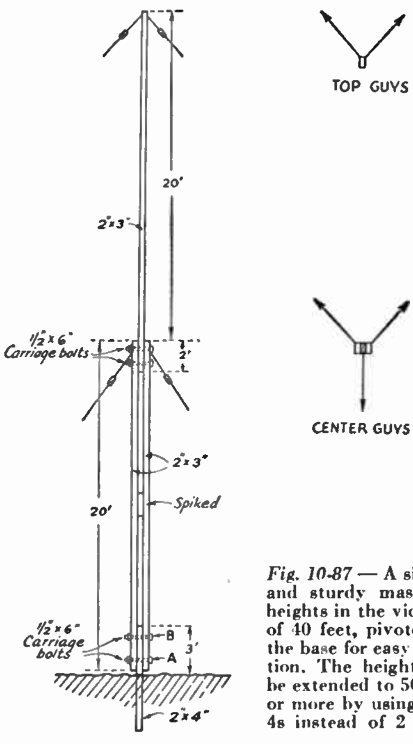


Fig. 10-87 — A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.

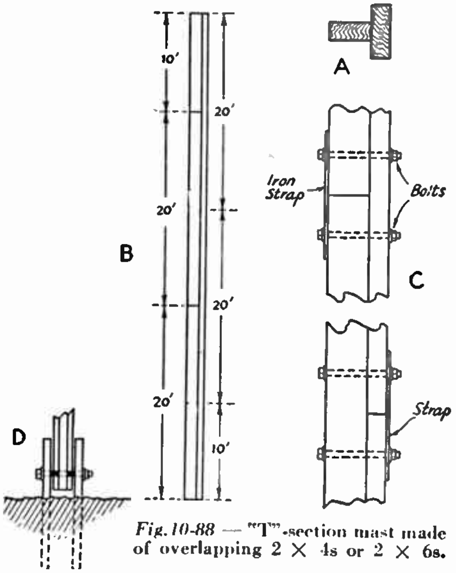


Fig. 10-88 — "T"-section mast made of overlapping 2 × 4s or 2 × 6s.

are 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.

"T"-SECTION MAST

A type of mast suitable for heights up to about 80 feet is shown in Fig. 10-88. The mast is built up by butting 2 × 4 or 2 × 6 timbers flatwise 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.

The method of making the joints is shown at C. Quarter-inch or 3/16-inch iron, 1½ to 2 inches wide, is recommended for the straps, with ½-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. The bolt goes through a hole in the mast so that it is pivoted at the bottom.

Half of the guys can be tightened up before the mast leaves the ground by using four sets of guys, one in front, one directly in the rear, and one on each side at right angles to the direction the mast will face. The mast should be guyed every twenty feet and at the top, at each of the joints in the edgewise sections, the guy wires being wrapped around the pole for added strength.

For heights up to 50 feet, 2 × 4-inch members may be used throughout. For greater heights, use 2 × 4s for the edgewise sections; 2 × 6-inch pieces will do for the flat sections.

POLE AND TOWER SUPPORTS

Poles, which often may be purchased at a reasonable price from the local telephone or power company, have the advantage that they do not require guying unless they are called upon to carry a very heavy load. The life of a pole can be extended many years by proper precautions before erecting, and regular maintenance thereafter.

Before setting the pole, it should be given four or five coats of creosote, applying it liberally so it can soak into and preserve the wood. The bottom of the pole and the part that will be buried in the ground should have a generous coating of hot pitch, poured on while the pole is warm. This will keep termites out and prevent rotting.

The pole should be set in the ground four to eight feet depending upon the height. It is a good idea to pour concrete around the bottom three feet of the base, packing the rest of the excavation with soil. The concrete will help hold the pole against strong winds. After filling the hole with dirt, a stream from a hose should be played on the dirt slowly for several hours.

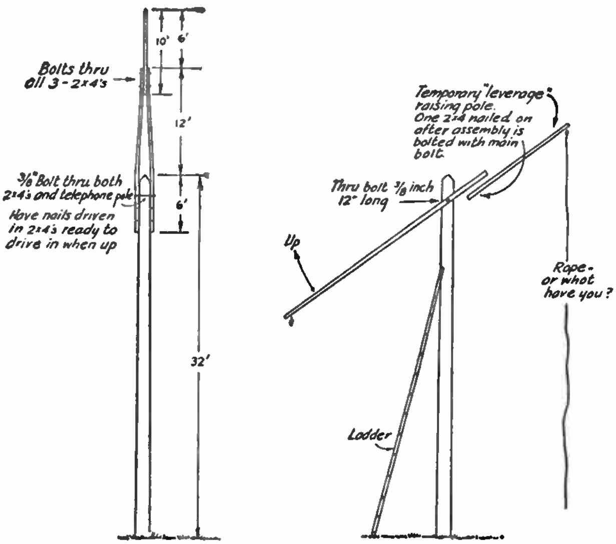


Fig. 10-89 — This type of mast may be carried to a height of fifty feet or more. No guy wires are required.

This will help to settle the soil quickly.

If desired, the pole may be extended by the arrangement shown in Fig. 10-89. Three 2 × 4s are required for the top section, two being 18 feet long and one 10 feet long. The 10-foot section is placed between the other two and bolted in place. A half-inch hole should be bored through the pole about 2 feet from its top and through both 18-foot 2 × 4s about 5 feet from their bottom ends, which are spread apart to fit the top of the pole. The bottom end of the extension is then hauled up to the top of the pole and bolted loosely so that the section can be swung up into place by the leverage of another 2 × 4 temporarily fastened to the section, as shown in Fig. 10-89.

Lattice towers built of wood should be assembled with brass screws and casein glue, rather than with nails which work loose in a short time. A tower constructed in this manner will give trouble-free service if treated with a coat of paint every year.

In painting outside structures, use pure white lead, thinned with three parts of pure linseed oil to one part of turpentine, for the first coat on new wood. The use of a drier is not recommended if the paint will possibly dry without it, since it may cause the paint to peel after a short time. For the second and third coats pure white lead thinned only with pure linseed oil is recommended. Plenty of time for drying should be allowed between coats. White paint will last fifty per cent longer than any colored paint.

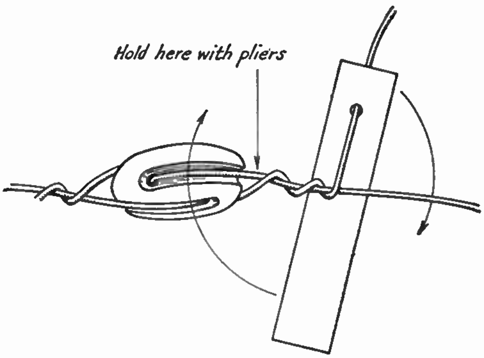


Fig. 10-90 — Using a lever for twisting heavy guy wires.

GUYS AND GUY ANCHORS

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material. Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

More than three guy wires in any one set usually are unnecessary. If a horizontal antenna is to be supported, two guy wires in the top set will be sufficient in most cases. These should run to the rear of the mast about 100 degrees apart to offset the pull of the antenna.

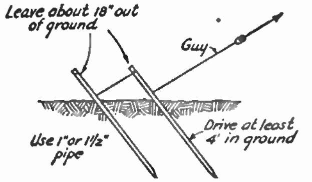


Fig. 10-91 — Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.

Intermediate guys should be used in sets of three, one running in a direction opposite to that of the antenna, while the other two are spaced 120 degrees either side. This leaves a clear space under the antenna. The guy wires should be adjusted to pull the pole slightly back from vertical before the antenna is hoisted so that when the antenna is pulled up tight the mast will be straight.

When raising a mast that is big enough to tax the facilities available, it is some advantage to know nearly exactly the length of the guys. Those on the side on which the pole is lying can then be fastened temporarily to the anchors beforehand, which assures that when the pole is raised, those holding opposite guys will be able to pull it into nearly-vertical position with no danger of its getting out of control. The guy lengths can be figured by the right-angled-triangle rule that "the sum of the squares of the two sides is equal to the square of the hypotenuse." In other words, the distance from the base of the pole to the anchor should be measured and squared. To this should be added the square of the pole length to the point where the guy is fastened. The square root of this sum will be the length of the guy.

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 part if the insulator breaks.

Twisting guy wires onto "egg" insulators may be a tedious job if the guy wires are long and of large gauge. The simple time- and finger-saving device shown in Fig. 10-90 can be made from a piece of heavy iron or steel by drilling a hole about twice the diameter of the guy wire about a half inch from one end of the piece. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in the sketch. By passing the wire through the hole in the iron and rotating the iron as shown, the wire may be quickly and neatly twisted.

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. 10-91.

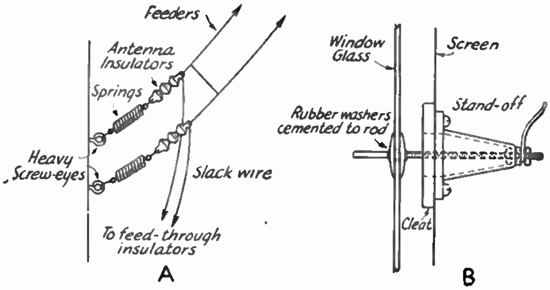


Fig. 10-93 — A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

HALYARDS AND PULLEYS

Halyards or ropes and pulleys are important items in the antenna-supporting system. Particular attention should be directed toward the choice of a pulley and halyards for a high mast since replacement, once the mast is in position, may be a major undertaking if not entirely impossible.

Galvanized-iron pulleys will have a life of only a year or so. Especially for coastal-area installations, marine-type pulleys with hardwood blocks and bronze wheels and bearings should be used.

An arrangement that has certain advantages over a pulley when a mast is used is shown in Fig. 10-92. In case the rope breaks, it may be possible to replace it by heaving a line over the brass rod, making it unnecessary to climb or lower the pole.

For short antennas and temporary installations, heavy clothesline or window-sash cord may be used. However, for more permanent jobs, ⅜-inch or ½-inch waterproof hemp rope should be used. Even this should be replaced about once a year to insure against breakage.

Nylon rope, used during the war as glider tow rope, is, of course, one of the best materials for halyards, since it is weatherproof and has extremely long life.

It is advisable to carry the pulley rope back up to the top in "endless" fashion in the manner of a flag hoist so that if the antenna breaks close to the pole, there will be a means for pulling the hoisting rope back down.

BRINGING THE ANTENNA OR FEEDLINE INTO THE STATION

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 10-93, 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 that 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. Either cement or rubber gaskets may be used to waterproof the exposed joints.

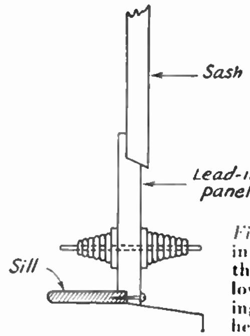


Fig. 10-94 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Sealing the overlapping joint will help make it weatherproof.

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. Rubber gaskets will render the holes waterproof. The lower sash should be provided with stops to prevent damage when it is raised. If the window has a full-length screen, the scheme shown in Fig. 10-93B 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. 10-94, and

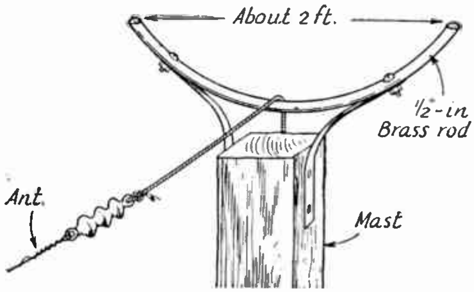


Fig. 10-92 — This device is much easier than a pulley to "rethread" when the rope breaks.

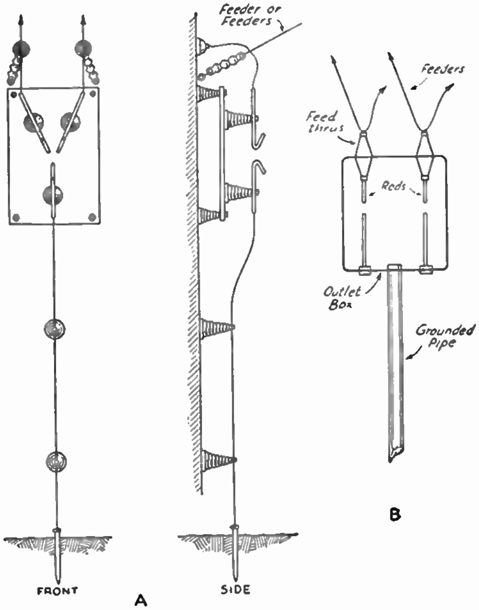


Fig. 10-95 — Low-loss lightning arresters for transmitting-antenna installations.

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. 10-95. 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.

Rotary-Beam Construction

It is a distinct advantage to be able to shift the direction of a beam antenna at will, thus securing the benefits of power gain and directivity in any desired compass direction. A favorite method of doing this is to construct the antenna so that it can be rotated in the horizontal plane. Obviously, the use of such rotatable antennas is limited to the higher frequencies — 14 Mc. and above — and to the simpler antenna-element combinations if the structure size is to be kept within practicable bounds. For the 14- and 28-Mc. bands such antennas usually consist of two to four elements and are of the parasitic-array type described earlier in this chapter. At 50 Mc. and higher it becomes possible to use more elaborate arrays because of the shorter wavelength and thus obtain still higher gain. Antennas for these bands are described in Chapter Fourteen.

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.

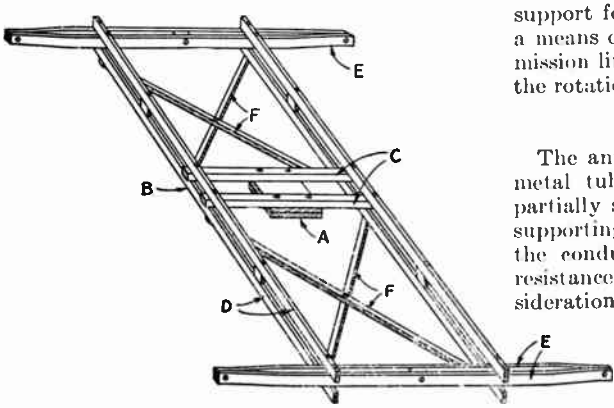


Fig. 10-96 — Easily-built supporting structure for horizontal rotary beams. Made chiefly of 1 × 2″ wood strip, it is strong yet lightweight. Antenna elements are supported on stand-off insulators on the arms, *E*. The length of the *D* sections will depend upon the element spacing, while the length of the *E* sections and the spacing between the *D* sections should be ¼ to ½ the length of the antenna elements.

Elements

The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the supporting structure. The large diameter of the conductor is beneficial also in reducing resistance, which becomes an important consideration when close-spaced elements are used.

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 for tuning quite easy. Electrician's thin-walled conduit also is suitable for rotary-beam elements.

If steel elements are used, special precautions

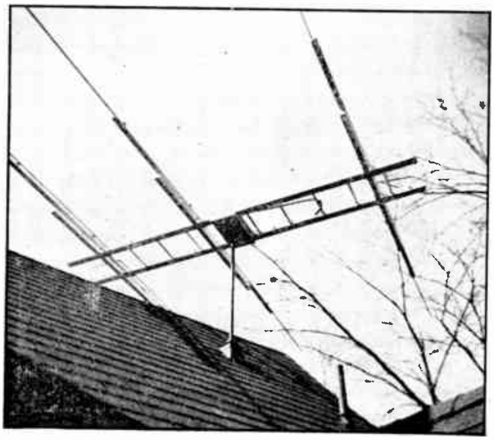


Fig. 10-97 — A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)

should be taken to prevent rusting. Even copper-coated steel does not stand up indefinitely, since the coating usually is too thin. The elements should be coated both inside and out with slow-drying aluminum paint. For coating the inside, a spray gun may be used, or the paint may be poured in one end while rotating the tubing. The excess paint may be caught as it comes out the bottom end and poured through again until it is certain that the entire inside wall has been covered. The ends should then be plugged up with corks sealed with glyptal varnish.

Supports

The supporting framework for a rotary beam usually is made of wood but sometimes of metal, using as lightweight construction as is consistent with the required strength. Generally, the frame is not required to hold much weight, but it must be extensive enough so that the antenna elements can be supported near enough to their ends to prevent excessive sag, and it must have sufficient strength to stand up under the maximum wind in the locality. The design of the frame will depend chiefly on the size of the antenna elements, whether they are mounted horizontally or vertically, and the method to be employed for rotating the antenna.

The general preference is for horizontal polarization, primarily because less height is required to clear surrounding obstructions when all the antenna elements are in the horizontal plane. This is important at 14 and 28 Mc. where the elements are fairly long.

An easily-constructed supporting frame for a horizontal array is shown in Fig. 10-96. It may be made of 1 × 2-inch lumber, preferably oak, for the center sections *B*, *C*, and *D*. The outer arms, *E*, and cross braces, *F*, may be of white pine or cypress. The square block, *A*, at the center supports the whole structure and may be coupled to the pole by any convenient means which permits rotation. Alternatively, the block may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

Another type of construction is shown in Fig. 10-97, with details in Figs. 10-98 and 10-99. This method, suitable for 28-Mc. beams, uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements. Fig. 10-98 also indicates a method of adjusting the lengths of the parasitic elements and bringing the transmission line down through the supporting pole from a delta match. The latter is especially adapted to construction in which the pole rather than the framework alone is rotated.

Metal Booms

Metal can be used to support the elements of the rotary beam. For 28 Mc., a piece of 2-inch diameter duraluminum tubing makes a good "boom" for supporting the elements. The elements can be made to slide through suitable holes in the boom, or special clamps and brackets can be fashioned to support the elements. The antenna of Fig. 10-84D shows one example of such construction.

Generally it is not practicable to support the elements of a 14-Mc. beam by a single-piece boom, because the size of the elements requires a stronger structure. However, by making use of tubing or duraluminum angle, a lightweight support for a 20-meter antenna can be built. The four-element beam shown in

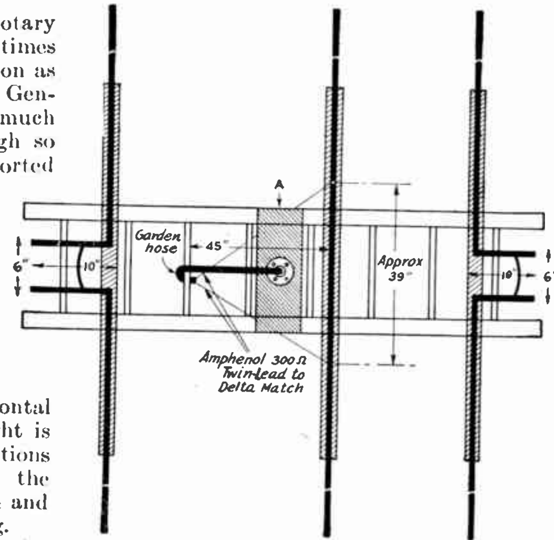


Fig. 10-98 — Top-view drawing of the ladder support and mounted elements. Lengths of director and reflector are adjusted by means of the shorting bars on the small stubs at the center. The drawing also shows a method for pulling off the wires of a delta match and feeding 300-ohm Twin-Lead transmission line through the pipe support.

Figs. 10-100, 10-101 and 10-102 is an example. It uses 1¾-inch angle for the main pieces and ¾-inch angle for the other members, and the entire framework plus elements weighs only forty pounds. This simplifies considerably the problem of supporting the beam.

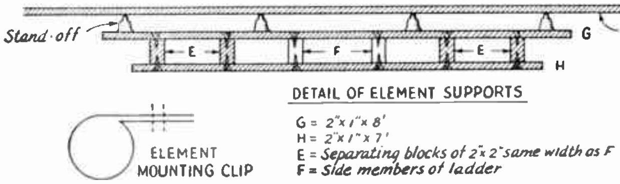


Fig. 10-99 — Detail of element supports for the ladder beam.

The following aluminum pieces are required:

4 — 1-inch diameter tubing, 12 feet long, $\frac{1}{16}$-inch wall
8 — ⅞-inch diameter tubing, 12 feet long, $\frac{1}{32}$-inch wall. Must fit snugly into 1-inch tubing.
2 — 1¾-inch angle, 21 feet long
2 — ¾-inch angle, 21 feet long
4 — ¾-inch angle, 1 foot long
2 — ½-inch diameter tubing, 6 feet long

Aluminum tubing and angle corresponding to the above sizes can possibly be bought from scrap dealers at reasonable prices, if not directly from the manufacturer. If the sections of the elements do not fit snugly, insert shims or make some other provision for a tight fit, since the appearance of the beam will be spoiled by sagging elements. Some amateurs reinforce their beam elements with copper-clad steel wire supported a foot above the elements at the boom and tied to the extreme ends of the elements.

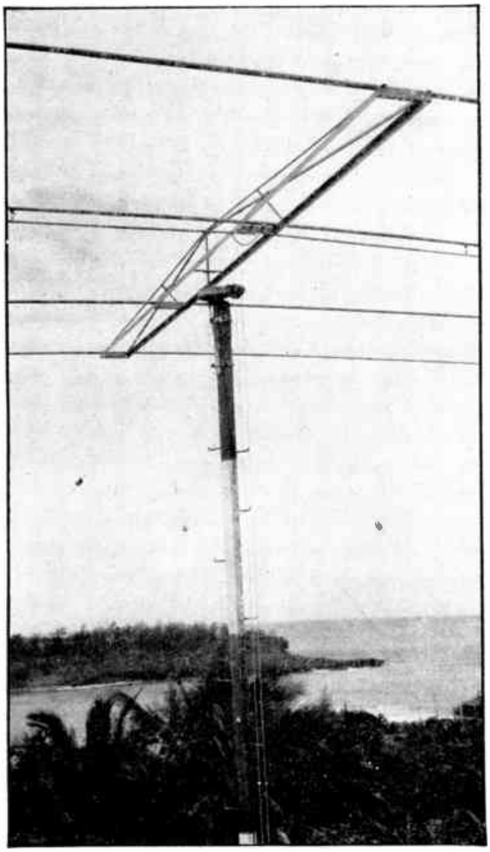


Fig. 10-100 — A four-element 14-Mc. beam of light-weight all-metal construction. Fed by coaxial cable and hand-rotated, the antenna and boom assembly weighs only 40 pounds. (KH6IJ, Dec., 1947, *QST.*)

As shown in Fig. 10-101A, two 1¾-inch aluminum angles 21 feet long serve as the main members of the boom. They are spaced one foot apart. The elements are spaced 7 feet apart. Wooden spacers of 2 × 2 are placed at the end of the boom and screwed on with brass screws. These spacers are also placed under each element where it crosses the boom. These spacers may be unnecessary if the elements are bolted to the boom, but if the construction is as in Fig. 10-101B the spacers are recommended.

The cross braces shown in Fig. 10-102 are put into position at the very last, after the beam is hung in position on the central pivot, since they offer a means for truing up minor sag in the elements.

The central pivot consists of a structure made from ¾-inch angle iron and ½-inch pipe, as shown in Fig. 10-101C. It has to be brazed. The crossbar rest is made separate from the boom and central pivot, and affords a means for tilting the beam when unbolted from these structures. The ½-inch pipe is drilled for the coaxial line that is fed through this pipe. The pinion gear on the ½-inch pipe should be brazed on.

A washing-machine gear train is well suited for this type of beam. Another possibility (used in this instance) is a discarded forge blower. It was fitted with a ½-inch pipe which serves as the central pivot. The gear train ends up in a "V"-pulley, and the beam is easily rotated by a system of ropes and pulleys that ends up in an automobile steering wheel at the operating position. A plumb bob attached to the shaft of the steering wheel serves as a direction indicator. A small cardboard scale mounted along the line of plumb-bob travel can be readily calibrated to show the direction of the beam.

The supporting structure for this beam consists of a 4 × 4 pole 30 feet long, with ten-foot extensions of 2 × 4 bolted to both sides of the bottom, making the total length about 36 feet. Two sets of guy wires should be used, approximately 2 feet and 15 feet from the top. As an alternative, the pole can be set against the side of the house, and only the top set of guys used to provide additional support.

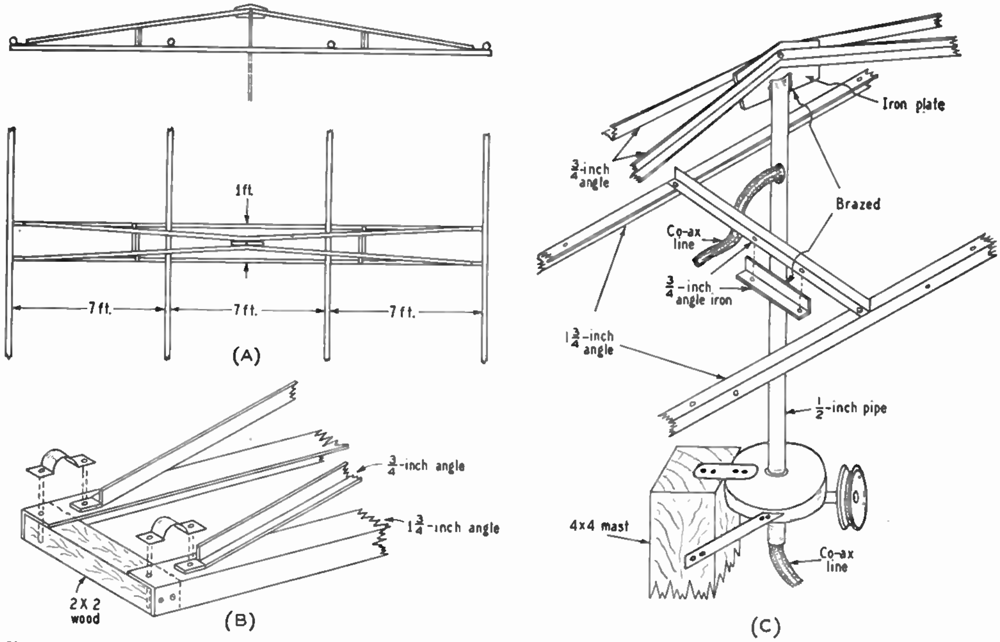


Fig. 10-101 — Details of the 4-element beam construction. The general dimensions and arrangement of the beam are given in A, the detail of the ends of the boom is shown at B, and C shows the construction of the central pivot. A discarded-forge blower gear train is used to drive the assembly.

With all-metal construction, delta match or "T"-match is the only practical matching method to use to the line, since anything else requires opening the driven element at the center, and this complicates the support problem for that element.

A Wooden Boom for 14 Mc.

Many amateurs prefer to build their beam booms from standard pieces of lumber, and the beam shown in Figs. 10-103 and 10-104 is an example of excellent design in wooden-boom construction. The boom members are two 20-foot 2 × 4s fastened to the 4 × 12 × 24-inch center block with six lag screws. The two center screws serve as the axis for tilting — the other four lock the boom in position after final assembly and adjustment have been completed. The blocks midway from each end are 2 × 4s spaced about six inches apart, with a long bolt between them. When this bolt is drawn tight, a very sturdy box brace is formed.

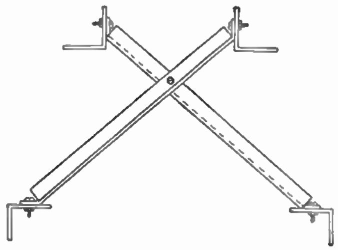


Fig. 10-102 — The boom for the 4-element beam is cross-braced at two points, about 6½ feet in from the ends.

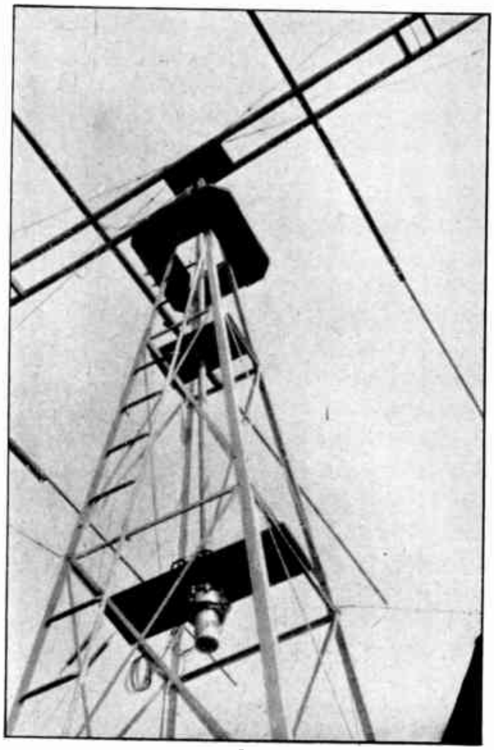


Fig. 10-103 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. This installation is made on a windmill tower, and the drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST*.)

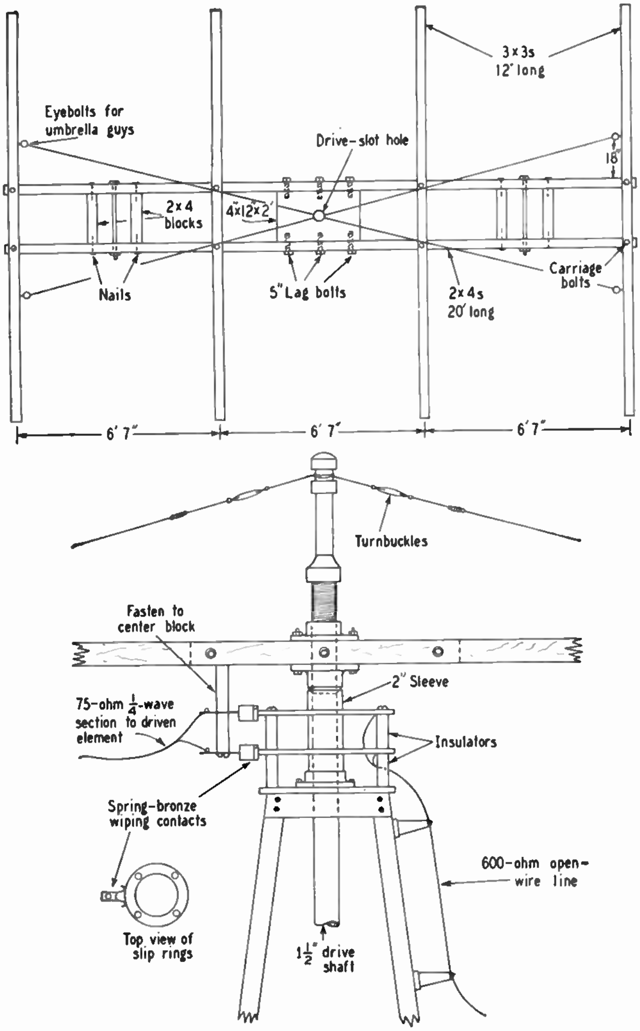


Fig. 10-104 — Details of the wooden boom, its method of support and the construction of the slip rings.

The crossarms are 3 × 3s twelve feet long, bolted to the boom with carriage bolts.

The umbrella guys should have turnbuckles in them, and the guys are fastened to the center support after the beam has been permanently locked in its horizontal position. With the turnbuckles properly adjusted, there will be no sag in the boom, the elements will be parallel and neat, and weaving in the wind will be eliminated.

The elements are 1⅜- and 1½-inch diameter duralumin tubing, supported by 1½-inch standoff insulators. Hose clamps are used to hold the elements on the insulators. Final adjustment of element lengths is possible through "hairpin" loops. The tower for the beam shown in Fig. 10-103 was a Sears-Roebuck windmill tower. The driving motor for the beam was located halfway down the tower, the torque being transmitted through a length of 1½-inch drive shaft. A pipe flange is welded to the drive shaft and bolted to the center block. A cone bearing is obtained by turning both the flange and a sleeve of 2-inch pipe to match, as shown in Fig. 10-104.

One method of matching the line to the antenna is to use a quarter wavelength of 75-ohm Twin-Lead between the radiator and the slip-ring contacts, to match a 600-ohm line from the slip rings to the transmitter.

A 600-ohm open-wire line is run to a point about halfway up on the tower, then up the side of the tower to the slip rings. The slip rings are mounted on the top of the tower, directly under the center block. A quarter-wavelength matching section of transmitting-type 75-ohm Amphenol Twin-Lead hangs in a loop between the driven element and the slip-ring contacts.

"Plumber's-Delight" Construction

The lightest beam to build is the so-called "plumber's delight" — an array constructed entirely of metal, with no insulating members between the elements and the supporting structure. Suggested constructional details are shown in Figs. 10-105, 10-106, 10-107, 10-108 and 10-109.

The boom can be built of two lengths of 3-inch diameter 24ST dural tubing of 0.072-inch wall thickness, as shown in Fig. 10-105. The two sections are spliced together with a three-foot length of 6 × 6 oak, turned down at each end to fit inside the tubing. The center of the block is left square to provide a flat surface to attach to the vertical rotating pipe. At each extremity of this boom is cut a hole the exact diameter of the parasitic elements. A two-foot length of ¾-inch pipe, complete with flange mounting plate,

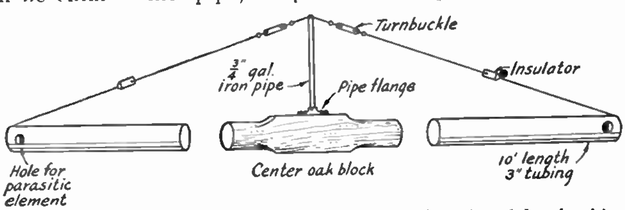


Fig. 10-105 — The boom is made of two 10-foot lengths of dural tubing slipped over a 3-foot oak block and held in place with 2-inch wood screws. Guy wires from the center add strength to the boom structure.

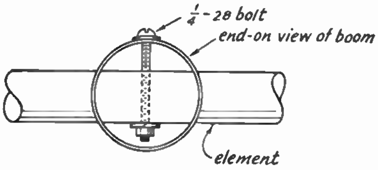


Fig. 10-106 — The center element section is held in the boom with a ¼-28 machine screw, nut and lock washer. The guy wire attaches to the head of the bolt.

is bolted to the top surface of the oak block, and a single guy wire is run to each end of the boom. An egg insulator and a turnbuckle are placed in each guy. The turnbuckles should be tightened until there is no sag in the boom when it is supported at the center, and then safety-wired. Finally the center block should be given a good coat of paint or varnish.

The elements can be made of three 12-foot lengths of dural tubing, the two outside lengths telescoping inside the center section. The ends of the center section should be slotted for a distance of about 4 inches with a hack saw, but it is advisable to do the slotting after the center sections have been assembled on the boom. The parasitic-element center sections are fastened to the boom with ¼-inch bolts, as shown in Fig. 10-106, while the driven element is secured in a cradle made of half sections of iron pipe welded together, as shown in Fig. 10-107. The cradle is bolted to the boom with three ¼-inch bolts, and the driven element is held fast with two bolts or with adjustable aircraft-tubing clamps.

The feedline for the antenna can be any balanced line, of from 200 to 600 ohms imped-

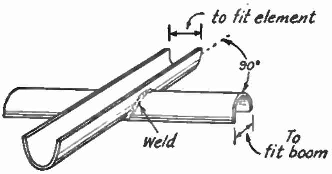


Fig. 10-107 — The clamp for the driven element is made by splitting 1-foot lengths of iron pipe and welding them as shown.

ance, and it is most conveniently coupled through a "T"-match. This "T"-match assembly can be made from two 4-foot lengths of dural tubing joined together by a piece of broomstick, as shown in Fig. 10-109. The "T" is connected to the antenna by two clamps fashioned of 1-inch-wide brass strip.

A convenient method for supporting the boom atop the pipe used to rotate the beam is shown in Fig. 10-108. A "U"-channel into which the boom will fit is welded to the end of the pipe. Holes are drilled in the side of the channel corresponding to holes in the boom. The boom is hoisted up and positioned between the two flanges and a bolt run through the flanges and the boom. The boom can then be swung into a horizontal position and the second bolt put in place.

Feeder Connections

For beams that rotate only 180 degrees, it is relatively simple to bring off feeders by making a short section of the feeder, just where it leaves the rotating member, of flexible wire. Enough slack should be left so that there is no danger of breaking or twisting. Stops should be placed on the rotating shaft of the antenna so that it will be impossible for the feeders to "wind up." This method also can be used with antennas that rotate the full 360 degrees, but again a stop is necessary to avoid jamming the feeders.

For continuous rotation, the sliding contact is simple and, when properly built, quite prac-

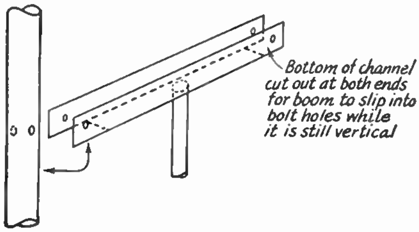


Fig. 10-108 — The mounting plate is made from a length of "U"-channel iron cut and drilled as shown. The boom is raised vertically until one set of bolt holes is in line and a bolt is slipped through. The boom is then swung into its horizontal position and the other bolt is put in place.

ticable. Fig. 10-110 shows two methods of making sliding contacts. The chief points to keep in mind are that the contact surfaces should be wide enough to take care of wobble in the rotating shaft, and that the contact surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition. Sliding contacts preferably should be used with nonresonant open lines where the characteristic impedance is of the order of 500 to 600 ohms, so that the line current is low.

The possibility of poor connections in sliding contacts can be avoided by using inductive coupling at the antenna, with one coil rotating on the antenna and the other fixed in position, the two coils being arranged so that the coupling does not change when the antenna is rotated. Such an arrangement is shown in Fig. 10-111, adapted to an antenna system in which the pole

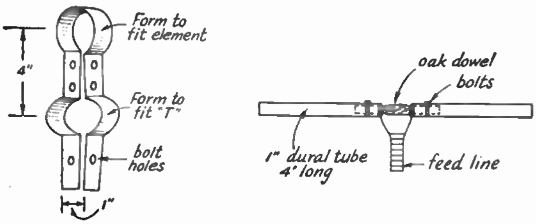


Fig. 10-109 — Details of the "T"-match assembly.

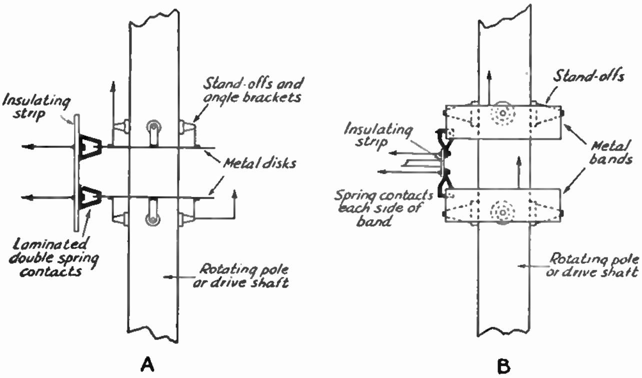


Fig. 10-110 — Ideas in sliding contacts for rotatable antenna feeder connection to permit continuous rotation. The broad bearing surfaces take care of any wobble in the rotating mast or driving shaft.

itself rotates. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. In the drawing, the link coil connects to a twisted-pair transmission line, but any type of line such as flexible coaxial cable can be used. The circuit would be adjusted in the same way as any link-coupled circuit, and the number of turns in the link should be varied to give proper loading on the transmitter. The rotating coupling circuit of course tunes to the transmitting frequency. The whole thing is equivalent to a link-coupled antenna tuner mounted on the pole, using a parallel-tuned tank at the end of a quarter-wave line to center-feed the antenna. To maintain constant coupling, the two coils should be quite rigid and the pole should rotate without wobble. The two coils might be made a part of the upper bearing assembly holding the rotating pole in position.

Other variations of the inductive-coupled system can be worked out. The tuned circuit might, for instance, be placed at the end of a 600-ohm line, and a one-turn link used to couple directly to the center of the antenna, if the construction of the rotary member permits. In this case the coupling can be varied by changing the L/C ratio in the tuned circuit. For mechanical strength the coupling coils preferably should be made of ¼-inch copper tubing, well braced with insulating strips to keep them rigid.

Rotation

It is convenient to use a motor to rotate the beam, but it is not always necessary, especially if a rope-and-pulley arrangement can be brought into the operating room. If the pole can be mounted near a window in the operating room, hand rotation of the beam will work out quite well, as has been proven by many amateur installations.

If the use of a rope and pulleys is impracticable, motor drive is about the only alternative. There are several complete motor-driven rotators on the market, and they are easy to mount, convenient to use, and require little or no maintenance. However, to many the cost of such units puts them out of reach, and a homemade unit must be considered. Generally speaking, lightweight units are better because they reduce the load on the mast or tower.

The speed of rotation should not be too great — one or two r.p.m. is about right. This requires a considerable gear reduction from the usual 1750-r.p.m. speed of small induction motors; a large reduction is advantageous because the gear train will prevent the beam from turning in weather-vane fashion in a wind. The ordinary structure does not require a great deal of power for rotation at slow speed, and a ⅛-hp. motor will be ample. Even small series motors of the sewing-machine type will develop enough power to turn a 28-Mc. beam at slow speed. If possible, a reversible motor should be used so that it will not be necessary to go through nearly 360 degrees to bring the beam back to a direction only slightly different, but in the opposite direction of rotation, to the direction to which it may be pointed at the moment. In cases where the pole is stationary and only the supporting framework rotates, it will be necessary to mount the motor and gear train in a housing on or near the top of the pole. If the pole rotates, the motor can be installed in a

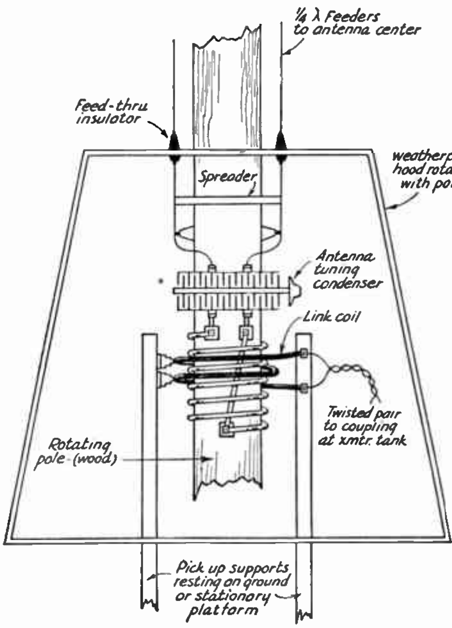


Fig. 10-111 — One method of transmission line-antenna system coupling which eliminates sliding contacts. The low-impedance line is link-coupled to a tuned line.

more accessible location (see Fig. 10-103).

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. Driving motors and gear housings will stand the weather better if given a coat of aluminum paint followed by two coats of enamel and a coat of glyptal varnish. Even commercial units will last longer if treated with glyptal varnish. Be sure, of course, that the surfaces are clean and free from grease before painting them. Grease can be removed by brushing it with kerosene and then squirting the surface with a solid stream of water. The work can then be wiped dry with a rag.

If hand rotation of the beam is used, or if the rotating motor drives the beam through a pulley system, bronze cable or chain drive is preferable to rope. However, if you must use rope, be sure to soak it overnight in pure linseed oil and then let it dry for several days before permanent installation.

The power and control leads to the rotator should be run in electrical conduit or in lead covering, and the metal should be grounded. Often r.f. appearing in power leads can be reduced by suitable filtering, but running wires in conduit is generally easier and more satisfactory. Any r.f. in the wiring can sometimes be responsible for feed-back in a 'phone transmitter. "Hash" from the motor is also reduced by shielding the wires, but it is often necessary to install a small filter at the motor to reduce this source of interference. Motor noise appearing in the receiver is a nuisance, since it is usual practice to determine the proper direction for the beam by rotating it while listening to the station it is desired to work and setting the antenna at the point that gives maximum signal strength.

The outside electrical connections should be soldered, bound with rubber tape followed by regular friction tape, and then given a coat of glyptal varnish.

About V.H.F.

In the days when DX activity first burgeoned on our lower frequencies the assignments above 30 Mc. were not too highly regarded. It was assumed that propagation on these frequencies was limited to distances only slightly beyond the visual horizon, and thus the bands allocated to amateurs in this region were used principally in areas where large concentrations of population brought hundreds of workers within local range of one another. In the early thirties activity boomed on 56 Mc. in the larger cities of the United States, but there were few stations elsewhere. Use of frequencies higher than 60 Mc. was confined to a few experimentally-inclined amateurs here and there.

In 1934, '35 and '36, new types of propagation were discovered by amateurs, and the opportunities for v.h.f. DX so brought to light caused a tremendous growth in activity, particularly in areas where it had not previously existed. Up to this time, practically all v.h.f. work had been done with the simplest sort of gear, mainly modulated-oscillator transmitters and superregenerative receivers; but when our available space began to fill with DX signals it became obvious that, if we were to realize anything like the possibilities inherent in this type of work, we must have improved techniques, whereby more stations could be accommodated in a given area. Crystal-controlled transmitters and superheterodyne receivers, permitting utilization of the 56-Mc. band on a scale comparable with that obtaining on lower frequencies, became the order of the day, and by the end of 1938 stabilization of transmitters used on all frequencies up to 60 Mc. became mandatory.

With the impetus of improved techniques, operating ranges on 56 Mc. grew by leaps and bounds. Meanwhile the use of the simplest form of equipment was transferred to the next higher band, then 112 Mc.; and this band, in turn, took over the burden of heavy urban occupancy formerly carried by the 5-meter band. Soon our principal cities were teeming with 112-Mc. activity, and before long it was found that this band, too, had much of interest to offer. Even more than had been the case on 56 Mc., it was found that weather conditions had a profound effect on 112-Mc. propagation, and before the close-down of amateur activity, at the entry of our country into the war, the record for 112-Mc. work had passed the 300-mile mark. There was a smattering of activity on the still higher frequencies of 224 and 400 Mc. as well.

In the postwar years the value of the very-high frequencies has been amply demonstrated. World-wide communication has been accomplished on 50 Mc.; two-way work on 144 Mc. has been extended to nearly 700 miles; and pioneering effort on 220 and 420 Mc. is establishing these bands as fields of great interest for the experimentally-inclined amateur. The v.h.f. worker need no longer apologize for his interests. His frequencies are among the most highly prized in the entire spectrum, and his is now regarded as one of the major fields of amateur endeavor.

Propagation Phenomena

A thorough understanding of the basic principles of wave propagation, outlined in Chapter Four, is a most useful tool for the v.h.f. worker. Much of the pleasure and satisfaction to be derived from v.h.f. endeavor lies in making the best possible use of propagation vagaries resulting from natural phenomena. Contrary to the impression of many newcomers to the field, a working knowledge of v.h.f. propagation is not difficult of attainment. Below are listed the principal ways by which v.h.f. waves may be propagated over abnormal distances.

F_2-Layer Reflection

The "normal" contacts made on 28 Mc. and lower frequencies are the result of reflection of the transmitted wave by the F_2 layer, the ionization density of which varies with solar activity, the highest frequencies being reflected at the peak of the 11-year solar cycle. The maximum usable frequency (m.u.f.) for F_2 reflection also rises and falls with other well-defined cycles, including daily, monthly, and seasonal variations, all related to conditions on the sun and its position with respect to the earth.

At the low point of the 11-year cycle, such as the period we were entering at the outbreak of war, the m.u.f. may reach 28 Mc. only during a short period each spring and fall, whereas it may go to 60 Mc. or higher at the peak of the cycle. The fall of 1946 saw the first authentic

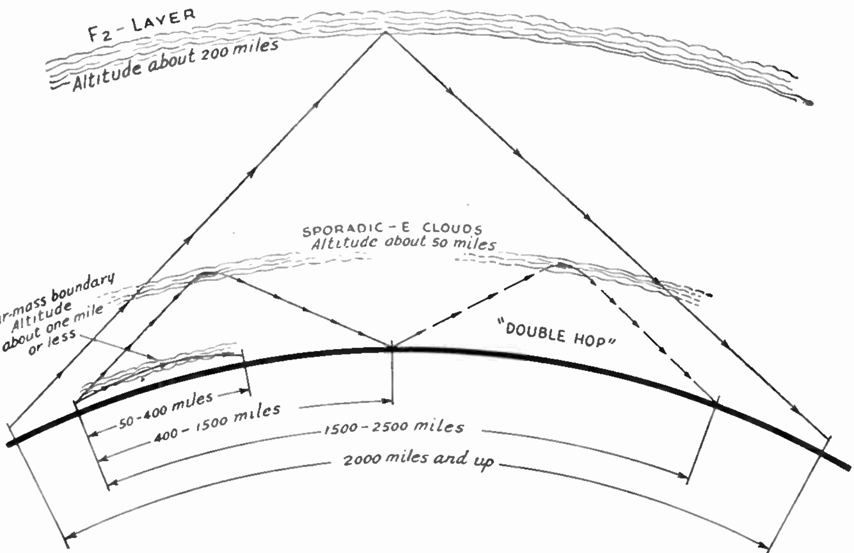


Fig. 11-1 — The principal means by which v.h.f. waves may be returned to earth. The F_2 layer, highest of the known ionospheric layers, is capable of reflecting 50-Mc. signals during the period around the peak of the 11-year solar cycle, and may support communication over world-wide distances. Sporadic ionization of the *E* layer produces "short-skip" contacts at medium distances. It is a fairly frequent occurrence regardless of the solar cycle, but is most common in May through August. Refraction of v.h.f. waves also takes place at air-mass boundaries in the lower atmosphere, making possible reception of signals at distances up to 300 miles or more without a skip zone.

instances of long-distance 50-Mc. work by this medium, and it is probable that F_2 DX will be workable on 50 Mc. until about 1950. In the northern latitudes there are peaks of m.u.f. each spring and fall, with a low period during the summer and a slight dropping off during the midwinter months. At or near the Equator conditions are more or less constant at all seasons.

Fortunately the F_2 m.u.f. is quite readily determined by observation, and means are available whereby it may be estimated quite accurately for any path at any time. It is predictable for months in advance,[1] enabling the v.h.f. worker to arrange test schedules with distant stations at propitious times. As there are numerous signals, both harmonics and fundamental transmissions, on the air in the range between 28 and 50 Mc., it is possible for the listener to determine the approximate m.u.f. by careful listening in this range. A series of daily observations will serve to show if the m.u.f. is rising or falling from day to day, and once the peak for a given month is determined it can be assumed that the peak for the following month will occur about 27 days later, this cycle coinciding with the turning of the sun on its axis. The working range, via F_2 skip, will be roughly comparable to that on 28 Mc., though the *minimum* distance is somewhat longer. Two-way work on 50 Mc. by means of reflection from the F_2 layer has been accomplished over distances ranging from 2200 to 10,500 miles. The maximum frequency for F_2 reflection is believed to be in the vicinity of 70 Mc.

Sporadic-E Skip

Patchy concentrations of ionization in the *E*-layer region are often responsible for reflection of signals on 28 and 50 Mc. This is the popular "short skip" that provides fine contacts on both bands in the range between 400 and 1300 miles. It is most common in May, June and July, during the early evening hours, but it may occur at any time or season. Since it is largely unpredictable, at our present state of knowledge, sporadic-*E* skip is of high "surprise value." Multiple-hop effects may appear, when ionization develops simultaneously over large areas, making possible work over distances of more than 2500 miles. So far as is known, no 144-Mc. effects have yet been observed, the known limit for sporadic-*E* skip being in the vicinity of 100 Mc.

Aurora Effect

Low-frequency communication is occasionally wiped out by absorption of these frequencies in the ionosphere, when ionospheric storms, associated with variations in the earth's magnetic field, occur. During such disturbances, however, 50-Mc. signals may be reflected back to earth, making communication possible over distances not normally workable on this band. Magnetic storms may be accompanied by an aurora-borealis display, if the

[1] *Basic Radio Propagation Predictions*, issued monthly, three months in advance, by the Central Radio Propagation Laboratory of the National Bureau of Standards. Order from the Supt. of Documents, Washington 25, D. C.; $1.00 per year.

disturbance occurs at night and visibility is good. When the aurora is confined to the northern sky, aiming a directional array at the auroral curtain will bring in 50-Mc. signals strongest, regardless of the true direction to the transmitting station. When the display is widespread there may be only a slight improvement noted when the array is aimed north. The latter condition is often noticed during the period around the peak of the 11-year cycle, when solar activity is spread well over the sun's surface, instead of being concentrated extensively in the region near the solar equator.

Aurora-reflected signals are characterized by a rapid flutter, which lends a "dribbling" sound to 28-Mc. carriers and may render modulation on 50-Mc. signals completely unreadable. The only satisfactory means of communication then becomes straight c.w. The effect may be noticeable on signals from any distance other than purely local, and stations up to about 500 miles in any direction may be worked at the peak of the disturbance. Unlike the two methods of propagation previously described, aurora effect exhibits no skip zone. It has been observed mainly on frequencies up to about 60 Mc., though there have been rare instances of it in our 144-Mc. band.

Reflections from Meteor Trails

Probably the least-known means of v.h.f. wave propagation is that resulting from the passage of meteors across the signal path. Reflections from the ionized meteor trails may be noted as a Doppler-effect whistle on the carrier of a signal already being received, or they may cause bursts of reception from stations not normally receivable. Sudden large increases in strength of normally-weak signals are another manifestation of this effect. Ordinarily such reflections are of little value in extending communication ranges, since the increases in signal strength are of short duration, but meteor showers of considerable magnitude and duration may provide fluttery 50-Mc. signals from distances up to 1000 miles or more. Signals so reflected have a combination of the characteristics of aurora and sporadic-*E* skip.

Tropospheric Bending

Refraction of radio waves takes place whenever a change in refractive index is encountered. This may occur at one of the ionized layers of the ionosphere, as mentioned above, or it may exist at the boundary area between two different types of air masses, in the region close to the earth's surface. A warm, moist air mass from over the Gulf of Mexico, for instance, may overrun a cold, dry air mass which may have had its origin in northern Canada. Each tends to retain its original characteristics for considerable periods of time, and there may be a well-defined boundary between the two for as much as several days. When such an air-mass boundary exists near the midpoint between two v.h.f. stations separated by 50 to 300 miles or more, a considerable degree of refraction takes place, and signals run high above the average value. Under ideal conditions there may be almost no attenuation, and signals from far beyond the visual horizon will come through with strength comparable to that of local stations.

Many factors other than air-mass movement of a continental character may provide increased v.h.f. operating range. The convection that takes place along our coastal areas in warm weather is a good example. The rapid cooling of the earth after a hot day in summer, with the air aloft cooling more slowly, is another, producing a rise in signal strength in the period around sundown. The early-morning hours, when the sun heats the air aloft, before the temperature of the earth's surface begins its daily rise, may frequently be the best hours of the day for extended v.h.f. range, particularly in clear, calm weather, when the barometer is high and the humidity low.

Any weather condition that produces a pronounced boundary between air masses of different temperature and humidity characteristics provides the medium by which v.h.f.

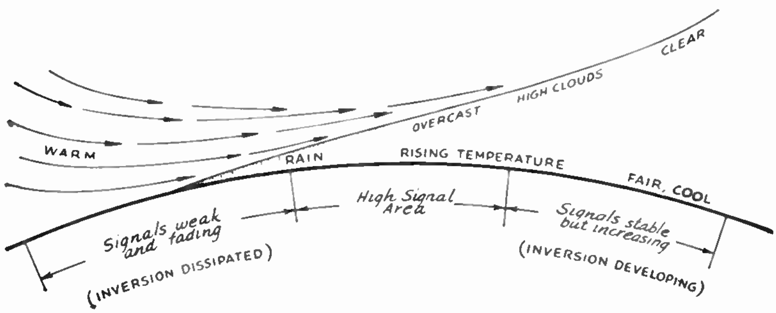


Fig. 11-2 — Illustrating a typical weather sequence, with associated variations in v.h.f. propagation. At the right is a cold air mass (fair weather, high or rising barometer, moderate summer temperatures). Approaching this from the left is a warm moist air mass, which overruns the cold air at the point of contact, creating a temperature inversion and considerable bending of v.h.f. waves. At the left, in the storm area, the inversion is dissipated and signals are weak and subject to fading. Barometer is low or falling at this point.

signals cover abnormal distances. The ambitious v.h.f. enthusiast soon learns to correlate various weather manifestations with radio-propagation phenomena. By watching temperature, barometric pressure, changing cloud formations, wind direction, visibility, and other easily-observed weather signs, he is able to tell with a reasonable degree of accuracy what is in prospect on the v.h.f. bands.

The responsiveness of radio waves to varying weather conditions increases with frequency. Our 50-Mc. band is considerably more sensitive to weather variations than is the 28-Mc. band, and the 144-Mc. band may show strong signals from far beyond visual distances when the lower frequencies are relatively inactive. It is probable that this tendency continues on up through the microwave range, and that our assignments in the u.h.f. and s.h.f. portions of the frequency spectrum may someday support communication over distances far in excess of the optical range. Already 144-Mc. communication by amateurs has passed the 600-mile mark, and even greater distances are believed possible on this and higher frequencies.

STATION LOCATIONS

In line with our early notions of v.h.f. wave propagation, it was once thought that only highly-elevated v.h.f. stations had any chance of working beyond a few miles. Almost all the work was done by portable stations operating from mountain tops, and only hilltop home sites were considered suitable for fixed-station work. It is still true that the fortunate amateur who lives at the top of a hill enjoys a certain advantage over his fellows on the v.h.f. bands, but high elevation is not the all-important factor it was once thought to be.

Improvements in equipment, the wide use of high-gain antenna systems, and an awareness of the opportunities afforded by weather phenomena have enabled countless v.h.f. workers to achieve excellent results from seemingly poor locations. In 50-Mc. DX work particularly, elevation has ceased to be an important factor, though it may help in extending the range of operation somewhat under normal conditions. A high elevation is somewhat more helpful on 144 Mc. and higher frequencies, particularly when no unusual propagation factors are present, as during the winter months. Other factors, such as close proximity to large bodies of water, may more than compensate for lack of elevation during the other seasons of the year, however.

Stations situated in sea-level locations along our coasts have been consistent in their ability to set distance records on 144 Mc.; weather variations provide interesting propagation effects over our Middle Western plain areas; and even the worker situated in mountainous country need not necessarily feel that he is prevented by the nature of his horizon from doing interesting work. Contacts have been made on 50 and 144 Mc. over distances in excess of 100 miles in all kinds of terrain.

The consistently-reliable nature of 50 and 144 Mc. for work over such a radius and more, regardless of weather, time or season, and the occasional opportunities these frequencies afford for exciting DX, have caused an increasing number of amateurs to migrate to the v.h.f. bands for extended-local communication, once thought possible only on the lower frequencies.

V.H.F. Techniques

Recognition of the value of the very-high frequencies has resulted in the development of many tubes and other components especially suited for use at these frequencies. Where, not so many years ago, it was necessary to remove the bases from available tubes, and otherwise cut down components designed for use at lower frequencies, we now have tubes and circuit components specifically designed for high-efficiency operation, not only on the v.h.f. bands, but on up through the microwave range. Examples of transmitting tubes especially made for v.h.f. work are shown in Fig. 11-3.

The higher frequencies are rapidly becoming the primary field of interest for those amateurs who like to design and build their own equipment. While there is an increasing tendency to the purchase of commercially-built equipment, both transmitting and receiving, for low-frequency operation, most gear used for v.h.f. and higher bands is still a product of the amateur's own ingenuity.

In the field of antenna design, too, the v.h.f. bands offer much to the amateur who is interested in experimental work. With their smaller physical size making for greater ease of construction and adjustment, the development of high-gain directional antennas continues to occupy much of the time devoted by the v.h.f. enthusiast to experimental work.

TRANSMITTER DESIGN

The use of crystal control, or its equivalent in stability, is standard for 50-Mc. work. The design of transmitters for this band differs hardly at all from that employed for lower amateur frequencies, except that much more care must be exercised in the selection of component parts and their placement in the equipment, in order to avoid more than the absolute minimum length in the connecting leads. Customary procedure is to start with a crystal or variable-frequency oscillator, operating at 6, 8, or 12 Mc., and follow with such frequency-multiplying stages as may be required to reach 50 Mc. The power level is not particularly important, as interference is not a critical factor in 50-Mc. communication. Much good work has, in fact, been accomplished with power inputs under 100 watts and even stations in the 10-watt class are quite capable of working out well, particularly if equipped with well-designed antenna systems.

At 144 Mc., crystal control, though not required by law, is now almost universally used, and it is being employed by the more advanced workers on 220 Mc. as well. Its use becomes increasingly difficult, but results obtained are more than worth the added effort. The number of tube types usable above 100 Mc. or so is limited, and only those designed specifically for v.h.f. service can be used successfully. Great care must be exercised to keep lead inductance and circuit capacitance to a minimum. Conventional coil-and-condenser combinations designed for lower frequencies are generally unsatisfactory, and only well-designed tank circuits will function efficiently at 144 Mc. and higher. High power is seldom used, most workers preferring to use high-gain antenna systems rather than high-powered transmitters in v.h.f. work.

Multistage-transmitter design may even be employed for 420 Mc., but the modulated oscillator still bears the brunt of operation on this and higher bands. Since occupancy is relatively low and the bands are wider than those lower in frequency, the broader signals radiated by such equipment, and the inefficiencies of the superregenerative receivers necessary to accommodate them, are not major problems.

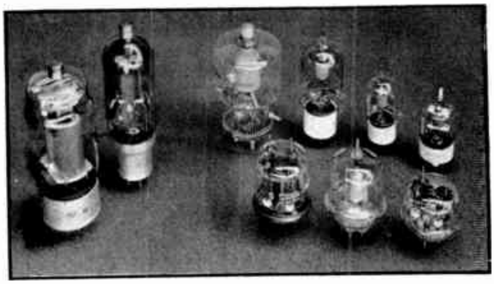


Fig. 11-3 — Vacuum tubes designed especially for high-efficiency operation at very-high frequencies are now available for amateur use. Several such tubes are shown above, in comparison to typical low-frequency tubes, the 813 and V-70-D at the left. The v.h.f. types are the GL-592, 35TG, 24G, HY-75-A, all triodes; and the 829-B, HK-57, and 832-A, all tetrodes.

RECEIVER CONSIDERATIONS

Even more than in work on lower frequencies, a good receiver is all-important in the v.h.f. station. Though commercial receivers that cover the 50-Mc. band are slowly appearing on the amateur market, the most satisfactory and inexpensive solution to the receiver problem is still that of a converter that works into a communications receiver designed for the lower frequencies. Such a combination is almost certain to give better results on 50 Mc. than a complete receiver, unless the latter is designed especially for v.h.f. use.

Converters are replacing the once-popular superregenerative receivers for 144-Mc. use also, particularly for fixed-station work in localities where the use of stabilized transmitters has become more or less standard procedure. Many types of superhet receivers used for radar and aircraft service during the war are convertible to amateur use, and hundreds of such surplus units are now employed by amateurs working on 220 and 420 Mc.

For portable or emergency use, where small size and low battery drain are important, the simple superregenerative receiver is still popular. For the number of tubes and parts required, it is still an efficient receiving system, especially in areas where there is not extensive activity. For frequencies higher than 148 Mc. it is still the principal receiving system, though the converter approach is practical for any frequency. To accommodate the broader signals generally found on these frequencies, a converter may be used in conjunction with a wide-band i.f. system, such as a receiver designed for FM broadcast reception.

V.H.F. Receivers

In its basic principles, modern receiving equipment for 50, 144 and even 220 Mc. differs very little from that used on lower amateur frequencies. Federal regulations impose identical restrictions on all frequencies below 54 Mc. as to stability of transmitted signals, and experience has shown that only through the use of stabilized transmitters and selective receivers can the full possibilities of 144 and 220 Mc. be realized. Thus, for AM or NFM reception, at least, receivers for the v.h.f. bands may have the same selectivity as those used for lower frequencies.

This order of selectivity is not only possible but desirable, since it permits a considerable increase in the number of stations that can work in a band without harmful interference. High selectivity also aids greatly in improving the signal-to-noise ratio, both as to noise originating in the receiver itself and its response to external noise. The effective sensitivity of a receiver having "communications" selectivity can be made considerably higher than is possible with nonselective receivers. First on the old 56-Mc. band in the late '30s, then on 144 Mc. in the early part of the postwar period, and currently on 220 Mc., the change to selective superheterodynes for use at the more progressive v.h.f. stations marked the beginning of real extensions of the effective operating radius of v.h.f. stations.

Superregenerative receivers, once the most popular type for v.h.f. work, are now used mainly for portable operation, or for other applications where maximum selectivity and sensitivity are not required. Its lack of these essential features, its inability to provide satisfactory reception of FM signals, and its tendency to radiate a strong interfering signal, rule the superregenerator out as a home-station receiver in areas where there is appreciable activity on the v.h.f. bands.

Superheterodynes for V.H.F.

Superheterodynes for 50 Mc. and higher should have fairly-high intermediate frequencies to reduce both image response and oscillator "pulling." 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 nearly as great as to the desired frequency. To obtain discrimination against the image equal to that obtainable at 3.5 Mc. would require an i.f. 16 times as high, or about 7 Mc. However, the Q of tuned circuits is less in the v.h.f. range than it is at lower frequencies, chiefly because the tube loading is considerably greater, and thus still higher intermediate frequencies are desirable. A practical compromise is reached at about 10 Mc., and the standard i.f. for converters and commercial v.h.f. receivers is 10.7 Mc.

To obtain the desired degree of selectivity with a reasonable number of i.f. stages, the double-conversion principle is often employed. A 10-Mc. intermediate frequency, for example, is changed to an i.f. of 1600 or 455 kc. by adding a second mixer-oscillator combination.

Most v.h.f. receivers are of this category, general practice being to use a conventional communications receiver to handle the i.f. output of a relatively-simple converter. Even a broadcast receiver which has a "short-wave band" may be used as an i.f. amplifier in this manner with good results. Only crystal-controlled or otherwise stabilized signals can be received with such combinations, but since nearly all v.h.f. amateur stations now employ stabilized transmitters this is not likely to be troublesome.

When a high-selectivity i.f. is employed in v.h.f. reception, the stability of the oscillator is a primary problem, and care must be taken to be sure that the converter oscillator is both mechanically and electrically stable. One satisfactory solution to this problem is the use of a crystal-controlled oscillator and frequency multiplier to supply the injection voltage, the method used in the 144-Mc. converter shown in Figs. 12-13–12-15.

Where reception of wide-band FM or unstable signals of the modulated-oscillator type is desired, a converter may be used ahead of an i.f. of the type used for FM broadcast reception, or with a complete receiver of the FM broadcast variety. A superregenerative detector operating at the intermediate frequency, with or without additional i.f. amplifier stages, also may serve as an i.f. and detector system for reception of wide-band signals. By using a high i.f. (10 to 30 Mc. or so) and by resistive loading of the i.f. transformers, almost any desired degree of bandwidth can be secured, providing good voice quality on all but the most unstable signals. Any of these methods may be used for reception in the u.h.f. and microwave regions, where stabilized transmission is extremely difficult at the current state of the art.

A Two-Tube Converter for 50 Mc.

The converter shown in Figs. 12-1–12-4 is designed to provide good performance on 50 Mc. with a minimum of complication. It employs a 6AK5 tuned r.f. stage and a dual-triode mixer-oscillator using a 12AT7. It has its own built-in power supply. To reduce tracking problems and simplify construction, only the oscillator is tuned by means of the vernier dial. The r.f. amplifier grid circuit and the mixer grid circuit, neither of which is critical in its tuning, are ganged together, and are adjusted by means of the knob at the left of the vernier dial, as seen in Fig. 12-1. In actual operation this control may be peaked at about 51 Mc. and the converter tuned over the lower half of the band or more before any readjustment of the knob is required.

Electrical and Mechanical Details

One section of the 12AT7 is used as a triode oscillator, employing a Colpitts circuit. It covers a range of six megacycles, in order to permit tuning well below the low end of the 50-Mc. band, a useful feature when one is interested in searching these frequencies for signs of DX. By resetting C_4 and C_5 to a higher capacitance the tuning range may be extended down to about 45 Mc., at the sacrifice of the top three megacycles of the 6-meter band.

The oscillator is tuned with a standard split-stator variable capacitor, C_3, the rotor of which is grounded. Parallel capacitance, for increased stability, is supplied by the small air padders, C_4, C_5. The oscillator tank coil, L_5, is attached directly to the stator terminals of the tuning condenser, and its center-tap provides three-point suspension, resistor R_8 being attached to a feed-through bushing directly below it. To prevent microphonics resulting from vibration, the turns of the coil are cemented together at four points.

The 6AK5 and 12AT7 tubes are mounted in an inverted position, with their sockets above the chassis, in order to provide the shortest possible leads. The socket for the 6AK5 has the baffle plate between the two tuned circuits passing across the middle of the socket in such a way that Pins 1 through 4 are on the enclosed side, while the others, which are concerned with the output circuits, are on the side toward the panel. The shield is made of sheet copper and is soldered to the cylindrical shield at the center of the 6AK5 socket. A high value of L/C ratio is maintained in the r.f. and mixer stages by the elimination of the parallel padder condensers which would have been required for tracking, if the oscillator had been tuned from the same shaft.

The intermediate frequency may be any convenient value, and the i.f. transformer is made plug-in so that the frequency may be changed over a wide range if desired. With the values of inductance and capacitance given the i.f. can be varied from about 7 to 11 Mc. Originally 10.7 Mc., the RMA standard i.f. for converters, was tried, but some image trouble was experienced from strong local 10-meter stations, so the i.f. transformer was retuned to 7.4 Mc. Note the position of C_6, the mica trimmer visible in the top-view photograph. It was found necessary to mount this trimmer directly on the 12AT7 terminals, as the mixer oscillated when the trimmer was connected across the coil terminals below the chassis.

Some precautions may also be required to prevent oscillation in the r.f. stage. Note the manner in which the screen and cathode circuits of the 6AK5 are by-passed. Referring to the schematic diagram, Fig. 12-2, it may be seen that both cathode terminals of the 6AK5 are by-passed, the bias resistor, R_1, and its by-pass, C_7, being on the input side, while C_8 is on the opposite side of the shield. Originally all the cathode connections were made to Pin 7,

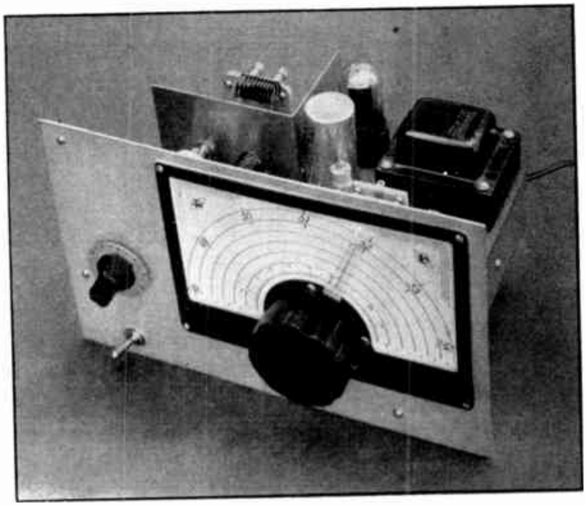


Fig. 12-1 — A two-tube converter for 50 Mc. Only the oscillator is tuned by the vernier dial, simplifying tracking problems. Mixer and r.f. circuits are adjusted by the knob at the left. The unit has a self-contained regulated power supply.

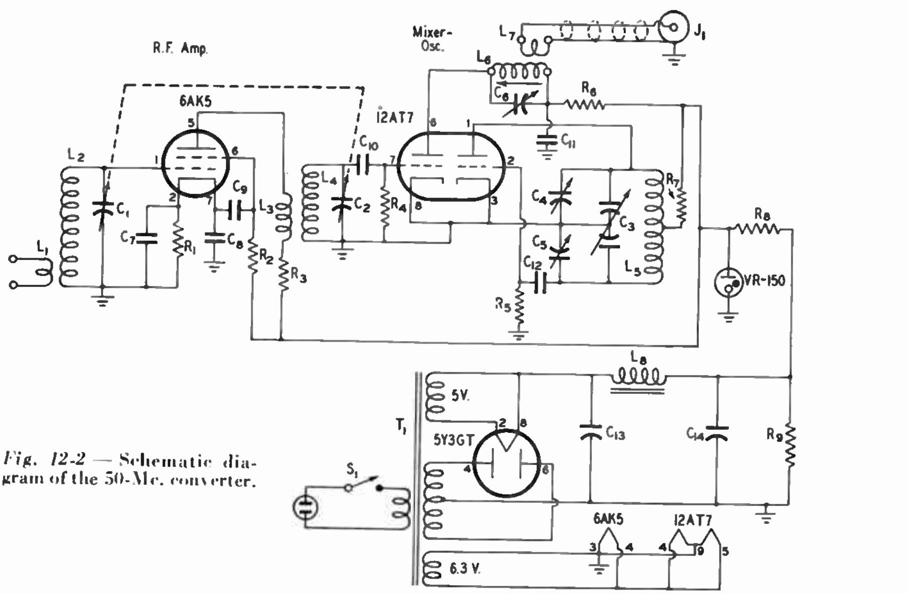


Fig. 12-2 — Schematic diagram of the 50-Mc. converter.

C_1, C_2 — 15-μμfd. variable (Hammarlund MC-20-S with one stator plate removed).
C_3 — 15-μμfd.-per-section split stator (Cardwell ER-15-AD).
C_4, C_5 — 3–30-μμfd. air trimmer (Silver type 619).
C_6 — 3–30-μμfd. mica trimmer (See text for mounting position).
C_7, C_8, C_9 — 680-μμfd. mica.
C_{10}, C_{12} — 100-μμfd. ceramic.
C_{11} — 0.001-μfd. mica.
C_{13}, C_{14} — 10-μfd. 450-volt dual electrolytic.
R_1 — 150 ohms, ½ watt.
R_2, R_5, R_6, R_7 — 10,000 ohms, ½ watt.
R_3 — 2200 ohms.
R_4 — 1.0 megohm.
R_8 — 5000 ohms, 10 watts.
R_9 — 25,000 ohms, 10 watts.
L_1 — 4 turns No. 22 enamel, interwound in cold end of L_2.
L_2 — 10 turns No. 16 enamel, ½-inch inside diam., 1⅛ inches long.
L_3 — 5 turns No. 22 enamel, interwound in cold end of L_4.
L_4 — 9 turns No. 16 enamel, ½-inch inside diam., ⅞ inch long.
L_5 — 11 turns No. 12 enamel, center-tapped, 1⅜ inches long. Cement turns together — see text.
L_6 — 28 turns No. 24 d.s.c., close-wound.
L_7 — 4 turns No. 24 d.s.c. close-wound over cold end of L_6 with one layer of insulating tape between windings.
L_8 — Midget filter choke.
J_1 — Coaxial fitting (Jones S-201).
S_1 — S.p.s.t. toggle switch.
T_1 — Midget power transformer (Thordarson T-13R19).

and the screen was by-passed to ground. With this arrangement the r.f. stage oscillated, so the capacitors were rearranged so that the screen was by-passed (by C_9) to the cathode, Pin 7, and both cathode pins by-passed separately. Even this way the stage may oscillate when no antenna is connected, but in normal operation it is completely stable.

Adjustments

Because of the separate tuning of the oscillator stage, alignment of the converter presents no great problem. First the oscillator should be set to cover the desired range, in this case 40.6 to 46.6 Mc., for a tuning range of 48 to 54 Mc. with an i.f. of 7.4 Mc. The i.f. transformer may then be set to the proper point, by moving the slug or adjusting C_6. A signal generator is convenient, but the adjustment can be made by listening for a noise peak, if no generator is available.

Next the flexible coupling between C_1 and C_2 should be disengaged so that the r.f. and mixer circuits can be adjusted separately. First, an antenna or signal source should be coupled to the mixer grid coil, L_4, and C_2 peaked for

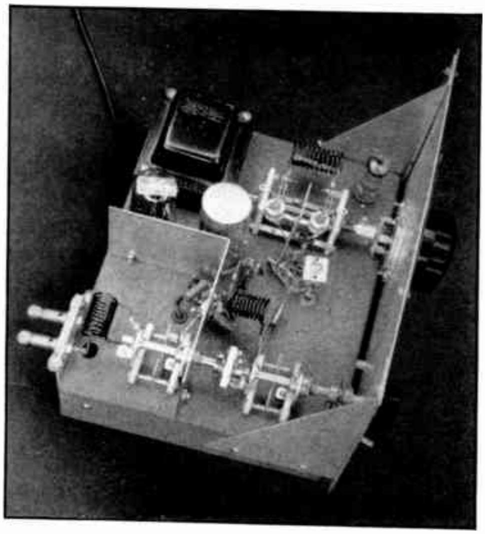


Fig. 12-3 — Top view of the 50-Mc. converter, showing placement of the principal r.f. components. The shielded plug-in coil near the middle of the chassis is the i.f. output transformer. The tube at the upper right of the photo is the voltage regulator.

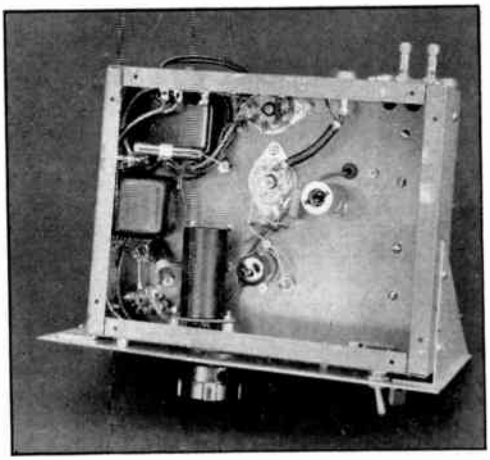


Fig. 12-4 — Bottom view of the 50-Mc. converter, showing the power-supply components. Note that the r.f. and mixer tubes are mounted in an inverted position below the chassis, permitting short leads to the r.f. components above.

maximum signal (or noise) near the middle of the band. Then connect the signal source to the antenna teminals and adjust C_1 in the same manner. If the two grid coils are the proper size the settings of C_1 and C_2 should come out the same. If they do not, the spacing of the turns in L_2 should be adjusted so that the setting of C_1 matches that of C_2.

If the physical arrangement of the converter components is different from that shown in the photographs, it may be necessary to add a small amount of oscillator injection for best performance. This may be done by connecting a short piece of insulated wire to the plate terminal of the oscillator, Pin 1, and running it over to the grid terminal of the mixer, Pin 7. Bend the wire near to the mixer terminal, and adjust its position to give the desired degree of injection. The wire may then be fastened in place with a drop of household cement. Oscillator injection may also be adjusted by using more capacity coupling than is needed, and then increasing the value of the oscillator plate dropping resistor until the desired performance is attained. The optimum degree of coupling is the largest value that can be used without resulting in a change in oscillator frequency when the mixer circuit is tuned.

Reducing Spurious Responses

In locations where there is broadcasting in the high FM band, 50-Mc. receivers and converters may experience severe interference resulting from these high-band signals beating with the second harmonic of the converter oscillator. The selectivity of the r.f. circuits is not sufficiently high at these frequencies to eliminate the unwanted signals, but the interference may be reduced by other means.

First, the output of the converter oscillator should be held to the minimum required to give satisfactory injection. In the case of the converter described above this may be accomplished by making the value of R_7 as high as possible, while still retaining satisfactory performance. This can best be checked by changing the resistor while listening to a very weak signal.

Second, if the above method does not cure the interference, a 100-Mc. trap may be inserted in series with the antenna pick-up coil, L_1. The trap may be made with an adjustable trimmer, or it may use a small fixed capacitor, in which case adjustment is accomplished by spreading or squeezing the turns.

If several interfering signals are present the trap should be adjusted on the strongest. Tune the signal in with the trap out of the circuit, then insert and tune the trap for maximum rejection of the interference. The sharpness of tuning of the trap (and its ability to reject interference) will depend on the L/C ratio. If only one strong signal must be eliminated, a high-C circuit should be used. If there are several, distributed over a considerable frequency range, more inductance and lower capacitance will provide a broader trap response, at the expense of some rejection at any one frequency.

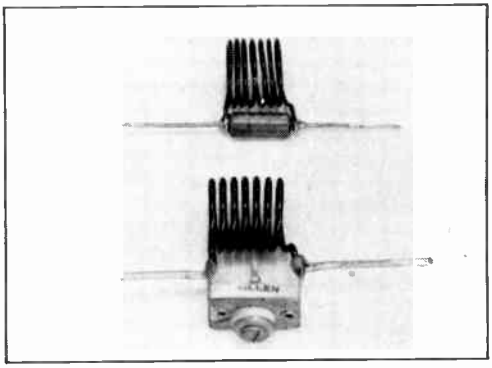


Fig. 12-5 — Examples of 100-Mc. traps used to reduce interference from high-band FM signals.

A usable average value is a coil of 7 turns of No. 14 to 18 wire, ½-inch diameter, spaced about the diameter of the wire. This may be mounted on a mica trimmer (3-30 $\mu\mu$fd.) or on a small 5-$\mu\mu$fd. ceramic condenser.

One-Tube Converter for 144 Mc.

A simple converter for 144 Mc. employing a single 7F8 tube is shown in Figs. 12-6–12-9. It is designed to work into a communications receiver on either 10.7 or 27.9 Mc., the latter frequency being provided so that the converter may be used with v.h.f. superheterodynes such as the Five-Ten, N1HU, S-27, S-36, and others which do not tune to the lower frequency. While it was designed for maximum simplicity, it is capable of outperforming the best super-regenerative receivers in weak-signal work. If greater sensitivity is desired, one or two stages of r.f. amplification (Figs. 12-10–12-12) may be added.

From the schematic diagram, Fig. 12-8, it may be seen that one section of the 7F8 dual triode is used as a mixer and the other as a Colpitts oscillator. Stability, an important factor in v.h.f.-converter design, is assured as the result of several precautions. The tuned circuit has a high C/L ratio, the coil is mounted rigidly on the tuning condenser, and the tube is mounted below the chassis to minimize heating effects. The Colpitts oscillator circuit permits grounding the cathode, preventing a.c. hum modulation, a common trouble when the cathode is operated above ground in v.h.f. circuits.

Mechanical Details

No attempt was made to gang the oscillator and mixer tuning controls, as the mixer setting is sufficiently broad so that it may be peaked at 146 Mc. and left in that position for the whole band. The oscillator tuning condenser, C_2, a split-stator variable, was made from a Millen type 21935, originally a 35-$\mu\mu$fd. single-section double-spaced midget variable. A section of the stator bars ¼ inch long is sawed out of the center of the condenser, leaving four stator and five rotor plates in each section. The three extra rotor plates, at the center of the rotor shaft, may be removed with long-nosed pliers. The condenser is mounted with the stator bars at the top, permitting the two-turn coil to be soldered directly to the sawed ends of the bars, for solid mounting.

If some other type of oscillator tuning condenser is substituted for the one specified above, care should be used to see that a physical arrangement similar to that shown at the left of Fig. 12-7 is achieved. There cannot be anything in the way of "leads" in a layout such as this, or the tuning range will be too low in frequency. The parallel padder, C_3, is an air trimmer of new design (Silver type 619), or a mica trimmer may be substituted, if necessary.

The vernier dial used is a National type K, but a large knob is substituted for the small one with which the dial is equipped, giving the converter tuning a communications-receiver quality. The appearance of the converter was further dressed up by giving the panel a "watch-case" finish. This is done with a small wad of steel wool in a drill press, or it may be done by hand with somewhat more effort.

The tube is mounted with its socket above the chassis, providing short r.f. leads. The arrangement of the smaller parts should be obvious from the photographs. Oscillator injection is controlled by the mica trimmer, C_4, which is mounted directly on the oscillator-plate and mixer-grid prongs of the tube socket. Its setting is not critical; it may be left near the minimum capacitance of the condenser.

The output coupling transformer is housed in a cut-down i.f. shield can, with the mixer plate lead coming out of a hole in the side. Winding data for both 10.7- and 27.9-Mc. transformers are given. The higher frequency is recommended for use wherever possible. The fixed padder, C_7, and the by-pass condenser,

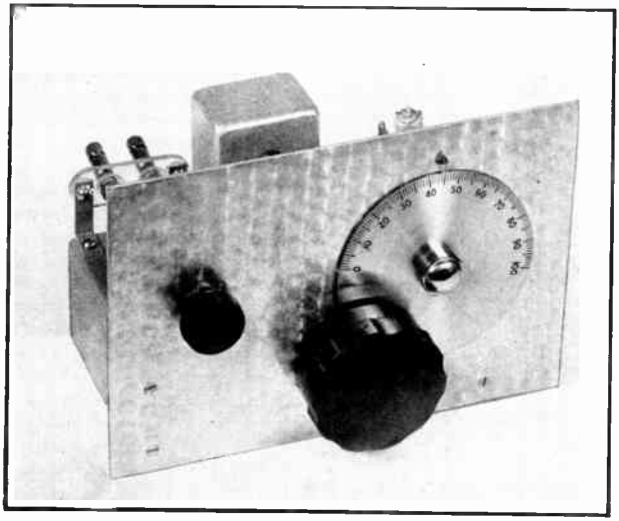


Fig. 12-6 — Front-panel view of the simple 2-meter converter.

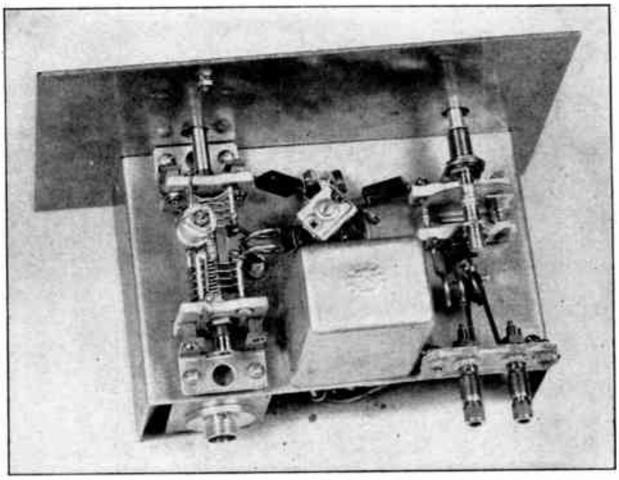


Fig. 12-7— Top view of the simple 2-meter converter. At the right are the mixer tuned circuit and antenna coupling coil. Oscillator components are at the left. The shield at the rear of the chassis houses the output coupling transformer. The trimmer attached to the 7F8 socket terminals is the oscillator injection condenser, C_4.

C_8, are mounted inside the i.f. shield. Converter output is taken off through a coaxial cable and fitting, though the latter may be eliminated if desired.

Ordinarily the receiver with which the converter is to be used will be capable of supplying the 6.3 volts a.c. at 0.3 amp. and 150 volts d.c. at 5 ma., but a separate supply may be used if desired. If the supply voltage is much over 150 volts, a dropping resistor should be used to bring it down to about that value.

The converter is built on a folded-aluminum chassis made from a sheet 4 by 10 inches in size, 2 inches being folded over on either end. The panel is 5 by 8 inches. The coaxial fitting and the antenna terminals are mounted on small aluminum brackets. The panel is fastened to the chassis by means of ¼-inch-square rods, but small angle brackets would serve equally well.

Adjustments

Adjustment and testing of the converter are simple enough, if a calibrated signal generator is available. Lacking this, harmonics of a VFO, or even the radiation from a receiver oscillator, may be used. A superregenerative receiver, the tuning range of which is known, may also be used as a signal generator. First, the i.f. output transformer should be tuned to 27.9 or 10.7 Mc. by means of its adjustable core. The exact frequency employed is, of course, unimportant, as the coaxial cable between the converter and receiver will prevent appreciable pick-up at the i.f. frequency. If any strong signals are present, the i.f. may be shifted to any clear frequency.

Next, the tuning range of the oscillator should be checked. When the converter is to be used with a 27.9-Mc. i.f., the tuning range of the oscillator will be 116.1 to 120.1 Mc. With a 10.7-Mc. i.f. it will be 133.3 to 137.3 Mc. Either of these ranges can be reached by adjustment of the parallel air trimmer, C_3. Bandspread, with the higher i.f., will be about 80 divisions. With the low i.f. it will be somewhat less. The oscillator may be checked with a calibrated absorption-type wavemeter, or by listening to it in a calibrated receiver.

A strong signal near 146 Mc. should then be tuned in, and the mixer condenser adjusted for maximum response. As the oscillator fre-

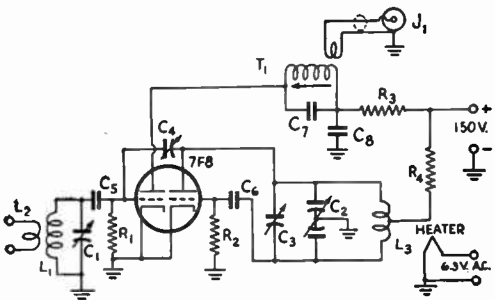


Fig. 12-8 — Schematic diagram of the 7F8 converter for 144 Mc.

C_1 — 10-μμfd. variable.
C_2 — 15-μμfd.-per-section split stator. See text.
C_3 — 3–30-μμfd. air trimmer (Silver type 619).
C_4 — 3–30-μμfd. mica trimmer.
C_5, C_6 — 47-μμfd. mica or ceramic.
C_7 — 27-μμfd. mica or ceramic.
C_8 — 470-μμfd. mica.
R_1 — 1 megohm, ½ watt.
R_2, R_3 — 10,000 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
L_1 — 3 turns No. 12, ⅜-inch i.d., ½ inch long.
L_2 — 2 turns "push-back," ⅜-inch i.d., inserted in cold end of L_1.
L_3 — 2 turns No. 12, ½-inch i.d., spaced ¼ inch, center-tapped.
J_1 — Coaxial jack (Jones S-201).
T_1 — 29.7-Mc. i.f.: 9 turns No. 22 d.s.c. wire, spaced one diameter, on National XR-50 form (slug-tuned). Coupling winding: 2 turns No. 22 d.s.c. wire, interwound in cold end of main winding.
— 10.7-Mc. i.f.: 22 turns No. 22 enameled wire, close-wound on National XR-50 form (slug-tuned). Coupling winding: 3 turns No. 22 enameled wire-wound over cold end of main winding. Insulate between two windings with polystyrene or other insulating tape.

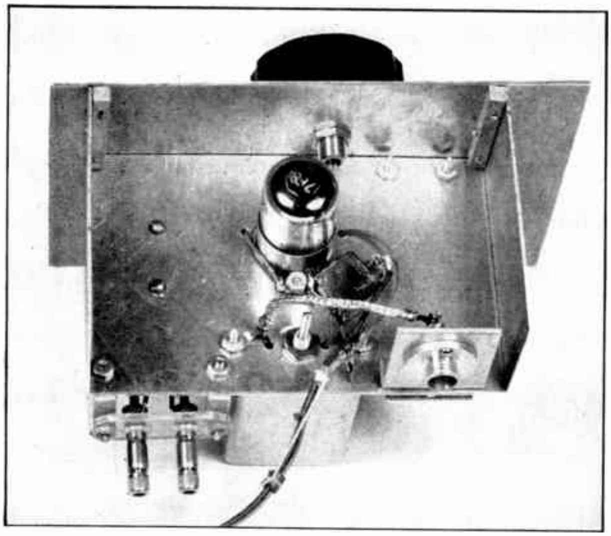


Fig. 12-9 — Bottom view of the converter shows the 7F8 tube mounted below the chassis. The i.f. core adjusting screw is in front of the tube. At the right, on a bracket, is the coaxial fitting for i.f. output.

quency varies when the mixer tuning is changed, it will be necessary to rock the oscillator dial back and forth across the signal as the mixer tuning is adjusted. Once the proper setting of C_1 has been determined, it may be left set for the entire band. When a sensitive receiver is used as an i.f. system quite good sensitivity will be obtained, and a signal of one microvolt or less will produce a plainly-audible response.

Choice of the intermediate frequency to be used will be determined by the tuning range and the performance characteristics of the receiver with which the converter is to be used. If the receiver does not include the 28-Mc. band some lower i.f. must be used. The performance of some receivers is so poor on 28 Mc. that much better results will be obtained with a lower i.f. With good-quality communications receivers, however, the higher i.f. is to be preferred. It makes for less pulling of the converter oscillator frequency when the mixer is tuned, and pick-up of interfering signals on the intermediate frequency is much less troublesome if 27.9 Mc. is used. The exact frequency is, of course, unimportant.

A 144-MC. GROUNDED-GRID R.F. AMPLIFIER

The two-stage r.f. amplifier unit shown in Figs. 12-10 through 12-12 may be used with the simple converter described above, or with any other 2-meter receiver where an improvement in sensitivity and image rejection is desired. It employs two 6J4 triodes in a grounded-grid circuit. Other tubes might also be used, but the 6J4 was designed especially for this service and has internal shielding which reduces the likelihood of self-oscillation troubles. In the grounded-grid circuit, the signal input is to the cathode of the tube instead of the grid, which is connected directly to ground, as the name implies.

In the unit shown, the antenna is coupled to a tuned circuit in the cathode of the first tube, the plate circuit of which is also tuned. Coupling between stages is effected by a small capacity, C_9, which is tapped down on the plate coil, L_3. The cathode of the second stage is returned to ground through an r.f. choke and a bias resistor, R_3. Both cathodes are maintained above ground for r.f. by insertion of r.f. chokes

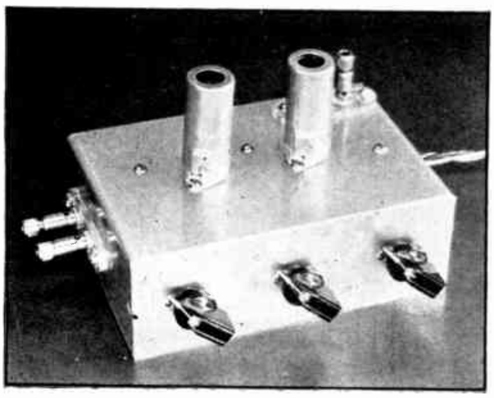


Fig. 12-10 — A two-stage grounded-grid r.f. amplifier for 144 Mc.

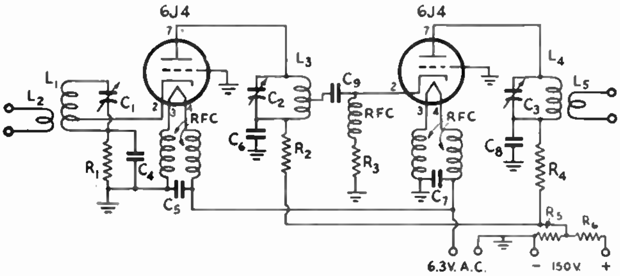


Fig. 12-11 — Schematic diagram of the 6J4 r.f. amplifier.

C_1, C_2, C_3 — 15-$\mu\mu$fd. midget variable (Millen 20015).
C_4, C_5, C_6, C_7, C_8 — 470-$\mu\mu$fd. midget mica.
C_9 — 47-$\mu\mu$fd. mica.
R_1, R_3 — 220 ohms, ½ watt.
R_2, R_4 — 470 ohms, ½ watt.
R_5, R_6 — 10,000 ohms, 10 watts.
L_1 — 4 turns No. 14, ⅜-inch inside diameter, ½ inch long, tapped 1½ turns from cold end.
L_2 — 2 turns "push-back," interwound in L_1.
L_3 — 3 turns No. 14, ⅜-inch inside diameter, ⅜ inch long, center-tapped.
L_4 — Same as L_3, but without tap.
L_5 — 2 turns "push-back," interwound in L_4.
RFC — No. 22 d.s.c. wire closewound on 1-watt carbon resistor. Winding length 1⅓₂ inch.

in the heater leads. The plate circuit of the second stage is link coupled to the mixer, the coupling line being brought out through two National FWG terminals at the back of the chassis. All three tuned circuits are provided with front-panel controls, for ease of adjustment, but only the plate tuning of the second stage will require any readjustment in tuning over the band. The other two controls may be set at 146 Mc. and no noticeable change in signal strength will be obtained if they are repeaked for a signal at either end.

The r.f. amplifier is mounted on a chassis similar to that used in the simple converter. It is 2 by 4 by 6 inches in size and was bent from a 4 × 10-inch piece of aluminum. The front panel is cut to fit, being approximately 2 by 6 inches in size. Interstage shields are sheets of copper, notched to fit closely over the centers of the 6J4 tube sockets, which are mounted in such a position that the cathode and plate terminals come on opposite sides of the shield. The three grid pins (1, 5, 6) are soldered directly to the shield itself, or grounded to soldering lugs under the screws with which the sockets are mounted. All leads should be as short as possible.

The amplifier unit may be operated from the same power supply as that used with the converter, but care should be taken to see that the plate voltage on the 6J4s does not exceed 150 volts. The performance of the unit does not change materially with a considerably-lower voltage, and it was found advisable to operate it from a bleeder tap (R_5R_6) as shown in the schematic diagram, when the amplifier was used in conjunction with the 7F8 converter and a 150-volt supply.

With the constants shown, the circuits will tune near minimum capacitance. Tuning the circuits to resonance produces a different result for each circuit. The noise output drops appreciably as the first circuit hits resonance; in the second there is only a slight noise change; while in the third the noise *increases* noticeably at resonance. Best results will be obtained if each circuit is adjusted while listening to a signal. If the receiver has an S-meter, the amplifier should be tuned for maximum reading on a medium-strength signal; if no S-meter is available, the adjustments should be made with the a.v.c. off, otherwise small changes in signal level will be difficult to detect. Once C_1 and C_2 have been set near the middle of the band they require no further adjustment, and C_3 may be repeaked for maximum noise.

Sharp tuning in any of these circuits indicates the presence of regeneration which will tend to nullify the desirable characteristics of the grounded-grid amplifier. Insufficient shielding, long ground leads, and too-light loading are possible causes of regeneration.

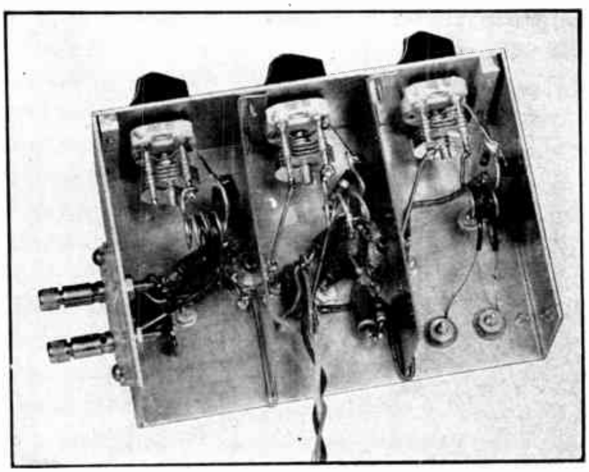


Fig. 12-12— Bottom view of the grounded-grid r.f. amplifier. At the left are the cathode input circuit and antenna coupling. The center compartment contains the plate circuit of the first stage and cathode circuit of the second. The right-hand compartment houses the plate circuit of the second stage, and the output coupling.

A Crystal-Controlled Converter for 144 Mc.

Stability in the oscillator is of utmost importance in achieving satisfactory performance in a 2-meter converter, particularly when a highly-selective communications receiver is to be used as the i.f. amplifier. Even a small amount of drift, or the slightest mechanical instability, will make constant readjustment of the converter tuning necessary. In addition it is difficult, if not impossible, to secure freedom from hum modulation of incoming signals, when ordinary tunable oscillators are used. The converter shown in Figs. 12-13, 12-14 and 12-15 eliminates these difficulties by the use of a crystal-controlled oscillator at 13 Mc., followed by two multiplier stages to bring the frequency up to that required for 144-Mc. reception. Tuning of the band is accomplished by varying the communications receiver (now the i.f. amplifier) from 14 to 18 Mc.

At first thought it would appear that the design of such a converter would be complicated and its construction difficult, but the photographs and schematic diagram show that this is not necessarily true. Since no tunable circuits are required at the signal frequency the mechanical construction is simplified, and the alignment problems usually associated with tracking of gang-tuned circuits are reduced. Only four tubes are required, two of them r.f. amplifiers.

Circuit Details

Two 6J6s perform the functions of oscillator, multiplier and mixer. Two 6AK5s are used as bandpass r.f. stages, self-resonant overcoupled plate and grid circuits being employed to achieve the bandpass characteristics. Referring to the schematic diagram, Fig. 12-14, it will be seen that the crystal oscillator is a simple triode circuit, using one section of a 6J6 and a 13-Mc. crystal. The second section of the tube is a doubler, which drives the first section of the second 6J6 as a quintupler to 130 Mc. Energy from this stage mixes with the signal in the second section of the tube, the output of which is at the intermediate frequency.

Two problems are presented by this approach. First, the r.f. and mixer circuits must be broadened out sufficiently so that the response of the converter will be substantially flat over the entire band; and second, signals at the intermediate frequency may cause considerable interference unless the unit is completely shielded. The needed bandpass characteristics in the r.f. stages are supplied by overcoupling the stages, adjustment of which is explained in a later paragraph. The i.f. output transformer in the mixer plate circuit is provided with an adjustable tuned circuit, which requires some readjustment if maximum sensitivity is to be maintained across the entire band. It is not critical in its setting, however, so it does not complicate the tuning process appreciably.

The only other adjustment used after the initial tune-up procedure is completed is an r.f. gain control, the setting of which may be employed to reduce possible cross-modulation from extremely-strong local signals. Normally it may be set at the optimum position and left there without further change.

Mechanical Construction

Structural details should be reasonably clear from the photographs. The chassis is a "U"-shaped affair folded from sheet aluminum. Another folded sheet is used as a shield, completely enclosing the components, all of which are mounted on the main chassis, which is 2 by 5 by 6 inches in size. Looking at the top view, Fig. 12-13, it may be seen that the gain control and i.f. tuning adjustment are mounted on the front wall of the chassis. Across the top are the crystal and the two 6J6s, and at the rear are the antenna input terminals, the two 6AK5s, and the coaxial fitting for the i.f. output cable. The power plug is between the two r.f. tubes, at the rear edge of the chassis. Looking at the bottom of the chassis, Fig. 12-15, the oscillator, multiplier and mixer components are in the upper portion of the photograph, with the r.f. stages across the bottom.

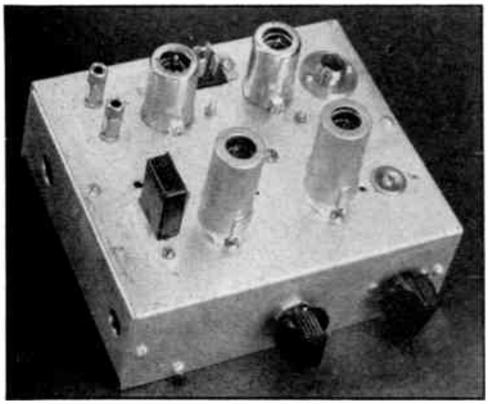


Fig. 12-13 — A crystal-controlled converter for 144 Mc. Two 6J6s serve as oscillator, multiplier and mixer. The two 6AK5s are bandpass r.f. stages.

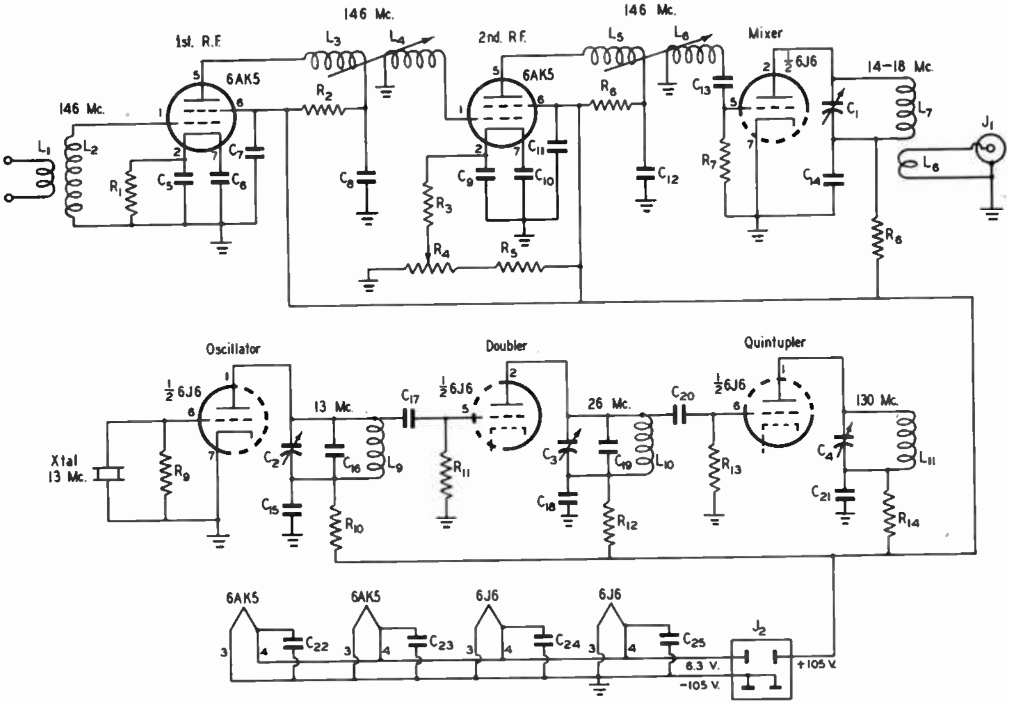


Fig. 12-14 — Wiring diagram of the 2-meter converter with crystal-controlled oscillator.

C_1 — 15-μμfd. variable (Millen 20015).
C_2, C_3, C_4 — 3–30-μμfd. mica trimmer.
C_5, C_6, C_7, C_9, C_{10}, C_{11}, C_{21}, C_{22}, C_{23}, C_{24}, C_{25} — 250-μμfd. ceramic.
C_8, C_{12} — 75-μμfd. ceramic.
C_{13} — 50-μμfd. ceramic.
C_{14}, C_{15} — 0.01-μfd. ceramic.
C_{16}, C_{17}, C_{20} — 100-μμfd. ceramic.
C_{18} — 0.0015-μfd. ceramic.
C_{19} — 27-μμfd. ceramic.
R_1, R_3 — 220 ohms, ½ watt.
R_2, R_6, R_8, R_{10}, R_{12}, R_{14} — 1000 ohms, ½ watt.
R_4 — 2000-ohm wire-wound potentiometer.
R_5 — 10,000 ohms, 1 watt.
R_7 — 8.2 megohms, ½ watt.
R_9 — 22,000 ohms, ½ watt.
R_{11}, R_{13} — 0.1 megohm, ½ watt.
L_1 — 2 turns No. 18 enameled wire, ⅜-inch inside diameter, tightly coupled to ground end of L_2.
L_2, L_4, L_6, L_{11} — 5 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_3 — 6 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_5 — 7 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_7 — 27 turns No 28 enameled wire, close-wound to a length of ⅜ inch on a ½-inch diameter form.
L_8 — 3 turns "push-back" wire, close-wound over cold end of L_7.
L_9 — 8 turns No. 18 tinned wire, ¾-inch diameter, ½ inch long.
L_{10} — 6 turns No. 18 tinned wire, ¾-inch diameter, ⅜ inch long.
Note: Coils L_9 and L_{10} are cut from a length of B & W "Miniductor" type 3011.
J_1 — Coaxial-cable connector (Jones S-201).
J_2 — 4-prong male plug (Jones P-304-AB).

The shield arrangement which isolates the r.f. stages may also be seen in the bottom view. It is made of sheet copper, to permit easy soldering to the cylindrical shields on the tube sockets. At the left is the grid coil, L_2, with the antenna winding, L_1, inserted between its first two turns. Inside the shield are the first r.f. plate (L_3) and second r.f. grid (L_4) coils mounted close together in the same plane. Just outside the upper right corner of the shield are the second r.f. plate coil, L_5, and the mixer grid coil, L_6.

At the upper left of the bottom view are the oscillator and doubler plate coils, L_9 and L_{10}, with their associated trimmers. The quintupler plate coil and its trimmer are between the gain control and the i.f. output tuning condenser. The i.f. output transformer, L_7L_8, is at the upper right. Because of their small size and greater effectiveness, ceramic capacitors were used throughout the unit, the particular values used now being available from Centralab and possibly others.

Adjustment and Operation

The first step in putting the converter into service is to set up the oscillator and multiplier stages for proper operation. This procedure is similar to that employed in multistage transmitters, except that at 105 volts the currents in the various stages are very low. It is recommended that the supply voltage be maintained at that figure with a VR-105, in which case the total drain for the two 6J6s is about 12 ma. The two 6AK5s draw approximately the same. A low-range milliammeter may be inserted in series with the plate decoupling resistors, R_{10}, R_{12} and R_{14}, to check on the operation of the oscillator and multiplier stages, as a good plate-current dip is noted as the 6J6 stages are tuned to resonance. A calibrated absorption-type wavemeter should be used to be certain that the doubler and quintupler stages are operating on the correct frequencies.

The r.f. circuits may be aligned by means of a grid-dip meter. Details of a suitable instrument, which is a useful tool for many other purposes, will be found in Chapter Sixteen. The inductance of the grid and plate coils may be adjusted to approximately the proper value by spreading or squeezing the turns to the point which produces maximum dip with the grid-dip meter set for 146 Mc.

The mixer plate circuit may be checked by feeding a signal into the mixer grid at the i.f. frequency, 14 to 18 Mc., making sure that the tuned circuit, C_1L_7, is capable of resonating across this range. In the absence of a signal generator, 20-meter amateur signals may be used for this check by connecting an antenna to the mixer grid. They should peak with the mixer plate condenser set near the middle of its tuning range.

A calibrated signal generator operating in the 144-Mc. range is helpful in checking the converter performance, but a low-powered oscillator, or even a superregenerative receiver, may be used. A fair idea of the performance can be obtained by merely connecting an antenna to the converter and aligning the r.f. circuits by noise pick-up. If the antenna in question is a simple dipole cut for 146 Mc., the noise level should remain nearly constant over the entire band, if the r.f. coils and the coupling between them has been adjusted properly. If necessary, the self-resonant coils may be "stagger-tuned" to achieve uniform response across the band.

Other Suggested Circuits

One of the problems in connection with the use of crystal-controlled oscillators in v.h.f. converters is the choice of a suitable crystal frequency. If a relatively low frequency is used in the crystal oscillator the crystal must be chosen carefully to avoid trouble from its harmonics falling in the band to be covered. If higher crystal frequencies are used the cost of the crystal becomes considerable, and some harmonic-type crystals have poor stability.

A crystal oscillator circuit which helps to by-pass these troubles is described in the transmitter chapter in connection with the v.h.f. exciter-transmitter pictured in Fig. 13-11. With this circuit, Fig. 13-12, ordinary 7–8-Mc. crystals are made to oscillate on their third harmonic, thus reducing the number of stages required, and permitting the use of inexpensive crystals.

This circuit may be substituted for that in Fig. 12-14, in case a less expensive or more readily-obtainable crystal is to be used. An example would be the use of a 7.22-Mc. crystal, oscillating at 21.66 Mc. in the first 6J6 triode section of Fig. 12-14. The second section would double to 43.32 Mc. The next triode section would operate as a tripler to 130 Mc. Except for the grid circuit of the first 6J6 the schematic would be similar to that shown in Fig. 12-14. The regenerative harmonic oscillator circuit is reproduced in Fig. 12-16. It is suggested that prospective users study the material on page 415, Chapter Thirteen, for further information before attempting to utilize this circuit for receiver purposes.

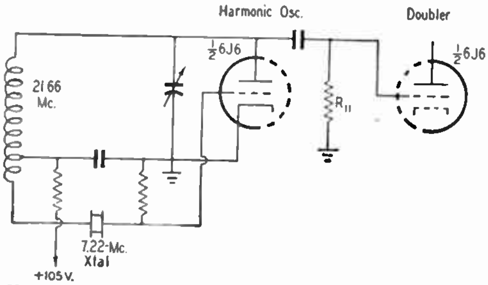


Fig. 12-16 — Schematic diagram of a regenerative harmonic oscillator circuit, which may be substituted for the 13-Mc. oscillator shown in Fig. 12-14. The second section then doubles to 43.32 Mc., and the second 6J6 operates as a tripler to 130 Mc., instead of a quintupler.

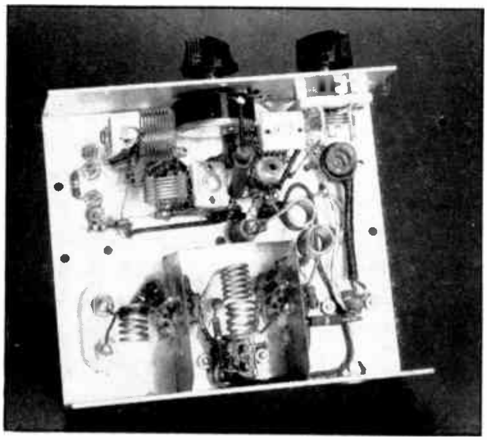


Fig. 12-15 — Bottom view of the crystal-controlled converter. Note the overcoupled circuits in the two r.f. stages.

Receiving Equipment for 220 Mc.

As much of the operation on 220 Mc. is with crystal-controlled transmitters, it will be possible for most workers on this band to use receiver techniques similar to those employed for 144 Mc. A converter for 220 Mc. need not differ greatly from one for 144, but the problem of oscillator instability is somewhat more troublesome. Even though stable signals are to be received, instability in the converter oscillator may make tuning difficult, if high selectivity is employed in the receiver i.f. system. For this reason it may be helpful to use the 220-Mc. converter with a communications receiver which has something higher than the customary 455-kc. i.f. The crystal-controlled converter is the most satisfactory solution to the stability problem.

● A SIMPLE 220-MC. CONVERTER

The one-tube converter shown in Figs. 12-17–12-20 is designed so that it may be used with any receiver which is capable of being tuned just outside the low end of the 10-meter band. This i.f. was selected to permit use of the converter with receivers such as the S-27, S-36 and SX-42, all of which are capable of broadband reception at this frequency. Receivers having 1600-kc. i.f. channels (NHU, Five-Ten, and others) will also provide somewhat greater ease of tuning than those having 455-kc. i.f. systems, though the latter may be used successfully, of course.

While the sensitivity of such a simple unit is not to be compared with more complicated converters employing r.f. stages, its performance is appreciably better than is obtainable with the superregenerative type of receiver. If improved sensitivity is desired an r.f. stage should be added. Choice of tubes for r.f. amplifier service at 220 Mc. is somewhat limited, but worth-while gain can be achieved with acorn-type pentodes, and with certain triodes, such as the 6J4 and 6J6.

The converter is built on the smallest-size utility box, with the bottom plate removed and used for a front panel. The 12AT7 dual triode is mounted in an inverted position, with its socket above the chassis, permitting short leads in the v.h.f. circuits, a prime necessity at these frequencies.

The circuit is similar to that of the 144-Mc. converter using the 7F8, shown in Figs. 12-6–12-9. One section of the tube is used as a mixer, its input circuit being self-resonant at

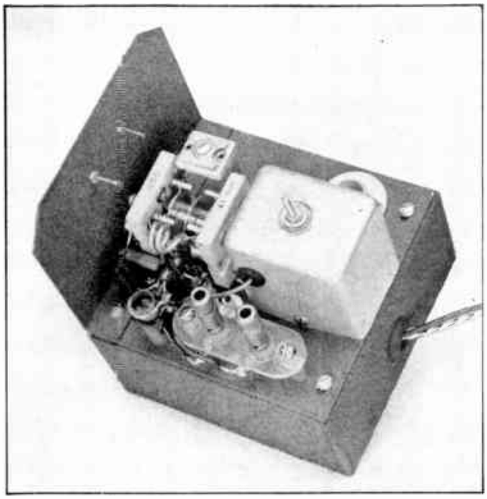


Fig. 12-18 — Top view of the 220-Mc. converter. At the left is the self-resonant mixer input circuit. The i.f. output transformer is at the right, with the coaxial output fitting just visible over the shield can.

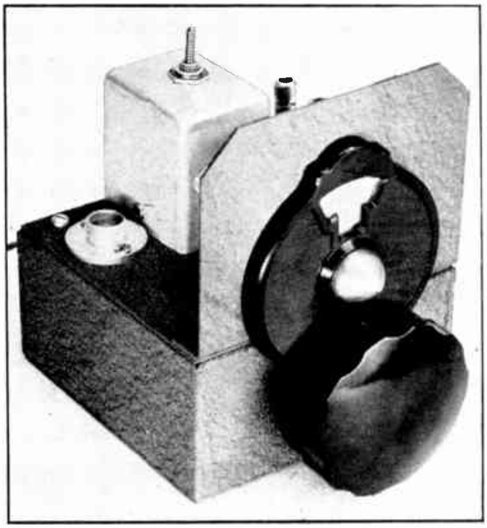


Fig. 12-17 — A one-tube converter for 220 Mc. The large knob was substituted for the customary small one to facilitate slow tuning.

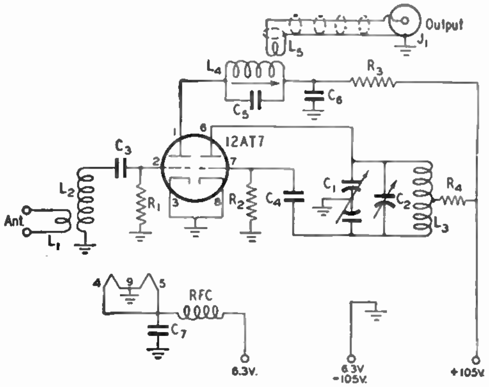


Fig. 12-19 — Schematic diagram of the one-tube 220-Mc. converter.

- C_1 — 15-μμfd.-per-section split stator (Bud LC-1660 reduced to 1 stator and 1 rotor plate per section).
- C_2 — 3–30-μμfd. mica trimmer (National M-30).
- C_3, C_4 — 22-μμfd. mica.
- C_5 — 27-μμfd. mica.
- C_6, C_7 — 470-μμfd. mica.
- R_1 — 1 megohm, ½ watt.
- R_2 — 10,000 ohms, ½ watt.
- R_3, R_4 — 1000 ohms, ½ watt.
- L_1 — 1 turn insulated wire, ⅜-inch diam., wound at cold end of L_2.
- L_2 — 1 turn No. 18 e. wire, ⅜-inch diam.
- L_3 — 2 turns No. 12 tinned wire, ¼-inch diam., turns spaced wire diam.
- L_4 — 9 turns No. 18 e. wire, ½-inch diam., ⅝ inch long.
- L_5 — 2 turns insulated wire, close-wound over bottom end of L_4. L_1 and L_2 are wound on a Cambridge Thermionic Corp. type L-3S slug-tuned coil form, ⅜-inch diameter. L_4 and L_5 are wound on a National Type XR-50 slug-tuned form.
- J_1 — Coaxial-cable connector (Jones S-201).
- RFC — 12 turns No. 18 wire, ¼-inch diameter, 1 inch long.

the signal frequency. No tuning is required, as the response is substantially flat across the entire band. The other triode section is the oscillator. The tuning condenser for this circuit, C_1, is controlled by the vernier dial. To aid in tuning a large knob is used, replacing the smaller knob normally supplied with the dial.

The general arrangement of parts should be clear from the photographs. Putting the converter into operation is similar to the process outlined for the 7F8 converter for 144 Mc., except that the mixer input circuit must be adjusted to resonance by squeezing or spreading the turns of the coil. This should be done on a signal near the middle of the band.

Improving 220-Mc. Receiver Performance

The considerable increase in working range of 144-Mc. stations which resulted when there was a general change to superheterodyne receivers for that band points the way to making 220 Mc. pay off. Though employment of advanced techniques becomes somewhat more difficult as we move to 220 Mc., it is only by the use of the best possible equipment that we will be able to appreciate fully the possibilities of the higher band. This means not only the use of converters but the employment of r.f. amplifier stages as well. All this is not as difficult as might be supposed, as the relatively narrow band of frequencies to be covered, and the low selectivity of tuned circuits at these frequencies, permit the use of stages which require little or no retuning in covering the band.

Three general types of r.f. stages may be used successfully at 220 Mc. Among the few r.f. pentodes which could be expected to provide appreciable gain in conventional circuits are the acorn types, 954 and 956. The r.f. amplifier shown in connection with the coaxial-line superregenerative receiver in the following pages could be adapted readily to use with a converter. Grounded-grid r.f. stages may provide worth-while gain in suitable circuits, though two stages would be required to achieve the desired result. The 6J4 grounded-grid amplifier of Fig. 12-11 may be adapted to 220-Mc. service. The use of push-pull r.f. stages has merit, and dual triodes such as the 6J6 are particularly well adapted to this type of circuit. Where triodes are employed in such a circuit they must be neutralized in a manner similar to that employed in triode-amplifier stages for transmitting purposes. This presents no great problem, for neutralization, at the voltages used in receivers, may be done with small trimmers or with short lengths of Twin-Lead.

Push-pull circuits may, in fact, be employed to good advantage in all stages of a 220-Mc. converter. An excellent example of the use of push-pull r.f. and mixer circuits at 220 Mc. was described by W1PMS and W1CTW in *QST* for October, 1948, page 31.

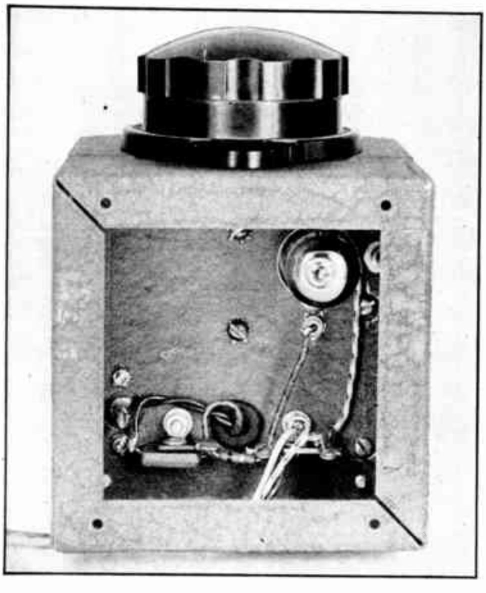


Fig. 12-20 — Bottom view of the 220-Mc. converter, showing the inverted 12AT7 tube.

The Superregenerative Receiver

The simplest type of v.h.f. receiver is the superregenerator, for many years the most popular receiver for v.h.f. work. It affords fair sensitivity with few tubes and elementary circuits, and though it has largely been replaced by the more effective superheterodyne for home-station use, it still has many v.h.f. applications. Its disadvantages are lack of selectivity, poor signal-to-noise ratio on weak signals, and its tendency to radiate a strong signal which causes severe interference.

Its selectivity may be improved somewhat and its interference capabilities reduced by the addition of an r.f. stage, a refinement which should be considered a necessity if the receiver is to be used in a locality where there are other stations operating on the same band. If no r.f. stage is used, as in portable applications where economy of space and battery drain are primary considerations, the detector should be operated with the lowest plate voltage that will permit superregeneration, in order to reduce its interference range.

From a practical aspect, superregenerative receivers may be divided into two general types. In the first the quenching voltage is developed by the detector itself, called a "self-quenched" detector. In the second, a separate low-frequency oscillator is used to generate the quench voltage. Self-quenched detectors have found wide favor, particularly for portable work; but it is possible to achieve better performance with the separately-quenched type, particularly as the frequency approaches the upper limit of the tube's capabilities.

Superregeneration Principles

The limit to which ordinary regenerative amplification can be carried is the point at which oscillation commences, 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, the value being between 20 and 200 kc. depending on the signal frequency. Because the oscillations are constantly being interrupted by this quenching voltage the regeneration can be greatly increased, and the amplified signal will build up to tremendous proportions. A one-tube superregenerative receiver 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 separately-quenched and self-quenched detectors are shown in Fig. 12-21, but the basic circuit may be any of the various arrangements used for straight regenerative detectors.

In the self-quenched 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.

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.

Because of the greater amplification, the hiss noise when a super-

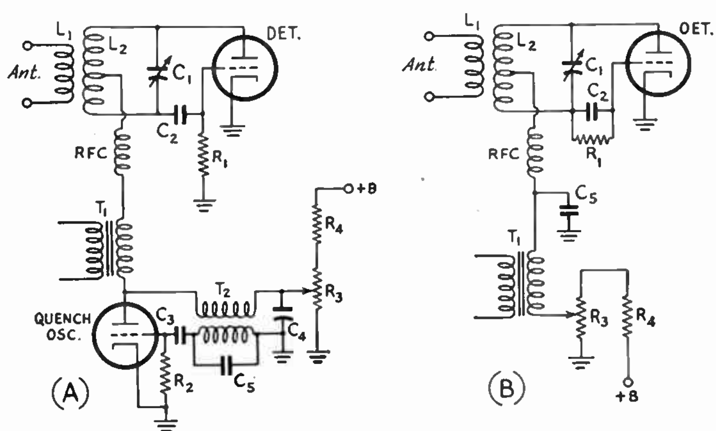


Fig. 12-21 — (A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:

C_2 — 47 μμfd.
C_3 — 470 μμfd.
C_4 — 0.1 μfd.
C_5 — 0.001–0.005 μfd.
R_1 — 2 — 10 megohms.
R_2 — 47,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 47,000 ohms.
RFC — R.f. choke, value depending upon frequency. Small low-capacitance chokes are required for v.h.f. operation.
T_1 — Audio transformer, plate-to-grid type.
T_2 — Quench-oscillator transformer.

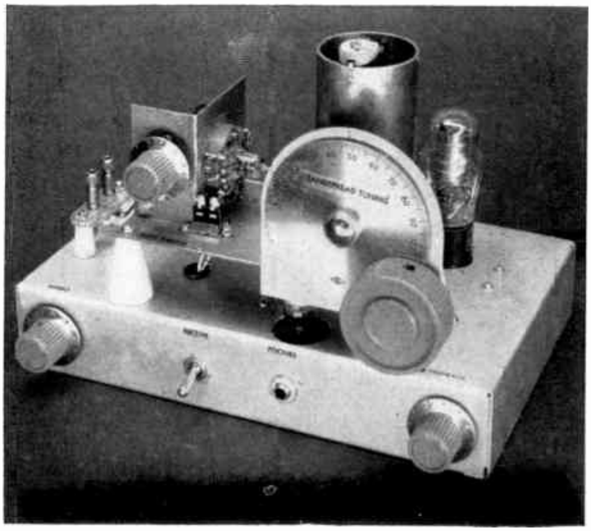


Fig. 12-22 — Front view of the coaxial-line receiver. The r.f. amplifier tuning control is at the left and the main control, for the concentric-line detector circuit, is at the right side of the unit. The audio gain control, send-receive switch, 'phone jack and regeneration control can be seen in that order, from left to right, across the front wall of the chassis.

regenerative 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 that 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 that 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.

A COAXIAL-LINE SUPERREGENERATIVE RECEIVER FOR 220 MC.

The performance of a superregenerative receiver, both as to selectivity and smoothness of operation, can be improved by the use of a coaxial line tank in the grid circuit of the detector, in place of the customary coil and condenser. Addition of an r.f. amplifier stage will improve sensitivity, reduce radiation, and make antenna coupling less critical. A superregenerative receiver for 220 to 240 Mc. incorporating these features is shown in Figs. 12-22–12-25.

The r.f. tube is a 954 acorn with a conventional tuned circuit in its grid. The plate circuit is a self-resonant loop, which is coupled to the concentric line grid circuit of the 6AK5 detector. The detector output is fed through a quench filter to a

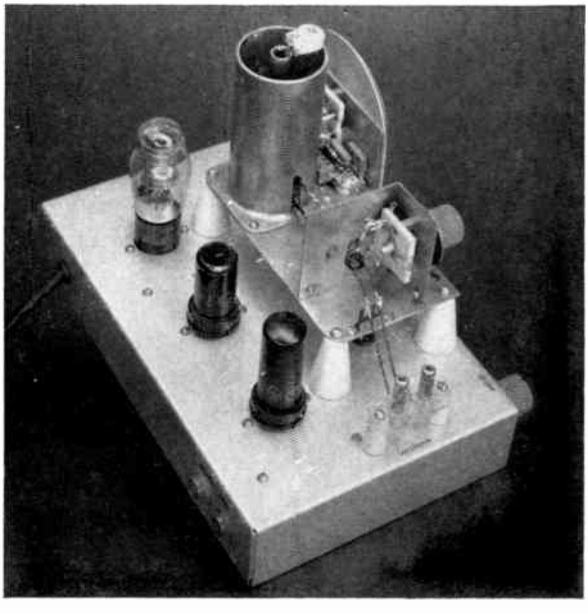


Fig. 12-23 — Rear view of the superregenerative receiver. The r.f. circuits are mounted on a copper shelf to the left of the antenna terminals. The detector tuning condenser is mounted on a small panel to the front of the coaxial line, and the band-set condenser is soldered across the open end of the line. The r.f. stage is mounted on an "L"-shaped bracket with the tube socket and plate-circuit components on the left side and the grid circuit on the right side. Audio tubes and voltage regulator are in line across the rear of the chassis.

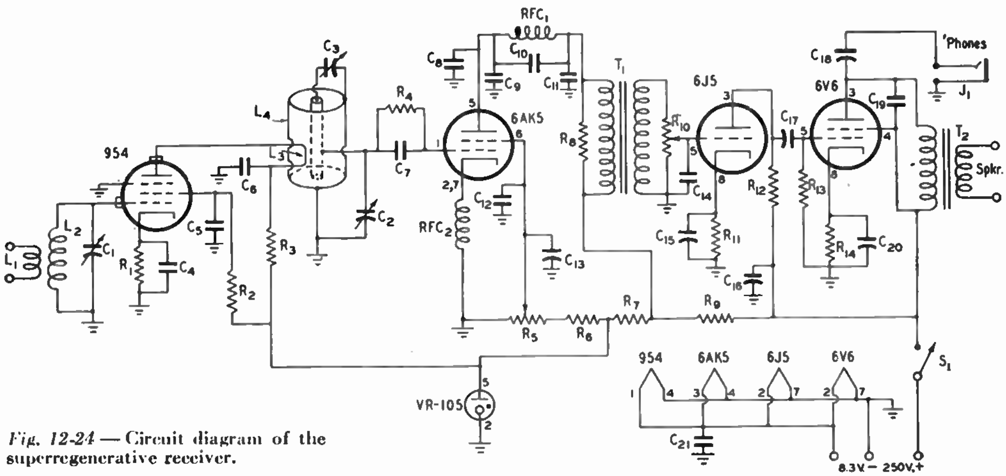


Fig. 12-24 — Circuit diagram of the superregenerative receiver.

C_1 — Midget variable condenser (Millen 20015 reduced to one stator and two rotor plates).
C_2 — Midget variable condenser (Millen 20015 reduced to one stator and one rotor plate).
C_3 — 5–20-$\mu\mu$fd. ceramic trimmer (Centralab 820-B).
C_4, C_5, C_6 — 100-$\mu\mu$fd. (National XLA-C).
C_7 — 22-$\mu\mu$fd. mica.
C_8, C_{12}, C_{21} — 470-$\mu\mu$fd. mica.
C_9 — 0.0022-μfd. mica.
C_{10}, C_{11} — 0.0068-μfd. mica.
C_{13} — 0.2-μfd. 400-volt paper.
C_{14} — 47-$\mu\mu$fd. mica.
C_{15} — 10-μfd. 25-volt electrolytic.
C_{16} — 8-μfd. 450-volt electrolytic.
C_{17}, C_{18} — 0.01-μfd. 400-volt paper.
C_{19} — 0.0047-μfd. mica.
C_{20} — 100-μfd. 25-volt electrolytic.
R_1, R_3 — 1000 ohms, ½ watt.
R_2 — 33,000 ohms, ½ watt.
R_4 — 0.1 megohm, ½ watt.
R_5 — 50,000-ohm potentiometer.
R_6 — 47,000 ohms, 1 watt.
R_7, R_9 — 1500 ohms, 10 watts.
R_8 — 22,000 ohms, ½ watt.
R_{10} — 0.25-megohm potentiometer.
R_{11} — 2200 ohms, ½ watt.
R_{12} — 0.1 megohm, ½ watt.
R_{13} — 0.47 megohm, ½ watt.
R_{14} — 270 ohms, 1 watt.
L_1 — 2 turns No. 18 e., ¼-inch inside diameter, close-wound.
L_2 — 2 turns No. 12 e., ¼-inch inside diameter, ⅛-inch space between turns.
L_3 — A 5¼-inch length No. 12 e., bent to form a U-shaped loop having a ¾-inch space between conductors. Plate side of loop is 1¾ inches long and the opposite side is 2¾ inches long.
L_4 — Concentric line. Inside conductor is a 4-inch length of ½-inch o.d. copper tubing. Grid tap 1 inch from grounded end for both 220- and 235-Mc. operation or ¾ inch from grounded end for 220 Mc. only. Outside conductor is a 4-inch length of 2-inch i.d. copper tubing.
J_1 — Open-circuit jack.
RFC_1 — 80-mh. choke (Meissner 19-5596).
RFC_2 — 1-mh. r.f. choke (National R-33).
S_1 — S.p.s.t. toggle switch.
T_1 — Interstage audio transformer (Stancor A-53C).
T_2 — Universal output transformer (Cinaudagraph U-85).

6J5 triode audio followed by a 6V6 second audio. Either 'phones or speaker may be used.

Constructional Details

The receiver is built on a standard aluminum chassis measuring 2 by 7 by 11 inches and the small panel for the detector tuning dial is cut from a sheet of $\frac{1}{16}$-inch aluminum measuring 3⅞ × 3⅞ inches. The shelf for the r.f. section is made from a piece of $\frac{1}{16}$-inch copper stock measuring 5½ × 6¼ inches which is cut and bent as shown in the photographs of the receiver. The horizontal section of the subchassis measures 3½ × 6¼ inches and the small vertical panel is 2 inches high and 2½ inches wide. The detector bandspread condenser and the aluminum panel for the detector tuning dial are both mounted on this upright member of the copper chassis. C_2 is mounted with the two stator terminals facing toward the right end of the chassis (as seen from the rear view) and the lower stator terminal is one inch up from the horizontal surface and 1¼ inches in from the left side of the copper panel. The tube socket for the 6AK5 is 2 inches in from the left end of the chassis and is located as far toward the front edge as possible.

The "L"-shaped bracket for the r.f. amplifier is 2½ inches high, has a depth of 2⅜ inches, and is 1½ inches across the front. Spade lugs are bolted, and then soldered, to the bottom of the partition to provide a method of mounting that is both electrically and mechanically sound. The National XLA tube socket is centered on the side of the partition at a point located 1⅜ inches in from the rear and top edges. A $\frac{5}{16}$-inch hole, drilled in the bracket at this point, allows the grid prong of the 954 to extend through to the grid-circuit components. The cathode and heater prongs of the socket face toward the front of the receiver and the XLA-C by-pass condensers are mounted inside the socket. The plate by-pass condenser, C_6, is mounted underneath socket prong No. 5 as this prong is used as the support point for the cold end of the plate loop, L_3. Note that the No. 5 prong is a spare so far as the 954 is concerned. A National XLA-S internal shield, designed for use with the XLA socket, provides a common path for the condenser ground connections and, of course, this soldering should be done before the socket is bolted to the copper partition. The heater, cathode and suppressor connections are also made to the internal shield and, after mounting, the shield is in turn soldered to the copper plate.

The r.f. amplifier tuning condenser is mounted with the shaft in line with the shaft of C_2. Stator terminals face to the left so that the bottom terminal is within ¼ inch of the 954 grid prong. L_2 is supported by the condenser terminals and the antenna coil, L_1, is supported by L_2 and by the two-terminal lug strip located to the right of the amplifier. Grid clips for the 954 were improvised by removing the prongs from a miniature tube socket.

Holes, large enough to clear 8⁄32 machine screws, are drilled at each corner of the copper mounting plate so that the unit may be mounted on 1½-inch stand-off insulators. Larger holes, equipped with rubber grommets, are adjacent to the detector and amplifier tube sockets so that power wiring may be passed down through the main chassis.

Construction of the concentric line is not difficult if the various operations are carried out as suggested below. The inner and outer conductors are 4 inches long, and the end plate is 2½ inches square. A ½-inch hole for the inner conductor of the line should be drilled at the center of the end plate, and the plate should also have a hole for a 6⁄32 machine screw at each corner. However, before the center hole is drilled, it is advisable to use the center-punch mark as the pivot for scribing a circle to indicate the position of the outside conductor. This will simplify the task of lining up the two pipes for the soldering operation.

A ⅜-inch hole should now be drilled in the large pipe at a point located 1 inch up from the bottom edge, and a second hole of 5⁄16-inch diameter should be drilled on line with the larger hole and around the pipe by 90 degrees. After the material between these two holes and the bottom of the tubing is removed by cutting with a hack saw, the finished slots will provide openings for the input coupling coil, L_3, and the detector-grid connection. The inner conductor should also be drilled and tapped for a 6⁄32 machine screw at this time. One hole, ¾ inch up from the bottom of the line, is required if the receiver is to be used to cover only one band. A second hole, ¼ inch above the first, is necessary if the receiver is to be tuned to both the 220- and 235-Mc.* bands. In either case, the tapped hole will be used as the connecting point for the lead running to the tuning condenser.

Unless extremely thin-walled tubing is used for the concentric line, it will be difficult to complete the soldering operation with an ordinary iron. Placing the assembly on an electric hot plate will heat the copper in a very few minutes and will allow the work to be done neatly and easily. The end plate should be laid on a flat level surface while the inner conductor is lined up perpendicular to the horizontal surface of the plate. This operation may be carried out with the metal resting on the hot plate if the latter is to be used. The outer conductor should be placed in the position indicated by the scribed circle. Heat may now be applied and the soldering completed. The metal is ready to accept solder when a rapid change in the color of the copper is noticed. A long piece of solder may be inserted through the open end of the line, and as the end is moved around the surfaces to be joined the solder will melt and run into place easily.

The remaining constructional work is straightforward and study of the three photographs will show the location of the various components. Since there is no crowding of parts, it should not be difficult to duplicate the original layout.

* The 235-Mc. band is still assigned in Canada at this writing.

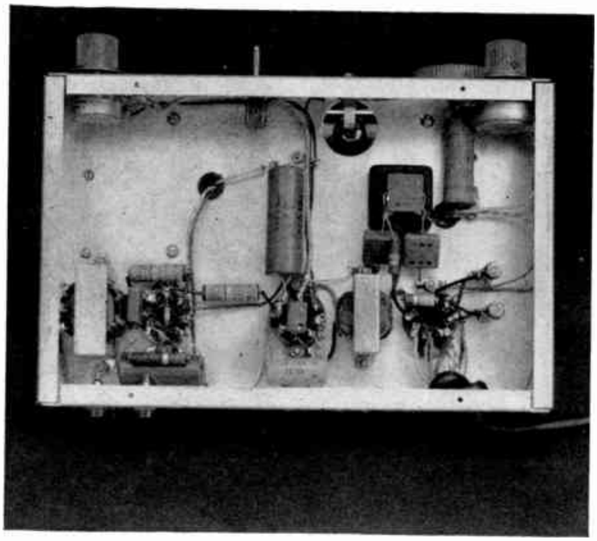


Fig. 12-25 — Bottom view of the coaxial-line receiver showing the output transformer, T_2, located at the lower left-hand corner of the chassis, and the audio transformer, T_1, mounted between the sockets for the audio tubes. The quench-filter choke, housed in a metal shield, is above and to the right of T_1. Resistors R_7 and R_9 are mounted on end to the right of the regulator-tube socket.

A Mobile Converter for 6, 10 and 11 Meters

The converter shown in Figs. 12-26–12-30 is designed for use with a mobile broadcast receiver. Two sets of plug-in coils cover 6, 10 and 11 meters, the set not in use being plugged into a pair of dummy sockets in the base of the converter. Power for the unit is obtained from the car-receiver power supply, and a switching arrangement allows a standard car antenna to be used for either broadcast or amateur reception. The performance of the converter may be somewhat below that of more advanced types employing r.f. stages, but it is adequate for mobile operation, where high noise level and the low power of the transmitter are limiting factors.

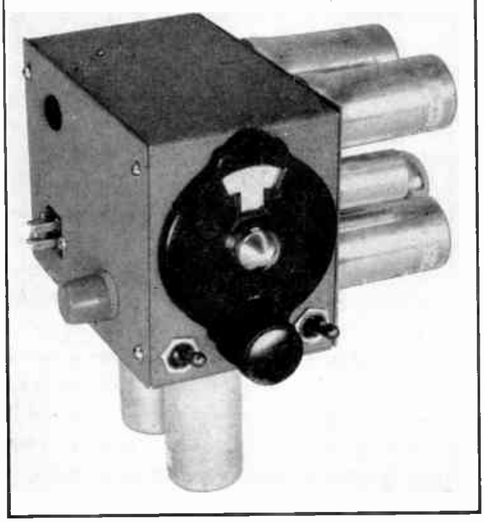


Fig. 12-26 — Front view of the mobile converter. The heater- and plate-voltage switches are to the left and right of the vernier-dial control knob. The pilot light and input connector are at the left side of the case. A hole for screwdriver adjustment of the output transformer is located above the input plug. Tube sockets, mounted in the bottom of the case, are used as holders for the extra set of coils.

Circuit Details

As may be seen from the diagram, Fig. 12-30, the converter uses a single 6BE6 pentagrid converter tube, employing electronic injection. A Colpitts oscillator is used, permitting the rotor of the tuning condenser to be grounded and doing away with the need for a cathode tap or tickler coil. The mixer section also uses a grounded-rotor condenser. The output transformer, C_5, L_4 and L_5, is made as a plug-in unit. Switch sections S_{1A} and S_{1B} transfer the antenna from the receiver input circuit to the converter, and, at the same time, connect the transformer output winding, L_5, to the receiver input. Toggle switches S_2 and S_3 are used for heater and stand-by purposes. A VR-105 regulates the oscillator plate voltage, preventing oscillator-frequency fluctuation which would otherwise result from the voltage variation usually encountered in mobile equipment.

Physical Layout

A side view of the converter, Fig. 12-27, shows most of the components mounted on one of the detachable plates of a 3 × 4 × 5-inch utility box. From top to bottom on the left side of the cover plate are the oscillator coil, the 6BE6 tube, and the mixer coil. The output transformer, the antenna switch, and the regulator tube are next in line, with the output jack, J_2, just to the right of the output transformer, and the input jack, J_1, below J_2.

The inside view of the unit, Fig. 12-29, shows the parts closely grouped around the tuning condenser, C_{1A}, C_{1B}, the mounting of which requires care in order that the control shaft can be made to line up with the vernier dial which is mounted on the small surface of the case. It is suggested that the layout drawing, Fig. 12-28, be followed as closely as possible. The condenser, mounted after other components have been fastened in place, is equipped with the brackets that are supplied with the Trim-Aire type of condenser. These brackets are attached to the outer surfaces of the ceramic end plates, with the mounting lips facing toward the back of the condenser. Be-

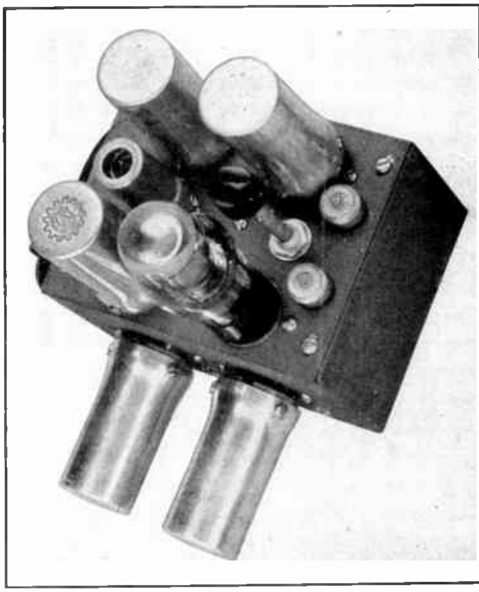


Fig. 12-27 — A side view of the mobile converter. The long shaft on the antenna change-over switch allows the switch to be operated without reaching down in among the other components.

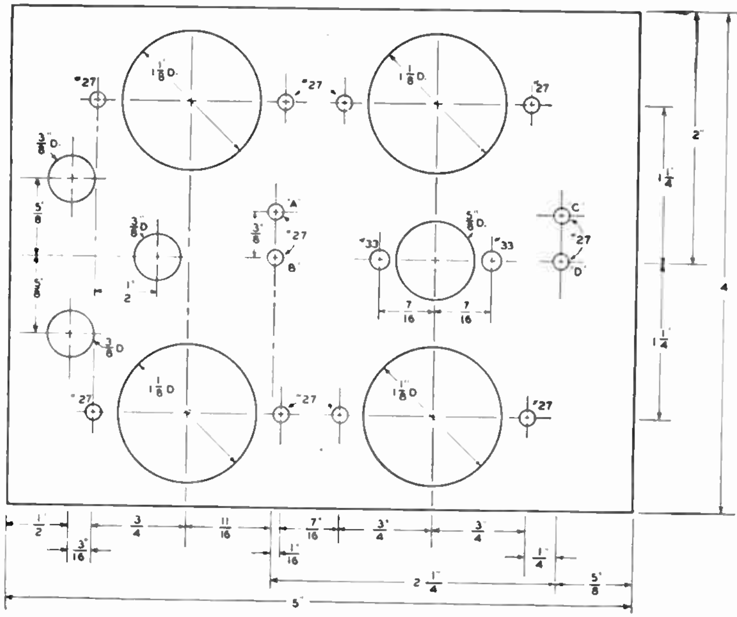


Fig. 12-28 — A mechanical drawing of the subchassis layout.

tween the brackets and the panel are metal pillars $1\frac{1}{16}$ inch long. The position of the condenser is such that the stator terminals are just above the tube- and coil-socket prongs.

It should be possible to follow the layout shown, as there is no excessive crowding of parts. Looking at Fig. 12-29, the oscillator components are at the left of the tuning condenser, with the mixer parts at the right. The output transformer is at the lower left and the regulator tube socket at the right, with the antenna switch in between. The band-set condensers are soldered directly to the coil-socket prongs. All four sockets have extra prongs, which may be used as tie points for other components.

The main part of the utility box needs some modification before the dial, switches and spare-coil sockets are mounted. It is necessary to fold back the two ½-inch flanges which are at the top and bottom edges of the right side of the case, and the flange at the rear must be filed out in two places to provide clearance for jacks J_1 and J_2.

Details of the coils are given in Fig. 12-30. The oscillator and mixer coils employ Millen No. 74002 shielded plug-in forms, and the output transformer uses the No. 74001 permeability-tuned form of similar construction. These were selected because of the protection their design affords against the rough handling to which such equipment is subjected. The output transformer is designed to resonate near 1500 kc., and is core-adjusted to this frequency.

Alignment and Installation

The converter may be lined up using an a.c. power supply and a communications receiver. A good signal generator simplifies this task, but if none is available the job may be done as follows: Turn on the receiver, set the volume control at maximum, and with the receiver dial at 1500 kc. adjust the output transformer on the converter for greatest noise. With the 10-meter coils in place and the tuning condenser set at about half capacitance, set the oscillator trimmer near maximum capacitance. With the aid of a signal, which may be noise from ignition or an electric razor, adjust the mixer trimmer for maximum response.

◆

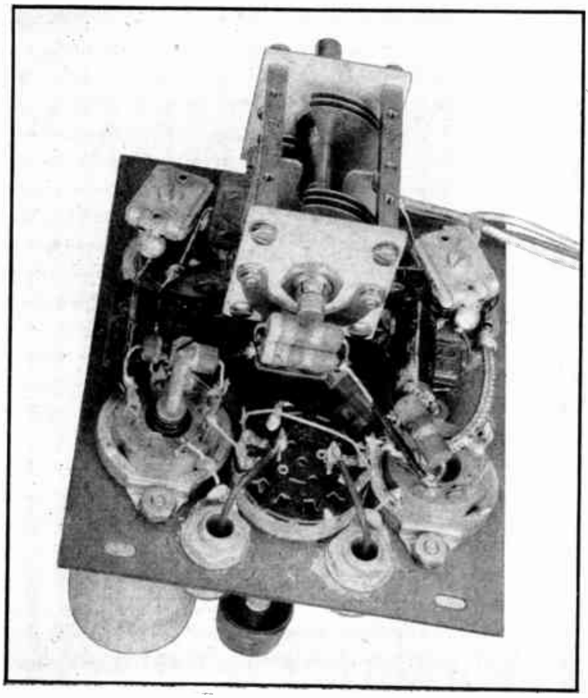


Fig. 12-29 — Inside view of the 6-, 10- and 11-meter mobile converter. Practically all of the construction and wiring can be completed before the subchassis is attached to the case.

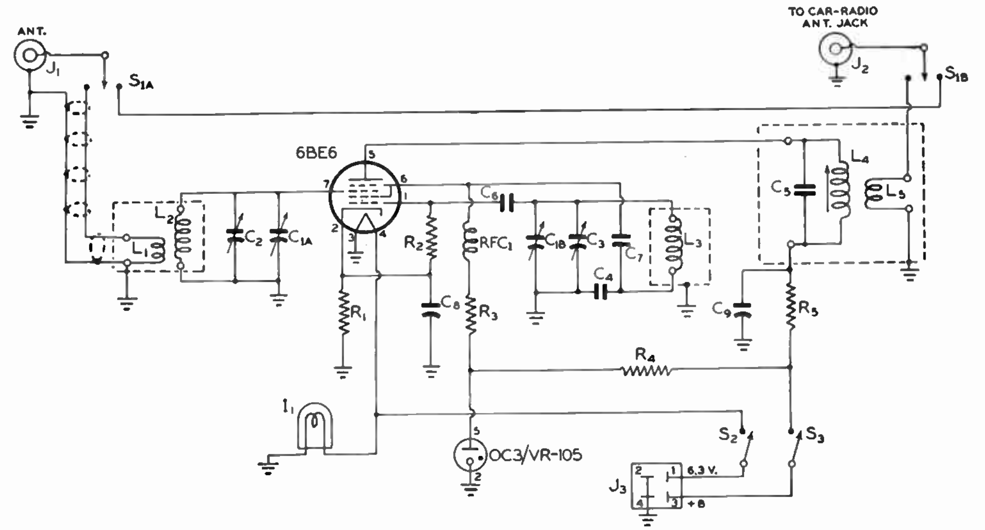


Fig. 12-30 — Wiring diagram of the mobile converter for 6, 10 and 11 meters.

C_{1A}, C_{1B} — Split-stator condenser, 15 μμfd. per section (Cardwell ER-15-AD).
C_2, C_3 — 3–30-μμfd. ceramic trimmer.
C_4, C_5 — 100-μμfd. midget mica.
C_6, C_7 — 0.0022-μfd. mica.
C_8, C_9 — 0.01-μfd. paper tubular.
R_1 — 150 ohms, ½ watt.
R_2 — 22,000 ohms, ½ watt.
R_3 — 680 ohms, 1 watt.
R_4 — 2200 ohms, 2 watts.
R_5 — 1000 ohms, 1 watt.
I_1 — 6.3-volt pilot lamp.
J_1, J_2 — Coaxial-cable jack.
J_3 — 4-prong male plug.
S_{1A}, S_{1B} — 2-pole 2-circuit selector switch.
S_2, S_3 — S.p.s.t. toggle switches.
L_1 — 50 Mc.: 2⅓ turns No. 18 d.c.c. wire, interwound in cold end of L_2.
— 28 Mc.: 3¼ turns No. 18 d.c.c. wire at cold end of L_2.
L_2, L_3 — 50 Mc.: 4½ turns No. 18 enameled wire, spaced one diameter.
— 28 Mc.: 11½ turns No. 18 enameled wire, ⅝ inch long.
L_4 — No. 36 d.s.c. wire, close-wound, one inch long.
L_5 — 15 turns No. 36 d.s.c. wire, close-wound over B-plus end of L_4. Note that condenser C_5 is mounted inside the coil shield.

Final adjustments can be made using amateur signals, resetting the oscillator trimmer as may be required to bring the band at the desired dial settings. The procedure outlined should be followed for the 6-meter range as well. Bandspread will be approximately 60 divisions for the 50-Mc. band and 85 for 27–29.7 Mc.

The exact anode voltage for the 6BE6 will not be known until the converter is actually tested in the car installation. The tube is designed to operate with 250 volts on the output plate and 100 volts on the oscillator plate. With a 150-ohm cathode resistor the cathode current is approximately 10 ma. It is important that the regulator tube be allowed to function normally at all times, and in cases where the receiver voltage is abnormally low it may be necessary to change the value of the limiting resistor, R_4. The proper value will permit the tube to regulate (as indicated by a constant glow) over the complete voltage range, as the input voltage varies with the condition of the car battery and the charging rate of the car generator.

In modifying the car receiver it is necessary to remove it from its case and install a power plug for carrying the heater and plate voltages and the ground lead to the converter. The heater lead should be connected directly to one of the receiver tube sockets, to take advantage of any 6-volt line filtering which may be incorporated in the receiver. Similarly, it is wise to tap for plate power at a point where the r.f. circuits are most heavily decoupled. This may result in a somewhat lower plate voltage than the maximum obtainable, but the extra filtering aids in eliminating hum and vibrator hash.

Antenna and power cables should be made up at this time. One coaxial cable must reach from the car antenna to the converter and the second from the converter to the receiver antenna connection. The 3-wire shielded power cable will connect between the receiver output plug and the 4-prong plug at the converter end.

In putting the converter into operation in the car it may be necessary to readjust the output transformer slightly, and retune the r.f. trimmer in the broadcast receiver slightly to compensate for loading effect of the converter. It may also be found that, although the broadcast receiver may be quiet in its operation, there may be considerable noise from ignition and the car generator when the converter is used. Even with the best available filters and suppressors the ignition noise may still be excessive, as the result of pick-up from other cars, in which case the best solution to the problem is the installation of some form of noise silencer in the car receiver. Suitable noise silencers are completely described in Chapter Five.

A 2-Meter Converter for Mobile Use

When 144-Mc. work was being done extensively with modulated-oscillator transmitters the mobile worker had no alternative to the superregenerative receiver, broadband superhets being too complex for most mobile installations. Now that practically all 2-meter stations are stabilized, it is possible to use the same technique for this band as for lower frequencies. The converter shown in Figs. 12-31 through 12-34 was designed primarily for mobile reception in connection with a car broadcast receiver. It may also be used for home-station work, with any receiver capable of tuning to the high end of the broadcast band.

Only one change from conventional practice is required: the image rejection at 144 Mc. when an i.f. of 1600 kc. is used is so low that double conversion must be employed for satisfactory results. Two 6J6 twin triodes are used, each serving as a mixer-oscillator. The first converts the signal frequency to 11.1 Mc., the second working from this frequency to 1600 kc. Only the high-frequency oscillator requires tuning during normal operation, the other circuits being present at fixed frequencies during the adjustment and testing of the converter. Plate voltage for all circuits is stabilized by an 0B2 regulator tube.

Circuit Details

The first mixer has a self-resonant grid coil which is tuned to the center of the 144-Mc. band by the tube and circuit capacitances. Its plate circuit is tuned to 11.1 Mc. by C_1 and L_3. The oscillator tunes from 132.9 to 136.9 Mc. to cover the band. It uses the second section of the first 6J6 and, beating with the incoming signal, produces an i.f. of 11.1 Mc. which is then capacity coupled by means of C_9 to the grid of the second mixer. Actually, the oscillator covers a somewhat greater range than that given above, in order that the converter may be tuned outside either end of the band. C_4 is the band-set condenser and C_5 is the band-spread capacitor. No coupling condenser is used between the oscillator and mixer, since stray coupling between grid pins at the socket gives adequate injection.

The second 6J6 serves as another mixer-oscillator combination, converting the 11.1-Mc. i.f. to 1600 kc. for working into a car radio at the high end of the broadcast band. Note that a trap (C_2L_4) is connected in series with the coupling condenser between the two mixer circuits. This trap is tuned to 14.3 Mc. and attenuates image response at a frequency removed from the signal frequency by 3200 kc. This image, which falls within the 2-meter band when the converter is tuned to the low edge, can be reduced by 35 to 40 db. through adjustment of the trap.

The plate circuit of the mixer is tuned to 1600 kc. by the trimmer, C_3, and a fixed capacitor, C_{11}, which supplies the additional capacitance required. A low-impedance output link, L_6, terminates at J_2, and a short length of coaxial cable is used between the jack and the receiver.

Circuit details of the low-frequency oscillator are nearly identical with those of the high-frequency oscillator, except that the low-frequency circuit uses only one capacitor, C_6, across the plate coil because the circuit operates at a fixed frequency of 12.7 Mc. Radiation from the oscillator, when the latter was operated with 108 volts applied to the plate of the

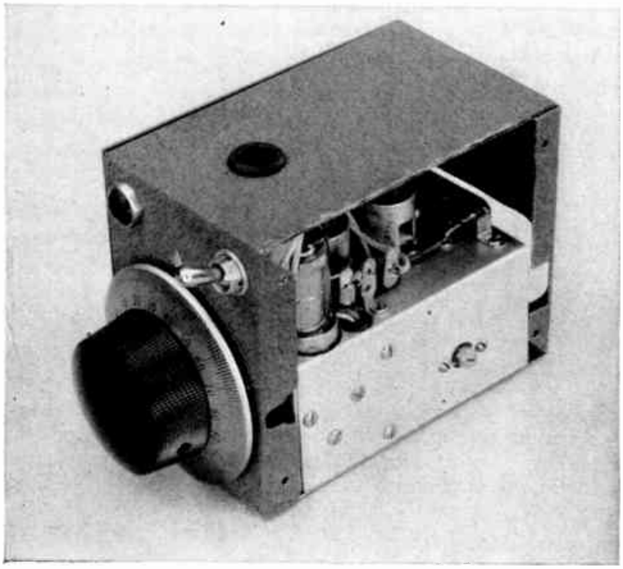


Fig. 12-31 — A two-tube 144-Mc. converter for use with a car broadcast receiver. The side plate was removed to show the modification of the utility cabinet.

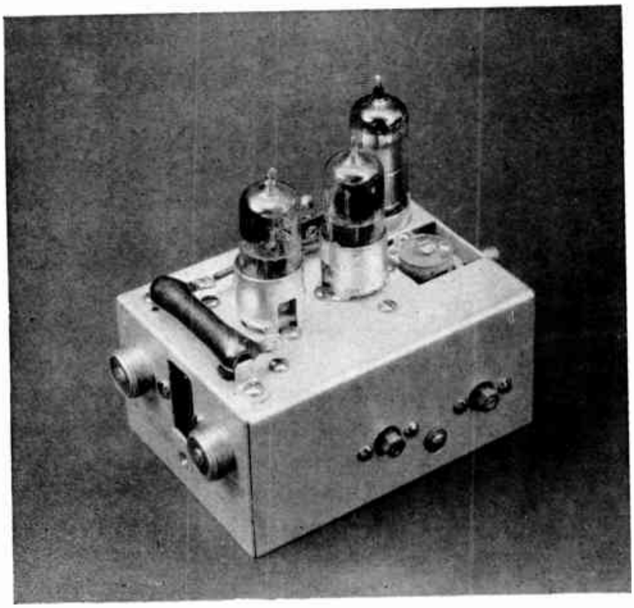


Fig. 12-32 — Top view of the 2-meter converter removed from its case.

6J6, reached the high-frequency mixer and caused numerous spurious responses as the converter was tuned through the band. This condition was eliminated by reducing the oscillator plate voltage (by means of the dropping resistor, R_5) and by placing a copper shield between the two circuits. The reduction in oscillator signal affected the mixer sensitivity and it was necessary to introduce a small amount of capacitive coupling between the oscillator and mixer. A 1½-inch length of 75-ohm Twin-Lead, identified as C_X on the circuit diagram, provides adequate coupling capacitance.

The 0B2 regulator tube is adjusted (by means of R_8) to pass approximately 12 ma. when the converter is connected to a 300-volt supply. The tube will be badly overloaded if the supply is turned on with the 6J6 tubes removed from their sockets. Otherwise, the tube will operate satisfactorily with a supply output voltage of 250 to 350 volts. The measured output potential of the regulator circuit is 108 volts.

Construction

The chassis for the converter measures 1⅞ by 2⅞ by 4 inches and is made from a 6⅝ by 7¾-inch sheet of 1/16-inch aluminum stock. A 1⅞-inch square is cut from each corner of the aluminum sheet so that the metal can be bent to form a boxlike chassis. It is recommended that the marking and drilling of mounting holes for parts be done before the chassis is bent into shape. The photographs of the converter show the location of most of the components. The hole for the oscillator band-set condenser (seen at the top of the chassis) is 1 inch square and is centered between the sides of the chassis. The mounting hole for the bandspread condenser is ¼ inch down on the front wall, and a ⅞-inch hole for the regulator-tube socket is centered to the left of the square hole. The high-frequency mixer-oscillator tube is centered on the chassis to the rear of the square hole, and the other r.f. tube is 1¼ inches to the right of the first tube. A mounting hole for the 11.1-Mc. coil is located ¾ inch in from the edge of the chassis directly to the right of the h.f. oscillator tube, and the 12.7-Mc. (second-oscillator) coil is ¾ inch in from the rear of the chassis and centered ⅞ inch away from the left edge. The form for L_5 is ⅝ inch from the right edge and ⅞ inch from the rear edge. R_8, J_1, J_2 and J_3 may be seen at the rear of the chassis and the location of these components is not critical. Holes, equipped with rubber grommets, are drilled adjacent to the limiting resistor and the regulator tube to provide feed-through points for the B-plus and heater wiring. A two-terminal lug strip is located to the rear of the regulator tube for the leads running to the filament switch and the pilot-lamp socket. Trimmer condensers, C_1, C_3 and C_6, are mounted on the side walls of the chassis with their shafts 1⅛ inches from the top of the box. C_1, mounted on the left side, is ¾ inch back from the front wall and C_3 is 1⅜ inches farther toward the rear. C_6 is 1¼ inches from the rear wall on the right side. The mounting hole for the 14.3-Mc. coil is 7/16 inch up from the bottom edge of the chassis and is centered between C_1 and C_3.

The bottom view of the converter, Fig. 12-34, shows how the regulator-tube socket is mounted on a small aluminum bracket which is in turn mounted on the side wall of the chassis. An aluminum strip, 1 inch wide, should be bent to form a right angle and the position of the socket mounting hole should be

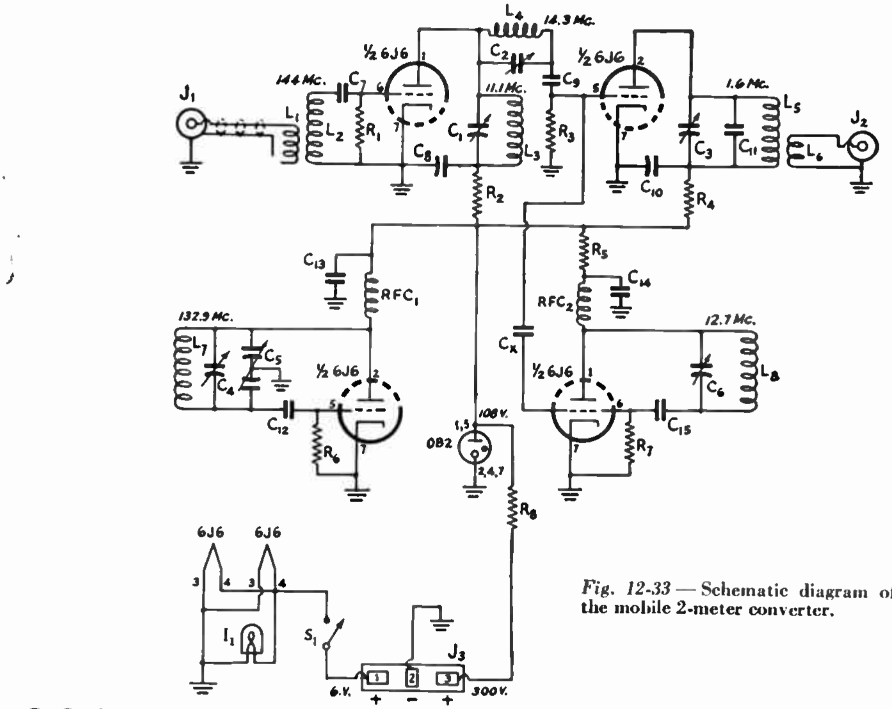


Fig. 12-33 — Schematic diagram of the mobile 2-meter converter.

C1, C3, C6 — 62-μμfd. trimmer (Centralab 823-AZ).
C2, C4 — 20-μμfd. trimmer (Centralab 820-B).
C5 — 5.27-μμfd. "butterfly" variable (Johnson 160-205).
C7, C15 — 47-μμfd. mica.
C8, C10 — 0.01-μfd. paper.
C9 — 100-μμfd. mica.
C11 — 150-μμfd. mica.
C12 — 15-μμfd. mica.
C13 — 470-μμfd. mica.
C14 — 0.0047-μfd. mica.
CX — Injection coupling, made from 75-ohm Twin-Lead — see text.
R1, R3 — 1.5 megohms, ½ watt.
R2, R4 — 1000 ohms, ½ watt.
R5 — 0.22 megohm, ½ watt.
R6, R7 — 15,000 ohms, ½ watt.
R8 — 3500 ohms, 10 watts.
L1 — 4 turns No. 22 enam., close-wound, 3⁄16-inch diam.
L2 — 6 turns No. 14 enam., 5⁄16-inch diam., ⅝ inch long.
L3 — 20 turns No. 28 enam., ½-inch diam., 5⁄16 inch long. Coil wound on a National PRD-2 form.
L4 — 28 turns No. 28 enam., ⅜-inch diam., ⅜ inch long. Coil wound on a National PRC-3 form.
L5 — 75 turns No. 28 enam., 9⁄16-inch diam., 1 inch long. Coil wound on a National PRE-3 form.
L6 — 10 turns No. 28 enam., close-wound over cold end of L_5.
L7 — 3 turns No. 14 enam., 5⁄16-inch diam., approx. ½ inch long. See text for adjustment of length.
L8 — 20 turns No. 28 enam., ½-inch diam., 5⁄16 inch long. Coil wound on a National PRD-2 form.
I1 — 6.3-volt pilot-lamp assembly.
J1, J2 — Coaxial-cable jack (Amphenol 75-PC1M).
J3 — Three-prong cable jack (Jones S-303-AB).
RFC1 — 1-μh. r.f. choke (National R-33).
RFC2 — 300-μh. r.f. choke (Millen 34300).
S1 — S.p.s.t. toggle switch.

marked after the bracket has been placed inside the chassis against the large clearance hole. Excess material may be cut from the bracket after it has been drilled for the socket. A three-terminal tie-point strip is mounted in a vertical position to the rear of the aluminum bracket, the bottom lug serving as a support point for the grid end of L_2. The coaxial cable and the antenna coupling loop are connected to the remaining two lugs.

A suitable shield for the low-frequency oscillator circuit can be made from a 1½ × 3¾-inch strip of 1⁄16-inch copper. The strip is bent to form a right angle having sides 1⅞ inches long and covering all of the components located at the top left-hand corner of the chassis. The shield is notched at the bottom corner to allow clearance for the coaxial cable which runs along the left edge of the chassis, and is equipped with a spade lug (the lug is soldered to the copper) for mounting.

The PRE-3 coil form for L_5 should be cut to 1¾ inches before the coil is wound. This and the other forms should then be marked and drilled to accommodate the windings. Terminal holes are drilled straight through the forms and the ends of the windings are passed through these holes. A coat of cement may be applied to the windings and allowed to dry while other operations are performed.

As shown by Fig. 12-31, some work must be done on the metal utility box before it can be used as a cabinet. This modification consists of removing the top and bottom flanges at the right side of the case and then notching the front and rear flanges to provide clearance for the condenser shaft and the jacks which are mounted on the aluminum chassis. A large slot must be cut in the rear of the case to allow access to the input and output jacks when the unit is assembled, and ¾-inch holes should be cut in the top, bottom and sides of the box so that the adjustment screws of the trimmer condensers may be reached with an alignment tool. The heater switch and the pilot lamp are mounted as far toward the top of the front panel as possible, and a ¾-inch hole is drilled up from the bottom of the panel for a distance of 1½ inches. This large hole will allow the National AM dial to be positioned correctly with respect to the tuning-condenser shaft after the chassis has been placed inside the cabinet.

The miniature Johnson condenser, C_5, may have a small-diameter control shaft which does not fit a standard dial coupling, in which case a bushing or shim is required. Fortunately, a ¼-inch length of easy-to-work ¼-inch soft-drawn copper tubing can be made to fit the shaft by working the inner surface with a rattail file.

Wiring

Construction and wiring are not difficult if the parts are mounted and wired in the following order: First, mount the tube sockets, the three jacks, and the lug strip (the one located on the top of the chassis). Next, complete the heater wiring and mount the grid-leak resistors in place. C_4 can now be soldered across the terminals of C_5 and L_7 can also be mounted on the condenser. This assembly is then mounted on the front wall of the chassis and, in turn, is connected to the tube socket by means of a short length of stiff tinned wire at the plate side and by C_{12} at the grid side. Now, mount the vertically-positioned lug strip on the side wall and connect a short piece of coaxial cable between the top lugs and J_1. C_7 can now be connected between the tube socket and the terminal strip and L_2 (with the small antenna winding slipped inside the cold end of the coil) may be mounted.

Condensers C_1, C_3 and C_6, and coils L_3, L_5 and L_8, are now mounted and wired into their respective circuits and, from here on, the wiring can proceed in any order. The 0.01-μfd. by-pass condensers are mounted in a vertical plane next to C_1 and C_3, respectively, and RFC_1 and R_2 are supported at the B-plus end by Pin 5 of the regulator-tube socket. The small metal post at the center of the rear tube socket is used as the tie point for the common connection between C_{14}, R_5, RFC_2 and the plate-voltage lead. L_4 is wired to C_2 after the padder condenser has been mounted between the coupling condenser, C_9, and a piece of No. 12 tinned wire which runs down to the stator terminal of C_1.

If the constructor wishes to use noise as a means of making a rough alignment of the converter, it is suggested that the injection-voltage condenser, C_X, and the dropping resistor, R_5, be left out of the circuit at this time. Of course, the plate of the 6J6 must be connected directly to RFC_2 in this case. The converter will have a much higher noise level when wired in this manner and alignment on noise is simplified. Actually, this is a poor method of aligning a double converter and should be used only as a last resort.

Testing

Power requirements for the converter are approximately 300 volts at 50 ma. and 6 volts at

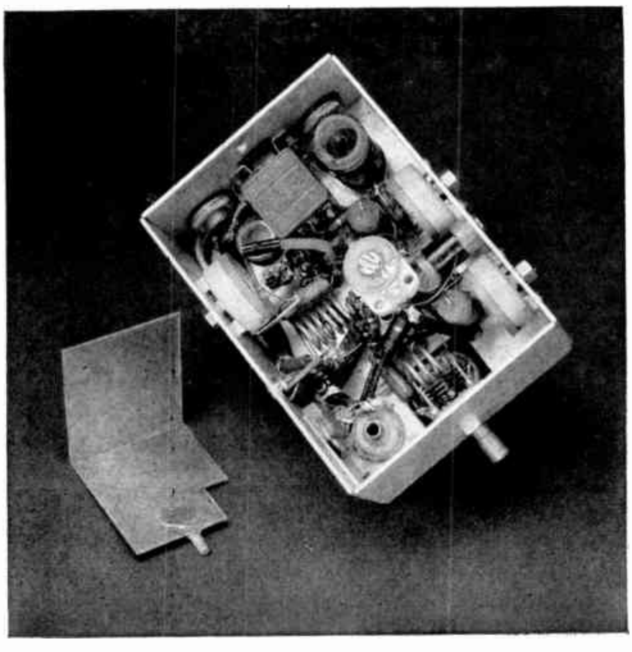


Fig. 12-34 — Bottom view of the 2-meter converter, showing the small copper shield used to reduce "birdies" from the low-frequency oscillator.

0.9 ampere. The first test consists of plugging the three tubes into the sockets and applying these potentials. In the absence of a voltmeter, it is safe to assume that the mixer and oscillator plate voltages are correct as long as the 0B2 glows when high voltage is turned on. A receiver capable of tuning to 1600 kc. should be coupled to the converter by a short length of coaxial cable and the receiver adjusted for normal operation at this frequency. If a signal generator is to be used, it is connected to the input jack, J_1, and if a generator is not available, the converter should be coupled to a low-impedance antenna system. Remember that C_X and R_5 should both be incorporated in the circuit if the converter is to be aligned with the aid of a test signal.

If preliminary testing is to be done with noise, the converter and the receiver are turned on and the converter output tuning condenser, C_3, adjusted until the noise level is at maximum. The low-frequency oscillator should now be adjusted by means of C_6 until a further increase in noise level is heard. C_4, the h.f. oscillator padder, should also be adjusted to produce maximum receiver output and this should occur with the padder adjusted to approximately half capacity.

At this stage of the game, it is necessary to introduce a test signal of known frequency, and it is helpful if the signal can be set at 146 Mc. — the center of the band. With such a signal fed to the converter, and with C_5 set at half capacity, C_4 is adjusted until the test signal is heard. It is advisable to check the frequency of the high-frequency oscillator at this point to make sure that it is adjusted to the low-frequency side of the input mixer circuit. Condensers C_1, C_3 and C_6 should now be tuned for maximum converter sensitivity. Incidentally, the frequency of the second oscillator can be checked by tuning the range around 12.7 Mc. with an all-band receiver.

The converter bandspread can be adjusted by changing the L/C ratio of the first oscillator, by altering the spacing between turns of L_7. Of course, C_4 must be reset each time the inductance of the coil is varied. Because the first mixer has a broad frequency response, it is only necessary to peak the input coil, L_2, at the center of the band by varying the length of the coil. The coupling between the antenna link and L_2 should be adjusted for maximum response.

When all of the circuits have been aligned, it is time to adjust the 14.3-Mc. trap. This is done by tuning to the high side of the signal frequency until the image is heard, and then adjusting C_2 until the image response is attenuated to the greatest degree.

It is to be expected that the various circuits will need slight readjustment after the chassis has been enclosed in the cabinet. However, this presents no difficulty as all of the tuning controls are accessible through small holes in the cabinet walls.

V.H.F. Transmitters

Beginning with the v.h.f. region, amateur frequency assignments are not in direct harmonic relationship with our lower-frequency bands. This fact, coupled with the necessity for extreme care in selection and placement of components for low circuit capacitance and minimum lead inductance, makes it highly desirable to construct separate gear for v.h.f. work, rather than attempt to adapt for v.h.f. use a transmitter designed for the lower amateur frequencies.

Transmitter stability requirements for the 50-Mc. band are the same as for lower bands, and proper design may make it possible to use the same rig for 50, 28, 21, and even 14 Mc., but incorporation of 50 Mc. and higher in the usual multiband transmitter is generally not feasible. Rather, it is usually more satisfactory to combine 50 and 144 Mc., since the two bands are close to a third-harmonic relationship. At least the exciter portion of the transmitter may be made to cover the requirements for both these bands very readily.

Though no stability restrictions are imposed by law on operation at 144 Mc. and higher amateur bands (other than that the entire emission must be kept within the limits of the band in question), experience has demonstrated the value of using crystal control or its equivalent for at least home-station operation in the 2-meter band. When large numbers of stations flock to a v.h.f. band, as occurred in the first months of operation on 144 Mc., severe interference soon develops if unstable transmitters and broadband receivers are employed. Conversion of this activity to crystal-controlled transmitters and receivers having the minimum bandwidth necessary for voice communication makes it possible for hundreds of stations to operate without undue interference, in the same band which appeared overcrowded with only a dozen or so stations working with inferior gear.

The use of narrow-band communications systems also pays off in the form of improved efficiency in both transmitter and receiver. It is this factor, perhaps more than the interference potentialities of the wide-band systems, which makes it desirable to employ advanced techniques at 220 and even 420 Mc. Stabilized transmitters for 220 Mc. are not too difficult to build, and their use at this frequency is highly recommended.

Construction of multistage rigs for 420 Mc. is not easy, and the choice of tubes suitable for this type of work is quite limited, but the advanced amateur who is interested in making the most of the interesting possibilities afforded by this developing field will be satisfied with nothing less. The 420-Mc. band is much wider than our lower v.h.f. assignments, however, and interference is not likely to become a limiting factor in this band for a long time to come. Thus it may be more important, in many localities, to get activity rolling with any sort of gear, leaving perfection in design to come along as the need develops.

At 420 Mc. and in the higher amateur assignments most standard tubes cannot be used with any degree of success, and special tubes designed for these frequencies must be employed. These types have extremely-close electrode spacing, to reduce transit-time effects, and are constructed with leads having virtually no inductance. Several more-or-less conventional tubes are now available which will operate with fair efficiency up to about 500 Mc., and the disk-seal or "lighthouse" variety will function up to about 3000 Mc. with specially constructed circuits. Above about 2000 Mc. the most useful vacuum tubes are the klystron and the magnetron. These are essentially one-band devices, the frequency-determining circuits being an integral part of the tube itself. Tuning over a small frequency range, such as an amateur band, is possible, usually by warping a built-in cavity, but the tubes are not independent of frequency in the conventional sense.

Frequency modulation may be used throughout the v.h.f. and higher bands, wide-band emission being permitted above 52.5 Mc. and narrow-band FM above 51 Mc. Where suitable receivers are available to make best use of such emissions, either wide-band or narrow-band FM can provide effective v.h.f. communication, and the latter is becoming increasingly popular, particularly in congested areas, where its freedom from broadcast interference permits operation under conditions which would be prohibitive for amplitude-modulated transmitters of any appreciable power.

In areas where there is television service in operation, the v.h.f. enthusiast must guard against interference to television reception. One way of keeping TVI to a minimum is the use of low power in the driving stages, building up the power level only after the operating frequency is reached. Extensive shielding and filtering, covered in other chapters, may also be required.

A 60-Watt Transmitter for 50, 28 and 14 Mc.

The transmitter-exciter shown in Figs. 13-1, 13-2 and 13-3 is a three-stage unit designed for use in the 50-, 28- and 14-Mc. bands. It employs a 6V6GT Tri-tet oscillator, a 6V6GT frequency multiplier, and an 815 operating as a straight amplifier without neutralization. It is capable of an input of 75 watts when being operated as an exciter or c.w. transmitter, but the power should be reduced to 60 watts input if the amplifier is modulated. Plug-in coils are employed for simplicity and flexibility.

Circuit Features

The Tri-tet oscillator has a fixed-frequency cathode circuit which resonates at approximately 21.5 Mc. With the cathode circuit so tuned, it is possible to employ a wide variety of crystal frequencies. For operation on 14 Mc., 3.5-Mc. crystals may be used, with the oscillator plate circuit tuned to 7 Mc., doubling in the second 6V6GT to 14 Mc. Most 3.5-Mc. crystals will deliver sufficient output from the oscillator on 7 Mc. to permit operating the second stage as a quadrupler to 28 Mc. also. Crystals between 7000 and 7200 kc. can be used with the oscillator working straight-through, provided that the oscillator plate circuit is not resonated at exactly the crystal frequency. Crystals from 7000 to 7425 kc. are used for operation of the amplifier in the 28-Mc. band, the oscillator doubling in this case.

For 50-Mc. operation crystals between 6250 and 6750 kc. are recommended. The oscillator is then operated as a quadrupler and the second stage as a doubler. The oscillator may also be operated as a tripler, using crystals between 8334 and 9000 kc., or as a doubler with 12.5–13.5-Mc. crystals. The complete unit may also be used as a driver for a 144-Mc. tripler stage by using the above crystal types, except that the ranges would then be 6000–6166, 8000–8222, or 12,000–12,333 kc.

Cathode bias is used on the oscillator stage, maintaining the plate current at a reasonable figure. The common-power-supply voltage is lowered to 300 volts by the dropping resistor, R_5, and the screen voltage of 250 is obtained from resistor R_3. The by-pass condenser, C_7, is inserted between the cold end of the plate coil, L_2, and ground, in order to permit grounding of the tuning-condenser rotor. Capacitive coupling, through C_8, is used between the oscillator and the multiplier stage.

The multiplier stage has a self-resonant plate circuit which is inductively coupled to the grid circuit of the final amplifier. This arrangement was found to give better balance of excitation to the final-stage grids than did the use of a self-resonant grid circuit in the final stage and a tuned plate circuit for the 6V6GT. Resistors R_9 and R_{15} drop the screen and plate voltages to the proper values. Grid bias is developed across R_6. The cathode by-pass condenser, C_9, is required because of the lengthening of the cathode lead by the insertion of the keying jack, J. No neutralization is required in this stage, since it is operated as a frequency multiplier at all times.

Bias for the final amplifier is obtained from a 45-volt "B" battery, permitting the preceding stage to be keyed for c.w. operation. The voltage-divider network, R_{12} and R_{13}, can be adjusted to provide the proper screen voltage for either type of operation. The plate tuning-condenser rotor is connected directly to ground. Power output is taken from the plate circuit by means of the adjustable link, L_6, which is part of the plug-in plate-coil assembly.

A switching system is provided for measuring all the necessary currents with one 50-ma. meter, connecting it across shunt resistors R_4, R_7, R_8, R_{10}, R_{11} and R_{14}. The range of the meter is extended to 300 ma. by means of R_{14}, which consists of about 31 inches of No. 30 insulated wire scramble-wound on a 100-ohm resistor. The length of wire required for this shunt may vary with different types of meters.

Mechanical Details

The front-view photograph, Fig. 13-1, shows the arrangement of the principal com-

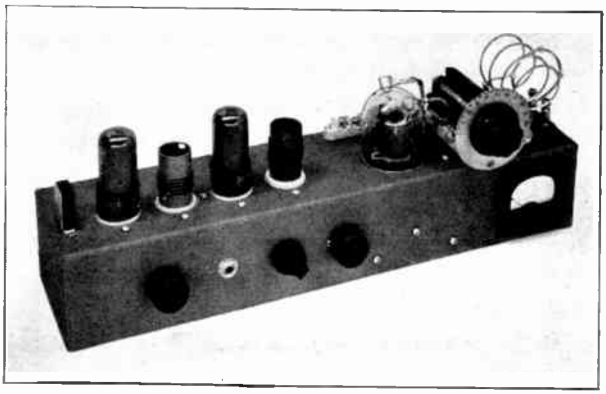


Fig. 13-1 — A front view of the 815 transmitter-exciter for 50, 28 and 14 Mc. Coils for 50-Mc. operation are in place.

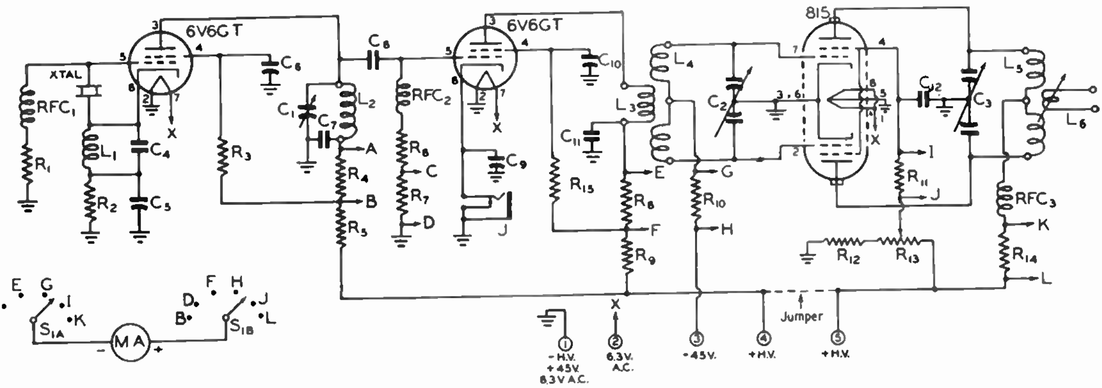


Fig. 13-2 — Circuit diagram of the 815 transmitter-exciter.

C_1 — 25-μμfd. variable (Cardwell ZR-25-AS).
C_2 — 50-μμfd.-per-section variable (Bud LC-1662).
C_3 — 35-μμfd.-per-section variable (Hammarlund MCD-35-SX).
C_4 — 68-μμfd. mica.
C_5, C_6, C_9, C_{10} — 0.01-μfd. paper.
C_7 — 0.0022-μfd. mica.
C_8 — 100-μμfd. mica.
C_{11}, C_{12} — 470-μμfd. mica.
R_1 — 0.12 megohm, ½ watt.
R_2 — 220 ohms, ½ watt.
R_3 — 15,000 ohms, 1 watt.
R_4, R_7, R_8, R_{10}, R_{11}, R_{14} — 100 ohms, ½ watt.
R_5, R_9 — 5,000 ohms, 10 watts, adjustable.
R_6 — 47,000 ohms, ½ watt.
R_{12}, R_{13} — 5,000 ohms, 10 watts, adjustable.
R_{15} — 40,000 ohms, 10 watts.
L_1 — 8 turns No. 18 enameled, close-wound, ½-inch dia.
L_2 — 7 Mc. — 25 turns No. 20 d.c.c., close-wound, 1 inch long.
— 14 Mc. — 12 turns No. 20 d.c.c., space-wound, 1 inch long.
— 28 Mc. — 5 turns No. 18 enameled, space-wound, ½ inch long.
L_3 — 14 Mc. — 9 turns No. 20 d.c.c., close-wound, ⅜ inch long.
— 28 Mc. — 6 turns No. 20 d.c.c., space-wound, ⅜ inch long.
— 50 Mc. — 3 turns No. 18 enameled, space-wound, ¼ inch long.
L_4 — 14 Mc. — 14 turns No. 20 d.c.c., close-wound, 7 turns each side of primary.
— 28 Mc. — 6 turns No. 20 d.c.c., spaced diam. wire, 3 turns each side of primary.
— 50 Mc. — 4 turns No. 18 enameled, spaced diam. wire, 2 turns each side of primary.

Above coils are wound on 1-inch diameter forms (Millen 45004 for L_2; Millen 45005 for L_3-L_4). Approximately ⅛ inch between L_3 and L_4.

L_5 — 14 Mc. — 14 turns No. 16, 1⅞-inch diameter, 2 inches long. (B & W 20-JVL.)
— 28 Mc. — 8 turns No. 12, 1⅞-inch diameter, 2 inches long. (B & W 10-JVL.)
— 50 Mc. — 4 turns No. 12, 1⅞-inch diameter, 2 inches long. (B & W 10-JVL with 2 turns removed from each end.)

Above coils are wound in two sections with half the total number of turns each side of center. A ½-inch space is left at the center to permit the use of a swinging link (L_6). The Barker & Williamson coils are mounted on five-prong bases of the type which plug into tube sockets.

J — Closed-circuit jack.
MA — 0–50 milliammeter.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke.
S_{1A}, S_{1B} — 2-circuit 6-position selector switch (Mallory 3226J).

ponents. The chassis measures 3 × 4 × 17 inches. It is suggested that this layout be followed closely, particularly as regards the arrangement of parts in the 815 plate and grid circuits, since this layout permits operation of the 815 without neutralization. Components along the top of the chassis, from left to right, are the crystal, oscillator tube, oscillator plate coil, multiplier tube, multiplier plate and final grid coil, final-amplifier tube, and amplifier tank circuit. A 5-connector terminal strip may be seen just to the rear of the 815. Across the front wall are the oscillator tuning knob, the keying jack, meter switch, multiplier tuning knob, and meter.

The 815 socket is mounted on a subchassis which can be seen in the bottom view, Fig. 13-3. This unit, Millen No. 80009, also includes the tube shield, which was cut off in this case so that, with the shield just flush with the chassis, the tube socket is 1½ inches below. Arrangement of other components is apparent from the bottom view. Resistors R_{12} and R_{13} are at the right end near the coil socket. The amplifier plate choke, RFC_3, is on a stand-off adjacent to the coil socket, and the meter shunt, R_{14}, is connected between the choke and one end of R_{13}. The dropping resistor, R_9, is parallel to the rear wall of the chassis, and R_5 is at right angles to it, both below, and to the left of, the sockets for the 6V6GT and its associated plug-in coil. RFC_1 and RFC_2 are mounted on stand-off insulators at the rear of the tube sockets. The cathode coil, L_1, is self-supporting, and is mounted directly on the tube-socket terminals. All other small parts are grouped closely adjacent to their respective tube sockets.

Coil information is given in full detail under Fig. 13-2. The number of turns specified should be satisfactory for the frequencies of operation referred to, though some adjustment of the coupling between windings of L_3 and L_4 may be required. Since variation in spacing of these windings affects the tuning range of the L/C combination to a noticeable degree, the setting of C_2 should be checked as the coupling adjustment is being carried on.

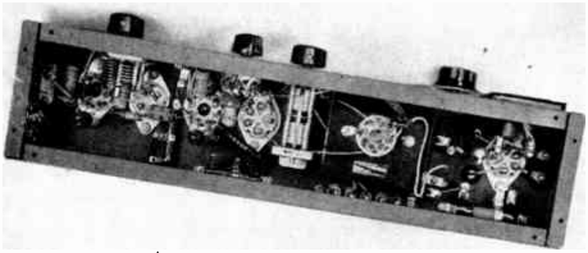


Fig. 13-3 — A bottom-inside view of the 815 transmitter.

Testing Procedure

The power supply for the transmitter should be capable of delivering 275 to 300 ma. at 400 or 500 volts. The filament transformer should supply 6.3 volts at 2.5 amperes or more. It is advisable to test the oscillator and multiplier stages before applying plate and screen voltage to the 815. The jumper between Terminals 4 and 5 on the terminal strip should be left off during this operation. With plate voltage applied to the oscillator and doubler, the oscillator plate circuit should be tuned to resonance as indicated by maximum grid current in the next stage. The amplifier grid circuit should then be tuned for maximum grid current. With the coil specifications given it is unlikely that the circuits will tune to an incorrect harmonic; nevertheless, it is wise to check with a calibrated absorption-type wavemeter to be sure that such is not the case. Dropping resistors R_5 and R_9 should be set at their full value of 5000 ohms during this operation, final setting of the adjustments being made after the power supply is loaded by the entire transmitter. Grid current to the final stage should be about 15 ma. for all bands at this point. This may be adjusted by changing the turns on L_3, or by detuning C_2, if the grid current is excessive. Detuning of C_2 is recommended as the more satisfactory of the two methods.

The possibility that neutralization may be required should be checked at this point. With the drive applied, but with no plate or screen voltage on the 815, the plate circuit of the amplifier should be tuned through resonance while watching the grid current. If there is an appreciable kick in the grid current it will be necessary to add neutralization, which may take the form of wires brought up alongside the tube envelope, in the manner shown in Fig. 13-14. With this particular layout no neutralization was needed.

The final amplifier may now be put into operation. The screen voltage should be tapped in between the screen divider, R_{12} and R_{13}, and the jumper connected between Terminals 4 and 5. With plate voltage and grid excitation applied, the off-resonance plate current should be about 250 ma., dropping to about 25 ma. when the plate circuit is tuned to resonance. A dummy load such as an electric lamp bulb should be connected across the output terminals, and the coupling adjusted to bring the plate current up to 150 ma. at resonance.

The oscillator and multiplier plate voltages and the amplifier screen voltage should now be adjusted to 300 and 200 volts, respectively, by adjusting the taps on R_5, R_9 and R_{13}. It is probable that the amplifier plate current will change appreciably at this point, so it will be desirable to readjust the coupling so that the current is again 150 ma., and readjust the resistor taps as required to secure the correct voltages on the various tube elements. Adjustment for 'phone operation is similar, except that the amplifier screen and plate voltages should be 175 and 400 volts, respectively.

With all voltages set for the proper values, the currents will run about as follows: oscillator and doubler plates — 35 ma.; doubler grid — 1 to 3 ma.; 815 grid — 5 to 6 ma.; 815 screen — 17 ma.; 815 plate — 150 ma. The 815 screen current should drop to 15 ma. for 'phone operation. A grid current of 4 to 6 ma. in the 815 is adequate. Though more grid current is usually available, an increase beyond 6 ma. does not improve the output. In c.w. operation, it will be noticed that the 815 plate current does not drop completely off when the excitation is removed. This is no cause for concern, so long as the plate and screen dissipation are held to recommended levels.

The amplifier plate coils are provided with links designed for working into a low-impedance line. The amplifier may thus be used for direct feed in the case of antenna systems having matched lines, or it may work into a line feeding an antenna-coupling unit, in case tuned-feeder antenna systems are employed.

The transmitter may be used by itself as a complete unit, or it may be made to serve as a driver for a final stage of 200 watts or more capacity, depending on the final-stage tubes used. The 24G amplifier unit shown on the succeeding pages is a typical example.

If variable-frequency operation is contemplated the output of a VFO unit may be plugged into the crystal socket. The "V.H.F. Man's VFO," described later in this chapter, is suitable for this purpose.

If more power is required than can be obtained from the 815, an 829 may be substituted in the output stage with only minor changes. Somewhat more grid drive will be required for the 829, but the output from the second stage should be sufficient for either tube.

200-Watt Driver-Amplifier for 50 and 144 Mc.

A medium-power driver-amplifier for v.h.f. use is shown in Figs. 13-4 to 13-7. The amplifier uses a pair of 24G triodes in push-pull, while the driver, a frequency tripler used for 144-Mc. operation only, is a single 829-B. If operation on 144 Mc. is not contemplated, all to the left of the final grid coil, L_3 in Fig. 13-6, may be omitted.

Looking at the front-panel view, Fig. 13-4, the two large dials are the plate tuning controls for both stages. The small dial at the left controls the swinging link, the center one is the grid tuning control for the final, and the one at the far right is the tripler grid tuning.

The rear view shows the general placement of parts. At the left rear is a jack bar, containing terminals *AA* and *BB*, into which the link from the exciter is plugged to furnish excitation for the tripler (terminals *AA*) or final stage (terminals *BB*). The tripler grid-circuit components, L_1C_1, are adjacent to the jack bar. C_1 is mounted with its shaft parallel to the panel, in order to permit short leads, and is tuned by means of a flexible shaft. The 829-B and its plate-circuit components are mounted in such position as to permit inductive coupling between the plate coil of the tripler and the grid coil of the final, when the transmitter is used on 144 Mc. For 50-Mc. work the 829-B is inoperative, and the 50-Mc. grid coil (the same coil is used in both tripler and final) just clears the plate coil of the 829-B. Care must be used in parts placement to work this out correctly.

Between the grid tuning condenser, C_3, and the 24G tubes are the two neutralizing condensers. These are triple-spaced midgets, mounted back-to-back, with coupled shafts. The stator plates had to be filed out to reduce the minimum capacitance to the small value needed to neutralize the 24Gs. The final tank condenser, C_5, and the jack bar for the final plate coil are positioned for the short leads that are required if satisfactory performance on 144 Mc. is to be attained. R.f. leads in the final stage are made with ¼-inch silver ribbon, which is appreciably better than braid at these frequencies. Thin copper strip may also be used. All connections in the plate circuit of the 24Gs should be made with bolts and nuts, as the tank circuit will heat sufficiently during 144-Mc. operation to melt soldered connections and increase losses. Plate connections to the 829-B are made by means of small Fahnestock clips. The tripler plate coil is supplied with two of these clips (Millen 36021), so that it may be removed separately from the condenser leads, in replacing the tube.

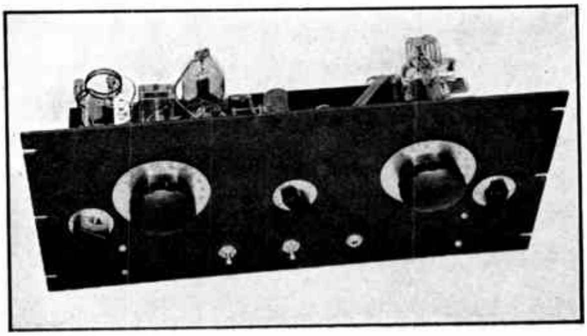
Operation on 50 Mc.

When the amplifier is to be used on 50 Mc. the switch, S_1, is left open, so that the heater of the tripler will not be energized when S_2 is closed. The link from the exciter is plugged into terminals *BB* in the jack bar, which is a Millen 41205 coil socket. The output of the exciter is thus connected to the link terminals of the final grid-coil socket, a National XB-16.

The final stage should be tested on 50 Mc. using an exciter having an output of 20 to 40 watts, before attempting 144-Mc. operation. With the proper coils inserted at L_3 and L_4, and with power on the exciter but no plate voltage on the final, rotate C_3 for maximum grid current. Set the neutralizing condensers at maximum capacitance and rotate C_5. If the plate circuit is capable of being resonated there will be a kick in the grid current as the circuit passes through resonance. The neutralizing condensers should be rotated, a small amount at a time, until the kick in the grid current disappears. This will probably occur close to the minimum-capacitance setting of C_{4a} and C_{4b}.

Power may now be applied to the plate circuit. It is advisable to make initial tuning adjustments at low voltage, preferably 750 volts or less. If everything is in order, the plate current will drop to about 20 ma. at resonance, at this voltage. A load of some sort should now be connected across the output terminals and the operation tested at increasingly higher voltages. The final tubes should be capable of an input of 250 watts or more on 50 Mc. without exceeding their normal plate dissipation of 50 watts for the pair, indicated by a bright orange color.

Fig. 13-4 — Front view of the 200-watt driver-amplifier for 50 and 144 Mc. The two large dials are the plate tuning controls. The small dial at the left adjusts the position of the output coupling link, the center dial is the grid tuning control for the final, and the third small dial is the tripler grid-tuning control. Across the lower center are the filament switches and grid-current meter jack.



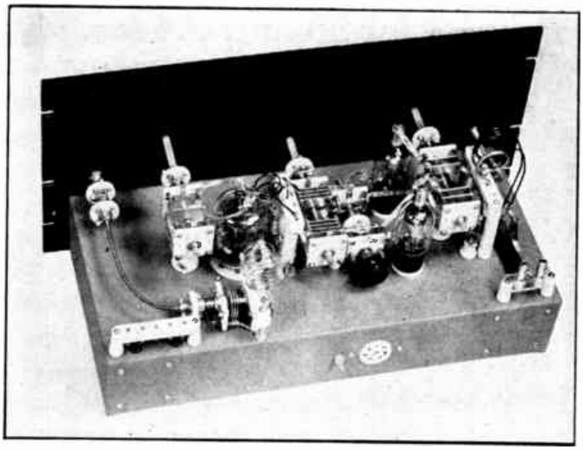


Fig. 13-5 — Rear view of the driver-amplifier for 50 and 144 Mc., with coils in place for 144-Mc. operation. At the lower left are the link socket and the grid circuit of the tripler stage. The plate coil of the tripler is inductively coupled to the final grid coil, a single turn. All parts are grouped as closely as possible, for efficient performance on 144 Mc. R.f. leads in the final stage are made of ¼-inch silver ribbon.

Checking on 144 Mc.

For operation on 144 Mc., the switch, S_1, should be closed, energizing the heater of the 829-B tripler. The link from the exciter should be plugged into terminals *AA* on the jack bar, so that drive is applied to the tripler grid circuit. The 144-Mc. coils should be inserted at L_3 and L_4. It will be noted that these coils have no bases. The grid coil (merely a center-tapped "U") is made of No. 12 wire, which fits snugly

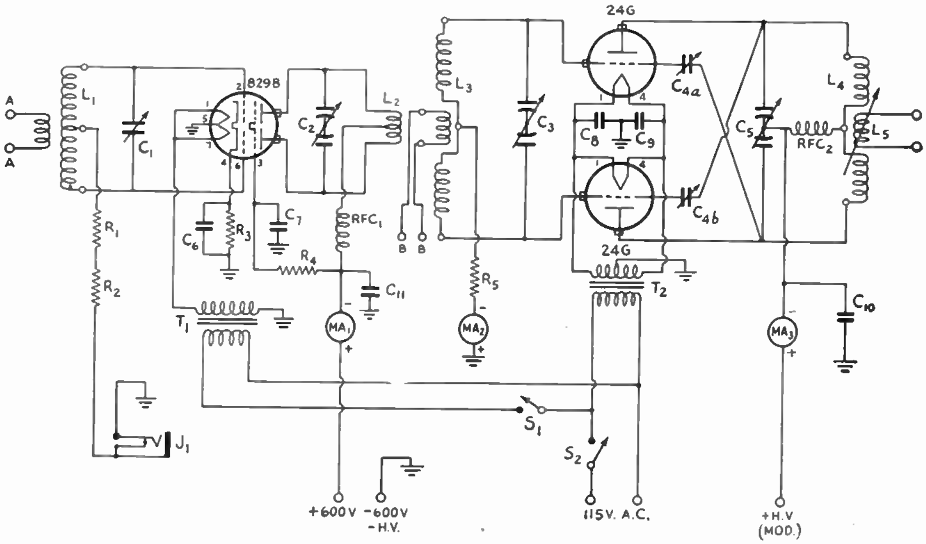


Fig. 13-6 — Schematic diagram of the 200-watt driver-amplifier.

C_1 — 15-μμfd. variable (Cardwell ZR-15-AS).
C_2 — 10-μμfd.-per-section split stator (Cardwell ER-10-AD).
C_3 — 15-μμfd.-per-section split stator (Cardwell ET-15-AD).
C_{4a}, C_{4b} — 2-plate triple-spaced midget variable (Cardwell ZS-4-SS with stator plates filed to reduce minimum capacitance).
C_5 — 4-μμfd.-per-section split stator (Cardwell ES-4-SD).
C_6, C_7 — 470-μμfd. midget mica.
C_8, C_9 — 0.001-μfd. mica.
C_{10} — 500-μμfd. 2500-volt mica.
C_{11} — 500-μμfd. 1000-volt mica.
R_1 — 4700-ohm 2-watt carbon.
R_2 — 50,000 ohms, 10 watts.
R_3 — 250 ohms, 10 watts.
R_4 — 15,000 ohms, 10 watts.
R_5 — 3000 ohms, 10 watts.
L_1 — 4 turns No. 18, 1¼-inch diameter, 1 inch long, 3-turn center link (National AR-16, 10-C, with 2 turns removed from each end).
L_2 — 2 turns No. 12 enameled, ⅞-inch diameter, spaced ¼ inch.
L_3 — 50 Mc.: Use L_1. 144 Mc.: Center-tapped "U" made from No. 12 enameled wire, ⅝ inch high, ¾ inch wide. See rear-view photograph. Sockets for L_1 and L_3 are National XB-16.
L_4 — 50 Mc.: 4 turns each side of center-tap, spaced diameter of wire, on Millen No. 40205 base.
— 144 Mc.: 2 turns ⅛-inch brass rod, spaced to fit into socket terminals, 1¼-inch inside diameter. May be silver-plated for best results.
L_5 — Swinging link: 3 turns No. 12 enameled wire, 1⅜-inch diam. Mount for adjustment on polystyrene rod; see rear-view photograph.
J_1 — Closed-circuit jack.
MA_1 — 0–200 d.c. milliammeter.
MA_2 — 0–100 d.c. milliammeter.
MA_3 — 0–300 d.c. milliammeter.
RFC_1, RFC_2 — Ohmite Z-0.
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — 6.3 volts, 2.5 amp.
T_2 — 6.3 volts, 6 amp.

in the coil-socket terminals. It should be bent slightly toward the 829-B plate coil, L_2, so that it fits between the turns at a position that provides optimum inductive coupling. The final plate coil is made of ⅛-inch brass rod, which provides a tight fit in the jack bar. The coil may be silver-plated, if desired.

The grid coil for the tripler (L_1) may be the same coil as is used for 50-Mc. operation at L_3. With this coil in place and the output of the exciter on 48 Mc., the operation of the tripler may be checked, using a voltage of around 400 on the plates of the 829-B initially. The grid circuit of the final should then be tuned to resonance as indicated by maximum grid current. The plate voltage on the tripler may be raised to 600 volts, if necessary, to assure adequate grid drive for the final stage. Typical operating conditions are as follows: tripler plate voltage — 600; plate current — 125 ma.; tripler grid current (read in J_1) — 10 ma. or more; final grid current — 35 to 50 ma.

Neutralization of the final should be rechecked, as the position of C_{4a} and C_{4b} may be slightly different from the 50-Mc. setting. Power may then be applied to the final stage, using low voltages at first. The final stage should not be operated without load, except with low plate voltages, as tank-circuit losses will cause excessive heating otherwise. The dip in plate current at resonance will be less than at 50 Mc., and minimum current at 1000 volts will probably not drop below about 65 ma. The maximum recommended plate potential for 144-Mc. operation is about 1250 volts, though tests have been made on this amplifier at voltages as high as 1700.

A safe check on the operation of the tubes is to adjust the plate voltage (with no excitation applied) until an input of 50 watts is being run to the pair. Note the color of the plates at this input — a bright orange. If this color is not exceeded in normal operation, one may be sure that the tubes are being operated within safe limits. An input of 200 watts can be handled safely on 144 Mc. if the stage is running properly.

The transmitter may be used for c.w. work on either band by keying the cathode of the driving stage. If the 829-B tripler cathode is to be keyed a jack will have to be added. Provision should be made to add fixed bias in series between the final grid meter, MA_2, and ground. A 45-volt "B" battery will suffice to hold the 24G plate current at a safe value when the excitation is removed.

The three meters shown in the schematic diagram, but not in the photographs, are mounted on a separate meter panel. Another useful addition, not shown, is a 3500-ohm 10-watt potentiometer, connected in series with R_5, so that the final-stage bias can be varied to suit different operating conditions. This is particularly useful if the transmitter is to be used on c.w. and FM, as well as AM voice operation.

Suggested Modifications

Extensive on-the-air experience with this particular piece of equipment has turned up a possible trouble which can be corrected readily. In v.h.f. equipment it is seldom possible to use commercial coils for the tank circuits, and thus the amateur must make his own. The coils for this transmitter are largely in that category. They were designed for maximum efficiency, but they lack satisfactory insulation at certain critical points.

Coupling between L_2 and L_3 must be adjusted carefully for maximum transfer of energy, if sufficient drive is to be obtained for operation of the final stage on 144 Mc. This presents the possibility that the two coils may come in physical contact with one another, in which case the enamel insulation may break down, allowing the 829 plate voltage to be applied to the final grids, causing almost instant failure of the 24G tubes if their plate voltage is applied at the time. The cure is to slip short lengths of high-grade spaghetti over the 144-Mc. coil for L_3, at the points where it might possibly come in contact with L_2.

The same precaution should be observed with the antenna coupling coil, L_5. The shape of L_4 should be made such that contact between the two is unlikely, and high-voltage insulation should be applied to L_5. This is important as a safety feature, to keep the final plate voltage off the antenna system, and it is absolutely necessary to prevent shorting final plate voltage to ground when all-metal arrays or other grounded antenna systems are used.

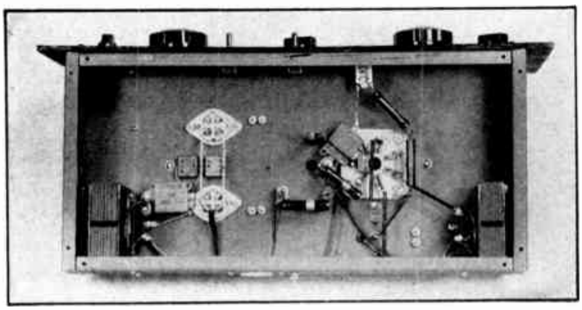


Fig. 13-7 — Bottom view of the 50- and 144-Mc. driver-amplifier.

A V.H.F. Man's VFO

Though a VFO is considered to be an almost indispensable part of an amateur station for lower frequencies. v.h.f. operation is still carried on mainly with crystal control. This is largely because of the relatively lower occupancy of the v.h.f. bands and the freedom from interference problems which results. It is also, in part, the result of the fact that, as we go higher in frequency, it becomes more difficult to generate an entirely satisfactory signal by means other than with crystal control.

With proper attention to the well-known factors affecting oscillator stability a frequency control unit for 80-, 40- or 20-meter use can be built with a minimum of complications, but many a signal which sounds acceptable on these frequencies becomes quite fuzzy by the time it is multiplied to the v.h.f. bands. Even on 10 meters it is not too easy to obtain a pure d.c. note, especially when the oscillator frequency is modulated for narrow-band FM.

The frequency-control unit described herewith has a degree of frequency stability that is adequate for the high-order frequency multiplication required in v.h.f. service, and the design of the audio portion is such that little or no hum is introduced in the reactance-modulation process. The unit has the reactance-modulator and speech amplifier built in, the gain of the latter being only just enough to provide sufficient deviation for 10-meter NFM. Much of the hum present on some FM signals comes from the use of excessive speech gain, or haywire patching systems in order to utilize the speech equipment in some other portion of the transmitter.

This unit, shown in Figs. 13-8–13-10, was designed with the needs of the v.h.f. man in mind. Since many v.h.f. operators also work on 10 and 11 meters the oscillator tuning range was extended to include these bands, as well as 2 and 6 meters. The actual output frequency of the VFO is 6.74 to 9 Mc. It is designed to serve as a crystal substitute, and may be plugged into the crystal socket of any transmitter employing crystals falling within its tuning range. Thus, though the dial is calibrated only for the bands from 11 to 2 meters, the unit may be used on 40 or 20, or on portions of the higher v.h.f. bands that are in harmonic relationship with the output frequency. The output is sufficient so that the unit may also be used as a driver for a low-powered amplifier or frequency multiplier whose grid circuit is on that frequency. It also includes a reactance modulator and speech amplifier, providing narrow-band FM on 27 Mc. and higher frequencies with only the addition of a crystal microphone.

Two 6AG7s are used in the r.f. portion. The first is an oscillator-doubler employing the highly-stable Clapp oscillator, the operating frequency of which is 3370 to 4500 kc., doubling in the plate circuit. The second is an amplifier operating on 6.74 to 9 Mc. By means of separate padders switched in by a front-panel control, a reasonable amount of bandspread is provided for each of the four bands from 2 to 11 meters. The 50-Mc. band covers 55 divisions on the vernier dial, 144 Mc. is covered in 25 divisions, the 10-meter band occupies 80 divisions, and 11 meters 20 divisions. By proper setting of the padders the 2- and 11-meter ranges can be made to come at the opposite ends of the National MCN dial, leaving the two other spaces on the dial card for the 10- and 6-meter calibrations.

Frequency modulation is accomplished by means of a reactance modulator and a speech amplifier, both using 6BA6 miniature tubes. Deviation of the oscillator frequency is approximately 500 cycles, providing adequate swing for 10-meter NFM as a result of the eight times multiplication. A deviation of approximately 10 kc. is possible in the 6-meter band, and as much as 30 kc. on 2 meters. This greater

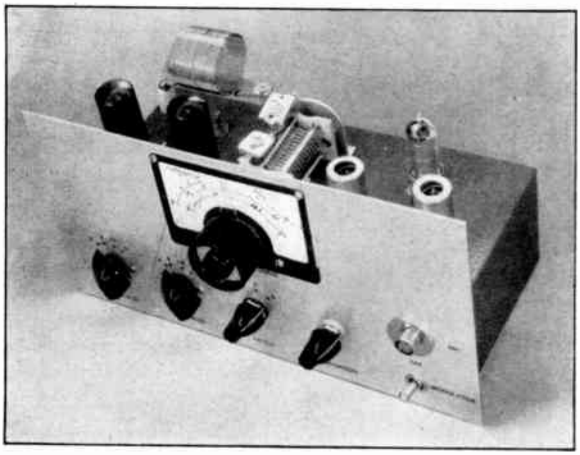


Fig. 13-8 — Panel view of the v.h.f. VFO with NFM modulator.

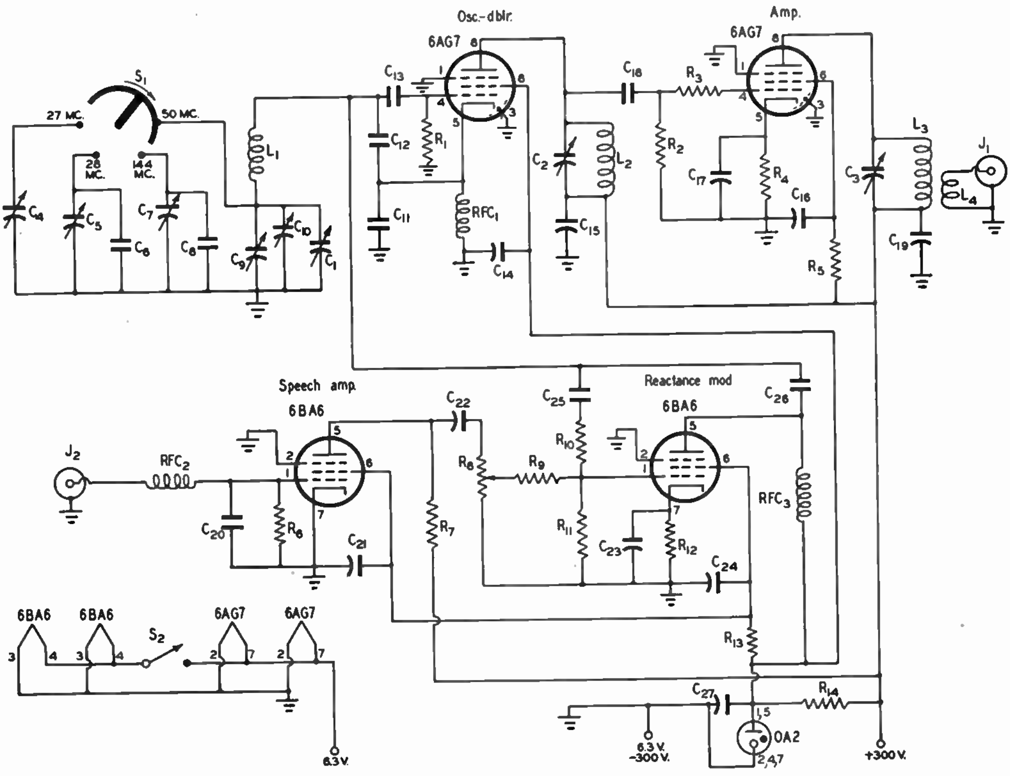


Fig. 13-9 — Circuit diagram of the NFM control unit for v.h.f. use.

C_1 — 35-μμfd. variable, double spaced (Millen 21935).
C_2, C_3 — 100-μμfd. variable (Millen 20100).
C_4, C_5, C_7, C_9, C_{10} — 2–30-μμfd. ceramic trimmer (Millen 27030).
C_6 — 33-μμfd. silver mica.
C_8 — 10-μμfd. silver mica.
C_{11}, C_{12} — 680-μμfd. silver mica.
C_{13} — 68-μμfd. silver mica.
C_{14}, C_{15}, C_{17}, C_{18}, C_{19}, C_{21}, C_{22}, C_{23}, C_{24}, C_{27} — 0.01-μfd. 400-volt paper.
C_{16}, C_{20} — 100-μμfd. mica.
C_{25}, C_{26} — 47-μμfd. mica.
R_1, R_9 — 0.1 megohm, ½ watt.
R_2, R_{10} — 10,000 ohms, ½ watt.
R_3 — 47 ohms, ½ watt.
R_4 — 330 ohms, 1 watt.
R_5 — 15,000 ohms, 2 watts.
R_6 — 1 megohm, ½ watt.
R_7, R_{13} — 0.22 megohm, ½ watt.
R_8 — 0.5-megohm potentiometer.
R_{11} — 0.47 megohm, ½ watt.
R_{12} — 470 ohms, ½ watt.
R_{14} — 7500 ohms, 10 watts.
L_1 — 24 turns No. 22 tinned wire, diameter 1½ inches, length 1⅛ inches (B & W 80 JCL with 18 turns removed).
L_2, L_3 — 14 turns No. 24 e. wire, diameter 1 inch, length ⅜ inch; wound on Millen 45000 form.
L_4 — 3 turns No. 24 e., close-wound at bottom end of L_3.
J_1, J_2 — Coaxial-cable jack (Jones S-101).
RFC_1, RFC_3 — 2.5-mh. r.f. choke (Millen 34100).
RFC_2 — 300-μh. r.f. choke (Millen 34300).
S_1 — 4-position progressive-shorting switch (Centralab GG modified; see text).
S_2 — S.p.s.t. toggle switch.

swing is useful on 144 Mc., where a considerable number of relatively-broad receivers is in use. The deviation is controllable to any required value below this, by means of the potentiometer, R_8. A switch is provided in the heater circuit of the speech section (S_2) so that this portion of the unit can be cut off when c.w. or amplitude modulation is being used. As operation of this switch affects the oscillator frequency appreciably it is usually preferable to leave the speech-section heaters on at all times, using the deviation control at its off position when emissions other than NFM are being used.

The arrangement of the parts should be clear from the photographs. The top view, Fig. 13-8, shows the microphone jack and heater switch at the right end of the panel. The deviation control, bandswitch, oscillator-plate and amplifier-plate tuning controls are in line across the bottom of the panel. The oscillator frequency setting is controlled by the vernier dial. Looking at the top of the chassis the two 6AG7s may be seen to the left of the tuning condenser, the first being the oscillator tube. The oscillator tank coil, L_1, is mounted on stand-offs, just in back of the 6AG7s. Two metal brackets are used to mount the tuning condenser, which should be the double-ended variety for greatest mechanical stability. The reactance-modulator and speech-amplifier tubes are at the right of the tuning condenser, with the regulator at the rear. The chassis is a standard 3 × 5 × 10-inch size and the panel is

6 by 11 inches. A 5 × 10-inch aluminum plate, with clearance holes for the trimmer adjustments, is attached to the bottom of the chassis.

The arrangement of components under the chassis is apparent from the bottom view, Fig. 13-10. The bandswitch and associated padders are at the middle, with the oscillator plate coil. The amplifier plate coil is at the left. The padder condensers are mounted with their grounded terminals soldered to metal pillars, in order to reduce sensitivity to vibration to a minimum.

The bandswitch requires some modification. In its original form it has a disk which shorts out all unused contacts. This disk must be cut through the center so that one half may be removed. As may be seen from the wiring diagram, Fig. 13-9, the connection between the oscillator coil and the switch is made to Number 1 terminal, rather than to the regular wiper contact.

The power supply for the VFO should be well-filtered and capable of delivering 300 volts d.c. at 60 to 70 ma., and 6.3 volts a.c. at 1.9 amp. Socket voltage measurements are approximately as follows: 20 volts on the audio-tube screens, 150 volts on the 6AG7 screens, 40 and 150 volts, respectively, on the speech-amplifier and reactance-modulator plates, and 300 volts on the 6AG7 plates. Cathode current for the oscillator should be about 10 ma., and the output stage, at resonance, 30 ma.

Calibration and Use

Calibration of the VFO dial can be accomplished with the aid of a receiver having an accurate dial calibration, as the readings on the VFO dial should not be relied upon for band-edge operation. The 50-Mc. range, requiring the least padder capacitance, should be calibrated first. Padders C_9 and C_{10} set at nearly full capacitance will provide the correct tuning range, which should be approximately 55 divisions spread over the middle of the dial scale. The 144-, 28- and 27-Mc. ranges should be calibrated in that order, their spread on the dial being approximately 25, 80 and 20 divisions respectively. If the NFM portion of the unit is to be used extensively it is recommended that the calibration procedure be carried out with the reactance-modulator heater on, as this tube affects the calibration appreciably.

When adjusting the plate circuits of the oscillator and amplifier stages it is recommended that the approximate settings of these controls for the middle of the band in question be marked on the panel. It will then not normally be necessary to readjust these controls when shifting frequency within a band. This broad-band effect is accomplished by slightly overdriving the amplifier tube at the center frequency, causing the screen voltage to drop and reduce the output. Tuning away from the center frequency reduces the drive and allows the screen voltage and output to rise. More than enough output is thus obtainable over the entire band, without too great a variation for proper operation of the succeeding stage. Two 250-ma. pilot lamps in parallel make a satisfactory dummy load for the amplifier.

Next the operation of the reactance modulator should be checked. The procedure for this operation is described in detail in Chapter Nine. It should also be pointed out that there is no excuse for radiation of an improperly-modulated FM signal, since it can be monitored readily in one's own receiver. With the receiver in operation on the band in which the transmitter is to be used, but with only the VFO turned on, it is a simple matter to tell exactly how the signal will sound on the air. Deviation requirements vary with different receivers, but a safe starting point is to set the deviation control so that the signal sounds well on a communications receiver with the crystal filter in the broadest "on" position.

Ordinarily a unit of this type may be used to replace the crystal stage of an existing transmitter by simply plugging it into the crystal socket. The output coupling is a low-impedance line, however, and it may be connected to a link winding on the grid coil of any low-power stage whose tuning range is 7 to 9 Mc. Although it is shown calibrated only for the frequencies above 27 Mc., it may be used as a c.w. exciter for 7- or 14-Mc. work. The deviation may, however, be insufficient for 20-meter NFM operation. Output, at 7 to 9 Mc., is about three watts.

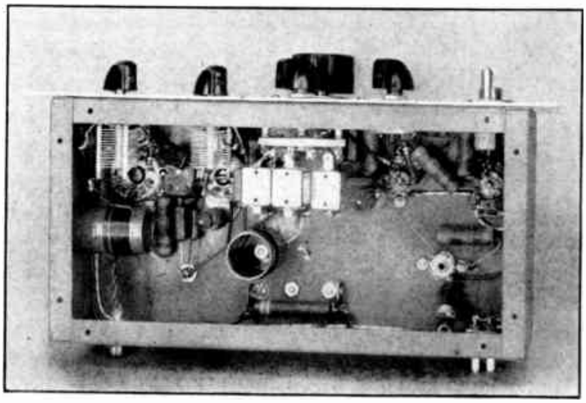


Fig. 13-10 — Bottom view of the VFO

A Simplified Exciter for 50 and 144 Mc.

Through the use of a special crystal-oscillator circuit, by means of which standard low-cost crystals are made to oscillate on their third harmonic in a simple triode regenerative oscillator, the transmitter-exciter shown in Figs. 13-11, 13-12 and 13-13 provides output on 50 and 144 Mc. with only two tubes and simple circuits. A dual-triode oscillator-multiplier is used, the first section oscillating on 24 to 27 Mc., depending on the frequency of the crystal, which may be anything from 8 to 9 Mc. The second section doubles to 48 to 54 Mc., providing more than enough output to drive an 832 amplifier or tripler. Plug-in coils are used in the 832 plate circuit, to permit output on 50–54 Mc. or 144–148 Mc. Output on the lower band is 20 watts or more, with three to five watts available on the higher frequency.

The rig may be modulated on 50 Mc., in either portable or fixed-station service, but it should be used as an exciter only for 144 Mc. The power output on the higher band is sufficient to drive another 832 or 829 stage, the design of which might follow that of the 829 amplifier described later in this chapter. It may also be used to drive a combination such as the rig in Fig. 13-6.

A standard 5 × 10 × 3-inch chassis is used, with the oscillator-multiplier components mounted below the deck and the 832 plate-circuit components above. A power-switching arrangement is included to permit use of the rig as a complete transmitter for 50 Mc. or as an exciter for an additional modulated stage on 144 Mc.

The Harmonic-Oscillator Circuit

Design is conventional except for the oscillator circuit, the key feature of which is the feed-back arrangement in L_1. The portion of the coil below the tap determines the proper functioning of the oscillator, the correct position of the tap being approximately one-third up from the crystal end of the coil when a 6J6 is used. With other dual triodes it may be necessary to alter this materially.

If too much inductance is included in the tickler portion of the coil the tube will oscillate at a frequency determined by the setting of C_2 rather than by the crystal. When the unit is ready for test the oscillator stage alone should be checked first. With a low-range milliammeter inserted temporarily in series with the multiplier grid resistor, R_2, about 150 volts should be applied to the oscillator plate. Rotate C_2 until grid current appears, indicating oscillation, the frequency of which should be checked in a calibrated receiver. Changing the setting of C_2 should not cause an appreciable change in the frequency of oscillation, and the crystal will oscillate only over a part of the tuning range of the condenser and at no other point. If the oscillator frequency shifts widely, indicating uncontrolled oscillation, the tap is too high on L_1. If the tap is too low the 6J6 will oscillate weakly or not at all, and will refuse to start when the condenser is tuned near the point of maximum output, as indicated by the grid-current peak in the succeeding stage.

It should be noted that pulling the crystal out of its socket is *not* a satisfactory check for uncontrolled oscillation, as the capacitance of the crystal and its holder is required to complete the feed-back circuit.

Provision is made for measuring the grid and cathode current of the amplifier stage by means of J_1 and J_2. The former is insulated from the panel, and connected in reverse, so that the meter leads need not be reversed in changing from one jack to the other. When the rig is operated on 50 Mc. the grid current in the 832 need not be more than 2 ma., and this amount of drive can be furnished by the 6J6 with 150 volts applied to the junction of R_1 and R_4. Amplifier cathode current, with no load, will be about 35 ma. at resonance, with a 400-volt supply. It may be loaded up to about 70 ma.

If 144-Mc. output is desired, the final stage should not be operated at more than 300 volts or so, but at this level it will provide more than

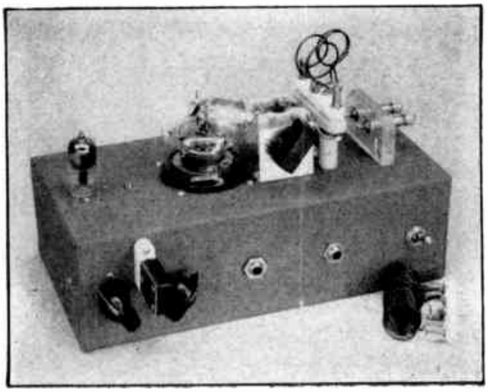


Fig. 13-11 — The two-tube exciter for 50 and 144 Mc. The 2-meter coil is plugged into the output stage, with the 6-meter one in the right foreground.

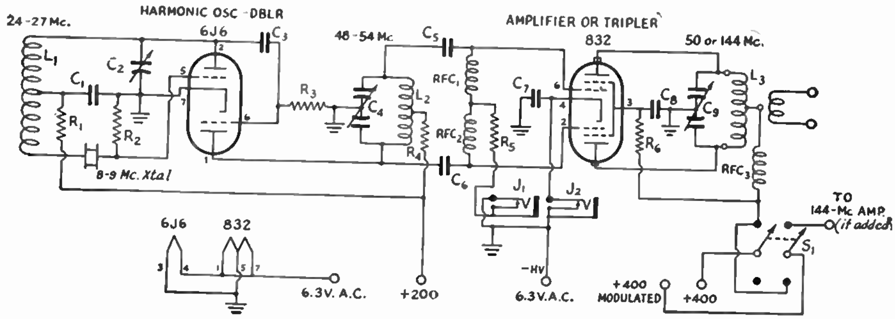


Fig. 13-12 — Schematic diagram of the 2-tube v.h.f. rig. The power-switching arrangement shown provides for later addition of a 144-Mc. amplifier stage.

C_1 — 680-μμfd. mica.
C_2 — 50-μμfd. variable.
C_3 — 15-μμfd. ceramic.
C_4 — 20-μμfd.-per-section split-stator, made by sawing the stator bars of a Millen 21050 and removing center plate.
C_5, C_6 — 75-μμfd. ceramic.
C_7, C_8 — 500-μμfd. ceramic.
C_9 — 6-μμfd.-per-section split-stator (Millen 21906D).
R_1 — 4700 ohms, ½ watt.
R_2 — 3300 ohms, 1 watt.
R_3 — 47,000 ohms, ½ watt.
R_4 — 3300 ohms, 1 watt.
R_5 — 22,000 ohms, 1 watt.
R_6 — 25,000 ohms, 10 watts.
L_1 — 14 turns No. 18, ½-inch diam., 1 inch long, tapped at 4½ turns.
L_2 — 12 turns No. 18, ½-inch diam., ⅞ inch long, center-tapped.
L_1 and *L_2* made from Barker and Williamson "Miniductor" type 3003.
L_3 — 50 Mc. — 14 turns No. 14 enamel, ⅞-inch diam., 2 inches long. Link: 3 turns No. 20 enamel, spaghetti-covered.
144 Mc. — 2 turns No. 14 enamel, 1-inch diam., spaced ½ inch. Link: 2 turns No. 16 enamel.
Base and plug assemblies are National XB-16 and PB-16.
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_2, RFC_3 — 25 turns No. 24 enamel on 1-watt resistor, or Millen 34300.
S_1 — D.p.d.t. toggle switch.

enough output to drive another 832 amplifier, or even an 829. For 144-Mc. use the whole unit may be operated from a single 300-volt supply, the additional voltage on the oscillator and doubler being helpful in securing sufficient drive to make the 832 triple effectively. It is not recommended that the 832 be modulated for 144-Mc. voice operation, as there is not enough drive for operation of the stage as a modulated tripler, and the functioning of such a stage would not be generally satisfactory under any conditions. Grid current, for tripling, should be 4 ma. or more.

In 50-Mc. service the over-all drain, with a 300-volt supply, is only about 85 to 90 ma., and under these conditions the amplifier delivers an output of about 10 watts, with a total load on the supply of less than 30 watts. On 144 Mc. the output is three to five watts.

A more complete description of the transmitter and the regenerative oscillator circuit used may be found in *QST* for October and November, 1948. The same technique could be employed to advantage in the construction of an exciter unit for 220 Mc., except that the second section of the 6J6 would be operated as a tripler to 75 Mc., instead of as a doubler to 50 Mc. More than enough output would be available to drive another 6J6 as a tripler from 75 to 225 Mc.

Another possibility in connection with the oscillator circuit used in this transmitter involves taking off the fifth harmonic instead of the third. Many 7-Mc. crystals can be used in this way, taking off the 5th harmonic from the first triode section, and then doubling in the second. Only an additional doubler stage is then needed to reach 144 Mc.

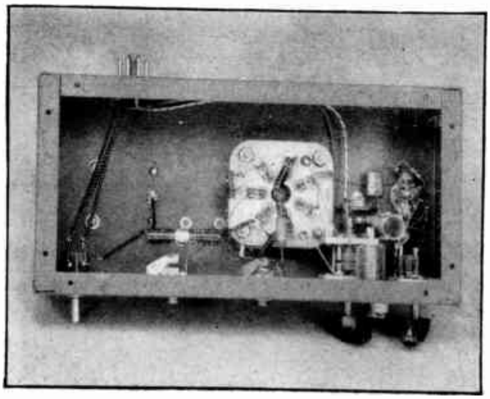


Fig. 13-13 — Bottom view of the simplified v.h.f. exciter.

144-Mc. Double Beam-Tetrode Power Amplifier

An amplifier set-up suitable for use with double beam-tetrode tubes is shown in Figs. 13-14, 13-15 and 13-16. The tube in the photographs is an 829, but an 815 or 832 can be used in the same layout. The only change that might be required would be in the inductances of the grid and plate coils, L_2 and L_3; these may have to be made slightly smaller or larger in diameter to compensate for the differences in input and output capacitances in the various types. When an 829 is used, the amplifier is well suited for use as an outboard unit with war-surplus transmitters such as the SCR-522.

The amplifier is built on an aluminum chassis formed by bending the long edges of a 5 × 10-inch piece of aluminum to form vertical lips ¾ inch high, so that the top-of-chassis dimensions are 3½ by 10 inches. The tube socket is mounted on a vertical aluminum partition measuring 3½ inches high by 3¼ inches wide on the flat face, with the sides bent as shown in the photographs to provide bracing. The partition is mounted to the chassis by right-angle brackets fastened to the sides. The socket is mounted with the cathode connection at the top, the cathode prong being directly grounded to the nearest mounting screw for the socket. The heater by-pass condenser, C_6, is mounted directly over the center of the tube socket, extending between the paralleled heater prongs at the bottom and the cathode prong at the top. The screen by-pass is connected with short leads between the screen prong and the nearest socket screw.

The grid coil, L_2, is supported by the grid prongs on the socket. The two turns of the coil are spaced about one-half inch to allow room for the input coupling coil L_1 to be inserted between them. The coupling is adjusted by bending L_1 into or out of L_2. The grid tuning condenser, C_1, is mounted between the socket prongs; although the condenser has mica insulation it is used essentially as an air-dielectric condenser since the movable plate does not actually contact the mica at any setting inside the band. The coupling link is soldered to lugs under binding posts on a National FWG strip, the strip being mounted on metal pillars 1½ inches high to bring the link to the same height as the grid coil.

Although the shielding between the input and output circuits of the tube is sufficiently good so that the circuit will not self-oscillate, tuning of the plate circuit will react on the grid circuit to some extent because the grid-plate capacitance, while small, is not zero. To eliminate this reaction it is necessary to neutralize the tube. The neutralizing "condensers" are lengths of No. 12 wire soldered to the grid prongs on the socket. The wires are crossed over the socket and then go through small ceramic feed-throughs at the top of the vertical shield, projecting over the tube plates on the other side as shown in Fig. 13-14.

Connections between the plate tank condenser, C_7, and the tube plate terminals are made by means of small Fahnestock clips soldered to short lengths of flexible wire. The tank coil, L_3, is mounted on the same condenser terminals to which the plate clips make connection. The output link, L_4, is mounted similarly to the grid link except that the posts are 1⅞ inches high. The plate choke, RFC_1, is mounted vertically on the chassis midway between the plate prongs of the tube, the mounting means being a short machine screw threaded into the end of the polystyrene rod. The "cold" lead of the choke is by-passed by C_5 underneath the chassis.

Supply connections are made through a 5-post strip on the rear edge of the chassis. The dotted lines between connections in Fig. 13-15 indicate

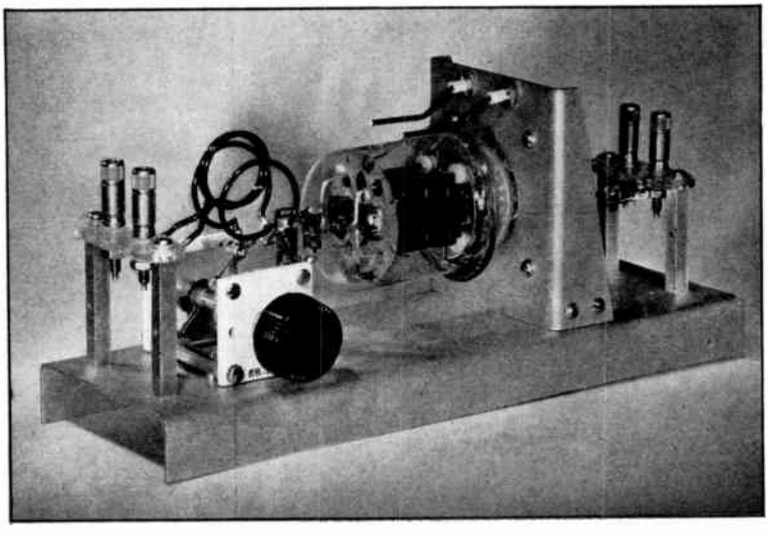


Fig. 13-14 — A 144-Mc. amplifier using a double beam tetrode. This type of construction is suitable for the 815 and 832 as well as the 829 shown. The vertical partition provides support for the tube as well as shielding between the input and output circuits. Note the neutralizing "condensers" formed by the wires near the tube plates.

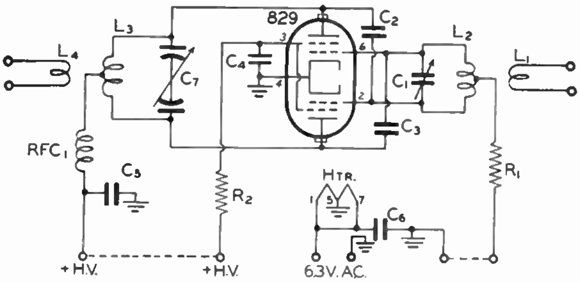


Fig. 13-15 — Circuit of the 829 amplifier for 144 Mc.

C_1 — 3–30-μμfd. ceramic trimmer.
C_2, C_3 — Neutralizing condensers; see text.
C_4 — 500-μμfd. mica, 1000 volts.
C_5 — 500-μμfd. mica, 2500 volts.
C_6 — 470-μμfd. mica.
C_7 — Split stator, 15 μμfd. per section (Cardwell ER-15-AD).
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.
L_1 — 2 turns No. 12, diameter ½ inch.
L_2 — 2 turns No. 12, diameter ½ inch, length ½ inch.
L_3 — 2 turns No. 12, diameter 1⅛ inches, length 1 inch.
L_4 — 2 turns No. 12, diameter 1 inch.
RFC_1 — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.

that these connections are normally short-circuited; leads are brought out so that the grid and screen currents can be measured separately.

In adjusting the amplifier, the plate and screen voltages should be left off and the d.c. grid circuit closed through a milliammeter of 0–25 or 0–50 range. The driver should be coupled to the amplifier input circuit through a link (Amphenol Twin-Lead is suitable, because of its constant impedance and low r.f. losses). Use loose coupling between L_1 and L_2 at first, and adjust C_1 to make the grid circuit resonate at the driver frequency, as indicated by maximum grid current. The coupling between L_1 and L_2 may then be increased to make the grid current slightly higher than the rated load value for the tube used — approximately 12 ma. for the 829. If the driver is an oscillator, the coupling between L_1-L_2 should be as loose as possible with proper grid current.

After neutralization, the procedure for which has been given in connection with other similar amplifiers, plate and screen voltage may be applied. If possible, the plate voltage should be low at first trial so there will be no danger of overloading the tube. Adjust C_7 to resonance, as indicated by minimum plate current (this should be measured independently of the screen); with the 829, the minimum plate current should be in the neighborhood of 80 milliamperes with 400 volts on the plate and no load on the circuit. A dummy load such as a 60-watt lamp should light to something near full brilliance when the coupling between L_3 and L_4 is adjusted to make the tube draw a plate current of 200 ma. When the loading is set, the grid current should be checked to make sure it is up to the rating for the tube.

Power-supply and modulator requirements will depend upon the particular tube used. For the 829, the plate supply should have an output voltage of 400 to 500 with a current capacity of 250 milliamperes. With a 400-volt supply the modulator power required is 50 watts, with an output transformer designed to work into a 1600-ohm load; with a 500-volt supply slightly over 60 watts of audio power is needed, the load being 2000 ohms.

This amplifier may also be used with the 832 rig described in the preceding pages. The output from the driver stage may be fed into the amplifier by means of a link, if the two units are to be operated remote from one another, or the grid circuit of the 829 may be arranged to provide direct inductive coupling to the 832, if the two are placed side by side.

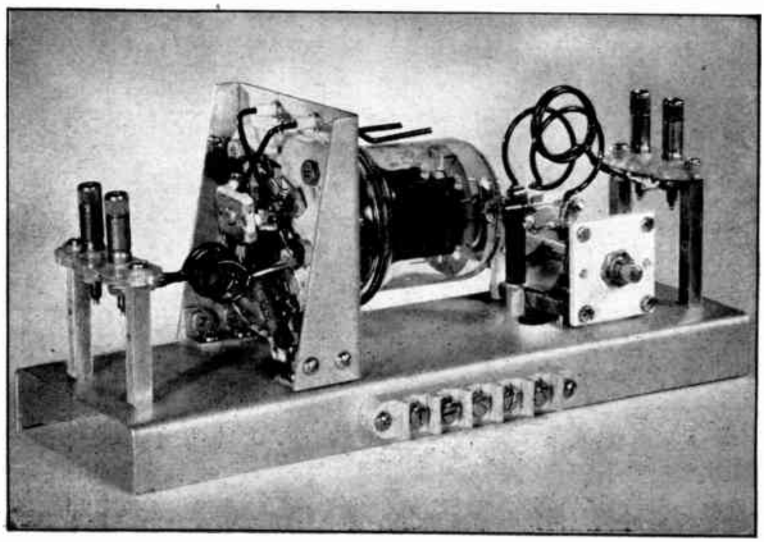


Fig. 13-16 — Another view of the 144-Mc. amplifier. The neutralizing wires are crossed over the socket before going through the feed-through insulators. The input circuit is designed for link coupling to the driver stage.

Crystal Control on 220 Mc.

Construction of a multistage transmitter for the 220-Mc. band is not as difficult as might be imagined, and the serious worker on this frequency will find the use of crystal control or its equivalent highly worth while. Fortunately the crystals used are also usable on 144 Mc., cutting down the total cost of building equipment for both bands, if the crystal frequencies are selected with this use in mind.

The transmitter-exciter shown in Figs. 13-17, 13-18 and 13-19 employs either 8- or 12-Mc. crystals, and if they are between 8148 and 8222 or 12,223 and 12,333 kc. they may also be used for operation in the upper portion of the 144-Mc. band. By using miniature tubes and components, and by arranging the parts for minimum lead length, efficient operation on 220 Mc. is obtained, with a simplicity of construction that puts the equipment well within the capabilities of the average experienced amateur.

Four 6J6 dual triodes are used. The first works as a triode oscillator and frequency multiplier, the second section doubling or tripling, depending upon which type of crystal is employed. Tuning is less critical, and the various stages operate somewhat more efficiently with 12-Mc. crystals, but 8-Mc. crystals may also be used. The next two stages are push-pull triplers, and the output stage is a neutralized amplifier. Capacitive coupling is used between stages. The chassis is 2½ inches wide, 2 inches high, and 12 inches long, with ½-inch edges folded over. It may be made from a piece of sheet aluminum 7½ by 12 inches in size. The first tube socket is 1½ inches in from the left end and the other sockets are spaced along the chassis, 2¼ inches center to center. The tuning condensers are spaced equally between the sockets, the last two, C_{13} and C_{17}, being mounted on the top surface of the chassis for minimum lead length and symmetrical layout. Pin jacks, labeled *a* and *b* on the schematic diagram, are mounted on the front wall of the chassis and may be used for metering or keying of the output stage.

Initial Adjustments

Meter jacks for the individual stages were not considered necessary, as there will normally be few occasions for shifting frequency and retuning, once the initial adjustment of the exciter is completed. For these first measurements the various circuits may be opened and tests made with a portable meter.

With a meter in series with R_2, set the core in L_1 at an intermediate position and adjust C_2 for oscillation, as indicated by a dip in plate current to about 10 ma. The frequency and note should be checked in a communications receiver, making sure that the oscillation is controlled by the crystal. Next, insert the meter in series with R_4 and tune C_4 for a dip at the proper frequency, which should be between 24.5 and 25 Mc. Adjustment of the multiplier tuning may be critical, if fundamental-type crystals are used, the crystal tending to "pop out" when C_4 is tuned on the nose. With "overtone" or harmonic-type crystals this trouble will not be in evidence, and the setting of C_4 (or the core in L_2) will not be fussy. Adjustment should be for maximum grid current in the second 6J6.

Adjustment of the push-pull tripler stages is merely a matter of resonating the circuits for maximum output as indicated by the grid current in the succeeding stage, being certain that the stages are tripling and not quintupling, which they will also do with fair efficiency. Each stage has cathode bias to prevent damaging the tubes during the adjustment period. Input to each will run about 25 ma. at 200 volts, when operating correctly.

Neutralization of the output stage is accomplished in the customary manner, except that the neutralizing capacitors are made from short lengths of 75-ohm Twin-Lead.

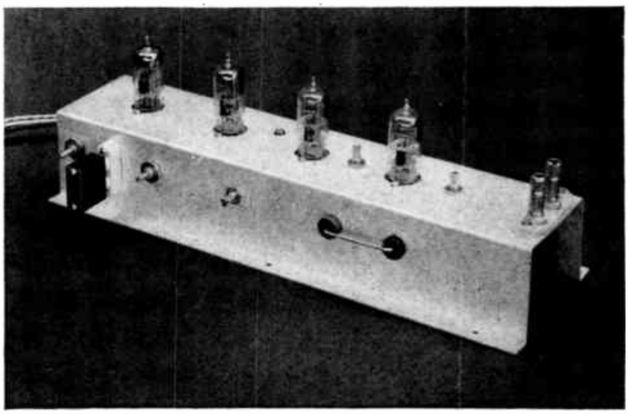


Fig. 13-17 — Front view of the 220-Mc. transmitter-exciter. Across the front of the chassis are the oscillator plate-coil adjustment, crystal, multiplier-coil adjustment, first-tripler plate condenser, and tip jacks for final cathode metering. Second-tripler and final plate condensers are mounted on the top portion of the chassis. Output terminals are at the far right.

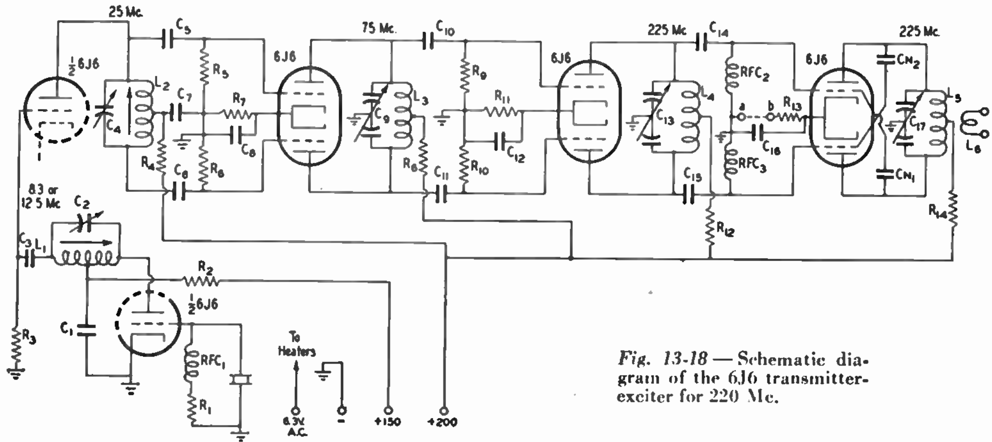


Fig. 13-18 — Schematic diagram of the 6J6 transmitter-exciter for 220 Mc.

C_1, C_7 — 680-μμfd. mica.
C_2, C_4 — 3-30-μμfd. mica trimmer.
C_3 — 68-μμfd. mica.
C_5, C_6 — 47-μμfd. mica.
C_8, C_{12} — 330-μμfd. mica.
C_9, C_{13} — 2.7-8.5-μμfd. midget butterfly variable (Johnson 160-208).
C_{10}, C_{11}, C_{14}, C_{15} — 50-μμfd. ceramic (National XLA-C).
C_{16} — 200-μμfd. ceramic.
C_{17} — 1.7-3.3-μμfd. midget butterfly variable (Johnson 160-203).
C_{N1}, C_{N2} — Neutralizing capacitors made of 75-ohm Twin-Lead; see text.
R_1, R_3 — 6800 ohms, ½ watt.
R_2 — 470 ohms, ½ watt.
R_4 — 3900 ohms, 1 watt.
R_5, R_6, R_9, R_{10} — 22,000 ohms, ½ watt.
R_7, R_{11}, R_{13} — 470 ohms, 1 watt.
R_8, R_{12}, R_{14} — 1500 ohms, 1 watt.
L_1 — 34 turns No. 28 d.s.c., close-wound on National XR-50 slug-tuned form, center-tapped.
L_2 — 12 turns No. 24 d.s.c., close-wound on National XR-50 slug-tuned form, center-tapped.
L_3 — 7 turns No. 16 enamel, ⅝-inch inside diameter, spaced wire diameter, center-tapped.
L_4 — 2 turns No. 16 enamel, ⅜-inch inside diameter, spaced ¼ inch, center-tapped.
L_5 — 1½ turns No. 12 enamel, ¾-inch inside diameter, center-tapped. Space turns about 3/16 inch apart. Coil 1½ inches long over-all. See bottom-view photograph.
L_6 — Hairpin loop No. 16 enamel inserted between turns of L_5.
RFC_1 — 250-μhy. r.f. choke (Millen 34300).
RFC_2, RFC_3 — Solenoid v.h.f. choke — No. 28 d.s.c. wire wound on ½-watt carbon resistor, ⅛-inch diameter, 5/16 inch long.

Starting with sections about two inches long, they should be trimmed a small amount at a time until tuning the final plate through resonance (with plate voltage removed) causes no downward kick in grid current.

Performance

With the voltages shown, the output on 220 Mc. will be about 2 watts, as indicated by a full-brilliance indication in a Number 46 (blue bead) pilot lamp. More output can be obtained by increasing the voltage above 200, but the increase is seldom worth the extra strain on the tubes. Operated as shown, the rig will give ample output to drive an 832 amplifier which will deliver about 12 watts, or the final 6J6 may be modulated and the unit operated as a complete low-powered transmitter.

The same general arrangement described above may be used to get to 220 Mc. with three tubes instead of four, if the regenerative harmonic-oscillator circuit shown in Fig. 13-12 is used to replace the more conventional crystal oscillator circuit of Fig. 13-18. An 8.3-Mc. crystal is then made to oscillate on 25 Mc. in the first 6J6 section. The second section triples to 75 Mc. The rest of the unit, from L_3 on, is the same as in Fig. 13-18. It is suggested that the description of the 6- and 2-meter transmitter of Fig. 13-12 be studied carefully before this substitution is attempted.

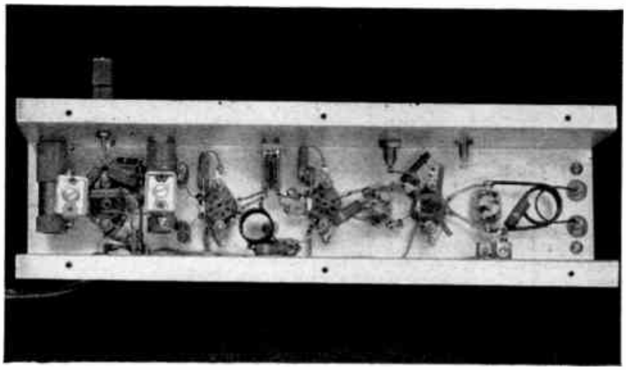


Fig. 13-19 — Bottom view of the 6J6 220-Mc. rig, showing the simplicity of the layout.

Simple Gear for 144 and 220 Mc.

Until recently, most stations operating in the higher v.h.f. bands employed simple transmitters of the modulated-oscillator type. Since the superregenerative receiver was also widely used, the instability of the transmitters was not a matter of great importance; but with the rapid swing to stabilized transmitters and selective receivers now in evidence, most of the modulated-oscillator signals are no longer readable. It is, however, still possible, by careful design and proper operation, to use the simple and economical oscillator rig and yet radiate a signal that can be copied on all but the most selective receivers. Two such transmitters, for 144 and 220 Mc., are shown in Figs. 13-20 through 13-27.

Oscillator Ills and Their Treatment

There are two principal faults in most simple oscillator-type transmitters. Many use filament tubes with a.c. applied to the filaments, causing severe hum modulation. Others, through poor design, have insufficient feedback (as evidenced by low grid current) so that they are unable to sustain strong oscillation under load. Lack of sufficient excitation also renders them incapable of maintaining oscillation at low plate voltages, causing them to go out of oscillation over a considerable portion of the modulation cycle. Such oscillators suffer from extreme frequency modulation, making their signals unreadable on all but the very broadest receivers, and even on these the quality is poor indeed.

A 2-METER UNITY-COUPLED OSCILLATOR

No simple transmitter can hope to overcome these faults entirely, but they are materially reduced in the rig described herewith. A.c. hum modulation is reduced through the use of indirectly-heated tubes; and stability is improved through the use of a high-*C* push-pull oscillator, employing the familiar "unity-coupled" circuit. This arrangement, wherein the grid coil is fed through the inside of a plate tank made of copper tubing, provides adequate excitation. Stability over wide ranges of plate voltage is quite good, and the degree of frequency modulation is not too severe if the modulation is held to 75 per cent or less. It is laid out so that it is stable mechanically, reducing possible frequency changes from vibration.

Mechanical Details

The transmitter is designed for use with a plate supply of 250 to 300 volts, making it useful for mobile or low-powered home-station

Fig. 13-20 — Front view of the simple 144-Mc. transmitter. The jacks at each side of the antenna terminals are for insertion of a meter in the oscillator grid (left) and plate (right) circuits. The microphone jack is at the lower left and the on-off switch is at the right. The calibration scale is drawn with India ink on heavy white paper.

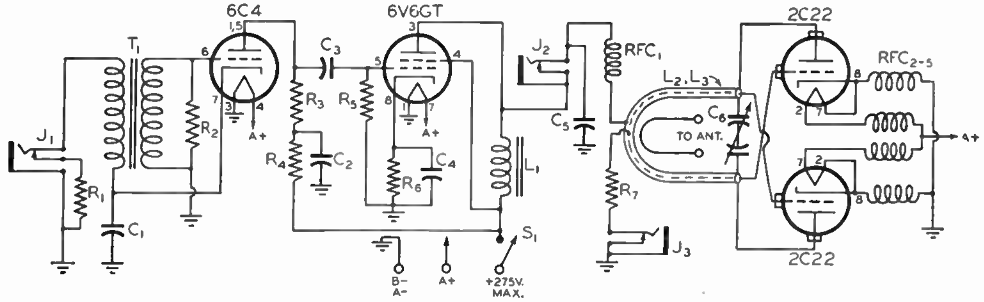


Fig. 13-21 — Schematic diagram of the simple 144-Mc. transmitter.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3, C_5 — 0.01-μfd. 600-volt paper.
C_6 — "Butterfly" variable (Cardwell ER-11-BF/S modified: see text).
R_1 — 470 ohms, 1 watt.
R_2 — 0.33 megohm, ½ watt.
R_3, R_4 — 5000 ohms, 5 watts.
R_5 — 0.47 megohm, ½ watt.
R_6 — 680 ohms, 1 watt.
R_7 — 10,000 ohms, 1 watt.
L_1 — Midget filter choke.
L_2, L_3 — Unity-coupled grid and plate coils. See text and Fig. 13-22.
J_1, J_2, J_3 — Closed-circuit jack.
RFC_1 — No. 28 d.s.c. wire, close-wound on 1-watt resistor, ¼-inch diam., ⅝ inch long.
RFC_2, RFC_3, RFC_4, RFC_5 — 20 turns No. 20 d.s.c. wire close-wound on ¼-inch polystyrene rod.
S_1 — S.p.s.t. toggle switch.
T_1 — Single-button microphone transformer (UTC "Ouncer" — surplus).

operation. It employs a pair of 2C22 tubes (also known as 7193s) as oscillators, a 6V6GT modulator, and a 6C4 as a speech amplifier and source of microphone voltage. It is housed in a standard 5 × 6 × 8-inch utility cabinet, the back and front of which are removable. The schematic diagram is shown in Fig. 13-21.

The plate tank "coil" is made of 3/16-inch copper tubing, bent into a "U" which is two inches long overall. The ends of the "U" are made into spade lugs, as shown in Fig. 13-22, the slotted ends providing a small range of inductance adjustment. The lug ends are fastened directly to two of the stator terminals of the butterfly-type tank condenser, C_6. Part of the "U" is cut out at the curved end, to provide an opening for the center-tap of the grid coil. An easy way to make the grid coil is to cut two pieces of flexible insulated wire, about four inches long, and feed them into the "U" through the center opening. The protruding tap, made by twisting the ends of the wires together, should be coated with household cement after the grid resistor has been soldered to it. Note that the grid leads are transposed. The 2C22s will not oscillate if these are improperly connected. The plate leads may be made of ¼-inch copper braid, or copper or silver ribbon is even better, if available. If braid is used, it may be made solid at the end by flowing solder over the last half inch, after which it may be drilled, to pass the stator terminal screw.

Provision is made for reading both grid and plate current to the oscillator, two meter jacks being mounted on either side of the plate tank. Their terminals make convenient mounting places for R_7 and RFC_1. Note that the jacks are connected so that the meter leads need not be reversed when changing from one jack to the other. The plate-meter jack must, of course, be insulated from the metal panel.

No battery is required for microphone current, this being obtained by running the cathode current of the 6C4 speech amplifier through the microphone transformer. The 6C4 cathode is by-passed with a large electrolytic condenser, and the plate is decoupled and bypassed to reduce hum. Since the 6C4 stage is used principally as a source of microphone current, resistance coupling to the 6V6GT modulator gives adequate drive. No gain control is included, as the full output of the modulator is insufficient for overmodulation.

Testing

Since the grid is the controlling element in the operation of any Class C stage, it is important that the grid current be observed in adjust-

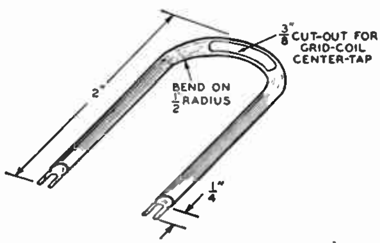


Fig. 13-22 — Detail drawing of the oscillator plate inductance. It is made from 3/16-inch copper tubing, bent into a "U" shape. Ends of the "U" are formed into spade lugs, the slots in which provide a means of slight inductance adjustment. It is mounted directly on the stator terminals of the tuning condenser.

ing the oscillator. The plate current may be almost meaningless, as an indication of the proper functioning of such a stage, but the grid current shows plainly if the oscillator is functioning correctly. If the grid current and bias are normal for the tubes used, the plate current can be ignored, except to see that the input is not excessive. Grid current in this os-

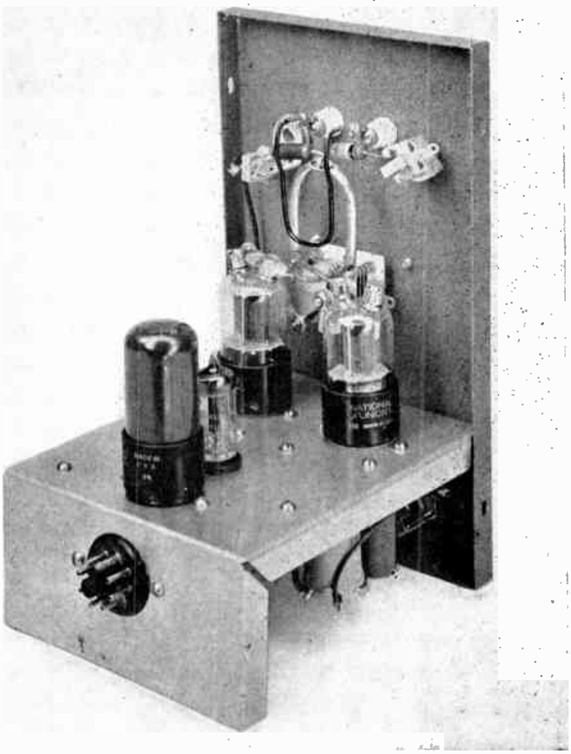


Fig. 13-23 — Back view of the 2-meter transmitter, showing the symmetrical arrangement of components. Note that the "U"-shaped tank inductance is mounted directly on the stator terminals of the butterfly tuning condenser.

cillator should run about 8 ma. when a plate voltage of 275 or so is used and the oscillator is loaded by a lamp or antenna. The "U"-shaped antenna-coupling loop, should be adjusted until the grid current is approximately this value. The plate current will be about 60 ma. with 275 volts on the plates.

The transmitter frequency should be checked with Lecher wires, or by listening to the signal in a calibrated receiver. In either case there should be a load across the antenna terminals, as the frequency may be appreciably different between loaded and unloaded operation.

The rough calibration scale shown was first roughed on a white card using pencil, and afterward drawn over in India ink. The calibration card is glued to the panel, and further held in place by the condenser mounting nut and two small machine screws.

A LINE OSCILLATOR FOR 220 MC.

A line oscillator which is suitable for low-power experimental work is shown in Figs. 13-25, 13-26 and 13-27. It is built entirely of readily-obtainable standard parts, and may be constructed at very low cost. The tube is a 7F8 dual triode, working as a push-pull oscillator, with parallel lines in the plate circuit. The frequency is varied by means of a mica trimmer which is connected across the line near the cold end, so that a vernier effect is attained. A rough adjustment of frequency is made by means of an adjustable shorting bar.

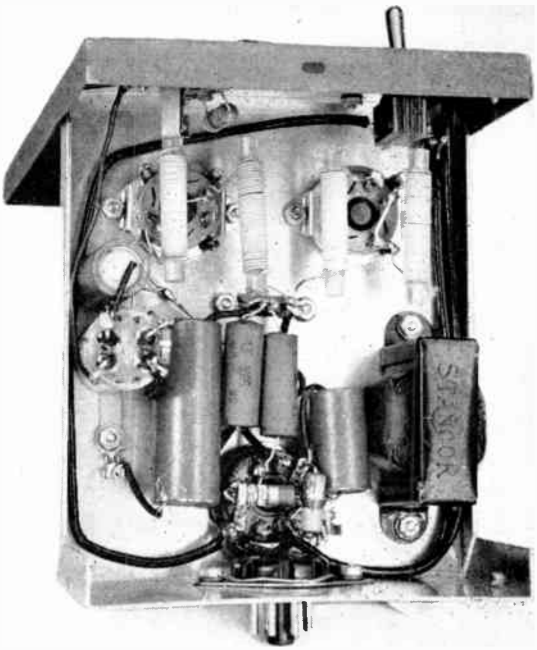


Fig. 13-24 — Under-chassis view shows the four heater chokes and audio components. The small round object, left center, is the microphone transformer, a surplus midget unit. The audio choke is at the right.

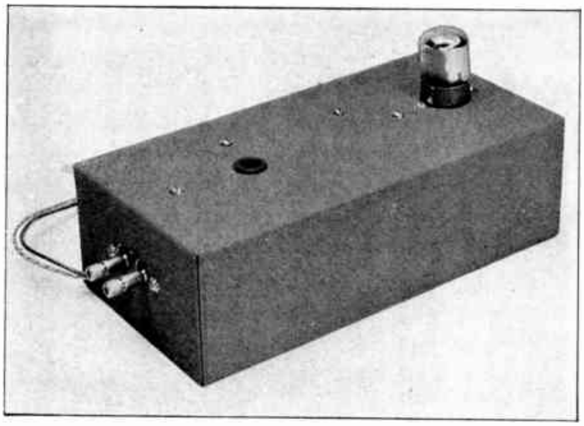


Fig. 13-25 — A one-tube oscillator for 220 Mc. using a 7F8 dual triode. Linear tank circuit and antenna coupling are under the chassis.

When the proper setting of the shorting bar is found, the 220–225-Mc. band will be covered by about two complete turns of the trimmer.

The transmitter is mounted on a 3 × 5 × 10-inch chassis. Only the oscillator tube is above the chassis, with the lines and antenna coupling below. The antenna coupling loop is connected to a National FWG terminal assembly which projects through the end of the chassis. The plate lines are 7½ inches long and made of ¼-inch copper tubing spaced ¾ inch, center to center. They are held in position by two halves of a National FWH or FWJ terminal block. These blocks are of low-loss insulating material, and the hole spacing is right for this application. The connection between the plates and the lines should be made with ¼-inch-wide copper strip. They are mounted on two cone stand-offs 6 inches apart. Self-supporting r.f. chokes, one in the cathode lead and the other in the B-plus lead, a 1000-ohm resistor from grids to ground, and a small by-pass condenser from the hot heater terminal to ground, complete the circuit. The antenna coupling is a "U"-shaped loop 4½ inches long.

The transmitter may be placed in operation by applying 6.3 volts a.c. and about 250 volts d.c. Plate current, under load, should be under 40 ma. A lamp load should be used across the antenna terminals until the frequency is adjusted to within the band limits. The shorting bar is made from two National No. 8 grid clips, which make a tight fit on the ¼-inch tubing, and the trimmer condenser is also connected to the line by means of a pair of these clips, making it possible to adjust the position of the condenser along the line to give the desired degree of frequency coverage. The shorting bar and the trimmer should be set in such positions that, with the trimmer set near maximum, the frequency of oscillation is near 220 Mc.

The antenna coupling may then be adjusted for maximum power transfer (a field-strength meter close to the antenna is a good indication), using the minimum coupling that will give satisfactory output. The frequency should be checked carefully with the antenna on and the coupling adjusted.

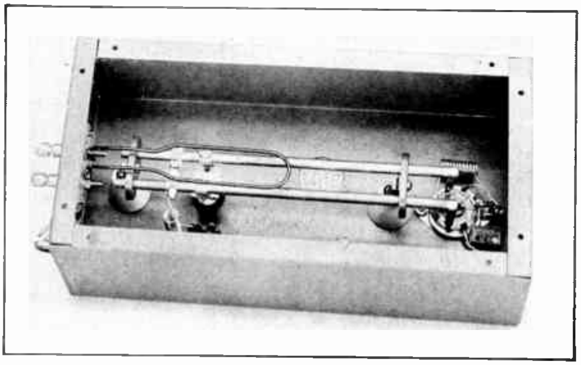


Fig. 13-26 — Under-chassis view of the 220-Mc. transmitter. Note the method of making the shorting bar and mounting the trimmer condenser — both by the use of spring grid clips, permitting adjustment of the position of either along the line.

The transmitter can be run at 10 watts input without endangering the tube. The useful output is in the vicinity of 2 watts. The rig may be modulated with a single 6V6 tube, a suitable modulator being that shown in Fig. 13-21.

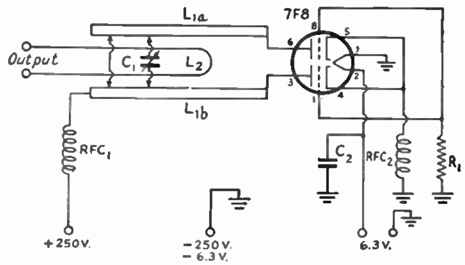


Fig. 13-27 — Schematic of the 220-Mc. transmitter. R.f. chokes are 12 turns No. 18 d.c.c. wire, ¼-inch diam. C_2 is 0.0047 μfd. See text for other values.

Mobile Transmitters

Each mobile installation is, of necessity, something of a custom-built job, since it must be designed to fit the available space. This may be almost any shape or size, and will vary with the make and model of the car in which the equipment is to be used. There are certain principles, however, which experience in amateur mobile work has shown to be common to nearly all mobile applications.

Transmitter location: If at all possible, install the equipment inside the car, preferably in such position that adjustments can be made from the driving position.

Power: The mobile station will be low-powered, at best. The difference between 10 watts and 50 watts input will make only a very slight improvement in signal strength, but the higher power involves many complications.

Choice of tubes: There are low-current equivalents of most of the tubes commonly used for the low-powered stages of amateur transmitters. Study the tube tables carefully — you may be able to save an appreciable amount of drain on the car battery.

Design the equipment so that tubes and other components operate well within ratings. It may not be possible to keep as close a watch over the operation of the equipment as is usually done with fixed-station gear.

Speech equipment: Make provision for enough audio power. Class A modulators are good for low-powered rigs, Class AB in the medium-power range, and Class B for the higher-powered jobs. Include a speech amplifier of some sort, if there is any doubt of the ability of the modulator unit to handle its job without it.

Microphone: Carbon microphones are generally most suitable. They require less speech amplification, and are more rugged than other types. Crystal microphones may melt in the heat of a closed car in summer, and dynamics are seldom able to withstand the rough usage associated with mobile work.

Antenna polarization: Vertical polarization is simplest for mobile, but if activity in the locality in which you expect to work is mainly horizontal it may be desirable to use some sort of horizontally-polarized system. The folded dipole arranged in a circular form, called the "halo," can be used for 50 Mc., and multielement arrays are feasible for the higher-frequency bands.

In most respects, gear designed for mobile operation is similar to that used for home-station service, except for the additional considerations imposed by space and current-drain limitations and the need to withstand constant vibration. Though there are various types of power supplies capable of delivering more power, the most satisfactory arrangement for most mobile installations is the generator or vibrator supply that furnishes 300 volts at 100 to 150 ma. This power is within the capability of the average family-car battery, making unnecessary the separate batteries and special generators usually needed when higher-powered systems are employed. The transmitters described in this section are designed for 300-volt service, though in several instances they may be modified readily for satisfactory use at higher power levels, if the car battery and generator will handle the extra current load imposed.

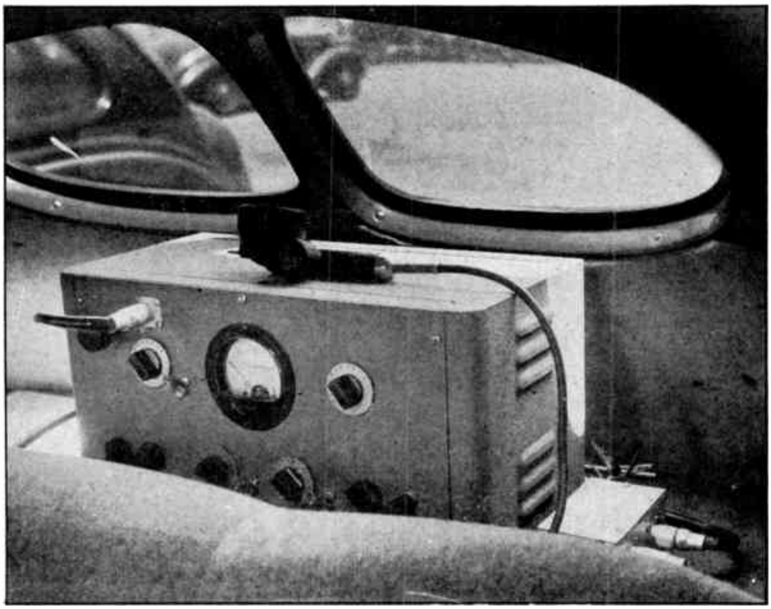


Fig. 13-28 — **A typical installation of a 6- and 10-meter mobile transmitter. The antenna relay is mounted at the right. Genemotor and starting relay are under the hood adjacent to the car battery. Operation is controlled entirely by the push-to-talk switch on the microphone.**

Fig. 13-29 — The plate circuit of the final stage of the mobile transmitter is the only r.f. circuit above the chassis. The three tubes at the left are the driver and audio stages, with the oscillator and multiplier tubes directly in back of the meter. The tube to the right of the modulation transformer is the 0A2 voltage regulator. Chassis size is 7 × 13 × 2 inches.

A Mobile Transmitter for 50 and 28 Mc.

Low over-all battery drain in mobile operation is best obtained through the use of filament-type tubes which are lighted only during transmission periods. The mobile unit for 6, 10 and 11 meters, shown in Figs. 13-28–13-32, employs filament-type beam tetrodes throughout. Five 2E30s are used, as crystal oscillator, frequency multiplier, Class A driver, and push-pull Class AB modulators. The final stage is a 2E25, a tube of somewhat larger design, having its plate connection at the top of the envelope. Two newer types, the 2E24 and 5516, may also be used. Total filament current is only 4.3 amperes, and there is no drain whatever when the rig is not actually on the air.

Mechanical Details

The transmitter is housed in a crackle-finished cabinet which may be mounted in back of the seat in coupé-type vehicles or in the trunk compartment of sedans.

Special attention is paid to ruggedness of construction, all leads being made as short and direct as possible. Small components are supported with terminal strips at each end where possible, and tuning controls are equipped with dial locks (National ODL). The meter (a Marion 0–10-ma. sealed unit) is back-of-panel mounted, with a sheet of Lucite serving as a protecting window. This method of mounting the meter, about ½ inch in back of the panel, also provides a convenient method for illuminating the meter face. Dial lights are mounted at either side of the meter, as shown in Figs. 13-29 and 13-31.

By using 100-$\mu\mu$fd. variable condensers for C_2 and C_3, the range of the oscillator and multiplier plate circuits is extended, so that it is unnecessary to change these coils in changing bands. Only the crystal and the final plate coil, L_5, need be changed. Complete push-to-talk operation is made possible through the use of two relays. Ry_1 starts the genemotor and applies the filament voltage to the transmitter. Ry_2 transfers the antenna from receiver to transmitter. Both are controlled by the switch on the microphone, which may be any single-button type that has a control switch. The Army T-17-B, now currently available as war surplus, is shown with the rig.

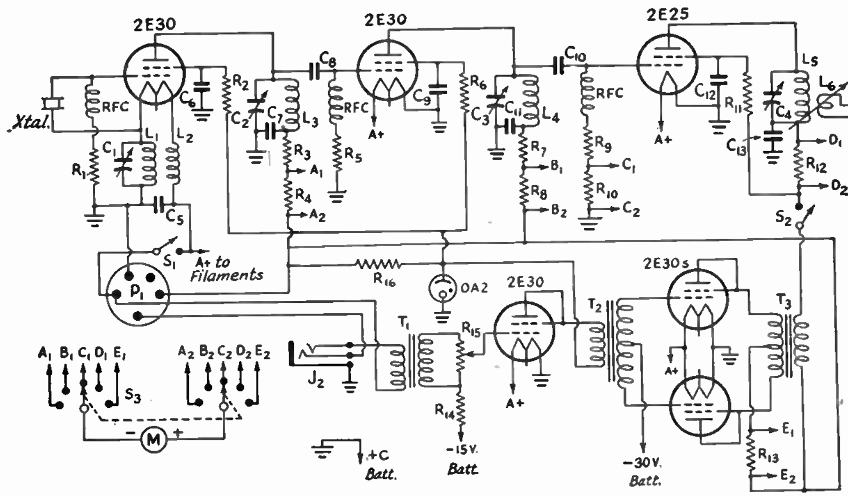
The Circuit

The crystal oscillator is a Tri-tet, modified for filament-type tubes. Interwound coils are inserted in the filament leads, and one of these is tuned. The setting of this adjustment is not critical and may be left near maximum setting, for both 7- and 8.4-Mc. crystals. The oscillator doubles in its plate circuit for both bands.

The stage following the oscillator is operated as a doubler for 27- and 28-Mc. work, and as a tripler for 50 Mc. The 2E30 is an effective frequency multiplier, and there is adequate excitation for the final in either case. Screen voltage on the exciter stages is stabilized with a miniature voltage-regulator tube, an 0A2. With a screen voltage of 150, the plate input to both 2E30s is held to about 6 watts per tube.

The final stage uses a 2E25, whose top-cap plate connection permits the mounting of the plate circuit above the chassis, well isolated from the other tuned circuits. A small shield, cut from an old-style tube shield to a length of about one inch, comes up to the bottom of the 2E25 plate assembly. These precautions are sufficient to provide completely-stable operation without neutralization.

The antenna coupling coil, L_6, is wound on a short length of polystyrene rod ½ inch in diameter, into which is inserted a ¼-inch rod of the same material. This shaft projects through the front panel, where a shaft-locking panel bushing (Bud PB-532 bushing, Millen 10061 shaft lock) holds it in the desired posi-



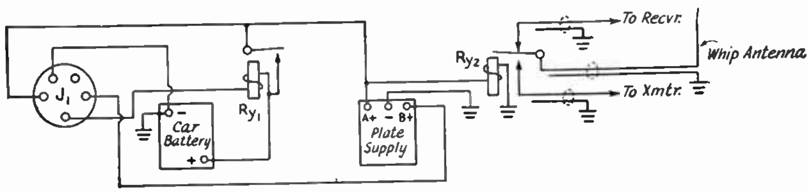


Fig. 13-30 — Wiring diagram of the mobile rig for 6 and 10 meters.

C_1 — 100-μμfd. midget, screwdriver-adjustment type (Hammarlund APC-100).
C_2, C_3 — 100-μμfd. midget, shaft type (Hammarlund HF-100).
C_4 — 15 μμfd., double spaced (Hammarlund HFA-15-E).
C_5 — 0.001-μfd. mica.
C_6, C_7, C_9, C_{11}, C_{12}, C_{13} — 470-μμfd. midget mica.
C_8, C_{10} — 100-μμfd. midget mica.
R_1 — 82,000 ohms, 1 watt.
R_2, R_6 — 1000 ohms, ½ watt.
R_3, R_7, R_{10} — 100 ohms, ½ watt.
R_4, R_8, R_{12}, R_{13} — Special shunts. (See text.)
R_5 — 0.15 megohm, 1 watt.
R_9 — 33,000 ohms, 1 watt.
R_{11}, R_{16} — 5000 ohms, 10 watts.
R_{14} — 10,000 ohms, ½ watt.
R_{15} — 0.5-megohm potentiometer.
L_1, L_2 — 7 turns each, No. 20 d.c.c., ⅝ inch long on 1-inch dia. form, windings interwound.
L_3 — 10 turns No. 12 enam., close-wound on 1-inch diam. form.
L_4 — 6 turns No. 12 enam., ¾ inch long, ½-inch inside diam., self-supporting.
L_5 — 28 Mc.: 10 turns No. 12 enam., 1½ inches long, 1-inch inside diam., self-supporting.
50 Mc.: 5 turns No. 12 enam., 1 inch long, 1-inch inside diam., self-supporting.
L_6 — 3 turns on ½-inch polystyrene rod — see text and detail photo.
J_1 — Socket on power cable, 5-prong.
J_2 — Double-button microphone jack. If T-17-B microphone is used, a special jack designed for this microphone must be obtained.
J_3 — Coaxial fitting (Amphenol 83-1R. Matching plug is 83-1SPN).
M — 0-10-ma. sealed unit (Marion).
P_1 — Power plug on transmitter chassis.
RFC — 2.5-mh. r.f. choke. National R-100.
Ry_1, Ry_2 — See text.
S_1, S_2 — S.p.s.t. snap switch.
S_3 — 2-section 5-position wafer-type switch.
T_1 — Single-button microphone transformer.
T_2 — Driver transformer (Stancor A-4752).
T_3 — Modulation transformer (UTC S-18).

tion. Coupling is adjusted by pushing or pulling the knob affixed to the shaft, following which the bushing may be locked. The bushing may also be set finger-tight, allowing the coupling to be adjusted, yet holding it firmly.

Three 2E30s are used for the modulator, one as a Class A driver and two in push-pull as Class AB modulators. All three are triode-connected. Bias is supplied by a 30-volt hearing-aid battery, which can be tapped at 15 volts by opening up the cardboard case and soldering on a lead at the point where the two 15-volt sections are joined together. This lead is brought out to the unused terminal on the battery socket and plug.

Metering of all circuits is provided by a 10-ma. meter, a 2-section 5-position switch, and a set of shunts. The shunts are made from small 100-ohm resistors, on which is wound about 7 feet of No. 30 enameled wire. The shunts should be wound with an excess of wire, the length of which may be reduced until the multiplication of the meter scale is just right. The resistor R_{10} in the final grid circuit is left without a shunt, giving direct reading on the 10-ma. scale for measuring the final grid current.

Testing

Except for the speech stages, the unit may be tested using 6.3 volts a.c. on the filaments and an a.c. power supply. A storage battery

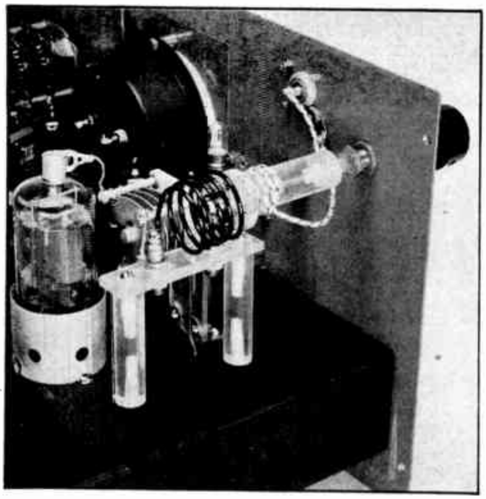


Fig. 13-31 — Detail photo of the 2E25 final stage, showing method of coupling to the antenna. The coupling coil, wound on a polystyrene rod, is adjustable from the front panel. The plate coil is mounted by means of GR plugs.

must be used for filament supply when the speech equipment is to be tested, as a.c. on the filaments will produce excessive hum. Initial testing should be carried on with about 200 volts on the tube plates, after which the potential can be raised to 300 volts.

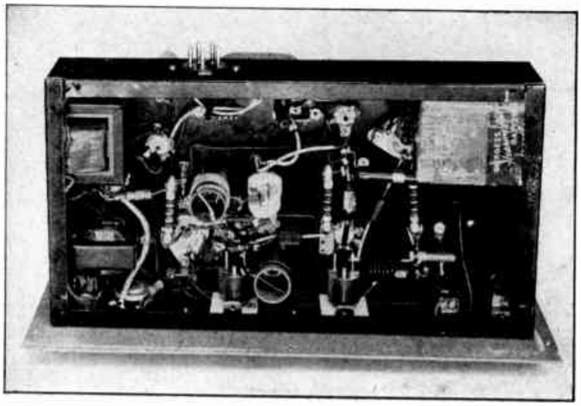
To place the unit in operation, set S_1 to the "on" position, leaving S_2 "off." With the meter switch in position A, apply plate voltage and note the meter reading, which is the oscillator plate current. This will be about 20 ma., dipping slightly at resonance as C_2 is adjusted. Next switch to position B and adjust C_3. The dip here may not be as pronounced as in the oscillator, and the final grid current, position C, 10-ma. scale, is the best indication of resonance in the preceding adjustments. This reading should be about 4 ma., dropping to 3 ma. under load. With S_2 turned on, the final plate current, position D, should drop to below 10 ma. at resonance, and coupling of the antenna should raise it to 50 to 60 ma. Modulator plate current will be about 20 ma., rising to 60 ma. or more on audio peaks. No metering position is provided for the Class A driver current, but this should be approximately 10 ma.

With the coil and condenser values given, it is impossible to get output from the final stage on a wrong frequency, but excitation to the final may be obtained on incorrect harmonics; hence it is advisable to check the frequency of each stage with an absorption wavemeter.

For maximum convenience, the same antenna should be used for both transmission and reception. Antenna change-over is handled with a conventional 6-volt antenna relay which is mounted in a small box made up for the purpose from folded sheet aluminum. Amphenol coaxial fittings, mounted on the sides of the relay box as close to the relay contacts as possible, provide for connection to the transmitter, the receiver, and the antenna by means of coaxial line. The relay case is grounded and only the inner conductor of the coaxial line is switched.

A headlight relay for genemotor starting may be purchased from any auto-accessory store, and this and the genemotor should be mounted as close to the car battery as possible, in order to minimize voltage drop. Battery wiring and filament cables should be as heavy wire as possible, with No. 12 as the minimum for the genemotor leads.

Fig. 13-32— Bottom view of the mobile rig. At the left center are the interwound coil and tuning condenser which are part of the oscillator filament circuit. Audio components are at the left, with oscillator and multiplier plate circuits near the front panel.



Mobile Transmitters for 50 and 144 Mc.

The two small transmitters shown in Fig. 13-33 were designed primarily for use together in mobile service on 50 and 144 Mc., but they may be used as a low-powered two-band home-station, or they may be built and operated separately, if only one of the bands is to be employed. The larger of the two is for 144 Mc., and this unit includes the modulator, though that part of the rig can very well be incorporated in the 50-Mc. unit, if that transmitter is to be used alone. When the two units are connected to a common power source, either one may be used by manipulation of the toggle switches, which apply the heater voltage to the desired circuits.

The r.f. sections are nearly identical, except for the inclusion of a 7F8 tripler stage between the oscillator and the final in the 2-meter unit. Both use Tri-tet oscillators with 6V6GT tubes and fixed-tuned cathode and plate circuits. Harmonic-type crystals are used, 24 to 24.66 Mc. for the 2-meter rig and 25 to 27 Mc. for the 6-meter job, the oscillator doubling in each case. The final stage in both units is an 832 amplifier, the only difference in the circuits being a small amount of neutralization required in the 2-meter rig.

When the two units are used together, 144-Mc. operation requires that switches S_1 and S_2 (Fig. 13-34) be closed, and S_1 in the 50-Mc. unit, Fig. 13-35, left open. For 50-Mc. operation, S_2 is opened, cutting off the r.f. heaters in the 144-Mc. unit, and S_1 in both units is closed. The terminal strips on the backs of the two units are connected in parallel, applying the plate voltages to both at all times, and the heaters of the desired circuits are energized by means of the toggle switches. Switching of the plate voltage is not necessary.

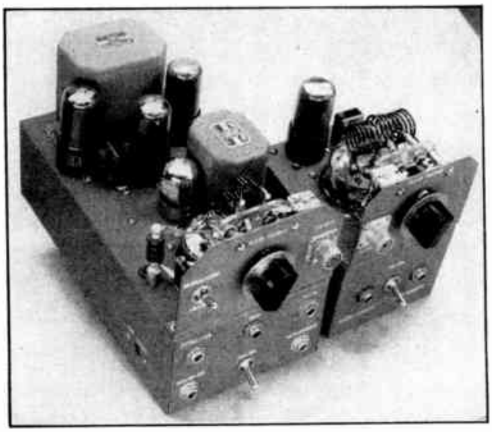


Fig. 13-33 — **A 2-band set-up for mobile or low-powered fixed-station operation on 50 and 144 Mc. At the left is the 2-meter unit, complete with modulator. The smaller is the 50-Mc. r.f. section. Toggle switches permit use of the modulator with either r.f. section.**

THE 144-MC. SECTION

The 144-Mc. unit, Figs. 13-34 and 13-36, includes the modulator and is designed to operate at about 15 watts input with a 300-volt power supply. Meter jacks are provided for measuring the cathode currents of all stages and the grid current of the final. The plate circuits of the oscillator and tripler stages are self-resonant, and are inductively coupled to their following grid circuits.

A small amount of neutralization was required to assure completely-stable operation of the final. The neutralizing condensers, C_{11} and C_{12} in the circuit diagram, are pieces of No. 12 wire extending from the grid of one section of the 832A to the vicinity of the plate of the other section. The wires are crossed at the bottom of the tube socket and go through Millen 32150 bushings mounted in the chassis between the 7F8 and the 832A sockets. It is possible that use of a shielded tube socket would eliminate the tendency toward oscillation in the 832A.

A series-tuned antenna circuit, consisting of C_4 and L_7, is intended for use with any of the low-impedance antenna feed systems commonly used for mobile work. The amount of loading is adjusted by varying the position of the pick-up link, L_7.

The modulator employs a pair of 6V6 or 6V6GT tubes working Class AB. A speech-amplifier stage is not required so long as a single-button carbon microphone is used. Voltage for the microphone is taken from the junction of the two cathode-biasing resistors, R_7 and R_8, thus eliminating the need for a microphone battery.

The microphone and modulation transformers used are both large and expensive for the job at hand and were used only because they happened to be available. The microphone transformer can be any single-button-microphone-to-push-pull-grids transformer and the modulation transformer need not be rated at more than 10 watts. It should be capable of matching a pair of 6V6 tubes to an r.f. load of 5000 to 7000 ohms, depending upon the input at which the 832A is operated.

The photographs of the transmitter show how the parts are mounted on a metal chassis measuring 3 × 5 × 10 inches. The front panel measures 3 × 5 inches and has a ½-inch lip for fastening to the chassis. The construction of the antenna assembly and the method of mounting the components on the panel are identical to the 50-Mc. transmitter. A recommended system of mounting the 832A tube socket is also detailed in the text referring to the 50-Mc. unit.

No special care need be given to the wiring of the audio circuit, but the r.f. leads should be kept as short as possible. The use of four

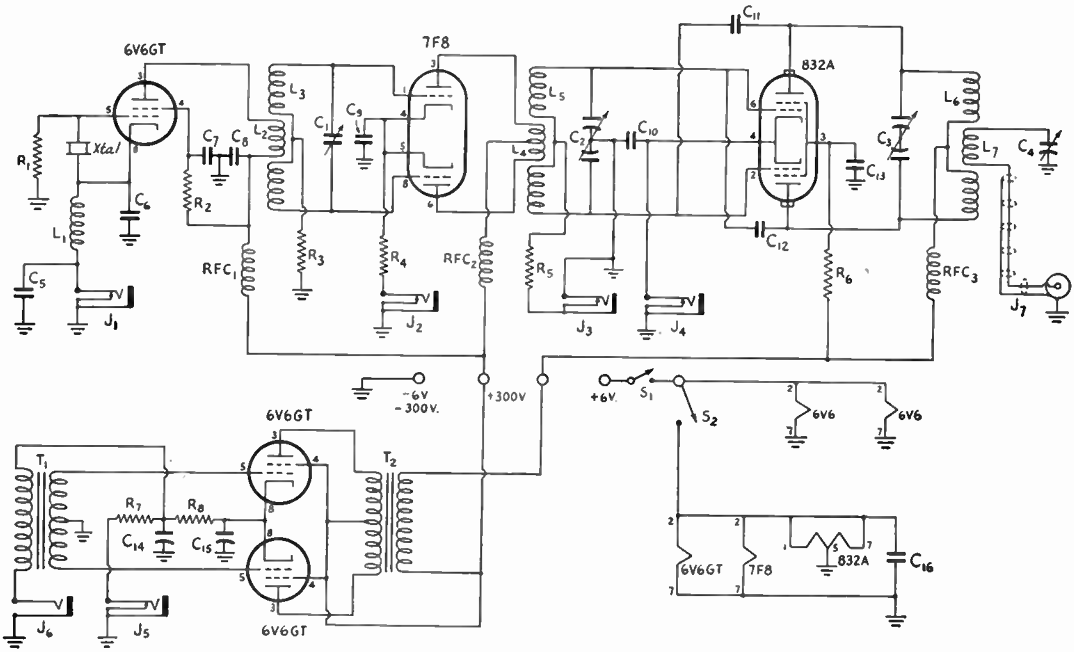


Fig. 13-34 — Circuit diagram of the 144-Mc. r.f. section and modulator

C_1, C_4 — 3–30-μμfd. mica trimmer.
C_2 — 15-μμfd.-per-section split stator (Bud LC-1660).
C_3 — "Butterfly" condenser, 6 μμfd. per section (Cardwell ER-6-BF/S).
C_5, C_7, C_8 — 0.0047-μfd. mica.
C_6 — 100-μμfd. midget mica.
C_9, C_{10}, C_{13}, C_{16} — 470-μμfd. midget mica.
C_{11}, C_{12} — Neutralizing wires. (See text.)
C_{14}, C_{15} — 10-μfd. 25-volt electrolytic.
R_1 — 0.1 megohm, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 33,000 ohms, ½ watt.
R_4 — 470 ohms, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 25,000 ohms, 10 watts.
R_7 — 100 ohms, 1 watt.
R_8 — 150 ohms, 1 watt.
L_1 — 3 turns No. 18 enam., close-wound, ½-inch diam.
L_2 — 4 turns No. 18 enam., ⅜ inch long.
L_3 — 10 turns No. 18 enam.; coil wound in two sections with 5 turns each side of L_2, each section ⅜ inch long. A ½ inch is left between windings. Form for L_2L_3 is a Millen 30003 Quartz-Q standoff insulator, ¾-inch diam.
L_4 — 3 turns No. 18 enam., ½ inch long, 9/16-inch diameter.
L_5 — 2 turns No. 18 enam., interwound with turns of L_4. L_4 and L_5 are wound on a National PRE-3 coil form.
L_6 — 4 turns No. 12 enam., ½-inch i.d., wound in two sections with 2 turns each side of center-tap and a ½-inch space at the center, turns spaced wire diameter.
L_7 — 3 turns No. 12 enam., ½-inch diam., turns spaced wire diameter.
J_1–J_5 — Closed-circuit jack.
J_6 — Open-circuit jack.
J_7 — Coaxial-cable connector.
RFC_1, RFC_2 — 300-μh. r.f. choke (Millen 34300).
RFC_3 — 2.5-mh. r.f. choke (Millen 34102).
S_1, S_2 — S.p.s.t. toggle.
T_1 — Single-button microphone transformer (UTC S-7).
T_2 — Modulation transformer (UTC S-19).

tie-point strips will simplify the mounting and wiring of parts. A single tie point is mounted to the rear of the oscillator tube socket and is used as the junction of R_7, R_8, C_{14} and the primary lead of the microphone transformer. A double tie-point strip is mounted to the right of the crystal socket (as seen in Fig. 13-36), one lug is used as the connecting point for the positive high-voltage lead and the bottom ends of RFC_1 and RFC_2; the bottom of L_1 and the top ends of C_5 and J_1 are connected to the second terminal. The cathode end of L_1 is connected to the cathode side of the crystal socket. The third tie-point strip is mounted on the 832A tube socket and serves as the connecting point between R_4 and J_2; the bottom end of R_6 connects to the high-voltage lead at the second lug. The fourth strip (single lug) is mounted on the frame of C_2 and the leads between R_5 and J_3 join at this point.

The construction of the driver-stage coils is not difficult if the coil forms are properly prepared in advance. A study of Fig. 13-36 will show how the windings are placed on the forms, and the lengths of the windings are given in the parts list. The forms should be marked and drilled to accommodate the windings with the holes for the ends of the windings passing directly through the forms. L_3 should be wound in two sections with the inside ends being soldered together after the winding of L_2 has been completed. The center-taps for L_4 and L_5 are made by cleaning and twisting the wire at the center of each winding. Condenser C_1 is soldered across the grid ends of L_3 before the coil is connected to the tube socket.

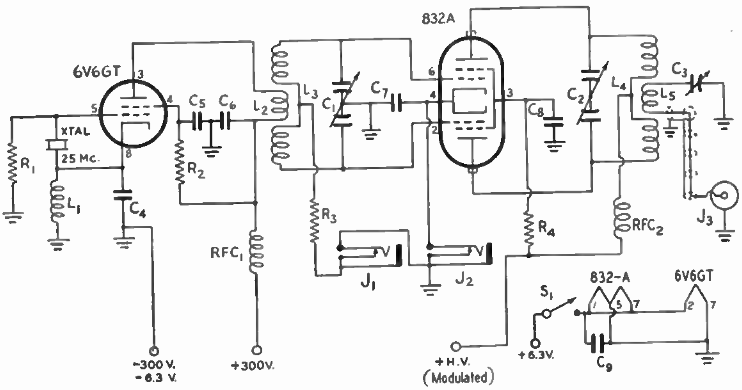


Fig. 13-35 — Circuit diagram of the 6-meter r.f. section.

C_1 — 15-μμfd. per section (Bud LC-1660).
C_2 — "Butterfly" condenser, 15 μμfd. total (Cardwell ER-15-BF/S).
C_3 — 3–30-μμfd. mica trimmer.
C_4 — 100-μμfd. midget mica.
C_5, C_6 — 0.0047-μfd. mica.
C_7, C_9 — 470-μμfd. midget mica.
C_8 — 0.001-μfd. mica.
R_1 — 0.12 meghom, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 22,000 ohms, ½ watt.
R_4 — 25,000 ohms, 10 watts.
L_1 — 3 turns No. 18 enameled wire, close-wound, ½-inch diam.
L_2 — 5 turns.
L_3 — 9 turns, 4½ each side of center, with a ⅞-inch space between sections.
L_4 — 10 turns, 5 each side of center, with a ¾-inch space between sections.
L_5 — 3 turns. L_2 through L_5 have an inside diameter of ¾ inch; No. 12 enameled wire, turns spaced wire diameter.
J_1, J_2 — Midget closed-circuit jack.
J_3 — Coaxial-cable connector.
RFC_1 — 10-μh. r.f. choke (Millen 34300).
RFC_2 — 2.5-mh. r.f. choke (Millen 34102).
S_1 — S.p.s.t. toggle switch.

Adjustment and Testing

When testing the transmitter, it is advisable to start with the high voltage applied to the first two stages only. With a 100-ma. meter plugged in J_1 the oscillator cathode current at resonance should be approximately 30 ma. A low-range milliammeter should now be plugged in J_3 and the final grid circuit should be brought into resonance by adjustment of C_2. Proper operation of the tripler stage will be indicated by a cathode current of approximately 20 ma. and a final-amplifier grid current of 2.5 to 3 ma. The tripler grid condenser, C_1, should be retuned after the amplifier grid circuit has been peaked, to assure maximum overall operating efficiency.

The amplifier should be tested for neutralizing requirements after adequate grid drive has been obtained. If a well-shielded tube socket has been used, it is possible that the amplifier grid current will not be affected by tuning the 832A plate circuit through resonance. However, if the grid current does kick down as the plate circuit is tuned, it will be necessary to add the neutralizing wires referred to in the text and parts list as C_{11} and C_{12}. After installation these wires should be adjusted until no kick in grid current is seen as the 832-A plate circuit is tuned through resonance.

Plate and screen voltages can now be applied to the 832A and the plate circuit tuned to resonance, as indicated by a dip in the cathode current to 40 ma. or less. Then a dummy load (a 15-watt light bulb will do) is connected to the antenna jack and the loading adjusted by varying the position of L_7 and the capacitance of C_4, to cause a cathode current of 60 to 70 ma. Approximately 10 ma. of the total cathode current will be drawn by the screen of the 832A and this value should be subtracted from the cathode current in determining the plate input. Amplifier grid-current should be 1.5 to 2 ma. under load.

Modulator cathode current should be 75 ma.; 85 ma. with modulation. The reading will decrease slightly when the microphone is plugged into the circuit. This is caused by the parallel current path that exists when the microphone circuit is completed.

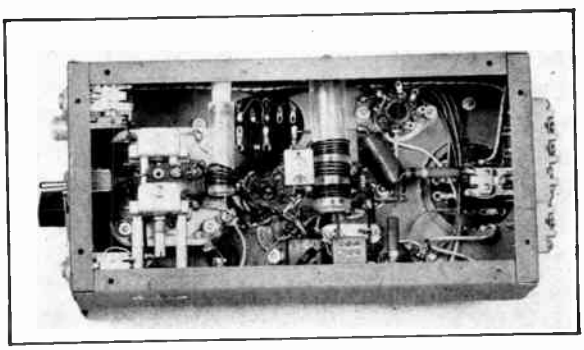


Fig. 13-36 — Bottom view of the 144-Mc. transmitter. The coil forms for L_2L_3 and L_4L_5 are mounted on the side wall of the chassis; the form for L_4L_5 is mounted on a small stand-off insulator so that the windings can be brought out to the center line of the chassis. C_1, the grid condenser for the frequency multiplier, is soldered across the grid ends of L_3. The amplifier grid tuning condenser, C_2, is mounted on metal pillars having a length of 1⅝ inches.

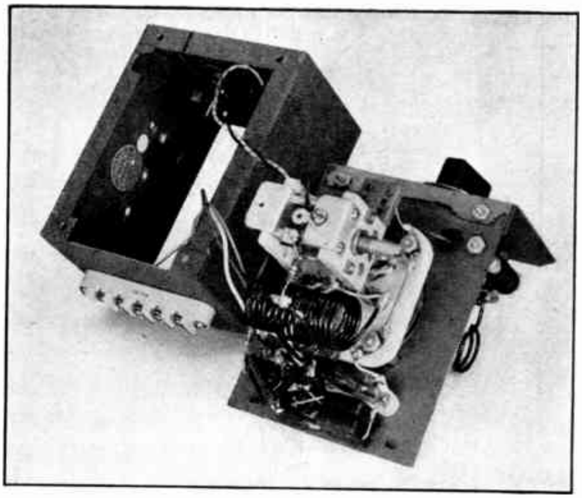


Fig. 13-37 — Bottom view of the mobile transmitter, showing all major components attached to the top plate.

THE 50-MC. PORTION

The 50-Mc. unit, shown in Figs. 13-35, and 13-37, is very similar to the 144-Mc. portion, but for the elimination of the tripler stage. Because of the somewhat lighter load on the power supply, slightly higher power can be run on 50 Mc. In addition, the amplifier operates more efficiently at the lower frequency, permitting inputs up to 30 watts or so if the power is available. Neutralization is not generally required in 50-Mc. operation, but this may not hold true for all physical layouts.

Jacks are provided for measuring the grid and cathode currents of the final stage, and the cathode jack may be used for keying, if c.w. operation is desired. Interstage and antenna coupling circuits are similar to the 144-Mc. section.

The photographs show how a metal box measuring 3 × 4 × 5 inches serves as the chassis for the transmitter. The bottom plate of the box is removed and used as a panel, and is held in place by the screws and nuts that hold the top cover and the box together. In Fig. 13-33 the condenser, C_2, and the antenna jack may be seen mounted on the panel. Metal pillars, ¼ inch long, are used to space the condenser away from the panel. A National FWB polystyrene insulator is used as a mounting support for the antenna coil, L_5, and the insulator is mounted on ¾-inch metal posts. C_3 is supported by its own mounting tabs, and is connected between one end of the pick-up link and ground.

The rear and bottom views of the transmitter show how the rest of the components are laid out on the top plate of the metal box. This plate should be removed from the box while the construction and wiring are being carried on. All of the wiring, with the exception of the d.c. leads to the metering jacks and the input terminals, can be completed in convenient fashion before the top plate is attached to the metal box.

The socket for the amplifier tube is centered on the chassis plate at a point 2⅜ inches in from the front edge, and is mounted below the plate on metal pillars ⅝ inch long. A clearance hole for the 832A, 2¼ inches in diameter, is directly above the tube socket. Sockets for the oscillator tube and the crystal are mounted toward the rear of the chassis.

The oscillator coil, L_2, is mounted on the 6V6 socket; the spare pin, No. 6, of the socket being used as the tie point for the cold end of the plate coil and the other connections that must be made at this part of the circuit. The oscillator cathode coil is mounted between the cathode pin of the 6V6 and a soldering lug placed under the mounting screw of the crystal socket C_5 and C_6 can be seen to the rear of the crystal socket, and RFC_1 is mounted between the tube socket and a bakelite tie-point strip located at the left of the chassis.

The method employed to assure good r.f. grounding of the amplifier components is visible in Fig. 13-37. Soldering lugs are placed beneath the mounting nuts of the 832A socket, and these lugs are joined together with a No. 12 lead which, in turn, is carried on to the common ground point for the oscillator circuit. The filament, cathode, and screen by-pass condensers for the amplifier are all returned to the common ground. These three condensers, C_7, C_8 and C_9, all rest on the 832A tube socket.

The amplifier grid coil, L_3, is self-supporting, with the ends connected to the grid pins of the 832A socket. The tuning condenser, C_1, is actually supported on metal pillars at the right-hand side of the metal box, but the condenser can be wired in place if the operation is carried out in the proper order. First, mount the chassis plate on the box and locate the proper place for the condenser. Next, determine the length of the leads to connect the condenser to the tube socket, and then remove the chassis from the case. The condenser may now be wired into the circuit, and the rigid mounting of C_1, by means of metal posts 1¼ inches long, can be done during the final assembly of the unit.

The grid leak, R_3, is connected between the center-tap of L_3 and a tie-point strip that is mounted on the condenser frame. RFC_2 is mounted toward the front of the chassis, and the grommet-fitted hole to the left of the choke (Fig. 13-37) carries the lead between the plate-voltage terminal and the choke.

The metering jacks and the power terminal strip may now be mounted on the front and rear walls of the metal box. Holes to permit mounting and adjustment of C_1 should also be drilled at this time. Portions of top flanges of the metal case must be cut away in order to provide clearance for the oscillator section and the mounting nut for the amplifier plate choke. After the case, chassis and panel have been fastened together, the wiring of the amplifier plate circuit may be completed.

Test Procedure

A power supply capable of delivering 300 volts at 100 ma. and 6.3 volts at 2 amp. may be used for testing the transmitter. The high voltage should not be applied to the 832A plates until the oscillator has been checked. For initial tests the input voltage can be reduced to approximately 150 volts while the circuits are checked for resonance and proper operation. Squeezing or spreading the turns of the coils should bring the circuits into resonance, as indicated by maximum grid current to the 832A. The grid current should fall to zero, and the plate current of the oscillator tube should rise considerably when the crystal is removed from the socket.

The amplifier plate and screen voltage can be applied at this point. The unloaded cathode current of the amplifier should be about 15 ma., rising to a maximum of 75 or 80 ma. under load, which may be a 15-watt light bulb connected to the antenna jack. C_3 should be adjusted along with the coupling between L_4 and L_5 until maximum output is obtained. The correct degree of loading has been obtained when the plate current at resonance is 10 to 15 ma. below the off-resonance value. The plate tuning condenser, C_2, should be reset each time that a loading adjustment is made.

A final check of voltages and currents should show the following: oscillator and amplifier plate, 300 volts; oscillator screen, 200 volts; amplifier screen, 150 volts; amplifier bias (read at the grid-coil center-tap with a high-resistance voltmeter), 65 volts, negative.

The oscillator plate current should be 28 to 30 ma. and amplifier grid current should be about 3 ma. Under load, the amplifier cathode current should be approximately 60 ma. with 8 or 10 ma. of this amount being drawn by the 832A screen.

Modulation can be supplied by the audio system used in the 2-meter rig shown in Fig. 13-34, or a similar unit may be added, if only 50-Mc. operation is desired.

Transceivers

The transceiver is a combination transmitter-receiver in which, by suitable switching of d.c. and audio circuits, the same tube and r.f. circuit functions either as a modulated transmitting oscillator or as a superregenerative detector. This makes for extreme compactness and light weight, making the transceiver popular for hand-carried portable equipment. It is a compromise with respect to other features, however. The transceiver can be a source of serious interference, and its efficiency is not equal to that of other types of gear wherein separate tubes and circuits are used for transmission and reception.

As a matter of good amateur practice the use of transceivers should be confined to very low-power operation — as in "walkie-talkie" or "handie-talkie" equipment — in the 144-Mc. band, and to experimental low-power operation in the higher-frequency bands. The use of transceivers should be avoided entirely for regular operation on the 144-Mc. band.

V.H.F. Antennas

While the basic principles of antenna operation are essentially the same for all frequencies, certain factors peculiar to v.h.f. work call for changes in antenna technique for the frequencies above 50 megacycles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in nearly every location, and experimentation with various types of arrays is an important part of the program of most progressive amateurs. The importance of high-gain antennas in v.h.f. work cannot be overemphasized. A good antenna system is often the sole difference between routine operation and outstanding success in this field. By no other means can so large a return be obtained from a small investment as results from the erection of a good directional array.

Design Factors

Beginning with the 50-Mc band, the frequency range over which antenna arrays should operate effectively is often wider in percentage than that required of lower-frequency systems; thus greater attention must be paid to designing arrays for maximum frequency response, possibly to the extent of sacrificing other factors such as high front-to-back ratio.

As the frequency of operation is increased, losses in the transmission line rise sharply; hence it becomes more important that the line be matched to the antenna system correctly. Because any v.h.f. transmission line is long, in terms of wavelength, it is often more effective to use a high-gain array at relatively low height, rather than to employ a low-gain system at great height above ground, particularly if the antenna location is not completely shielded by heavy foliage, buildings, or other obstructions in the *immediate* vicinity.

This concept is in direct contrast to early notions of what was most desirable in a v.h.f. antenna system. An appreciable clearance above surrounding terrain is desirable, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained by many v.h.f. workers, especially on 50 and 144 Mc., with antennas not more than 25 to 40 feet above ground. DX can be worked on 50 Mc. with arrays as low as a half-wave above the ground level.

Polarization

Practically all the early work on frequencies above 30 Mc. was done with vertical antennas, probably because of the somewhat stronger field in the immediate vicinity of a vertical system. When v.h.f. work was confined to almost pure line-of-sight distances, the vertical dipole produced a stronger signal at the edge of the working range than did the same antenna turned over to a horizontal position. With the advent of high-gain antennas and extended operating ranges, horizontal systems began to assume importance in v.h.f. work, especially in parts of the country where a considerable degree of activity had not already been established with verticals.

Numerous tests have shown that there is very little difference in the effective working range with either polarization, *if* the most effective element arrangements are used and the same polarization is employed at both ends of the path. Vertical polarization still has its adherents among 50-Mc. enthusiasts and much fine work has been done with vertical antennas, but an effective horizontal array is somewhat easier to build and rotate. Simple 2-, 3- or 4-element horizontal arrays have proven extremely effective in 50-Mc. work, and the postwar era has seen an increase in the use of such arrays which has amounted to standardization on horizontal polarization.

The picture is somewhat different when one goes to 144 Mc. and higher. At these frequencies, the most effective vertical systems (those having two or more half-wave elements, vertically stacked) are more easily erected than on 50 Mc. Important, in considering the polarization question, is the existence of countless 144-Mc. mobile stations whose antenna systems must, of necessity, be vertical. While horizontal polarization will undoubtedly find increased favor at 144 Mc. and higher, particularly for point-to-point work in rural areas, it is probable that vertical polarization will continue in use for some years to come, particularly in areas where activity has been established with vertical systems. Under certain conditions, notably a station directly in the shadow of a hill, there may be a considerable degree of polarization shift, but ordinarily it may be assumed that best results in 144-Mc. work will be obtained by matching the antenna polarization of the stations one desires to contact.

Impedance Matching

Because line losses tend to be much higher in v.h.f. antenna systems, it becomes increasingly important that feedlines be made as nearly "flat" as possible. Transmission lines commonly used in v.h.f. work include the open-wire line of 500 to 600 ohms impedance, usually spaced about two inches; the polyethylene-insulated flexible lines, available in impedances of 300, 150, 100 and 72 ohms; and coaxial lines of 50 to 90 ohms impedance. These may be matched to dipole or multi-element antennas by any of several arrangements detailed below.

The "J"

Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave matching section, as shown at A, Fig. 14-1. The spacing between the two sides of the matching section should be two inches or less, and the point of attachment of the feedline will depend on the impedance of the line used. The feeder should be slid along the matching section until the point is found that gives the best operation. The bottom of the matching section may be grounded for lightning protection. A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 14-1. The "J" is also useful in mobile applications.

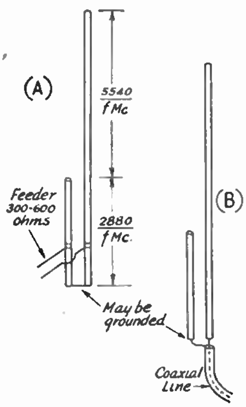


Fig. 14-1 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

The Delta or "Y"-Match

Probably the simplest arrangement for feeding a dipole or parasitic array is the familiar delta, or "Y"-match, in which the feeder system is fanned out and attached to the radiator at a point where the impedance along the element is the same as that of the line used. Information on figuring the dimensions of the delta may be found in Chapter Ten. Chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multi-element array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The "Q" Section

An effective arrangement for matching an open-wire line to a dipole, or to the driven element in a 2- or 3-element array having wide (0.25 wavelength or greater) spacing, is the "Q" section (Chapter Ten). This consists of a quarter-wave line, usually of ½-inch or larger tubing, the spacing of which is determined by the impedance at the center of the array. The parallel-pipe "Q" section is not practical for matching multielement arrays to lines of lower impedances than about 600 ohms, nor can it be used effectively with close-spaced parasitic arrays. The impedance of the "Q" section required in these cases is lower than can be obtained with parallel sections of tubing of practical dimensions. A quarter-wave section of coaxial or other low-impedance line is a commonly-used means of matching a line of 300 to 600 ohms impedance to the low center impedance of a 3- or 4-element array. The length of such a line will depend on the velocity of propagation (propagation factor) of the line used. The propagation factors of all the commonly-used lines are given in table form in Chapter Ten.

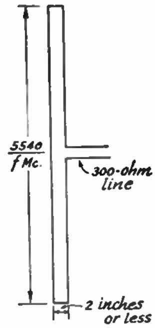


Fig. 14-2 — Details of the folded dipole.

In some installations it may be more convenient to use a line of greater length than a single quarter wave for matching purposes, in which case any odd multiple of a quarter wavelength may be used. The exact length required may be determined experimentally by shorting one end of the line and coupling it to a source of r.f., and trimming the line length until maximum loading is obtained at the center frequency of the operating range.

The "T"-Match

The principal disadvantages of the delta system can be overcome through the use of the arrangement shown in Figs. 14-5 and 14-13, commonly called the "T"-match. It has the advantage of providing a means of adjustment (by sliding the clips along the parallel conductors), yet the radiation from the matching arrangement is lower than with the delta, and its rigid construction is more suitable for rotatable arrays. It may be used with coaxial lines of any impedance, or with the various other forms of transmission lines up to 300 ohms. The position of the clips should, of course, be adjusted for maximum loading and minimum standing-wave ratio, the latter being most important as an indication of proper setting. The "T" system is particularly well suited for use in all-metal "plumbing" arrays.

The Folded Dipole

Probably the most effective means of matching various lines to the wide range of antenna impedances encountered in v.h.f. antenna work is the folded dipole, shown in its simplest form in Fig. 14-2. When all portions of the dipole are of the same conductor size, the impedance at the feed-point is equal to the *square* of the number of elements in the folded dipole times the normal center impedance which would be present if only a conventional split half-wave radiator were used. Thus, the simple folded dipole of Fig. 14-2 has a feed-point impedance of 4 × 72, or approximately 288 ohms. It may be fed with the popular 300-ohm line without appreciable mismatch. If a three-wire dipole were used, the step-up in impedance would be *nine* times. Note that this step-up occurs *only* if all portions of the folded dipole are the same conductor size.

The impedance at the feed-point of a folded dipole may also be raised by making the fed portion of the dipole smaller than the parallel section. Thus, in the 50-Mc. array shown in Fig. 14-4 the relatively low center impedance of a 4-element array is raised to a point where it may be fed directly with 300-ohm line by making the fed portion of the dipole of ¼-inch tubing, and the parallel section of 1-inch. A 3-element array of similar dimensions could be matched by substituting ¾-inch tubing in the unbroken section. Conductor ratios and spacings may be obtained from Fig. 10-83, Chapter Ten.

Antenna Systems for 50 Mc.

Since the same basic principles apply to all antennas regardless of frequency, little discussion is given here of the various simple dipoles that may be used when nondirectional systems are desired. Details of such antennas may be found in Chapter Ten, and the only modification necessary for adaptation to use on 50 Mc. or higher is the reduction in length necessary for increased conductor diameter at these frequencies.

A Simple 2-Element Array

A simple but effective array which requires no matching arrangement is shown in Fig. 14-3. Its design takes into account the drop in center impedance of a half-wave radiator when a parasitic element is placed a quarter wavelength away. A director element is shown, as the drop in impedance using a slightly-shortened parasitic element is just about right to provide a good match to a 50-ohm coaxial line. The element lengths are not extremely critical in such a simple system, and the figures presented may be used with satisfactory results.

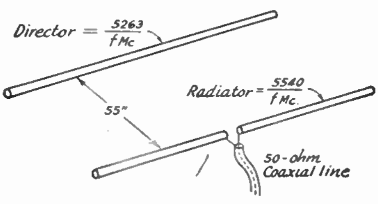


Fig. 14-3 — A simple 2-element array for 50 Mc. No matching devices are needed with this arrangement.

A 4-Element Array

The importance of broad frequency response in any antenna designed for v.h.f. work cannot be overlooked. The disadvantage of all parasitic systems is that they tend to tune quite sharply, and thus are often effective over only a small portion of a given band. One way in which the response of a system can be broadened out is to increase the spacing between the parasitic elements to somewhat more than the 0.1 or 0.15 wavelength normally considered to provide optimum front-to-back ratio. Some broadening may also be obtained by making the directors slightly shorter and the reflector slightly longer than the optimum value. The folded dipole is useful as the radiator in such an array, as its over-all frequency response is somewhat broader than other types of driven elements.

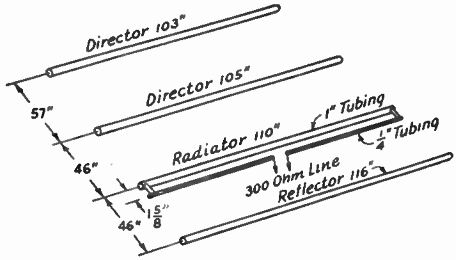


Fig. 14-4 — Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

A 4-element array for 50 Mc. having an effective operating range of about 2 Mc. is shown in Figs. 14-4 and 14-5. It employs a folded dipole having nonuniform conductor size. Reflector and first director are spaced 0.2 wavelength from the driven element, while the forward director is spaced 0.25 wavelength. The spacing and element lengths given were derived experimentally, and are those that give optimum forward gain at the expense of some front-to-back ratio. As the latter quality is not of great value in 50-Mc. work, it can be neglected entirely in the tuning procedure for such an array.

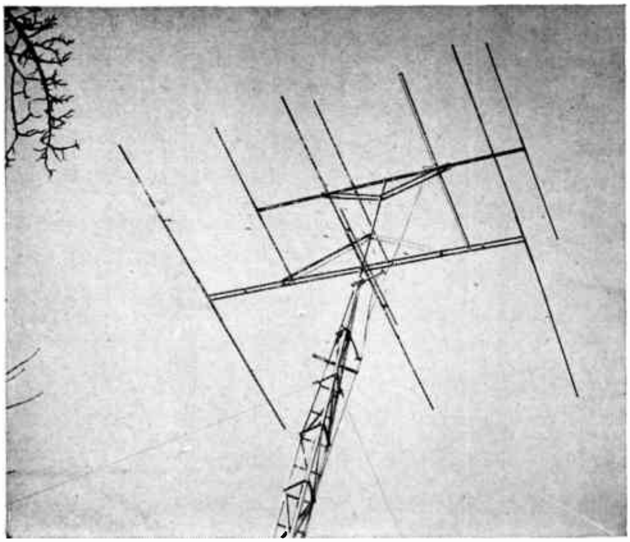


Fig. 14-5 — An example of stacking two arrays for different bands on the same support. The top section is a 4-element array for 50 Mc.; the lower a 3-element system for 28 Mc. All-metal construction is employed.

The dimensions given are for peak performance at 50.5 Mc. For other frequencies, the length of the folded dipole in inches should be figured according to the formula

$$L = \frac{5540}{f_{Mc.}}$$

The reflector will be 5 per cent longer, the first director 5 per cent shorter, and the second director 6 per cent shorter than the driven element. A broadening of the response may be obtained, at a slight sacrifice in forward gain, by adding to the reflector length and subtracting from the director lengths. For those interested in experimenting with element lengths, slotted extensions may be inserted in the ends of the various elements, other than the dipole, as shown in Fig. 14-7. A 3-element array may be built, using the same general dimensions, except that the unbroken section of the folded dipole, in this case, should have a ¾-inch diameter element in place of the 1-inch tubing used in the 4-element array.

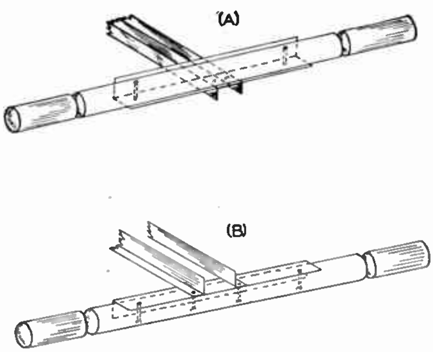


Fig. 14-6 — Detail sketches showing method of mounting elements in the dual array for 28 and 50 Mc. A — The 50-Mc. boom is comprised of two pieces of angle stock mounted edge-to-edge to form a channel. The elements are fastened to the boom by means of a cradle, also of angle stock. B — In the 10-meter array, the two portions of the boom are separated, and mount on either side of the vertical support. The elements and their supporting crossarms are attached to the lower surface of the boom.

Stacked Antennas

Excellent results in long-distance work are being obtained by 50-Mc. stations using various more-complex directional arrays than the ones described above. The most important factor in such work is the attainment of the lowest-possible radiation angle, and this purpose is well served by stacking of elements, in either vertical or horizontal systems. The use of two parasitic arrays, one a half-wavelength above the other, fed in phase, provides a gain of 3 db. or more over that of a single array. The system shown (for 144 Mc.) in Fig. 14-8 is excellent for either vertical or horizontal polarization, as is the "H" array, using four half-wave elements, with or without parasitic elements.

Stacked Arrays for Two Bands

As many 50-Mc. enthusiasts also operate on 28 Mc., it is often desirable to stack arrays for the two bands on a common tower and rotating device. Such a dual array, combining a 4-element system for 50 Mc. with a 3-element array for 28 Mc., is shown in Fig. 14-5.

If space limitations make it absolutely necessary, the two arrays may be mounted with but a few inches separating them, but experience has shown that some effectiveness is sacrificed, particularly in the array for the higher frequency. A separation of at least three feet is recommended as the minimum for avoiding harmful interaction. In the example shown the separation is six feet, at which distance each array performs equally as well as it would if mounted alone.

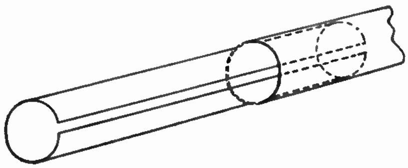


Fig. 14-7 — Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.

In this dual array all-metal construction is employed, doing away with the use of insulators in mounting the elements. The booms are made of two pieces of 1-inch angle stock (24ST aluminum), with supporting braces of the same material. The method of assembling the booms and mounting the elements is shown in Fig. 14-6. The booms are 150 inches and 160 inches in length for the 6- and 10-meter arrays respectively. To prevent swaying of the 10-meter elements, they are braced with guy wires, which are broken up with small insulators. These sway-brace wires are attached to the elements at approximately the midpoint between the boom and the outer end, and are brought up to the vertical support at the point of attachment of the horizontal fore-and-aft braces.

The 50-Mc. portion of the array is similar in element length and spacing to the 4-element array already described. The element spacing for the 10-meter array is 0.2 wavelength, or 80 inches. The driven element is 198 inches long, the director 188 inches, and the reflector 208 inches. It is fed by means of a "T"-match and a 300-ohm line. These dimensions give quite uniform performance and low standing-wave ratio over the range from 28 to 29.1 Mc., and the array will take power and show appreciable gain over a half-wave from 27.2 to 29.7 Mc. Complete details of this dual array will be found in *QST* for July, 1947.

Antenna Systems for 144 Mc.

The antennas already described may, of course, be used for 144 as well as 50 Mc., but since they are designed for maximum effectiveness in a horizontal position, other designs may be used more effectively where vertical polarization is desired. With either polarization, the stacking of elements vertically lowers the radiation angle and extends the operating range. The smaller size of 144-Mc. arrays makes such stacking of elements a much simpler procedure than on 50 Mc. Another advantage of the array employing elements fed in phase is that its frequency response is likely to be less critical than an array that achieves the same gain with but one driven element and parasitic directors and reflectors. Thus the element lengths, even in such complex systems as the 16-element array shown in Fig. 14-9, are not at all critical.

Plane reflectors are usable at 144 Mc., their size at this frequency being within reason. An interesting possibility in connection with this type of reflector is its use with two different sets of driven elements, one on each side of the reflecting screen. A set of elements arranged for vertical polarization may be used on one side, and a set of horizontally-polarized elements on the other, or the plane reflector may be made to serve on two different bands by a similar arrangement of elements for two frequencies, on opposite sides of the reflector. The screen need not be a solid sheet of metal, or even a close-mesh screen. A set of wires or rods arranged in back of the driven elements will work almost equally well. The dimensions of the reflector are not critical. For maximum effectiveness, the plane reflector should extend at least one-quarter wavelength beyond the area occupied by the elements, but reflecting curtains no larger than the space occupied by the reflectors shown in Fig. 14-9 have been used with good results.

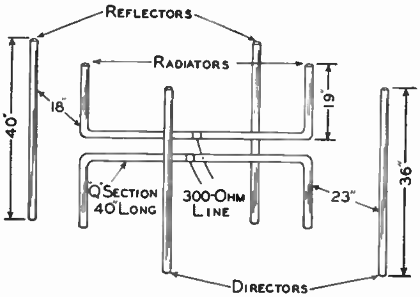


Fig. 14-8 — A double-"Q" array for 144 Mc. The horizontal portion of the half-wave "H" acts as a "Q" section, matching the antenna impedance to the 300-ohm line attached at the center of the array. This array works well in either vertical or horizontal positions.

A 6-Element Array

In designing directional arrays having more than one driven element it is advisable to arrange for feeding the array at a central point. A simple 6-element array of high performance, incorporating this principle, is shown in Fig. 14-8. All the elements may be made of soft-copper tubing, ¼ inch in diameter. The driven elements are comprised of two pieces which are bent into two "U"-shaped sections and arranged in the form of a half-wave "H." The horizontal portion of the "H" is then a double quarter-wave "Q" section, matching the impedance of the two radiators to that of the feedline. With the wide spacing used, the position of the parasitic elements is not particularly critical, except as it affects the impedance of the system, and the spacing of the elements may be varied to provide the best

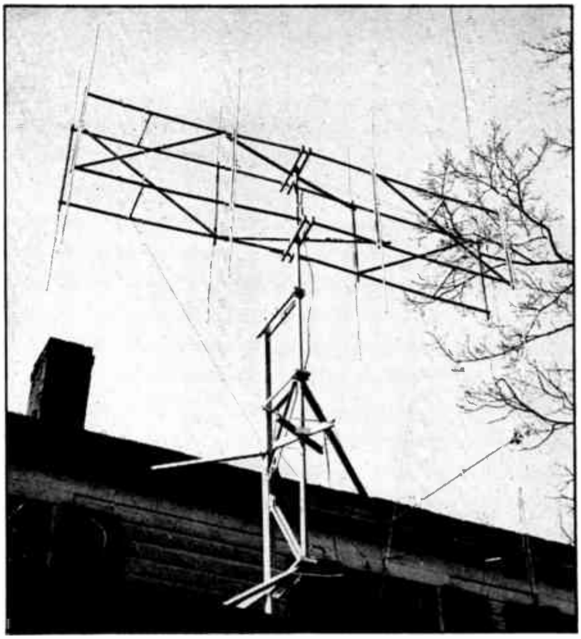


Fig. 14-9 — A 16-element array for 144 Mc., showing supporting structure and "rotating mechanism." Sash cord wrapped three times around the crisscross pulley permits 360-degree rotation.

match. The spacing of the horizontal section may be varied for the same purpose. With the dimensions given, a spacing of one inch between centers is about right for feeding with a 300-ohm line. The radiation pattern of this array is similar in both horizontal and vertical planes; thus it will work with equal effectiveness in either position, provided the polarization is the same as that of the stations to be contacted.

A 16-Element Array

By using a curtain of eight half-wave elements, arranged as shown in Figs. 14-9 and 14-10, backed up by eight reflectors, a degree of performance can be obtained which is truly outstanding. A gain of as much as 15 db. can be realized with such an arrangement, effecting an improvement in operating range which could never be obtained by any other means. Such an array is neither difficult nor expensive to construct, and its performance will more than repay the builder for the trouble involved in its construction.

The cumbersome nature of the structure required to support such an array would make its construction out of the question for a lower frequency, but for 144 Mc. the outside dimensions are only 1½ × 7 × 10 feet, and the supporting frame can be made quite light.

The center pole (a 1½-inch rug pole 10 feet long) turns in three bearings which are mounted on braced arms extending out about two feet from a "two by three," which is braced in a vertical position. An improvised pulley made of two pieces of 1 × 2-inch "furring" notched in the ends and fastened crisscross fashion near the bottom of the center pole serves as a "rotating mechanism." Sash cord wrapped three times around this "pulley" and run over to the window on small pulleys allows the beam to be rotated more than 360 degrees before reversal is required. To keep the array from twisting in high winds light sash cords are attached near each end of the supporting structure. These cords are brought through the window near the rotating ropes and are pulled up tight and fastened when the antenna is not in use.

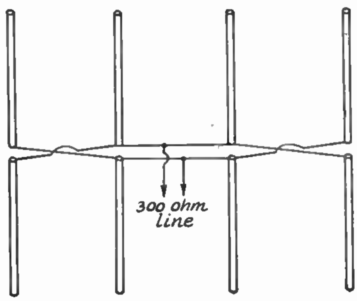


Fig. 14-10 — Schematic of the radiating portion of the 16-element 144-Mc. array. Reflectors are omitted for clarity. Radiators are 38 inches long, reflectors 40.5 inches. Crossover or phasing sections are also 40.5 inches long. Reflectors are mounted 17 inches in back of each radiator.

The elements are of 7⁄16-inch soft-aluminum tubing for light weight. To stiffen the structure, and to help maintain alignment, inserts were turned down from ½-inch polystyrene rod to fit tightly into the elements at the point where the crossover or phasing wires are connected. Similar inserts are used for the reflector elements also. The interconnecting phasing sections are of No. 16 wire, spaced about 1½ inches. The feedline, connected at the center of the system, is Amphenol 21-056 Twin-Lead, 300 ohms impedance. The impedance at the center of the array is about right for direct connection of the 300-ohm line, without the necessity for a matching section of any kind. It is probably somewhat lower than 300 ohms, actually, and if a perfect match is desired, a "Q" section may be used. The performance is not greatly affected by such a change, however, as the standing-wave ratio is relatively low with the connection as shown.

The center section of the array may be used without the outside 8 elements, if space is limited, and a simpler array of good performance is desired. The simple "H" with reflectors may also be fed with 300-ohm line without the necessity for special matching devices.

Desiring still further gain, ambitious 144-Mc. enthusiasts have doubled and even tripled the 16-element array with worth-while improvements in gain. The same general arrangement, but using four rows of 4 driven elements, with 16 reflectors, has been particularly effective in improving the gain and sharpening the pattern.

Long-Wire Antennas

Where space permits their use, the possibilities of long-wire antenna systems, particularly the rhomboid, should not be overlooked. Design problems are similar to those for lower frequencies, and data contained in Chapter Ten may be applied to systems for the v.h.f. bands. A vertical rhomboid for 144 Mc. or higher is of practical size, and such antennas have been used with outstanding results. At still higher frequencies the stacking of such arrays a half-wave apart presents interesting possibilities.

Mobile and Portable Antennas

A common type of antenna employed for mobile operation on 50 and 144 Mc. is the quarter-wave radiator which is fed with a coaxial line. The antenna, which may be a flexible telescoping "fish pole," is mounted in any of several places on the car. The inner conductor of the coaxial line is connected to the antenna, and the outer conductor is grounded to the frame of the car. Quite a good match may be obtained by this method with the 50-ohm coaxial line now available; however, it is well to provide some means of tuning the system, so that all variables can be taken care of. The simplest tuning arrangement consists of a variable condenser connected between the low side of the transmitter coupling coil and ground, as shown in Fig. 14-11. This condenser should have a maximum capacitance of 75 to 100 $\mu\mu$fd. for 50 Mc., and should be adjusted for maximum loading with the least coupling to the transmitter. Some method of varying the coupling to the transmitter should be provided.

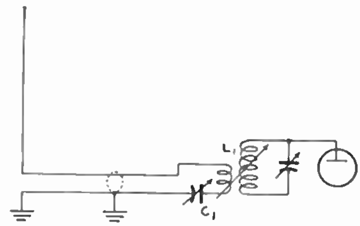


Fig. 14-11 — Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacitance of 75 to 100 $\mu\mu$fd. for 28- and 50-Mc. work. L_1 is an adjustable link.

The short antenna required for 144 Mc. (approximately 19 inches) permits mounting the antenna on the top of the car. Such an arrangement provides good coverage in all directions, the car body acting as a ground plane. When the antenna is mounted elsewhere on the car, it is apt to show quite marked directional characteristics. Because of this it is desirable to make provisions for the use of the same antenna for both transmitting and receiving.

A Collapsible Array for 50 Mc.

The best antenna possible for operation under mobile conditions is not particularly effective, as compared with antenna systems normally used in fixed-station work. To make the most of the fine opportunities for DX work afforded by countless high-altitude locations which are accessible by car, it is helpful to have some sort of collapsible antenna array which can be assembled "on the spot." Even a simple array like the one shown in Figs. 14-12 and 14-13 will effect a great improvement in the operating range of the low-powered gear normally used for mobile operation. This one is designed for 50-Mc. use, but similar arrangements can be made for other frequencies.

The array shown is a 2-element system, comprised of a radiator which is fed with coaxial line by means of a "T"-match, and a reflector which is spaced 0.15 wavelength in back of the

Fig. 14-12 — A 2-element collapsible array for 50-Mc. portable use.

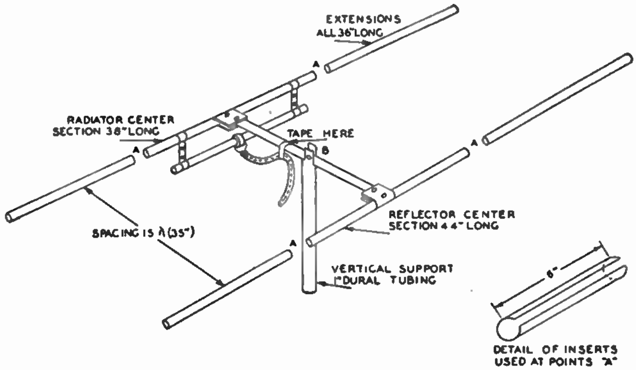


Fig. 14-13 — Detail drawing of the collapsible 50-Mc. array shown in Fig. 14-12. All parts except the vertical support, which is 1 inch in diameter, are made of ¾-inch duralumin tubing. For carrying purposes, it is taken apart at points *A* and *B*, inserts of slotted dural tubing being used at points *A* to hold the sections together. All extensions are the same length, the difference in element length being provided by the length of the center sections.

driven element. It is made entirely of ¾-inch dural tubing, except for the vertical support, which is 1-inch tubing of the same material. A suggested method of mounting is shown in Fig. 14-12. A short length of 1 × 2-inch or larger wood is bolted to the car bumper. A piece of ¾-inch dural tubing is bolted to this upright, and the 1-inch vertical section of the array slips over the top of the ¾-inch section. The array is turned by means of ropes attached to the reflector element. Height of the array may be increased over that shown by using a longer wooden support, in which case it is desirable to use a 2 × 2 for greater strength. An anchoring pin made from a spike inserted in the bottom end of the wooden support is helpful to prevent tilting of the array. With such a device embedded in the ground, the whole assembly will remain rigid, which is helpful in the high winds usually encountered in mountain-top locations. Portability is provided by making the elements in three sections, with the end sections all the same length. The center section of the radiator is 6 inches shorter than that of the reflector.

The fed section of the "T" matching device is composed of two pieces of ¾-inch dural tubing about 14 inches long. The two sections are held together mechanically, but insulated electrically, by a piece of polystyrene rod which is turned down just enough to make a tight fit in the tubing. The inner and outer conductors of the coaxial line are fastened to the two inside ends of the matching section. Clips made of spring bronze are used for connection between the radiator and the "T." The position of these should be adjusted for maximum loading and minimum standing-wave ratio on the line.

This antenna system may be used as a dipole on 29 Mc. by plugging the reflector sections into the driven element, thus bringing its over-all length to approximately that of a half-wave for the high end of the 10-meter band.

Miscellaneous Antenna Systems

Coaxial Antennas

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. 14-14 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 nonresonant, since its characteristic impedance is the same as the center impedance of the half-wave 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.

Cylindrical Antennas

Radiators such as are used for television and broad-band FM are of interest in amateur v.h.f. operation because they work at high efficiency without adjustment throughout the width of an amateur band.

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 whole system. However, in the case of antennas fed by matched-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 reduction in power input to the antenna.

A properly-designed-and-constructed wide-band antenna, on the other hand, will exhibit very nearly constant input impedance over 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 use of a relatively-large diameter-to-length ratio lowers the Q of the antenna, thus broadening the resonance characteristic.

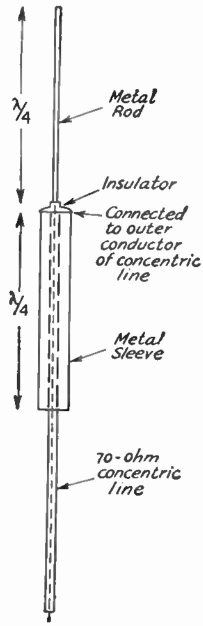


Fig. 14-14 — Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

As the diameter-to-length ratio is increased, end effects also increase, with the result that the antenna must be made shorter than a thin-wire antenna resonating at the same frequency. The reduction factor may be as much as 20 per cent with the tubing sizes commonly used for amateur antennas at v.h.f.

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 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. 14-15.

Plane Sheet Reflectors

The small physical size 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 one-quarter or three-quarter 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 closely-spaced parallel-wire spines, is more easily erected and offers lower wind resistance.

Parabolic Reflectors

A plane sheet may be formed into the shape of a parabolic curve and used with a driven radiator situated at its focus, to provide a highly-directive antenna system. 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 of the reflector 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. With an aperture of the order of 10 or 20 wavelengths, a beam-width of 5 degrees 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 length of the parabola is that which places the radiator along the plane of the mouth; this length is equal to one-half the mouth radius. At other focal distances interference fields may deform the pattern or cancel a sizable portion of the radiation.

Corner Reflectors

The "corner" reflector consists of two flat conducting sheets which intersect at a designated angle. 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 degrees, 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. 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. 14-16, 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 degrees can be

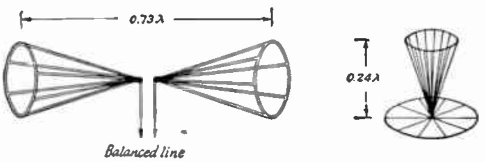


Fig. 14-15 — Conical broad-band antennas have relatively constant impedance over a wide frequency range. The three-quarter wavelength dipole at left and the quarter-wave vertical with ground plane at right have the same input impedance — approximately 65 ohms. Sheet-metal or spine-type construction may be used.

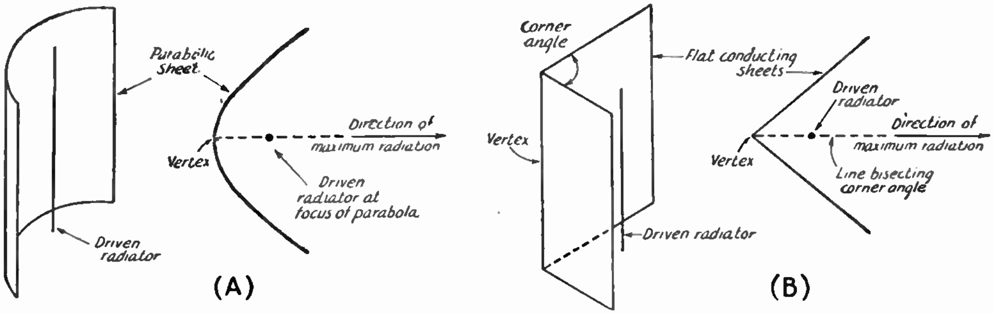


Fig. 14-16 — Plane sheet reflectors for v.h.f. and u.h.f. A shows a parabolic sheet and B a square-corner reflector.

used, with some decrease in gain. A 180-degree "corner" is equivalent to a single flat-sheet reflector. With angles smaller than 90 degrees, 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 144- and 50-Mc. square-corner reflector 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.

CHAPTER 15

U.H.F. and Microwaves

Once the amateur passes the 220-Mc. band on the way up through the radio-frequency spectrum, he encounters a distinct change of technique. So far he has been operating in a region where various modifications make usable the familiar coils and condensers, the crystal-controlled transmitters, selective superhet receivers, and other more-or-less standard items of the amateur field.

The boundary line beyond which such conventional gear is no longer usable has moved ever higher and higher in frequency as new developments and improvements in existing equipment have come along. In the early '30s the boundary line was our 28-Mc. band; then, as that band filled, the line moved up to 56 Mc., which remained border territory until 1938, when stabilization of transmitters used was made a legal requirement of operation in the old 5-meter band. For some years, then, the 112-Mc. band, and since the war the 144-Mc. band, constituted the dividing line, but even the latter band has now swung into the stabilized-transmitter-and-superhet-receiver field, and the 220-Mc. band is rapidly achieving the same status.

In the light of current developments, it may be said that the 420-Mc. band is now true borderline territory. The multistage transmitter can be used successfully, as can the superheterodyne receiver of semiconventional design, but special tank circuits must be employed and extreme care in mechanical layout must be used, in order to achieve satisfactory results.

The 420-Mc. band is fruitful territory for the experimentally-minded amateur. Most of the gear used will have to be made by the worker himself, but the techniques employed are such that construction of the necessary equipment will not be outside his capabilities. There is enough interest in a number of areas to support regular activity in this band, and more can be generated with a little organizational effort.

Antenna work on these frequencies is particularly intriguing. The antenna systems are so small in size that arrays having a gain of 10 db. or more can be erected in almost any location. Experimentation with models built for 420 Mc. is a fine way of checking the performance of arrays for lower frequencies. The experimenter who starts to work with u.h.f. antenna systems is bound to find himself spending many interesting hours checking his pet antenna ideas. Since u.h.f. or microwave experimentation is best accomplished in groups of interested workers, it is a fine project for coöperative effort by radio clubs.

The communication possibilities of the u.h.f. region should not be overlooked. Recent experience in the 144-Mc. band has demonstrated the possibilities of that band for long-distance work, and it is reasonable to assume that propagation vagaries, as regards tropospheric effects, will continue on up through the microwave range. With suitable antenna systems, it is probable that operating ranges on the frequencies above 200 Mc. may equal or approach those now being covered in the 70–160-Mc. region.

At least some amateur work has been done in all the microwave bands now assigned. The work of the pioneers in adapting these frequencies to communication purposes has been in line with the best amateur tradition, and it is hoped that the almost unknown territory from 500 Mc. up will see much amateur exploration in the near future.

U.H.F. Tank 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 needed capacitance. The fact that the coil itself has a certain amount of self-capacitance, as well as some resistance, while the condenser also possesses a small self-inductance, can usually be disregarded.

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" may in part be radiated. 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

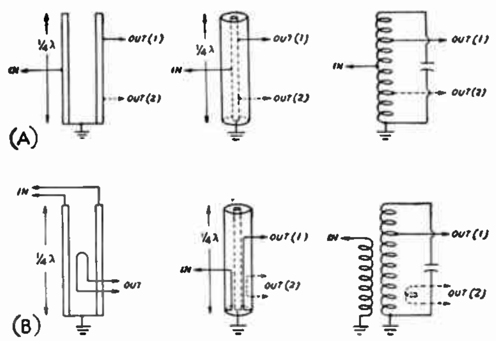


Fig. 15-1 — Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

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 greatly reduced.

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. The input impedance may be as high as 0.4 megohm for a well-constructed line.

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. 15-1. 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 circuits operating above 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.

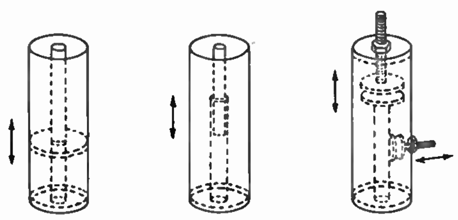


Fig. 15-2 — Methods of tuning coaxial resonant lines.

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 this and other 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. 15-2. At the left, a sliding shorting disk 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 thereby 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 capacitor at the open end of the line has the greatest tuning effect per unit of capacitance; the alternative method, which is equivalent to "tapping" the condenser down on the line, has less effect on the Q of the circuit. Lines with capacitive "loading" of the sort illustrated will be shorter, physically, than an unloaded line resonant at the same frequency.

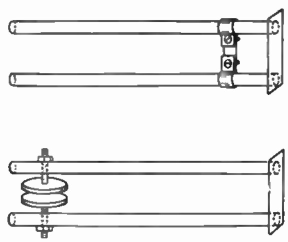


Fig. 15-3 — Methods of tuning parallel-type resonant lines.

The short-circuiting disk 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 insulating films. Usually a soldered connection or a tight clamp is used to secure good contact. When the length of line

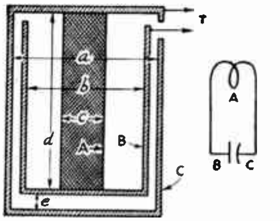


Fig. 15-4 — Concentric-cylinder or "pot"-type tank for v.h.f. The equivalent circuit diagram is also shown. Connections are made to the terminals marked *T*. For maximum *Q* the ratio of *b* to *c* should be between 3 and 5.

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 conductor.

Two methods of tuning parallel-conductor lines are shown in Fig. 15-3. The sliding short-circuiting strap can be tightened by means of screws and nuts to make good electrical contact. The parallel-plate condenser in the second drawing may be placed anywhere along the line, the tuning effect becoming less as the condenser is located nearer the shorted end of the line. Although a low-capacitance variable condenser of ordinary construction can be used, the circular-plate type shown is symmetrical and thus does not unbalance the line. It also has the further advantage that no insulating material is required.

Equivalent impedance points, for coupling or impedance-transformation purposes, are shown in Fig. 15-1 for parallel-line, coaxial-line, and conventional coil-and-condenser circuits.

Lumped-Constant 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, and radiation from unshielded lines may reduce their effectiveness materially.

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 disks combining half-turn inductance loops with semicircular 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, commonly termed the "hat" resonator. Two copper shells with wide, flat "brims" are mounted facing each other on an axially-aligned copper rod. The capacitance in the circuit is that between the wide shells, while the central rod comprises the inductance.

"Pot"-Type Tank Circuits

The lumped-constant concentric-element tank in Fig. 15-4, commonly referred to as the "pot" circuit, is equivalent to a very short coaxial line (no linear dimension should exceed 1/20 wavelength), loaded by a large integral capacitor.

The inductance is supplied by the copper rod, *A*. Capacitance is provided by the concentric cylinders, *B* and *C*, plus the capacitance between the plates at the bottoms of the cylinders.

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

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

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

where the symbols are as indicated in Fig. 15-4, and dimensions are in inches. The left-hand term for capacitance applies to the concentric cylinders, *B* and *C*, while the second term gives capacitance 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. 15-5 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

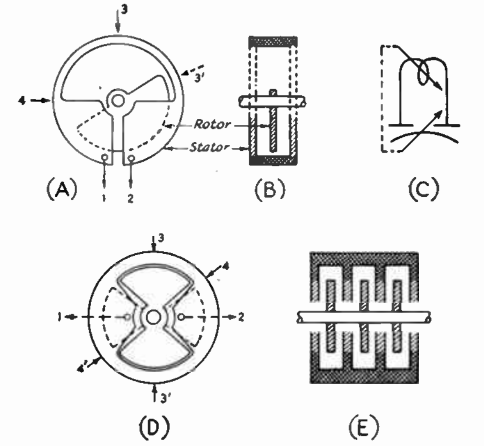


Fig. 15-5 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

from point *3* to point *3'* as the rotor is turned. Constant magnetic coupling may be obtained by a coupling loop located at point *4*, however.

In the modification shown at D, two sectoral stators are spaced 180 degrees, 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 capacitance of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in Chapter Twenty-Four. 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 conventional 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 sections may either be made in a single piece or with separate sectoral stator plates and spacing rings assembled with machine screws.

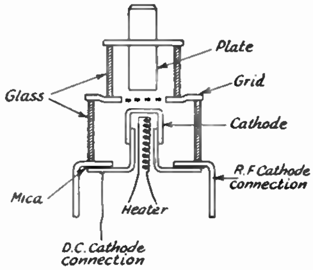


Fig. 15-6 — Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disk electrode connections reduce lead inductance.

V.H.F. and U.H.F. Tubes

At very-high frequencies, interelectrode capacitance 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. 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. But at 100 Mc., this same transit time represents 1/10 of a cycle and a full cycle at 1000 Mc. These limiting factors establish about 3000 Mc. as the upper frequency limit for negative-grid tubes.

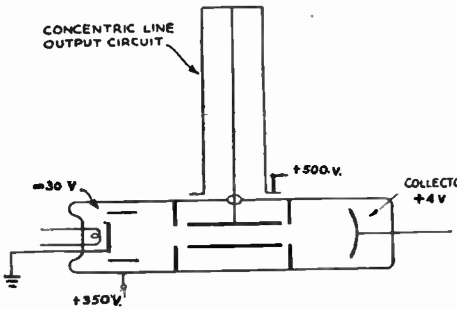


Fig. 15-7 — Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a superregenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50-cm. region.

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 capacitance 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 capacitance 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, as shown in Fig. 15-6, instead of coaxially. The uniform coplanar electrode design and disk-seal terminals permit low interelectrode capacitance.

Velocity Modulation

In negative-grid operation the potential on the grid tends to reduce 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 them. 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

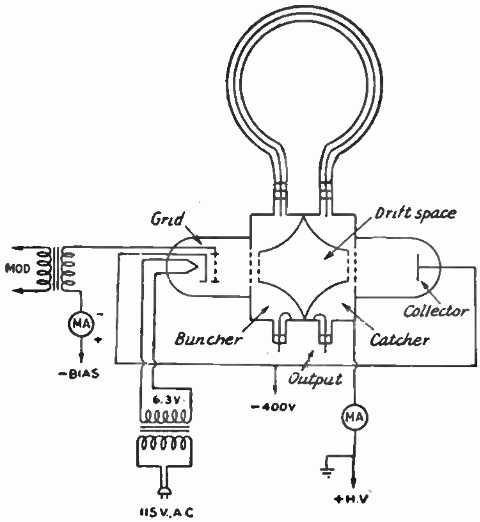


Fig. 15-8 — **Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.**

power output at the operating frequency, the efficiency is reduced in direct proportion to the variation in velocity, the output reaching a value of 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 the grid is used to change the velocity of the electrons in a constant-current electron beam, rather than to vary the intensity of a constant-velocity current flow as is the method in ordinary tubes.

A simple form of velocity-modulation oscillator tube is shown in Fig. 15-7. 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 space at each end between the inner tube, through which the electron beam passes, and the disks at the ends of the larger tube portion. With r.f. voltage applied across these gaps, which are small compared to the distance traveled by the electrons in one half-cycle, 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 distance covered by the electrons in one-half cycle, 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 electric 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. The electrons emerging from the pair of grids 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 r.f. current created by the bunching of the electron beam induces an r.f. voltage between the grids. The catcher cavity is made resonant at the frequency of the

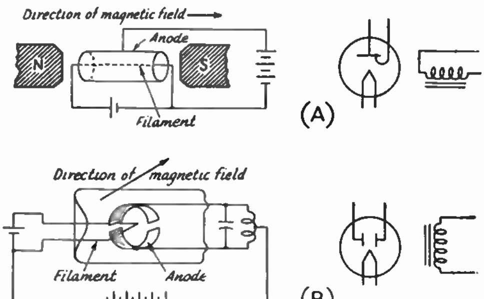


Fig. 15-9 — **Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.**

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. 15-8, 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 bunched beam current is rich in harmonics, but the output waveform is remarkably pure because the high Q of the catcher rhumbatron suppresses the unwanted harmonics.

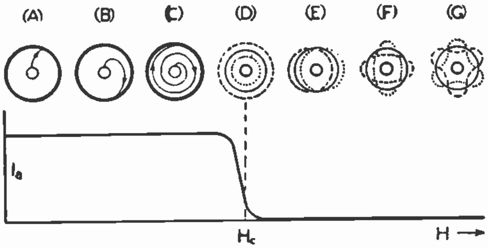


Fig. 15-10 — Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, I_a. Oscillations commence when *H* reaches a critical value, H_c; progressively higher-order modes of oscillation occur beyond this point.

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 cylinder is divided longitudinally.

Magnetron oscillators are operated in two different ways. Electrically the circuits are similar, the difference being in the relation between electron transit time and the frequency of oscillation.

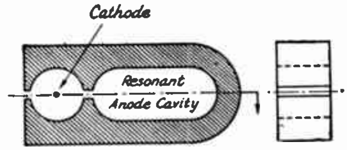
In the negative-resistance or dynatron type of magnetron oscillator, the element dimensions and anode voltage are such that the transit time is short compared with the period of the oscillation frequency. Electrons emitted from the cathode are driven toward both halves of the anode. If the potentials of the two halves are unequal, the effect of the magnetic field is such that the majority of the electrons travel to that half of the anode that 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. The magnetron consequently exhibits negative-resistance characteristics. 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.

In the transit-time magnetron the frequency is determined primarily by its dimensions and by the electric and magnetic field intensities 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. With increased magnetic field the electrons tend to spiral around the filament, their radial component of velocity being much smaller than the angular component. Under critical conditions of magnetic field strength, a cloud of electrons rotates about the filament. It extends up to the anode but does not actually reach it.

The nature of these electron trajectories is shown in Fig. 15-10. Cases A, B and C correspond to the nonoscillating 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

Fig. 15-12 — Split-anode magnetron with integral resonant anode cavity for use at u.h.f.



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.

Fig. 15-10E, F and G depicts 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 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 those in a positive-grid oscillator. If the period of the alternating voltage is made equal to the

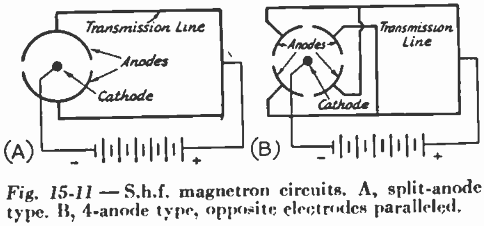


Fig. 15-11 — S.h.f. magnetron circuits. A, split-anode type. B, 4-anode type, opposite electrodes paralleled.

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 that lose energy remain in the interelectrode space longer than those that 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 connected 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. 15-12. 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. 15-13.

The efficiency of multisegment 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.

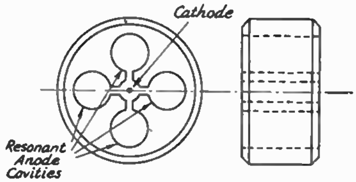


Fig. 15-13 — Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.

Equipment for 420 Mc.

Though it is entirely possible to use crystal control, even on 420 Mc., it is highly improbable that many amateurs will care to go to the trouble necessary to accomplish it. For some time to come, at least, most work on this band will be done with simple oscillators supplying the r.f. power. Superheterodyne-receiver design is not too difficult, and a considerable number of experimenters will find the superhet the most satisfactory receiving system, especially when broad-band i.f. systems are employed. The i.f. strips used in radar work are readily adapted to 420-Mc. amateur work, as their extreme broadness is not important at the present state of activity on this frequency. Broadness in both the transmitter and receiver need not be troublesome in this band, which is considerably wider than any of our lower v.h.f. assignments.

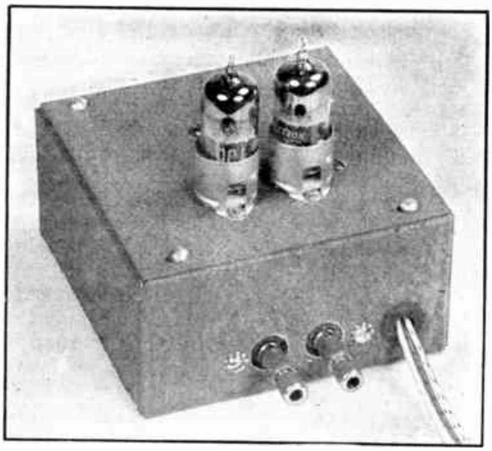


Fig. 15-14 — A transmitter for 420 Mc. using two 6J6 tubes in push-pull-parallel. Antenna coupling terminals project through the front of the chassis.

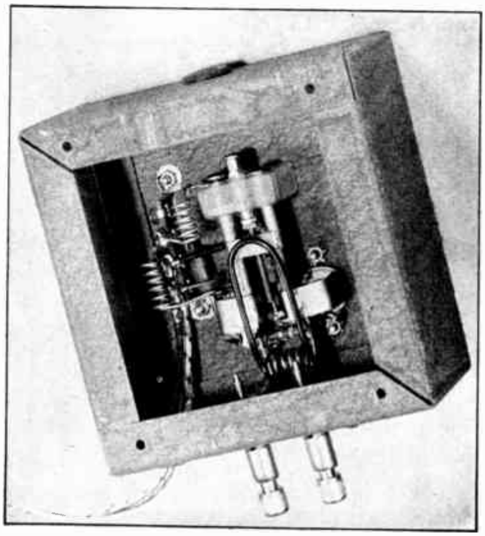


Fig. 15-15 — Bottom view of the 420-Mc. transmitter, showing the plate-line assembly. Note wide strips used to make connection to the tube plates.

● A 420-MC. TRANSMITTER

Not too many tubes are available that can be made to function at 420 Mc. Of these, the 6J6 is a logical choice. The transmitter shown in Figs. 15-14 through 15-17 uses a pair of 6J6s in push-pull-parallel. It can be operated at 15 watts input and is capable of delivering about 2 watts output at 420 Mc., the output dropping off slightly toward the 450-Mc. end of the band.

The circuit and shorting-bar details were suggested by a war-surplus transmitter-receiver known as the AN/APS-13. Details of the tank circuit may be seen in the under-chassis photo-

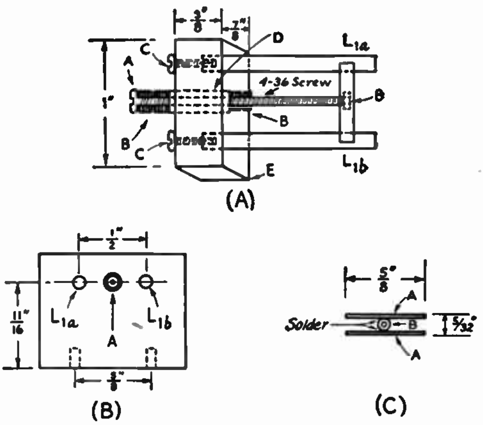


Fig. 15-16 — Detail drawing of the plate-line assembly used in the 420-Mc. transmitter. A shows the lines and method of mounting in the polystyrene block, which is shown in detail at B. The shorting bar, for frequency adjustment, is shown at C. In view A parts *B* are 4-36 nuts which are tightened on shaft *A*, a 1½-inch 4-36 screw. Part *D* is a bushing which is embedded in the polystyrene block. It serves as a bearing. Lines L_{1a} and L_{1b} are 1½ inches long.

graph, and the drawing, Fig. 15-16. It is made of two pieces of 3/16-inch copper tubing, this size being used because it can be tapped merely by running a 6-32 tap into the end of the tube, providing a simple means of mounting. The shorting bar is made of two pieces of spring bronze which are soldered to opposite faces of a 4-36 nut. A long screw, mounted in the polystyrene block which also acts as a support for the two portions of the line, serves as a means of frequency adjustment, its head being reached through a rubber-grommetted access hole in the end of the chassis. With the dimensions shown, movement of the shorting bar the whole length of the line just about covers the 420-450-Mc. range. Output drops off slightly at the high end, as might be expected, with most of the tank circuit shorted out. Note that the two plates of each 6J6 are connected to the tank circuit by means of 3/8-inch-wide copper strip, for negligible lead inductance.

The unit is mounted on the top plate of a small-sized utility box, with antenna terminals (National FWG) coming out through one side. Small self-supporting r.f. chokes are inserted in each heater lead, and the cathode is connected to the heater (be sure it's the grounded side) at the socket. Another r.f. choke is in the B-plus lead to the cold end of the tank circuit. A grid resistor is the only additional circuit component required.

The transmitter should be checked at low voltage, preferably 200 or under, until it is determined that it is operating correctly. At 250 volts it can run as high as 60 ma. without damaging the 6J6s, provided a load is kept across the antenna terminals. Since few experimenters will have an absorption-type wavemeter for 420 Mc., it is best to check the frequency with Lecher wires. Antenna coupling should be adjusted until maximum output is obtained, the least amount of coupling possible being used. Frequency checks should be made with the antenna attached, once the frequency of the oscillator is found to be close to the band.

Though the output of this little transmitter is less than two watts, it is sufficient for much interesting work. Since it is a simple matter to attain antenna gains of 15 db. or more at this frequency, a small amount of power can be made to produce surprising results.

SUPERREGENERATIVE RECEIVER FOR 420 MC.

Though the advanced experimenter will wish to use something more elaborate, the simple superregen is a good start on 420 Mc. The tuning system used in the receiver, see Fig. 15-19, could, however, be adapted to the tuning of the oscillator in a converter, if a suitable broad-band i.f. system is available.

It is obvious that coils and condensers, in the normal sense, are incapable of operation at 420 Mc. The tuning arrangement used in this receiver is the vane type, wherein a self-resonant loop of wire has its inductance reduced by passing a copper vane across the plane of the loop. The three-tube receiver shown in Figs. 15-18–15-21 uses a 955 superregenerative detector, with a loop of stiff wire connected between the plate and the grid-blocking condenser, in a manner similar to the conventional circuit used on lower frequencies. In details other than the tuning arrangement, the receiver is similar to self-quenched detectors used for years on lower v.h.f. bands.

The 955 socket is mounted on one side of a "U"-shaped bracket, with the grid and plate terminals at the top, so that the inductor extends above the top of the mounting bracket. The vane is attached to a 1/4-inch polystyrene rod which extends through a shaft bearing on the opposite side of the bracket from the tube socket. The vane is moved across the coil by means of a vernier dial, providing slightly more than 30 Mc. tuning range. The vane is

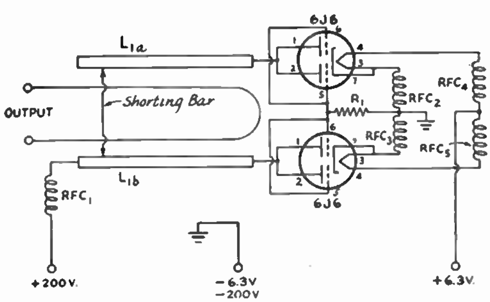
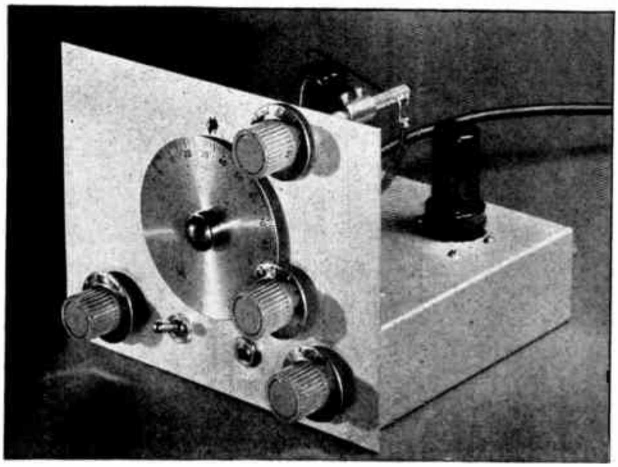


Fig. 15-17 — Circuit diagram of the 420-Mc. push-pull-parallel transmitter.

R_1 — 2700 ohms, 1/2 watt.

L_{1a}, L_{1b} — Plate line; see Fig. 15-16.

RFC_1, RFC_2, RFC_3, RFC_4, RFC_5 — 6 turns No. 20. 3/16-inch inside diameter, 5/8 inch long.



♦

Fig. 15-18 — A superregenerative receiver for 420 Mc. The two lower controls are for variation of detector voltage (left) and audio gain. The vernier dial controls the position of the tuning vane, while the knob at the top adjusts the antenna coupling.

♦

not grounded, so that if it accidentally touches the coil it will not short the B-plus to ground.

Adjustable antenna coupling, absolutely necessary at this frequency, is provided by a hairpin loop which is attached to a front-panel control in a manner that will be clear from the front- and rear-view photographs. A length of 300-ohm line runs from the hairpin through a hole in the chassis to the antenna terminals on the back of the chassis.

Critical points in attaining smooth operation of the detector are the values of R_1, C_2 and R_2, and the condition of the 955 tube. Tubes which work satisfactorily on lower frequencies may fail to operate entirely on 420 Mc. The addition of r.f. chokes in the heater and cathode leads may be necessary in some cases. They should be similar to those used in the 420-Mc. transmitter.

● A TWO-TUBE CONVERTER FOR 420 MC.

Though the selectivity of a communications receiver is too great to permit such a receiver to be used with a converter for 420-Mc. reception, the converter approach may be used with good results if the receiver or i.f. amplifier with which the converter is used has a reasonably-wide bandwidth. Wideband F.M. receivers such as the S-27, S-36, SX42 and SX43; superregenerative receivers such as the One-Ten and the HFS; and various home receivers designed for F.M. broadcast reception fall into this category. The i.f. strips used in many radar and altimeter receivers may also be used with converters to provide reception of 420-Mc. signals of the modulated-oscillator variety.

A suitable converter is shown in Figs. 15-22 through 15-25. It is extremely simple in design and is similar to the one-tube converters for 144 and 220 Mc. described in Chapter Twelve, except that two acorn-type triodes are used, instead of a dual triode combining the functions of mixer and oscillator. The mixer input circuit is self-resonant. Oscillator tuning is by means of a vane, the position of which is controlled by the vernier dial. The i.f. transformer is designed for approximately 27 Mc., but by suitable alteration of it the converter could be made to work into a receiver or i.f. amplifier on any frequency above about 10 Mc. or so. The converter has its own built-in power supply.

Mechanical Details

The photographs of the converter show how the parts are laid out on a chassis measuring 2 by 7 by 7 inches. The front panel is 8 inches wide and 7

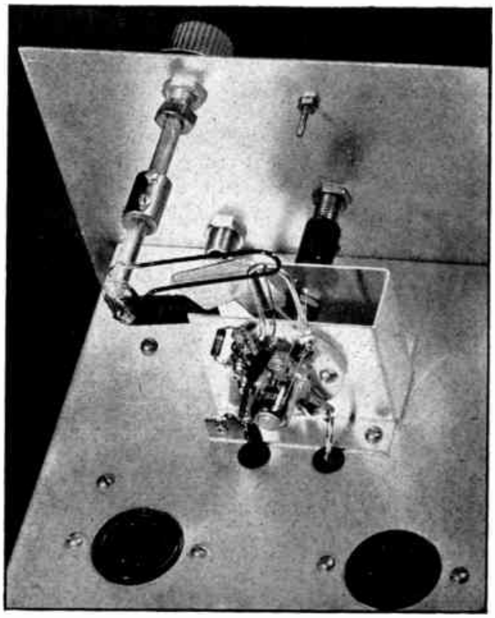


Fig. 15-19 — Close-up view of the detector assembly used in the 420-Mc. receiver. The frequency coverage is attained by moving a copper vane across the "U"-shaped detector tank circuit, maximum frequency being reached when the loop is covered by the vane. Antenna coupling is varied by means of the hairpin loop at the left. The two audio tubes were removed from their sockets to permit a clear view of the tuning system.

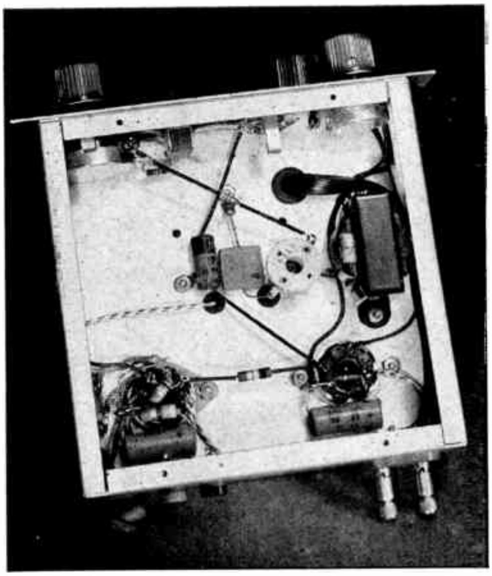


Fig. 15-20 — Bottom view of the 420-Mc. receiver. Loudspeaker terminals are at the lower left. At the right are the antenna terminals, from which a length of 300-ohm line runs up through the chassis to the antenna coupling loop.

inches high. The mixer-oscillator unit can be constructed and wired as a subassembly and, as a result, this section will be treated as a separate constructional operation. The "U"-shaped chassis for this assembly is formed from a piece of 1/16-inch aluminum stock measuring 3 inches by 8¾ inches which has been bent to form a channel having a height of 3½ inches and a distance of ¾ inch between upright members. Flanges having a width of ½ inch are bent to form right angles with the vertical surfaces and are used as mounting feet for the finished bracket. It is recommended that the mounting holes for the tube sockets, as well as clearance holes for the envelopes of the 955s, be drilled before the subchassis is bent into shape. These holes should be located so that the centers of the sockets will be down one inch from the top of the bracket. The sockets should be mounted with the cathode pins facing the left side of the bracket (as seen from the front-view photograph of the converter). Holes for the feed-through bushings should also be drilled at this time. These should be as close as possible to Pins 2 and 6 of the oscillator socket and Pin 1 of the mixer socket. The pin numbers referred to above apply only when the National type XLA socket is used. This type of socket is convenient to use because of the two extra prongs (the socket has seven prongs instead of the usual five) which may be used as tie points for the circuit wiring.

At this frequency leads must be practically nonexistent, a condition achieved by grouping the components around the tube sockets as shown in the views of the converter. The oscillator blocking condenser, C_6, is soldered directly to the grid-prong of the socket and the plate coil, L_3, is mounted between this condenser and the plate pin. C_1, the band-set condenser, is soldered across the open end of L_3. The feed-through bushing located adjacent to Pin 2 of the tube socket is used as the junction for RFC_1, C_7, and the 150-volt lead to the oscillator. Incidentally, it should be noted that the tube-socket prong numbers do not always cor-

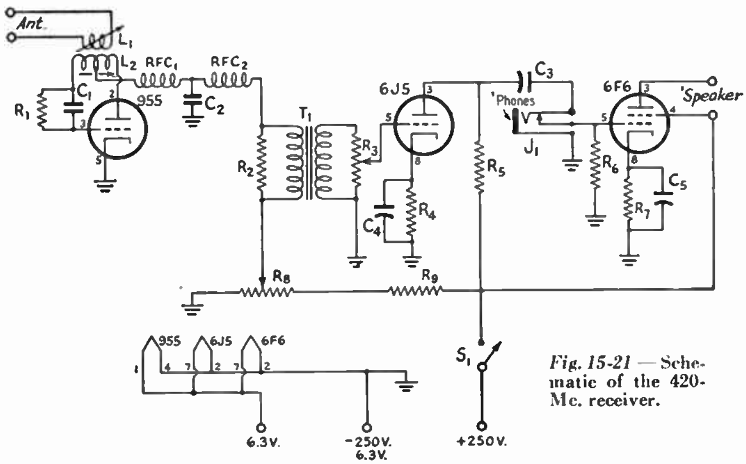


Fig. 15-21 — Schematic of the 420-Mc. receiver.

C_1 — 47-μμfd. trimmer (National XLA-C).
C_2 — 0.0033-μfd. mica.
C_3 — 0.01-μfd. 400-volt paper.
C_4, C_5 — 10-μfd. 25-volt electrolytic.
R_1 — 1.2 megohms, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 2200 ohms, ½ watt.
R_5, R_6 — 0.1 megohm, ½ watt.
R_7 — 470 ohms, 1 watt.
R_8 — 50,000-ohm potentiometer.
R_9 — 22,000 ohms, ½ watt.
L_1 — Hairpin loop: No. 12 wire with an inside diameter of ⅜ inch and a length of 2½ inches.
L_2 — Hairpin loop; No. 12 wire with an inside diameter of ½ inch and a length of 2 inches. Vane-tuned — see text and photograph.
J_1 — Closed-circuit jack.
RFC_1 — 4-μh. r.f. choke (Millen 34300).
RFC_2 — 10-mh. r.f. choke (Millen 34210).
S_1 — S.p.s.t. toggle switch.
T_1 — Interstage transformer.

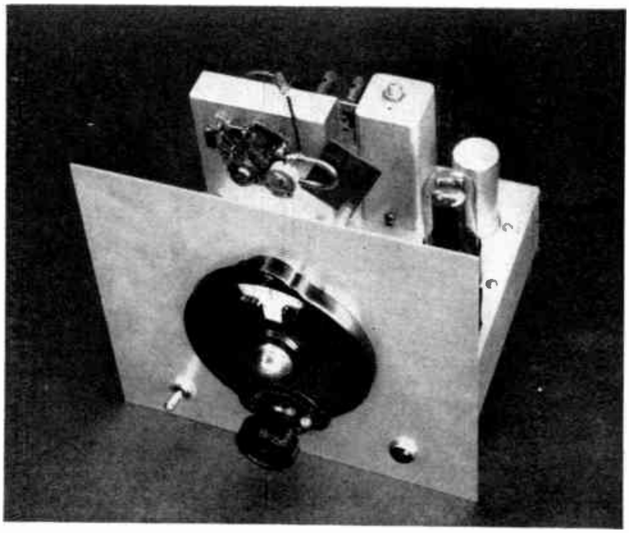


Fig. 15-22 — Panel view of the acorn converter for 420 Mc. Note the tuning vane used to vary the oscillator frequency.

respond with the tube pin numbers inasmuch as 7-prong sockets are used with 5-prong tubes. Heater voltage is fed to the tube through the bushing located next to Prong 6, and the heater by-pass condenser, C_{10}, is connected between the bushing and ground. RFC_2 is soldered between Prongs 4 and 5, and the grid-leak, R_3, is connected between Prong 5 and ground. Prong 7 has been made the common ground point for the entire circuit.

On the mixer-circuit side of the assembly, the grid coil, L_2, is soldered between Prongs 4 and 5, with the latter point serving as common ground for the circuit. Heater wiring and by-passing is identical with that of the oscillator tube with the one exception that the feed-through bushing is adjacent to Prong 1. The cathode resistor and by-pass condenser, R_1 and C_3 respectively, are connected between Prongs 5 and 7. A jumper connection is made between Prongs 7 and 2 and the oscillator-to-mixer coupling condenser, C_2, is soldered between the latter point and the oscillator tube socket. The rear view of the converter shows this condenser up above the horizontal portion of the subchassis.

Leads, each one having an approximate length of 12 inches, should now be soldered to the heater and plate-voltage feed-through bushings — it is much easier to complete this operation before the subassembly is mounted on the main chassis.

The output transformer may now be made and mounted in any convenient i.f. shield can, with the slug protruding through the top. If a slug-tuned form is not available, it is possible to make the transformer tunable by using a variable condenser of approximately 35 $\mu\mu$fd. as a substitute for the mica unit recommended in the parts list.

The front view of the converter shows the main tuning dial flanked by the power switch, S_1, and the pilot-light jewel. The panel and chassis are held together by the two components just mentioned. The output transformer is to the rear and right of the oscillator-mixer assembly, and the filter condenser and the rectifier tube at the upper right-hand corner of the chassis. The subassembly is mounted at the extreme left-hand edge of the main chassis at a point 2½ inches in from the rear edge. A 5/16-inch hole must be drilled in the lower chassis at a point directly below the center point of the oscillator-mixer unit. This hole is equipped with a rubber grommet and is used as a feed-through point for the heater and B+ wiring.

The rear-view, Fig. 15-23,

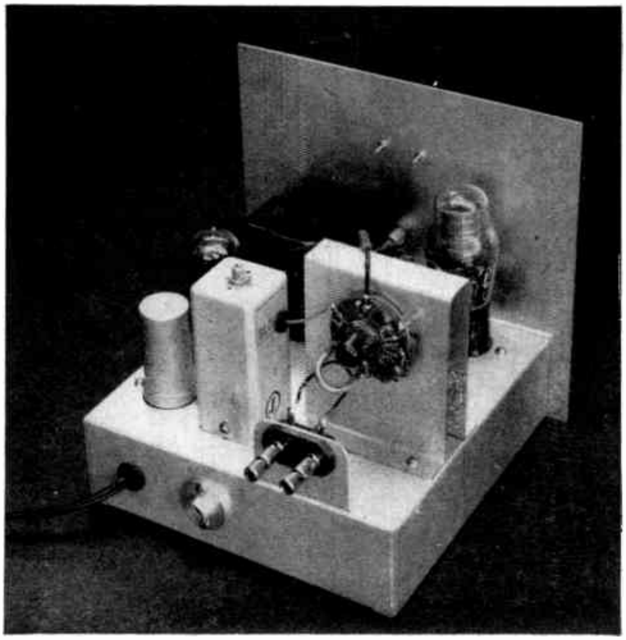


Fig. 15-23 — Rear view of the 420-Mc. converter, showing the mixer side of the r.f. subassembly. Power-supply components and i.f. output transformer are at the left.

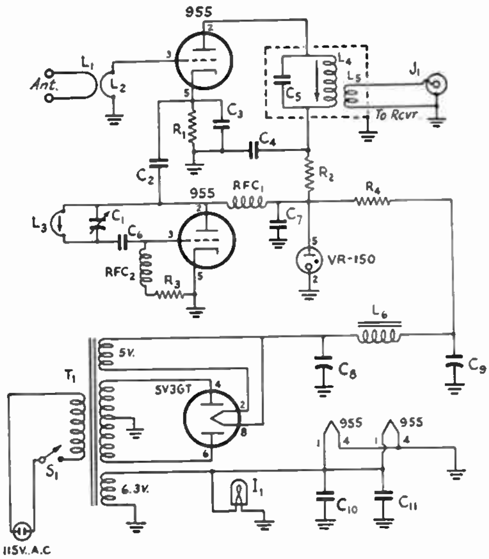


Fig. 15-24 — Wiring diagram of the 420-Mc. acorn converter.

C_1 — 5–20-μμfd. ceramic trimmer (Centralab 820-B).
C_2 — 5-μμfd. ceramic.
C_3, C_7, C_{10}, C_{11} — 27-μμfd. ceramic.
C_4 — 470-μμfd. midget mica.
C_5 — 15-μμfd. midget mica.
C_6 — 50-μμfd. ceramic (National XLA-C).
C_8, C_9 — 10-μfd. 450-volt electrolytic (Mallory FP-231).
R_1 — 10,000 ohms, ½ watt.
R_2 — 1000 ohms, 1 watt.
R_3 — 22,000 ohms, ½ watt.
R_4 — 10,000 ohms, 10 watts.
L_1 — 1 turn No. 12 enam., hairpin shape; over-all length 1½ inches; inside diameter, ½ inch.
L_2 — 1 turn ⅛-inch copper tubing, hairpin shape; over-all length, ⅞ inch; inside diameter, ½ inch.
L_3 — 1 turn of ⅛-inch copper tubing, hairpin shape; over-all length, 1⅛ inch; inside diameter, ½ inch.
L_4 — 12 turns No. 22 enam., ½-inch diameter, 1⁄16-inch long. Coil wound on a National XR-50 form.
L_5 — 3 turns No. 22 enam., interwound with bottom three turns of L_4.
L_6 — 8.5-henry 50-ma. filter choke (Stancor C-1279).
I_1 — 6.3-volt pilot-lamp-and-socket assembly.
J_1 — Coaxial-cable jack.
RFC1, RFC2 — 1-μhy. r.f. choke (National R-33).
S_1 — S.p.s.t. toggle switch.
T_1 — 325–325 volts, 40 ma., with 5- and 6.3-volt windings (UTC R-1).

shows the antenna terminals mounted on a small aluminum bracket to the rear of the mixer tube. The pick-up link, L_1, is rigid enough to be supported by its own leads and is soldered to the antenna terminal lugs. The 115-volt cord runs through a hole in the rear wall of the main chassis and the coaxial output fitting, J_1, is mounted to the right of the line cord. The power transformer and the regulator tube are located at the front of the chassis, as near the front panel as possible. Arrangement of other power-supply components may be seen in Fig. 15-25.

The National Type B dial is centered on the panel and is coupled to a metal panel-bearing-and-shaft assembly. The chassis side of the shaft is cut down to a length of 1¾ inches and is coupled to a 1¼-inch length of ¼-inch diameter polystyrene rod. One end of this rod is drilled and tapped to accommodate a 6–32 machine screw. A piece of 1⁄16-inch copper measuring 1¼ inches by 1¾ inches serves as the oscillator tuning vane and this vane is drilled at the bottom corner and then fastened to the polystyrene rod by a short machine screw. When the dial is rotated, the vane should come as close to the coil as possible without making actual contact. Its position may be varied by sliding the polystyrene rod either forward or backward in the shaft coupler.

Alignment

After the antenna coupling loop, L_1, has been soldered in place, it should be bent up toward L_2 so as to provide maximum coupling (minimum separation without touching) between the two coils. Power may now be applied to the circuits and, if a voltmeter is available, a quick check on the various voltages should be made. The supply output voltage (between

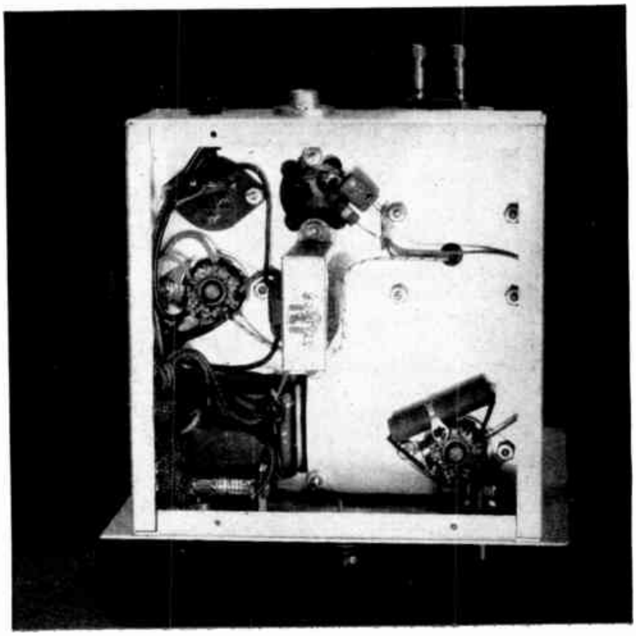


Fig. 15-25 — Below-chassis view of the acorn converter.

the positive side of C_9 and ground) should be approximately 350 volts and 150 volts should appear at Prong 5 of the regulator tube. A drop of 6 volts should be indicated with the meter connected across the cathode resistor, R_1. The total current drain, measured in series with the limiting resistor, R_4, should be approximately 20 ma.

First the i.f. transformer should be set on the proper frequency. Next the oscillator tuning range should be checked, by means of Lecher wires or an absorption-type wavemeter. For an i.f. of 27 Mc. the oscillator should tune from 393 to 423 Mc. A test signal of some sort is necessary to check the mixer performance.

Wave Guides and Cavity Resonators

A wave guide is a conducting tube through which energy is transmitted in the form of electromagnetic waves. The tube is not considered as carrying a current in the same sense that the wires of a two-conductor line do, but rather as a *boundary* which confines the waves to the enclosed space. Skin effect prevents any electromagnetic effects from being evident outside the guide. The 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.

The difficulty of visualizing energy transfer without the usual closed circuit can be relieved somewhat by considering the guide as being evolved from an ordinary two-conductor line.

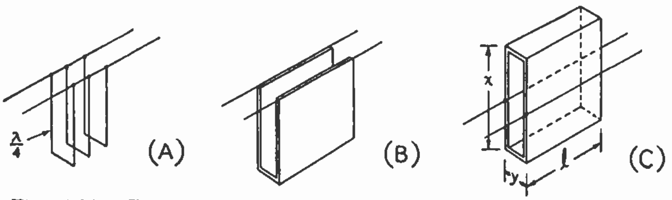


Fig. 15-26 — Evolution of a wave guide from a two-wire transmission line.

In Fig. 15-26A, 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. 15-26C. 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 analogy to wave-guide operation is not exact, and therefore should not be taken too literally. In the evolution from the two-wire line to the closed tube the electric- and magnetic-field configurations undergo considerable change, with the result that the guide does not actually operate like a two-conductor line shunted by an infinite number of quarter-wave stubs. If it did, only waves of the proper length to correspond to the stubs would be propagated through the tube, but the fact is that such waves do *not* pass through the guide. Only waves of shorter length — that is, higher frequency — can go through. The distance x represents half the *cut-off wavelength*, or the shortest wavelength that is unable to go through the guide. Or, to put it another way, waves of length equal to or greater than $2x$ cannot be propagated in the guide.

A second point of difference is that the apparent length of a wave along the direction of propagation through a guide always is greater than that of a wave of the same frequency in free space, whereas the wavelength along a two-conductor transmission line is the same as the free-space wavelength (when the insulation between the wires is air).

Operating Principles of Wave Guides

Analysis of wave-guide operation is based on the assumption that the guide material is a perfect conductor of electricity. Typical distributions of electric and magnetic fields in a rectangular guide are shown in Fig. 15-27. It will be observed that the intensity of the electric field is greatest at the center along the x dimension, diminishing to zero at the end walls. The latter is a necessary condition, since the existence of any electric field parallel to the walls at the surface would cause an infinite current to flow in a perfect conductor. This represents an impossible situation.

Zero electric field at the end walls will result if the wave is considered to consist of two separate waves moving in zigzag fashion down the guide, reflected back and forth from the end walls as shown in Fig. 15-28. Just at the walls, the positive crest of one wave meets the negative crest of the other, giving complete cancellation of the electric fields. The angle of reflection at which this cancellation occurs depends upon the width x of the guide and the length of the waves; Fig. 15-28A illustrates the case of a wave considerably shorter than the cut-off wavelength, while B shows a longer wave. When the wavelength equals the cut-off value, the two waves simply bounce back and forth between the walls and no energy is transmitted through the guide.

The two waves travel with the speed of light, but since they do not travel in a straight line the energy does not travel through the guide as rapidly as it does in space. A further conse-

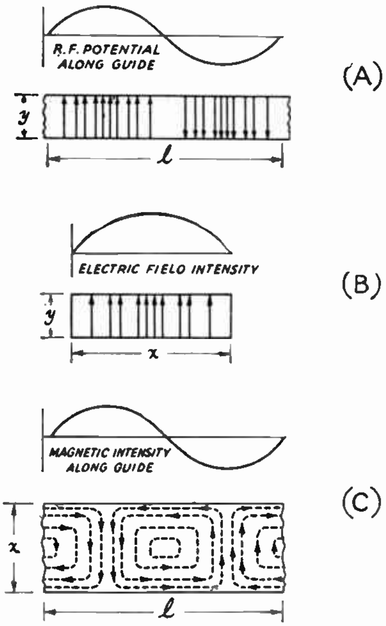


Fig. 15-27 — Field distribution in a rectangular wave guide. The $TE_{1,0}$ mode of propagation is depicted.

quence of the repeated reflections is that the points of maximum intensity or wave crests are separated more along the line of propagation in the guide than they are in the two separate waves. In other words, the wavelength in the guide is greater than the free-space wavelength. This is also shown in Fig. 15-28.

Modes of Propagation

Fig. 15-27 represents a relatively simple distribution of the electric and magnetic fields. There is in general an infinite number of ways in which the fields can arrange themselves in a guide so long as there is no upper limit to the frequency to be transmitted. Each field configuration is called a *mode.* All modes may be separated into two general groups. One group, designated *TM* (*transverse magnetic*), has the magnetic field entirely transverse to the direction of propagation, but has a component of electric field in that direction. The other type, designated *TE* (*transverse electric*) has the electric field entirely transverse, but has a component of magnetic field in the direction of propagation. TM waves are sometimes called *E* waves, and TE waves are sometimes called *H* waves, but the TM and TE designations are preferred.

The particular mode of transmission is identified by the group letters followed by two subscript numerals; for example, $TE_{1,0}$, $TM_{1,1}$, etc. The number of possible modes increases with frequency for a given size of guide. There is only one possible mode (called the *dominant* mode) for the lowest frequency that can be transmitted. The dominant mode is the one generally used in practical work.

Wave-Guide Dimensions

In the rectangular guide the critical dimension is x in Fig. 15-26; this dimension must be more than one-half wavelength at the lowest frequency to be transmitted. In practice, the y dimension usually is made about equal to $\frac{1}{2}x$ to avoid the possibility of operation at other than the dominant mode.

Other cross-sectional shapes than the rectangle can be used, the most important being the circular pipe. Much the same considerations apply as in the rectangular case.

Wavelength formulas for rectangular and circular guides are given in the following table, where x is the width of a rectangular guide and r is the radius of a circular guide. All figures are in terms of the dominant mode.

	Rectangular	*Circular*
Cut-off wavelength.......	$2x$	$3.41r$
Longest wavelength transmitted with little attenuation................	$1.6x$	$3.2r$
Shortest wavelength before next mode becomes possible.................	$1.1x$	$2.8r$

Cavity Resonators

At low and medium radio frequencies resonant circuits usually are composed of "lumped" constants of L and C; that is, the inductance is concentrated in a coil and the capacitance concentrated in a condenser. However, as the frequency is increased, coils and condensers must be reduced to impracticably small physical dimensions. Up to a certain point this difficulty may be overcome by using linear circuits but even these fail at extremely high frequencies. Another kind of circuit particularly applicable at wavelengths of the order of centimeters is the *cavity resonator*, which may be looked upon as a section of a wave guide with the dimensions chosen so that waves of a given length can be maintained inside.

The derivation of one type of cavity resonator from an ordinary *LC* circuit is shown in Fig. 15-29. As in the case of the wave-guide derivation, this picture must be accepted with some reservations, and for the same reasons.

Considering that even a straight piece of wire has appreciable inductance at very-high fre-

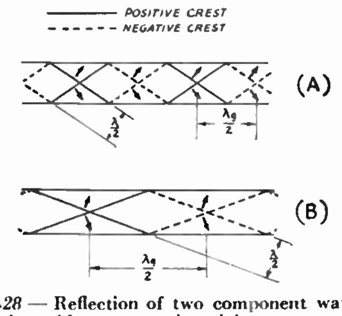


Fig. 15-28 — Reflection of two component waves in a rectangular guide. λ = wavelength in space, λg = wavelength in guide. Direction of wave motion is perpendicular to the wave front (crests) as shown by the arrows.

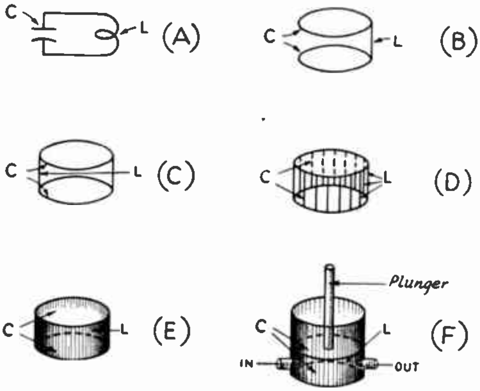


Fig. 15-29 — Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.

quencies, it may be seen in Fig. 15-29A 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. 15-29C, 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, 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 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 vacuum tube of suitable design, to form the main components of an oscillator circuit which will be capable of functioning at frequencies considerably beyond the maximum limits possible when conventional tubes, coils and condensers are employed.

Other shapes than the cylinder may be used as resonators, among them the rectangular box, the sphere, and the sphere with re-entrant cones, as shown in Fig. 15-30. The resonant frequency depends upon the dimensions of the cavity and the mode of oscillation of the waves (comparable to the transmission modes in a wave guide). For the lowest modes the resonant wavelengths are as follows:

Cylinder	$2.61r$
Square box	$1.41l$
Sphere	$2.28r$
Sphere with re-entrant cones	$4r$

The resonant wavelengths of the cylinder and square box are independent of the height when the height is less than a half-wavelength. In other modes of oscillation the height must be a multiple of a half-wavelength as measured inside the cavity. Fig. 15-29F shows how a cylindrical cavity can be tuned when operating in such a mode. Other tuning methods include placing adjustable tuning paddles or "slugs" inside the cavity so that the standing-wave pattern of the electric and magnetic fields can be varied.

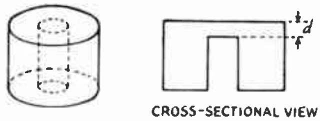


Fig. 15-31 — Re-entrant cylindrical cavity resonator.

A form of cavity resonator in wide practical use is the re-entrant cylindrical type shown in Fig. 15-31. It is useful in connection with vacuum-tube oscillators of the types described for u.h.f. use earlier in this chapter. In construction it resembles a concentric line closed at both ends with capacitance loading at the top, but the actual mode of oscillation may differ considerably from that occurring in coaxial lines. The resonant frequency of such a cavity depends upon the diameters of the two cylinders and the distance *d* between the ends of the inner and outer cylinders.

Compared to ordinary resonant circuits, cavity resonators have extremely-high Q. A value of Q of the order of 1000 or more is readily obtainable, and Q values of several thousand can readily be secured with good design and construction.

Coupling to Wave Guides and Cavity Resonators

Energy may be introduced into or abstracted from a wave guide or resonator by means of either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods for coupling to which

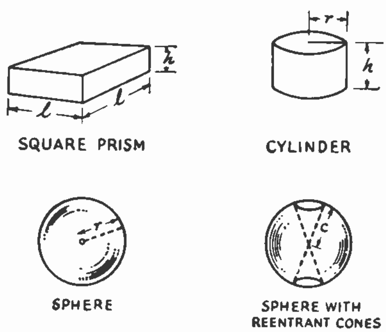
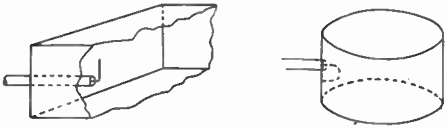


Fig. 15-30 — Forms of cavity resonators.



(A) (B)

Fig. 15-32 — Coupling to wave guides and resonators.

are shown in Fig. 15-32. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

Coupling can be varied by turning either the probe or loop through a 90-degree angle. When the probe is perpendicular to the electric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its least possible value.

Amateur Microwave Technique

All the microwave bands allotted to amateurs have been used experimentally for communication purposes. Complete description of the equipment used is beyond the scope of this text, but reference is made to various articles which have appeared in *QST*, describing the gear devised by the amateur pioneers in this field.

For the experimentally inclined, our microwave assignments represent a challenge to amateur ingenuity. Who can say but greater use of these frequencies will repeat past history, turning up propagation peculiarities and potential uses which will make these bands as coveted a region as our "communication frequencies" are considered today?

The first amateur microwave communication was carried on by Merchant and Harrison, W6BMS/2 and W2LGF, who assembled the gear shown in Fig. 15-33 in time to communicate with each other on November 15, 1945 — the date that the microwave bands were officially opened to amateur experimentation. They used two klystron tubes, one as a frequency-modulated transmitter oscillator, the other as a local oscillator for receiving. The latter worked in conjunction with a crystal mixer, into a 30-Mc. i.f. in the form of an FM receiver.

The 2300-Mc. amateur assignment was first used for communication by Koch and Floyd, W9WHM/2 and W6OJK/2, who used lighthouse tubes in simple transceivers, both of which are shown in Fig. 15-34. Their antenna systems used parabolic reflectors, one being made of wire screening attached to a wooden frame, and the other, also shown in the photograph, was simply an electric-heater assembly, with the microwave dipole substituted for the heater element.

Amateur communication on 10,000 Mc. was first accomplished by Atwater and McGregor, W2JN and W2RJM, who modified 723-A/B klystrons to permit their operation in the amateur band. They are shown, with one of their equipment set-ups, in Fig. 15-35. A somewhat similar arrangement was used by W4HPJ/3 and W6IFE/3 to extend the distance record, and has since been employed by W6IFE in opening the 3300-Mc. band to amateur use, except that the tube used in the latter instance was a 707-B with an external cavity.

The highest frequency ever used in amateur work is 21,000 Mc., first employed by Sharbaugh and Watters, W1NVL/2 and W9SAD/2, whose laboratory set-up is shown in Fig. 15-36. The r.f. generator, for transmitting and receiving, was a developmental tube designated as the Z-668, a velocity-modulated tube of the reflex type. Communication was carried on, two-way, over a distance of 800 feet.

A list of *QST* references, arranged according to the amateur band concerned, follows. It should be emphasized that the equipment

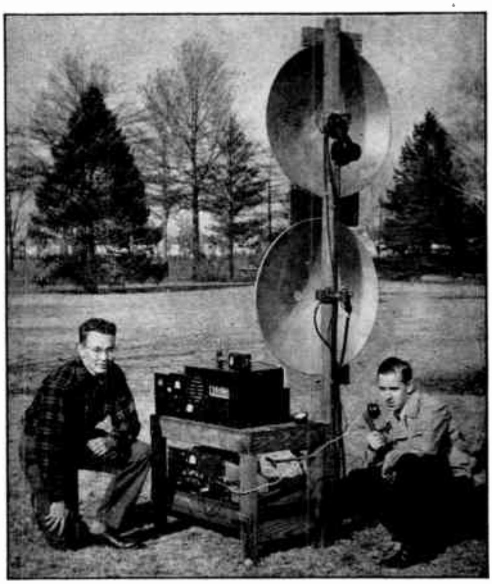


Fig. 15-33 — The first amateur microwave communication was accomplished by W6BMS (left) and W2LGF, who used two sets of similar equipment to open the 5300-Mc. amateur band on November 15, 1945, the date that the first microwave bands were released for amateur use.

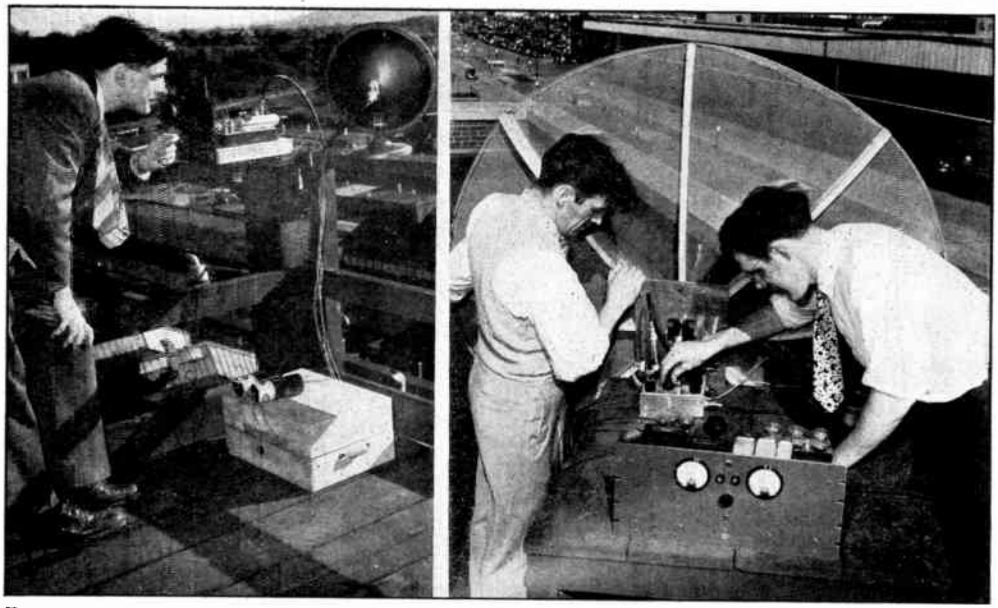


Fig. 15-34 — The 2300-Mc. band was first employed for amateur communication by W6OJK (left) and W9WHM (extreme right). Antenna systems employed a standard electric-heater unit and a handmade screen-lined parabola.

described in these reports is experimental in nature. In most instances it represents only one of several ways in which microwave communication equipment might be built. The distances covered in the pioneering work just mentioned are not, for the most part, indicative of the maximum working range, since exploration of the particular band in question was the end in view when the experiments were conducted, rather than the covering of any long distances.

Bibliography

1215 Mc. — "World Above 50 Mc." (W1BBM), May 1947 *QST*, page 136; also July 1947 *QST*, page 136. Sulzer and Ammerman, "An Oscillator for the 1215-Mc. Band," April 1948 *QST*, page 16.

2300 Mc. — Koch and Floyd, "CQ — 2400 Mc.," July 1946 *QST*, page 32. Also "World Above 50 Mc." (W6IFE), Aug. 1947 *QST*, page 128.

3300 Mc. — "World Above 50 Mc." (W6IFE), Aug. 1947 *QST*, page 128.

Fig. 15-35 — W2RJM (left) and W2JN, with one of the equipments used in pioneering work on 10,000 Mc.

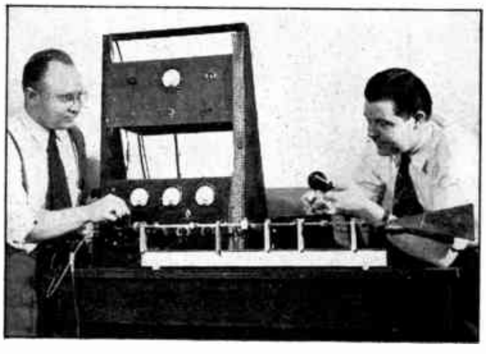


Fig. 15-36 — W1NVL (left) and W9SAD with the equipment used to work a distance of 800 feet on the highest frequency ever used for amateur communication — 21,000 Mc. Antenna systems employed a parabolic reflector at one end and a horn radiator at the other.

5250 Mc. — Merchant and Harrison, "Duplex 'Phone on 5300 Mc.," Jan. 1946 *QST*, page 19.

10,000 Mc. — McGregor and Atwater, "Dishing Out the Milliwatts on 10 KMc.," Feb. 1947 *QST*, page 58. Also "World Above 50 Mc." (W4HPJ/3, W6IFE/3), Sept. 1946 *QST*, page 152.

21,000 Mc. — Sharbaugh and Watters, "Our Best DX — 800 Feet!" Aug. 1946 *QST*, page 19.

Measuring Equipment

To comply with FCC regulations it is necessary that the amateur station be equipped to make a few relatively simple measurements. For example, the regulations require that means be available for checking the transmitter frequency to make sure that it is inside the band. This means must be independent of the frequency control of the transmitter itself; it is not enough to depend on, say, the calibration of a crystal in the crystal-controlled oscillator that drives the transmitter. In addition, it is necessary to make sure that the plate power input to the final stage of the transmitter does not exceed one kilowatt. The regulations also impose certain requirements with respect to plate-supply filtering, stability and purity of the transmitted signal, and depth of modulation in the case of 'phone transmission.

In many cases all these measurements can be made to a satisfactory degree of accuracy with no more auxiliary equipment than the regular station receiver. However, a better job usually can be done by building and calibrating some relatively simple test gear. Too, the progressive amateur is interested in instruments as an aid to better performance.

Frequency Measurement

Frequency-measuring equipment can be divided into two broad classes: oscillators of various types generating signals of known frequency that can be compared with the signal whose frequency is unknown, and adjustable resonant circuits.

Instruments in the first classification are the more accurate. Two types are commonly used by amateurs, the **secondary frequency standard** and the **heterodyne frequency meter.** The secondary frequency standard, nearly always crystal-controlled, usually generates a frequency of 100 kc. and employs a circuit that is rich in harmonic output. As a result, it supplies a series of frequencies, all multiples of 100 kc., which provides accurate calibration points throughout the communications spectrum. The more elaborate instruments of this type are provided with frequency dividers (multivibrators) to supply intermediate calibration points; a divisor commonly used is 10, thus furnishing signals at intervals of 10 kc. when the fundamental frequency is 100 kc.

The heterodyne frequency meter is a variable-frequency oscillator which is calibrated in frequency against a secondary standard or by other means. The oscillator usually is designed to cover the lowest frequency band in which measurements are to be made; measurements then can be made in higher frequency bands by using the harmonic output of the oscillator. For example, when the oscillator is set to 3560 kc. its second harmonic is 7120 kc., its fourth harmonic is 14,240 kc., and so on. The proper frequency reading is determined by knowing the fundamental frequency of the oscillator and the number of the harmonic that falls in the desired frequency range.

Both the secondary standard and the heterodyne meter are ordinarily used in conjunction with a receiver, the signals from the instruments being picked up just as though they were from distant stations. In the case of the secondary standard, the frequency of the unknown signal can be determined by locating it between two known 100-kc. or 10-kc. multiples. With the heterodyne meter, the frequency is measured by adjusting the frequency meter until its signal is at zero beat with the signal of unknown frequency, after which the frequency can be read from the frequency-meter calibration.

Since the secondary standard operates on a fixed frequency and can be crystal-controlled, its accuracy can be quite high. However, it simply establishes a series of known frequencies at regular intervals, and thus auxiliary methods must be used for determining frequencies between the known points. The series of fixed frequencies, when they mark the edges of amateur bands (as they do if they are multiples of 100 kc.), is quite sufficient for amateur work because the information that is required is whether or not the transmitter frequency is *inside* the band limits, rather than the exact frequency itself. On the other hand, the heterodyne frequency meter, while capable of giving readings at any point in its calibrated range, is inherently less accurate than the crystal-controlled standard because of the lower stability of the variable-frequency oscillator.

In the absence of more elaborate frequency-measuring equipment, a calibrated receiver may be used to indicate the approximate frequency of the transmitter. If the receiver is well made and has good inherent stability, a

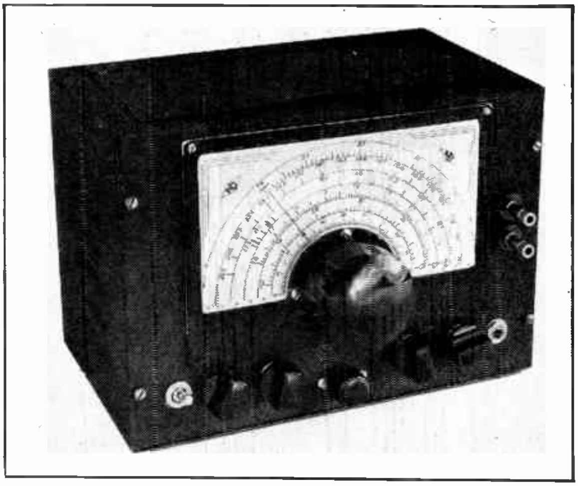


Fig. 16-1 — Heterodyne frequency meter with built-in harmonic amplifier, crystal calibrator, and detector, usable on all amateur bands up to 144 Mc. Controls along the bottom of the panel are, from left to right, crystal-oscillator on-off switch, 100/1000-kc. crystal-selector switch, calibration range switch, drift compensator, harmonic-amplifier range switch, output control, headphone jack. The two output terminals are along the right-hand edge.

bandspread dial calibration can be relied upon to within perhaps 0.2 per cent. For most accurate measurement maximum response in the receiver should be determined by means of a carrier-operated tuning indicator (such as an S-meter), the receiver beat oscillator being turned off.

When checking the transmitter frequency the receiving antenna should be disconnected, so that the signal will not overload or "block" the receiver. If the receiver still blocks without an antenna the frequency may be checked by turning off the power amplifier and tuning in the oscillator alone.

HETERODYNE FREQUENCY METER AND CRYSTAL CALIBRATOR

The basis of the heterodyne frequency meter is a completely-shielded oscillator with a precise frequency calibration. The oscillator must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

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 the mechanical construction is flimsy. Inherent frequency stability can be improved by avoiding the use of phenolic compounds and thermoplastics (bakelite, polystyrene, etc.) in the oscillator circuit, employing only high-grade ceramics instead. Plug-in coils ordinarily are not acceptable; instead, a solidly-built and firmly-mounted tuned circuit should be permanently installed. The oscillator panel and chassis should be as rigid as possible.

A stable oscillator circuit suitable for use in a heterodyne frequency meter is the electron-coupled circuit. It is possible to take output from the plate with but negligible effect on the frequency of the oscillator, and strong harmonics are generated in the plate circuit.

The heterodyne frequency meter shown in Figs. 16-1 to 16-4, inclusive, combines a number of features that make it suitable for accurate frequency measurement in the amateur bands from 3.5 to 144 Mc. As shown in the circuit diagram, Fig. 16-3, it consists of a 6SK7 electron-coupled oscillator followed by a 6AC7 amplifier that is used to intensify the higher-frequency harmonics. A second 6SK7 oscillator, using a crystal of the type that operates at either 100 or 1000 kc., provides check-points and a means for calibration of the frequency meter. A 6SL7 is incorporated to amplify the crystal harmonics and to provide a detector circuit in which the outputs of the crystal and e.c. oscillators can be mixed for calibration purposes. The detector also enables direct checking of the transmitter frequency.

The fundamental tuning range of the heterodyne oscillator is from 3500 to 4000 kc. By means of S_1 this range can be changed to 3500–3720 kc., approximately, so that the eighth harmonic just covers the 28–29.7-Mc. band. This avoids excessively critical tuning at the higher frequencies. The main tuning condenser, C_2, is connected across all of L_1 for the larger range and is connected to a tap on L_1 for the smaller to increase the bandspread. Simultaneously, an adjustable padding condenser, C_1, is switched in so that the oscillator frequency will be exactly 3500 kc. with C_2 set at maximum capacitance regardless of the switch position. C_4 is a fixed padding condenser to make the circuit fairly high C, and C_5 is the band-setting condenser. C_3 is a small padder adjustable from the panel; its function is to permit resetting the oscillator frequency to the calibration points provided by the crystal oscillator and thus take care of drift effects.

The 6AC7 plate circuit is broadly tuned by means of switched coils resonating, with the circuit capacitances, at 144, 50 and 28 Mc., and thus increases the harmonic strength on those bands. A radio-frequency choke is connected to the fourth switch position; this gives ample signal strength at 14 Mc. and lower frequencies. R_5 makes it possible to reduce the strength of the signal from the meter to the value desired for measurement purposes.

In the crystal-oscillator circuit, S_2 changes the frequency from 100 to 1000 kc. or vice versa. In the 100-kc. position C_{14} is connected across the crystal to provide means for adjusting the frequency to exactly 100 kc.

WWV SCHEDULES

Standard radio and audio frequencies are broadcast continuously, day and night, from WWV, the station of the Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C., on the following frequencies:

Mc.	*Power (kw.)*	*Audio Freq. (cycles)*
2.5	0.7	1 and 440
5.0	8.0	1 and 440
10.0	9.0	1,440 and 4000
15.0	9.0	1,440 and 4000
20.0	8.5	1,440 and 4000
25.0	0.1	1,440 and 4000
30.0	0.1	1 and 440
35.0	0.1	1

The 1-c.p.s. modulation is a 0.005-second pulse, the beginning of which marks each second to an accuracy of one part in 1,000,000. The pulse is omitted on the 59th second of every minute.

The accuracy of the radio and audio frequencies is within one part in 50,000,000. The audio frequencies are interrupted at precisely one minute before each hour and each five minutes thereafter (59th minute, 4 minutes past hour, etc.): they are resumed in precisely one minute. During each silent interval the time is given in telegraphic code. A station announcement is given in voice on the hour and half hour.

As shown in Figs. 16-2 and 16-4, the frequency meter is built on a chassis folded from a piece of sheet aluminum, the dimensions being 9 inches wide by 5½ inches deep by 2 inches high. Half-inch lips are bent along the bottom edges of the walls to make the chassis more rigid. The cabinet into which the meter fits is 10 by 7 by 6 inches. The main tuning condenser, C_2, is mounted on an aluminum bracket above the chassis and the coil, L_1, is similarly mounted below it. The bandsetting condenser, C_5, is mounted on the chassis behind the coil, with its shaft protruding through the chassis for screwdriver adjustment. Trimmer C_3 is mounted on the panel and is adjusted by a knob underneath the main tuning dial. The coil is shielded from the amplifier section by the small aluminum baffle shown in Fig. 16-4. The bandspread padder, C_1, is mounted to the left of the oscillator range switch and, like C_5, is screwdriver-adjusted from the top of the chassis. Wiring in the oscillator tuned circuit, including the switch, should be short, direct, and as rigid as possible.

The 100-kc. oscillator trimmer, C_{14}, does not require frequent adjustment and is therefore mounted on the rear edge of the chassis, close to the crystal unit. C_{16}, the plate tuning condenser for 1000 kc., is adjusted from the top of the chassis and is mounted to the right of the crystal-oscillator socket in Fig. 16-4.

In putting the instrument into operation, the crystal oscillator should be checked first. Connect a length of wire to the crystal output terminal (from C_{18}) and listen on a receiver over the range from 3.5 to 5 Mc. With S_2 in the 1000-kc. position, signals should appear at 4000 and 5000 kc., and with S_2 in the 100-kc. position signals should be heard every 100 kc. Tune in WWV on 5000 kc., wait for the modulation to go off, and then adjust C_{14} for zero beat. This sets the oscillator to precisely 100 kc. In the 1000-kc. position there may be a difference of a few kilocycles between the frequency of WWV and the 5-Mc. harmonic, but this is not a serious condition since the 1000-kc. crystal-oscillator section is used only as an aid in identification of the 100-kc. harmonics.

To set the range of the e.c. oscillator, put S_2 in the 1000-kc. position, plug a pair of 'phones into J_1, set S_2 on the maximum range position (C_2 across all of L_1), and set C_2 near minimum capacitance. Adjust C_5 until the 4000-kc. harmonic is heard. Then switch S_2 to 100 kc. and tune C_2 toward maximum, counting off five additional 100-kc. signals. C_5 may then be readjusted to bring the 3500-kc. marker close to the end of the tuning-dial scale. The 100-kc. points may then be marked off on the scale or the readings recorded. The second tuning range is adjusted by setting C_2 at 3500 kc. on the *first* range, then setting S_1 so that C_2 is connected to the tap, and adjusting C_1 (without touching C_2) so that the 3500-kc. marker is brought to the same point on the dial. The second range may be calibrated by the 100-kc. points in the same way as the first.

Calibration points may be obtained between the 100-kc. markers on both ranges by

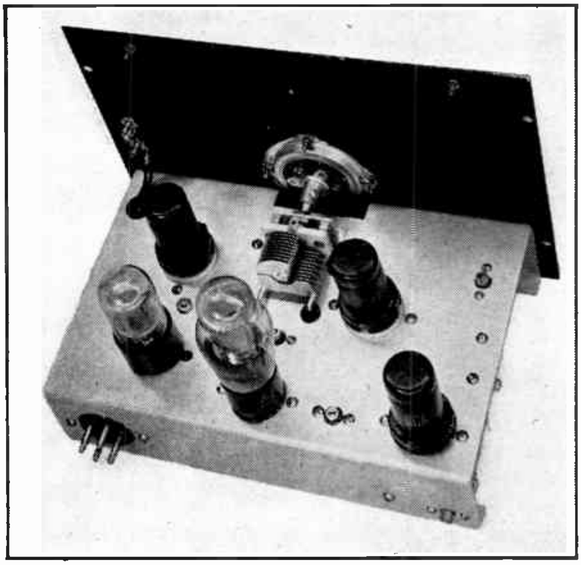


Fig. 16-2 — Inside view of the heterodyne frequency meter. The main tuning condenser is in the center with the e.c. oscillator tube to the right and the 6AC7 to the left. Along the rear edge, left to right, are the 6SL7 harmonic amplifier-detector, voltage regulator, and crystal-oscillator tube.

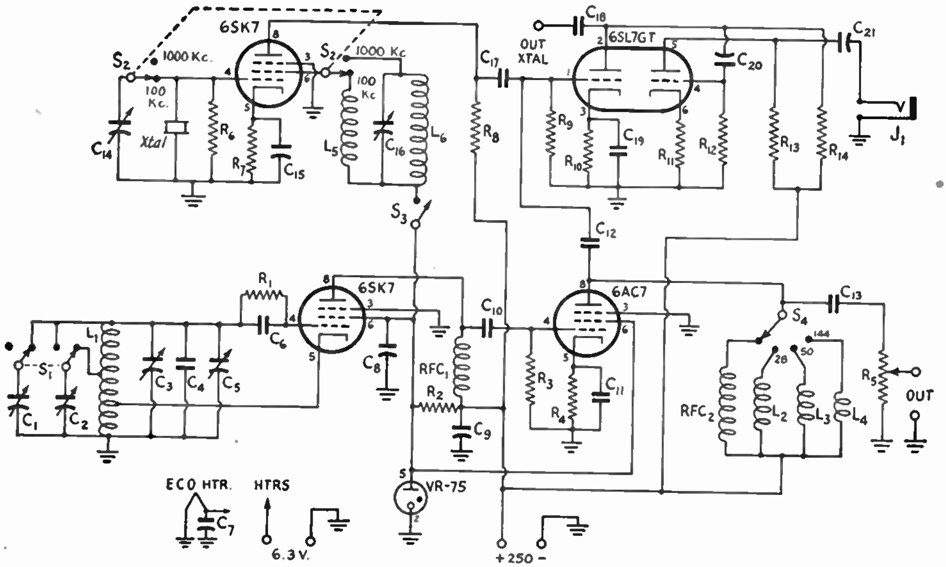


Fig. 16-3 — Circuit diagram of the heterodyne frequency meter.

C_1, C_5 — 75-μμfd. variable.
C_2, C_{16} — 100-μμfd. variable.
C_3, C_{14} — 25-μμfd. variable.
C_4 — 220-μμfd. mica.
C_6, C_{10}, C_{13} — 100-μμfd. mica.
C_7, C_8, C_9, C_{15}, C_{20}, C_{21} — 0.01-μfd. paper.
C_{11}, C_{19} — 470-μμfd. mica.
C_{12} — 10-μμfd. mica.
C_{17} — 0.001-μfd. mica.
C_{18} — 47-μμfd. mica.
R_1, R_3, R_9, R_{12} — 0.47 megohm, ½ watt.
R_2 — 10,000 ohms, 1 watt.
R_4 — 330 ohms, 1 watt.
R_5 — 25,000-ohm potentiometer.
R_6 — 4.7 megohms, ½ watt.
R_7 — 470 ohms, 1 watt.
R_8 — 0.22 megohm, 1 watt.
R_{10} — 10,000 ohms, 1 watt.
R_{11} — 1500 ohms, 1 watt.
R_{13}, R_{14} — 0.1 megohm, ½ watt.
L_1 — 18 turns No. 18 on 1-inch form, length 1½ inches. Cathode tap 5 turns from ground end; bandspread tap 11 turns from ground.
L_2 — 24 turns No. 18 enam. close-wound on ¼-inch form.
L_3 — 11 turns No. 18 enam. close-wound on ¼-inch form.
L_4 — 2 turns No. 16 spaced ½ inch, diameter ¼ inch.
L_5 — 8-mh. coil (r.f. choke).
L_6 — 1 pie of 4-pie 2.5-mh. r.f. choke.
J_1 — Open-circuit jack.
RFC_1, RFC_2 — 2.5-mh. r.f. choke.
S_1 — 2-position 2-pole ceramic wafer switch.
S_2 — 2-position 2-pole switch (bakelite insulation satisfactory).
S_3 — S.p.s.t. toggle.
S_4 — 4-position 1-pole ceramic wafer switch.
XTAL — 100/1000-kc. crystal unit (Bliley SMC-100).

using a receiver as an auxiliary. For example, if the receiver is adjusted to pick up the fifth harmonic of the e.c. oscillator (17.5 to 20 Mc.) and the harmonic is beat against 100-kc. points from the crystal oscillator in that range, 100-kc. intervals on the fifth harmonic will give 20-kc. intervals on the fundamental. With a straight-line capacitance condenser at C_2, the relationship between dial divisions and frequency is almost linear, and marking off the dial at the proper intervals between actual calibration points will result in a calibration of sufficient accuracy.

The various amateur bands are covered by the following harmonics: 3.5–4 Mc., fundamental; 7–7.3 Mc., 2nd harmonic; 14–14.4 Mc., 4th; 27.185–27.245 Mc., 7th; 28–29.7 Mc., 8th; 50–54 Mc., 14th; 144–148 Mc., 40th. At lower frequencies a short length of wire connected to the output terminal will give ample signal strength under average conditions, but in the v.h.f. range closer coupling — such as running the wire in close proximity to the receiving antenna lead, or actually connecting it to the antenna post through a small fixed condenser — may be necessary to get a good signal.

With an instrument of this type the edges of amateur bands may be quite accurately determined, if care is used in setting the 100-kc. oscillator to WWV and equal care is used in setting the e.c. oscillator scale to the 100-kc. crystal points. C_3 may be used for the latter purpose each time the meter is used, and particularly during the first 30 minutes or so of operation when the temperature of the equipment is rising. The accuracy at intermediate points will depend upon the accuracy of the original calibration; it should be possible to read within 0.05 per cent under normal conditions by using the "drift corrector," C_3.

● ABSORPTION FREQUENCY METERS

The simplest possible frequency-measuring device is a resonant circuit, tunable over the desired frequency range and having its tuning dial calibrated in terms of frequency. Such a frequency meter operates by extracting a small amount of energy from the oscillating circuit to be measured, the frequency then being determined by tuning the frequency-meter circuit to resonance and reading the frequency from the calibrated scale. This method is not capable of as high accuracy as the heterodyne methods for two reasons: First, the resonance indication is relatively "broad" as compared to the zero beat of a heterodyne; second, the necessarily close coupling between the frequency meter and the circuit being measured causes some detuning in both circuits, with the result that the calibration of the frequency-meter circuit depends to some degree on the coupling to the circuit being measured.

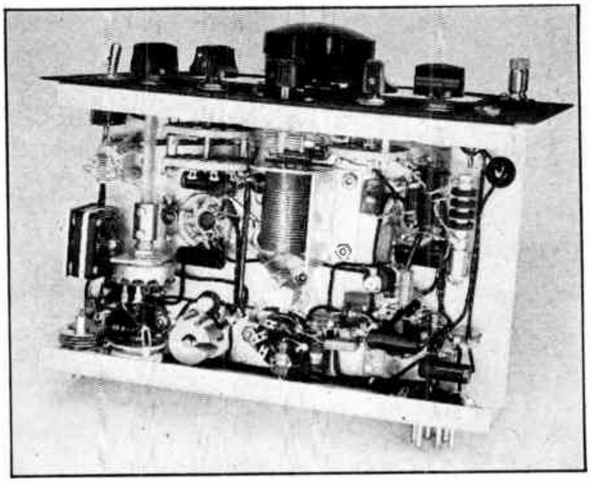


Fig. 16-4 — Underneath the chassis of the heterodyne frequency meter. The parts layout is discussed in the text.

It is necessary to have some means for indicating resonance with an absorption frequency meter. When such a meter is used for checking a transmitter, the plate current of the tube connected to the circuit being checked can provide the resonance indication. When the frequency meter is tuned through resonance the plate current will rise, and if the frequency meter is loosely coupled to the tank circuit the plate current will simply give a slight upward flicker as the meter is tuned through resonance. The greatest accuracy is secured when the loosest possible coupling is used.

A receiver oscillator may be checked by tuning in a steady signal and heterodyning it to give a beat note as in ordinary c.w. reception. When the frequency meter is coupled to the oscillator coil and tuned through resonance the beat note will change. Again, the coupling should be made loose enough so that a just-perceptible change in beat note is observed when the meter is tuned through resonance.

Although the absorption-type frequency meter should not be depended upon for accurate measurement, it is a highly-useful instrument to have in the station even when better frequency-measuring equipment is available. Since it generates no harmonics itself, it will respond only to the frequency to which it is tuned. It is therefore indispensable for distinguishing between fundamental and various harmonics, and for detecting harmonics and parasitic oscillations. When provided with a sensitive resonance indicator it is also useful for detecting r.f. in undesired places such as power wiring, for making rough measurements of field strength in adjustment of antennas, and can likewise be used as a modulation monitor.

An approximate calibration — usually sufficient — may be obtained by comparison with a calibrated receiver. The usual receiver dial calibration is sufficiently accurate. A simple oscillator circuit covering the same range as the frequency meter will be useful in calibration. Set the receiver to a given frequency, tune the oscillator to zero beat at the same frequency, and adjust the frequency meter to resonance with the oscillator as described above. This gives one calibration point. When a sufficient number of such points has been obtained a graph may be drawn to show frequency *vs.* dial settings on the frequency meter.

A SENSITIVE ABSORPTION FREQUENCY METER

Figs. 16-5 to 16-7, inclusive, show an absorption frequency meter or "wavemeter" with a crystal-detector/milliammeter resonance indicator that provides a relatively high degree of sensitivity. As shown in Fig. 16-6, a

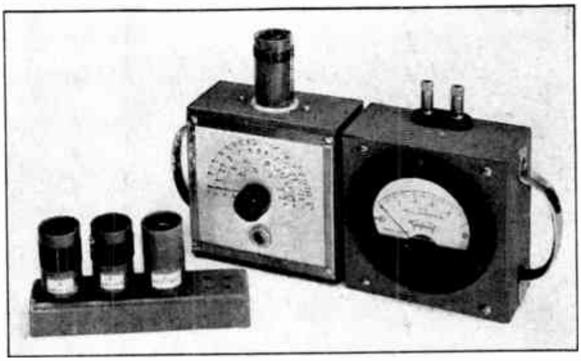


Fig. 16-5 — A sensitive absorption-type frequency meter with a crystal-detector rectifier and a d.c.-milliammeter indicating circuit. The meter is housed in a separate compartment so that it may be used with other measuring devices. The cabinet and front cover are drilled and tapped to accommodate the mounting screws for a large-size chart frame; frequency calibrations are marked on cardboard held in place by the chart frame. A short strip of wood, drilled to match the coil-form prongs, is used as a rack for the coils. Meter-box connections are shown in Fig. 16-15.

resonant circuit is connected in series with a crystal detector and a 0–1 milliammeter. The tank coil, L_1, serves as the pick-up coil, and the crystal is tapped down on the inductance in order to improve the sensitivity and selectivity of the meter. Plug-in coils are provided so that the unit covers a frequency range from about 1 megacycle to 165 megacycles. Any type of fixed crystal detector may be used, but the v.h.f. types are recommended. The meter box shown at the right in Fig. 16-5 is the same unit

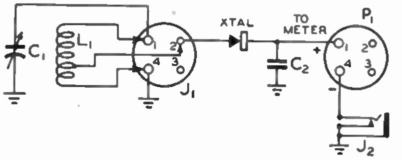


Fig. 16-6 — Circuit diagram of the absorption-type frequency meter.

C_1 — 140-μμfd. variable (Millen 22140).
C_2 — 0.0015-μfd. midget mica.
L_1 — 1.22–4.0 Mc.: 70 turns No. 32 enameled wire, 1-inch diam., 5/8 inch long. Tap 12½ turns from grounded end.
— 4.0–13.5 Mc.: 20 turns No. 20 enameled wire, 1-inch diam., 9/16 inch long. Tap 4½ turns from grounded end.
— 13.2–44.0 Mc.: 5 turns No. 20 enameled wire, 1-inch diam., 5/16 inch long. Tap 1½ turns from grounded end.
— 39.8–165 Mc.: Hairpin loop of No. 14 wire, ½-inch spacing, 2 inches long (total length including ends which fit down into the coil-form prongs). Tap 1⅝ inches from grounded end.
All four coils wound on Millen 45004 coil forms.
J_1 — 4-prong tube socket.
J_2 — Closed-circuit jack.
P_1 — 4-prong male plug.
XTAL — Type 1N34.

that is used with the volt-ohm-milliammeter described later in this chapter.

The frequency meter is housed in a 2 × 4 × 4-inch metal box, the milliammeter being mounted in a separate box of the same size. The coil socket is on the top near the front edge, with the tuning condenser just below it inside the case. This arrangement keeps the tuned-circuit leads short. A headphone jack is provided for monitoring 'phone transmissions. The unit may be calibrated as described in the preceding section.

A two- or three-foot antenna rod may be added to the unit to permit using the instrument for field-strength measurements. The antenna should be connected to the top end of the tank coil, L_1. The rod antenna may be undesirable when the frequencies of individual simultaneously-operating circuits are to be checked — as in the case of a multistage transmitter with frequency multipliers — because the antenna increases the sensitivity to such an extent that it may be difficult to identify the output of a particular circuit. It may be convenient to interconnect the two units by means of a length of lamp cord or coaxial cable of any reasonable length (up to several hundred feet) when the meter is being used as a field-strength measuring device.

In addition to the uses mentioned in the preceding section, a meter of this type may be used for final adjustment of neutralization in r.f. amplifiers. For this purpose it may be loosely coupled to the plate tank coil. Alternatively, L_1 may be removed and the final-amplifier link output terminals connected to Prongs 2 and 4 in the coil socket. The latter method tends to ensure that the pick-up is from the final tank coil only.

LECHER WIRES

At very-high and ultrahigh 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 wires.** Such a line shows pronounced resonance effects, and it is possible to determine quite accurately the current loops (points of maximum current). The physical distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read directly in meters (39.37 inches = 1 meter; 0.3937 inch = 1 cm.), or in centimeters for the very-short wavelengths.

The Lecher-wire line should be at least a wavelength long — that is, 7 feet or more on 144 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. The spacing between wires should be about 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 per-

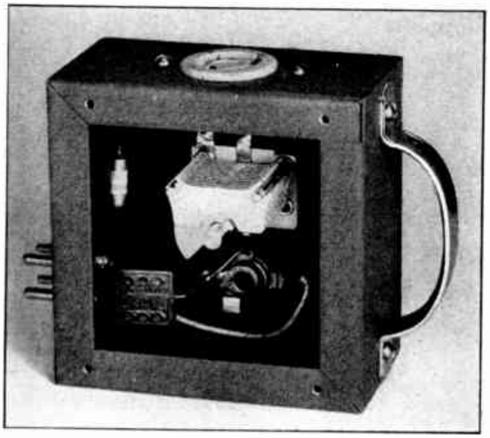


Fig. 16-7 — A rear view of the absorption-type frequency meter. The crystal is wired between the connector plug at the left and the coil socket at the top. The meter by-pass condenser is mounted between the plug and the grounded side of the 'phone jack. The variable-condenser terminals are connected directly to the coil socket.

manent fashion and provided with a shorting bar maintained at right angles to the wires (Fig. 16-8). The support may consist of two pieces of "1-by-2" 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, 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.

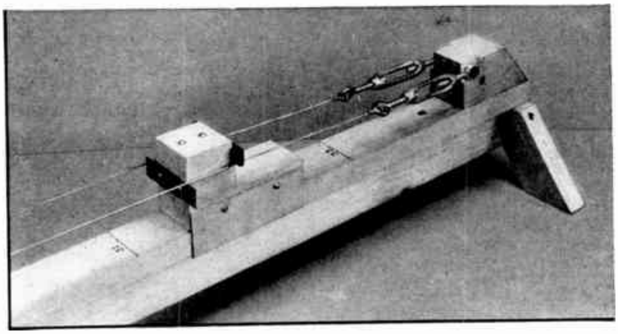


Fig. 16-8 — **One end of a typical Lecher wire system. The feet at each end keep the assembly from tipping over when in use. The wires terminate in airplane-type strain insulators at one end, and at the other in small turnbuckles for maintaining tension. The wire is No. 16 bare solid-copper antenna wire (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.**

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. 16-9. 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 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

$$F_{\text{Mc.}} = \frac{5905}{\textit{length} \text{ (inches)}}$$

If the length is measured in meters,

$$F_{\text{Mc.}} = \frac{150}{\textit{length} \text{ (meters)}}$$

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.

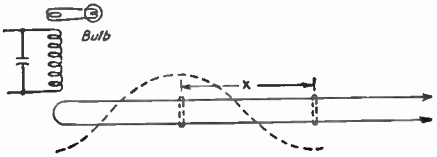


Fig. 16-9 — **Coupling a Lecher-wire system to a transmitter tank coil. Typical standing-wave distribution is shown by the dashed line. The distance, X, between the positions of the shorting bar at the current loops equals one-half wavelength.**

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.

Signal Monitoring

Every amateur station should make provision for checking the quality of the transmitter output. This requires that some means be available in the station for reproducing the conditions existing at a distant receiving station; that is, for reducing the strength of the signal from the transmitter to such a point that its characteristics can be examined without danger of false indications from overloading the receiving equipment.

The simplest method of checking the quality of c.w. transmissions is to use the regular station receiver. If the receiver is a superheterodyne the process may simply be that of reducing the r.f. gain to minimum and tuning to the transmitter frequency. If distant signals are stable and have "pure-d.c." tone in normal reception, then the local transmitter should too, when the receiver gain is reduced to the point where the receiver does not overload. If the signal is too strong with the r.f. gain "off," shorting the receiver antenna input terminals may reduce it to suitable proportions, or the mixer circuit in the receiver may be temporarily detuned to arrive at the same desired result.

An alternative method is to set the receiver on the next lower-frequency band than the one in use, then tune the receiver so that the second harmonic of its oscillator beats with the transmitter signal to produce the intermediate frequency. Higher-order harmonics also may be used for this purpose. With this harmonic method there is ordinarily no danger that the receiver will overload, because the r.f. and mixer tuned circuits are so far from resonance with the transmitter frequency. The setting of the tuning dial bears no direct relation to the transmitter frequency under these conditions, since the oscillator harmonic must maintain a constant difference with the transmitter to produce the i.f. beat.

A 'phone signal may be monitored in the same way, provided a headset is used for reception. Use of a loudspeaker is not usually practicable because the sound output feeds back to the microphone and causes howling. A crystal detector and headset may also be used for the same purpose, as described in preceding sections. In monitoring a 'phone signal the best plan is to have another person speak into the microphone rather than to listen to one's own voice. It is difficult to judge quality when speaking and listening at the same time.

C.W. SIGNAL MONITOR

One trouble with checking a c.w. signal on the station receiver is that receivers frequently have hum and instability, particularly on the higher frequencies, that cause the signal to sound worse than it really is. The best way to get a true picture of the signal is to monitor it with a battery-operated oscillator. With battery plate and filament supplies, the monitor cannot introduce hum. If the oscillator is mechanically and electrically stable and is well shielded, a true replica of the signal can be obtained.

The construction of a shielded, battery-operated monitor is described in Chapter Eight. Instructions for its use also will be found in that chapter.

MODULATION MONITOR

Fig. 16-10 is the circuit of a 'phone monitor that can be used both for aural checking and for measuring modulation percentage. When a small r.f. voltage is applied to the input circuit it is rectified by the crystal. With switch S_1 in the "r.f." position the average value of the rectified current is measured by the 0–1 milliammeter, *MA*. With the switch in the "a.f." position, the audio modulation on the signal is transferred through T_1 to a second rectifier. The average value of the rectified audio is again read by the milliammeter. The circuit constants are chosen so that if the input is adjusted to make the meter read full scale on r.f., the a.f. meter readings will be directly proportional to percentage of modulation (for voice modulation), 100-percent modulation being represented by a cur-

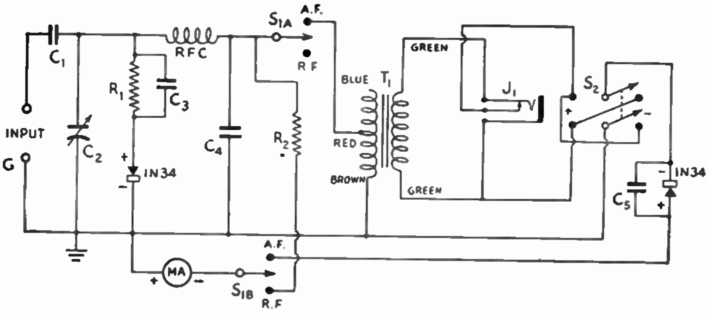


Fig. 16-10 — Circuit of direct-reading modulation meter.
C_1, C_4 — 1000-μμfd. ceramic.
C_2 — 100-μμfd. variable midget.
C_3 — 12-μμfd. mica.
C_5 — 470-μμfd. mica.
R_1 — 1100 ohms, 5%, 1 watt.
R_2 — 16,000 ohms, 5%, 1 watt.
J_1 — Closed-circuit jack.
MA — 0–1 ma., 100 ohms.
RFC — 20 μh.
S_{1A-B}, S_2 — D.p.d.t. toggle.
T_1 — Push-pull interstage transformer, 1:1 ratio (Stancor A-4711).

rent of 1 milliampere. Switch S_2 provides for reversing the "polarity" of the modulation, giving a qualitative indication of the up- and down-peaks. A headphone jack, J_1, is provided for listening to the quality of the modulation. (The percentage modulation cannot be read with 'phones plugged into J_1, so the 'phones must be removed when readings are to be taken.)

In constructing such an instrument, care should be used to prevent r.f. pick-up in the audio rectifier circuit. This can be checked by testing the instrument on an unmodulated carrier (which must be substantially hum-free); with a full-scale reading when S_1 is in the "r.f." position, the meter should read zero when S_1 is switched to "a.f." The values of R_1 and R_2 are critical and should be within 5 per cent of the recommended values.

A sample of the modulated carrier may be coupled into the instrument through a one-turn link and a length of Twin-Lead, the link being placed within a few inches of the final tank circuit of the transmitter. The coupling between the link and final tank coil must be adjusted to give a full-scale r.f. reading, after C_2 has been set for maximum reading. Alternatively, a coil that will resonate with C_2 at the operating frequency may be connected to the input terminals and the instrument located so that a suitable full-scale reading will be obtained.

Besides indicating modulation percentage, the instrument will show carrier shift (as shown by a change in the reading, when modulating, with S_1 in the "r.f." position) and thus detect nonlinearity in the modulated amplifier.

Measurement of Current, Voltage and Power

The amateur regulations require that when the power input to the final stage is above 900 watts, means must be provided for measuring the power input. This may be done by measuring the d.c. voltage applied to the final-stage plates and the d.c. current flowing to them. The instruments required are a milliammeter and voltmeter.

Although in lower-power transmitters power-input measurements are not required, it is nevertheless true that a milliammeter is an almost indispensable instrument in the amateur station. It is invaluable in the adjustment of transmitting amplifier stages; tuning a transmitter without measuring grid and plate currents is like working in the dark. A d.c. voltmeter, although not essential, is useful in conjunction with the milliammeter in determining whether tube ratings are being exceeded or not and thus is helpful in prolonging tube life.

Besides d.c. measurements, it is also well to measure the filament voltages applied to transmitting tubes. Tube performance is dependent upon proper cathode emission, which in turn depends upon the voltage applied to the filament or heater. Also, the life of some transmitting tubes, particularly the thoriated-tungsten filament types, is critically dependent upon maintaining the filament voltage within rather close limits. Since most transmitting-tube filaments are operated on a.c., an a.c. voltmeter is a worth-while addition to amateur transmitting equipment.

Adjustment of a transmitter for maximum power output to the antenna or transmission line is facilitated by the use of instruments which measure radio-frequency current. Such instruments, although not actually essential, round out the measuring equipment used in transmitter adjustment.

D.C. Instruments

D.c. 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 that 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.

If a single instrument must be used for measuring widely-different values of current or voltage, it is advisable to purchase one which will read, at about 75 per cent of full scale, the *smallest* value of current or voltage

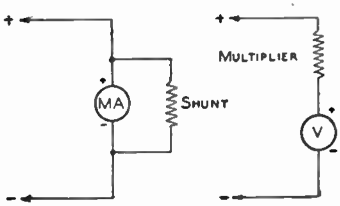


Fig. 16-11 — How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.

to be measured. Small currents cannot be read with any degree of precision on a high-scale instrument; on the other hand, the range of a low-scale instrument can be extended as desired to take care of larger values. The ranges of both voltmeters and ammeters can be extended by the use of external resistors, connected in series with the instrument in the case of a voltmeter or in shunt in the case of an ammeter. Fig. 16-11 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

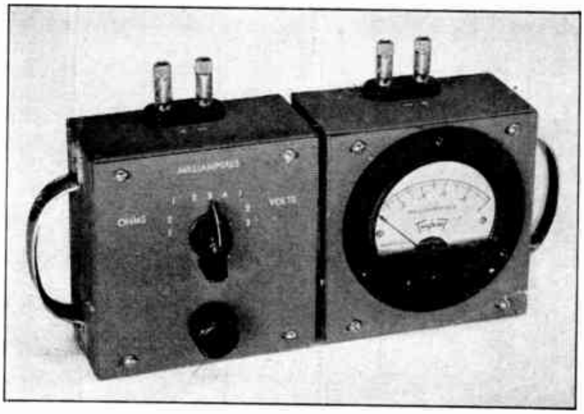


Fig. 16-12 — An inexpensive multirange volt-ohm-milliammeter. The 2 × 4 × 4-inch cabinet at the left houses the multipliers, shunts, switch and zero-adjustment resistor. The meter is mounted in the metal cabinet shown at the right. The units are provided with plugs and jacks so that the meter can be used independently or as the indicator component for other instruments. Connections to the volt-ohm-milliammeter, or to the meter alone, are made to the terminals mounted at the top of both boxes. Handles are mounted on the cabinets to facilitate handling.

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:

$$R = \frac{1000E}{I}$$

where E is the desired full-scale voltage and I the full-scale reading of the instrument in milliamperes.

To increase the current range of a milliammeter, the resistance of the shunt is

$$R = \frac{R_m}{n - 1}$$

where the symbols have the same meanings 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 in Chapter Twenty-Four gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that 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 enough current to flow through the meter to make it read full scale without the shunt; connecting the shunt should then give the correct reading on the new full-scale range.

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 fixed resistors may be used. Such resistors are supplied in tolerances of 5, 10 or 20 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 attributable to manufacturing tolerances.

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: $P = EI$. Thus the voltmeter and ammeter are also the instruments used in measuring d.c. power.

Multirange Voltmeters and Ohmmeters

A combination voltmeter-milliammeter having various ranges is extremely useful for experimental purposes and for trouble shooting in receivers and transmitters. As a voltmeter

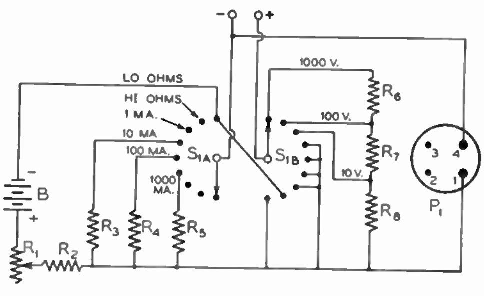


Fig. 16-13 — Diagram of the volt-ohm-milliammeter.

R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½ watt.
R_3 — 10-ma. shunt, 6.11 ohms (see text).
R_4 — 100-ma. shunt, 0.555 ohm (see text).
R_5 — 1000-ma. shunt, 0.055 ohm (see text).
R_6 — 1000-volt multiplier, 0.9 megohm, ½ watt.
R_7 — 100-volt multiplier, 90,000 ohms, ½ watt.
R_8 — 10-volt multiplier, 10,000 ohms, ½ watt.
B — 4.5-volt dry battery (Burgess 5360).
P_1 — 4-prong male plug (for milliammeter).
S_{1A-B} — 9-point 2-pole selector switch (Mallory 3229J).

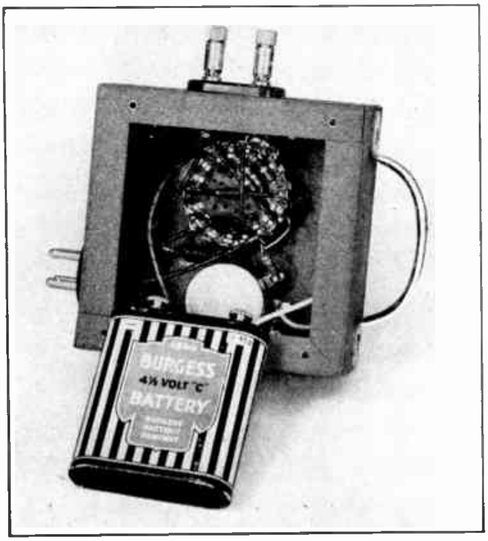


Fig. 16-14 — A rear view of the volt-ohm-milliammeter. The range-selector switch is mounted above the zero-adjustment potentiometer, and the shunts and multipliers are connected across the switch terminals. A four-prong male plug, for connection to the meter box, is shown at the left of the cabinet. The ohmmeter battery fits inside the case; the battery terminals should be insulated with tape or paper before the battery is installed in the box.

such an instrument should have high resistance so that very little current will be drawn in making voltage measurements. A voltmeter taking considerable current will give inaccurate readings when connected across a high-resistance source — as is often the case in various parts of a receiver circuit. For such purposes the instrument should have a resistance of at least 1000 ohms per volt; a 0–1 milliammeter or 0–500 microammeter (0–0.5 ma.) is the basis of most multirange meters of this type. Microammeters having a range of 0–50 μa., giving a sensitivity of 20,000 ohms per volt, also are used.

The various current ranges on a multirange instrument can be obtained by using a number of shunts individually switched in parallel with the meter. Care should be used to minimize contact resistance in the switch.

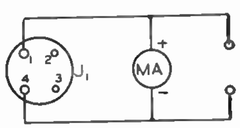


Fig. 16-15 — Wiring diagram of the 0–1 milliammeter shown in Figs. 16-5 and 16-12. J_1 is a 4-prong tube socket.

It is often necessary to check the value of a resistor or to find the value of an unknown resistance, particularly in receiver servicing. An **ohmmeter** is used for this purpose. The ohmmeter is simply a low-current d.c. voltmeter provided with a source of voltage (usually dry cells), the meter and battery being connected in series with the unknown resistance. If a full-scale deflection is obtained with the connections to the external resistance shorted, insertion of the resistance under measurement will cause the meter reading to decrease. The meter scale can be calibrated in ohms. When the resistance of the voltmeter is known, the following formula can be applied:

$$R = \frac{eR_m}{E} - R_m$$

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 resistance of the voltmeter.

Since the resistance of a voltmeter is usually rather high, this method is not well adapted to measuring low values of resistance. For very low resistances, the unknown may be connected in shunt with the instrument (not including the multiplier) instead of in series.

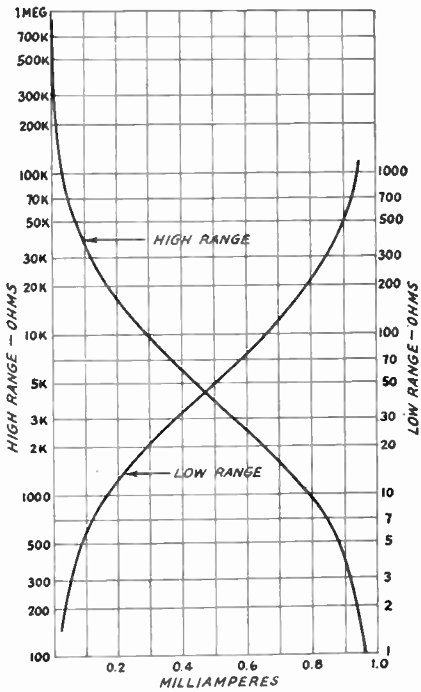


Fig. 16-16 — Calibration curve for the high- and low-resistance ranges of the volt-ohm-milliammeter.

AN INEXPENSIVE V.O.M.

A combination multirange volt-ohm-milliammeter, reduced to simple and inexpensive terms, is shown in Figs. 16-12 to 16-15. Using a 0–1 milliammeter, the voltmeter has three ranges at 1000 ohms per volt: 0–10, 100 and 1000 volts. Current ranges of 0–1, 10, 100 and 1000 ma. are provided. There are two resistance-measurement ranges, a series range that is useful up to about 0.5 megohm, and a shunt range of 0–1000 ohms.

For economy, ordinary carbon resistors are used as voltmeter multipliers. These can be obtained with an accuracy within 5 per cent. However, standard resistors of 10-per-cent tolerance can be used without introducing undue error. The 1000-volt multiplier, R_6, is two 1.8-megohm resistors in parallel, and the 100-volt multiplier, R_7, is two 0.18-megohm resistors in parallel.

The 10-, 100- and 1000-ma. shunts are made of ordinary copper magnet wire wound on ½-watt resistors of high resistance value — 10,000 ohms or higher. The approximate lengths and sizes of the wire for the shunts are as follows: R_3, 9 inches No. 38 enameled; R_4, 5 inches No. 30 enameled; R_5, 3 inches No. 18.

A calibration curve for the ohmmeter ranges is given in Fig. 16-16. With instruments having different internal resistance than the one shown in the photograph (Triplett Model 0321-1) the "low-ohms" curve will not apply exactly.

Test Oscillators and Grid-Dip Meters

A useful and inexpensive general-purpose instrument is an oscillator covering a wide frequency range. When it generates frequencies in the audio range it can be used as a signal source for checking the performance of audio amplifiers. As a radio-frequency oscillator it may be made to generate signals that can be used for receiver alignment, for calibrating absorption wavemeters as described earlier in this chapter, and for furnishing small r.f. voltages for whatever purpose may be required. When equipped with a low-range milliammeter connected to read the oscillator grid current, it becomes a grid-dip meter and may be used for checking the resonant frequencies of tuned circuits, and as a means for measuring inductance and capacitance as described in a later section.

The grid-dip meter is so called because when its oscillator is coupled to a tuned circuit, the oscillator grid current will show a decrease or "dip" when the oscillator is tuned through resonance with the unknown circuit. The reason for this is that the external circuit will absorb energy from the oscillator when both it and the oscillator are tuned to the same frequency, and the loss of energy from the oscillator circuit causes the feed-back to decrease. The decrease in feed-back is accompanied by a decrease in grid current. The dip in grid current is quite sharp when the circuit to which the oscillator is coupled has reasonably high Q.

GENERAL-PURPOSE OSCILLATOR AND GRID-DIP METER

A general-purpose test oscillator is shown in Figs. 16-17 and 16-19. This simple unit can be used as an audio oscillator, an r.f. signal generator, a grid-dip meter, a field-strength indicator, or as an absorption wavemeter, and has a frequency range of 200 cycles to 56 Mc.

As shown by the circuit diagram, Fig. 16-18, a Type 6SN7GT tube is used in a cathode-coupled oscillator circuit. The only critical values are those of the cathode resistor, R_2, and the coupling condenser, C_4. Use of a cathode resistor of less than 1000 ohms will result in a poor waveform at audio frequencies, and the oscillator output will be greatly reduced if the cathode resistor is larger than 3000 ohms. The audio attenuator, R_4, loads the circuit to some

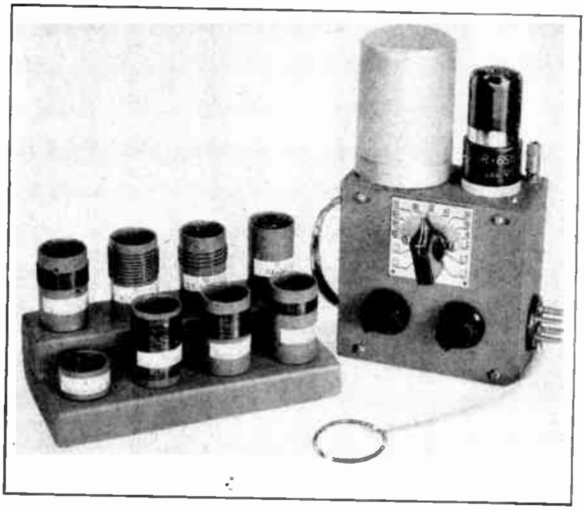


Fig. 16-17 — The general-purpose test oscillator. The variable condenser, sensitivity control and audio-output potentiometer are mounted on the front panel of a 2 × 4 × 4-inch metal box. A handle at the left of the box, and a meter jack at the right, are provided so that the oscillator may be used with the meter unit shown in Fig. 16-12. Sockets for the oscillator tube and the plug-in coils are mounted on the top of the box. The coil shield is necessary only when the oscillator is being used as a signal generator. The r.f. coils are shown mounted in a wooden rack which has been drilled to fit the prongs of the coil forms.

extent, and will prevent oscillation if the coupling condenser, C_4, is made much smaller than 0.1 μfd. The oscillator requires a filament supply of 6.3 volts at 0.6 amp., and the plate supply should deliver 150 to 250 volts at 14 ma. (the plate-current drain will be less than 10 ma. at the audio and the low radio frequencies).

The frequency of the oscillator is controlled by the values of inductance and capacitance connected across points *C* and *D* in Fig. 16-18. The coil chart lists values of capacitance and inductance that can be used for either audio- or radio-frequency output. At audio frequencies, the *LC* combination is connected across terminals *C* and *D*, and at radio frequencies the coils plug into the socket provided for this purpose. The variable condenser, C_1, is used as the frequency adjustment over the r.f. range of the oscillator, and can be used as a vernier adjustment at audio frequencies.

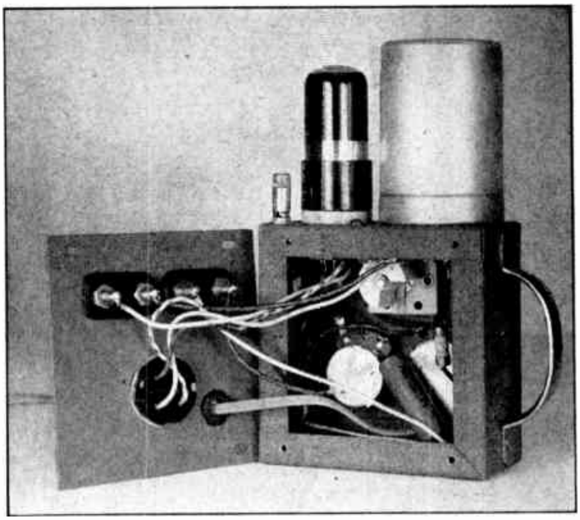


Fig. 16-19 — **A rear view of the general-purpose test oscillator. The low impedance link for the pick-up loop feeds through a rubber grommet mounted in the rear cover. Input and output jacks are also located on the backplate. The arrangement of components inside the box is not critical. The 1N34 crystal and the meter by-pass condenser, C_6, are mounted inside the box.**

A potentiometer, R_3, serves as the sensitivity control when the oscillator is used as a grid-dip meter, an absorption wavemeter, or as a field-strength indicator. This control also acts as an output attenuator when the oscillator is used as a signal generator. It should be noticed that the audio-output attenuator, R_4, in series with the output coupling condenser, C_5, is connected to the resonant circuit by means of a jumper connected between terminals *A* and *C*. This jumper must be removed when the oscillator is used at radio frequencies; otherwise the circuit will not oscillate.

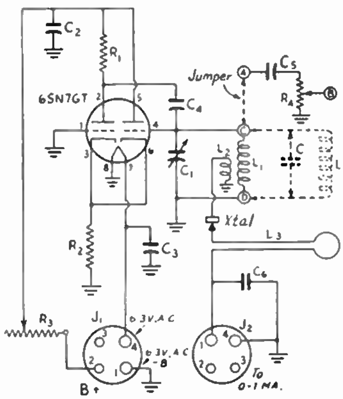


Fig. 16-18 — **Simple general-purpose test oscillator.**
C_1 — 100-μμfd. midget variable (Millen 20100).
C_2, C_3 — 0.01-μfd. 400-volt paper.
C_4, C_5 — 0.1-μfd. 400-volt paper.
C_6 — 100-μμfd. midget mica.
R_1 — 68,000 ohms, 1 watt.
R_2 — 1500 ohms, 1 watt.
R_3 — 0.2-megohm carbon potentiometer.
R_4 — 50,000-ohm carbon potentiometer.
L_1, L_2, L_3 — See text and coil table.
J_1, J_2 — 4-prong plug (Amphenol 86-CP4).
XTAL — Type 1N34.

The rectifier system consists of coil L_2, the 1N34 crystal, a pick-up loop, and a meter by-pass condenser. R.f. current generated by the oscillator circuit is coupled to the crystal rectifier by means of L_2. The current is rectified by the crystal and flows through the pick-up loop to an external meter (any 0–1 milliammeter will do). The pick-up loop, if placed in series with the crystal as shown, will not overload the oscillator circuit.

For output in the audio-frequency range, the desired *LC* combination is connected across the circuit and output connections are made to terminals *B* and *D* (the jumper is connected between *A* and *C*). If a test signal of radio frequency is desired, the jumper connection is removed and a coil is plugged in the coil socket. Terminal *C* will then serve as a short antenna from which the signal is radiated. The strength of the signal being radiated can be increased by adding a short length of wire to the output terminal. It is advisable to shield the r.f. coil for this type of operation so that hand-capacity effects will be minimized.

To use the oscillator as a grid-dip meter, connect a 0–1 milliammeter to the meter socket, J_2. With the proper r.f. coil in place, and with power turned on, adjust the sensitivity control for maximum meter deflection (full deflection will be approximately 0.5 ma. on the highest frequency range). The pick-up loop can then be coupled to the circuit under measurement and, if the oscillator frequency is varied by means of the tuning condenser, C_1, there will be a pronounced dip in grid current as the resonant frequency of the external circuit is passed.

COIL CHART FOR THE TEST OSCILLATOR

Audio Frequencies			Radio Frequencies				
Frequency (Cycles)	*Inductance (hy.) (L)*	*Capacitance (μfd.) (C)*	*Freq. (Mc.)*	*Coil*	*No. Turns*	*Wire Size*	*Length of Winding (Inches)*
200	1.2	0.02	0.955–1.75	L_1	120	No. 30 enam.	1 5/16
400	"	0.06		L_2	12	" 18 "	
600	"	0.15	1.72–3.1	L_1	50	" 30 "	9/16
1000	"	0.5		L_2	6	" 18 "	
1300	0.125 (Meissner)	0.1	3.0–5.4	L_1	23	" 30 "	1/4
	No. 19–6848.			L_2	2	" 18 "	
1800	"	0.05	5.3–9.8	L_1	15	" 22 "	7/16
2000	"	0.04		L_2	6	" 18 "	
2300	"	0.03	9.7–17.8	L_1	9	" 22 "	5/8
2800	"	0.02		L_2	2	" 18 "	
3300	"	0.015	17.6–31.5	L_1	4	" 22 "	3/8
4000	"	0.01		L_2	2	" 18 "	
5200	"	0.0068	31.0–56.0	L_1	3	" 12 "	1/2
6250	"	0.005		L_2	1	" 18 "	
10,000	"	0.002					

L_1 and L_2 for 0.33 to 0.6 Mc. are the primary and secondary windings of a 456-kc. i.f. transformer, closely coupled, and mounted inside the coil form. L_1 for 31.0 to 56.0 Mc. has a diameter of 5/8 inch and is mounted inside the form. For all other frequencies L_1 is wound on the outside of the coil form. All L_2 windings are close-wound, have a diameter of 7/8 inch, and are mounted inside the forms. Millen type 45004 forms are used throughout. L_3, the pick-up coil, is one turn of No. 12 enameled wire, 1½ inches in diameter, connected to a length of 75-ohm Twin-Lead.

The 0–1 milliammeter is also required when the unit is used as an absorption wavemeter or as a field-strength indicator. However, the power supply is not required for these types of operation. It is only necessary that the proper r.f. coil be selected and that the coil (or the pick-up loop) be placed in the field of the frequency-generating device that is being measured. The sensitivity, during field-strength measurements, can be increased by attaching a short antenna to terminal C.

Because of the large frequency range covered by this instrument, it is not practical to employ a dial calibrated directly in terms of frequency. Therefore, an ordinary 0–100 degree dial is used and the actual frequency calibrations are marked on a separate chart. The r.f. ranges can be calibrated by using a calibrated receiver for listening to the oscillator output signal. An alternative method is to use the unit as a grid-dip meter coupled to a calibrated absorption wavemeter. The audio-frequency range can be calibrated by feeding the oscillator output to the vertical amplifier of an oscilloscope while the horizontal amplifier of the 'scope is being excited by the output of a calibrated audio oscillator. A circular pattern will be registered on the screen of the 'scope when the outputs of the two audio oscillators are adjusted to the same frequency. An alternative method of a.f. calibration is to connect a headset to the oscillator output terminals and compare the tone to the notes of a piano, determining which piano note is nearest. The approximate frequency then can be found from the table in Chapter Twenty-Four.

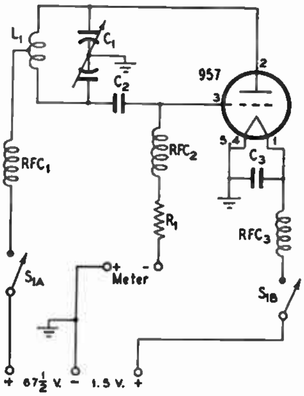


Fig. 16-21 — Schematic diagram of the v.h.f. grid-dip meter.

C_1 — 11-μμfd. "butterfly" variable (Johnson 160–211).
C_2 — 50-μμfd. ceramic (National XLA-0).
C_3 — 680-μμfd. mica.
R_1 — 68,000 ohms, ½ watt.
L_1 — 2 turns No. 12 wire, 1 1/16-inch i.d., turns spaced 1/32 inch. Ends of coil extend 5/8 inch past o.d. of coil.
RFC_1, RFC_2, RFC_3 — 1-μhy. r.f. choke (National R33).
$S_{1A\text{-}B}$ — D.p.s.t. "push-button" type toggle switch, normally open.

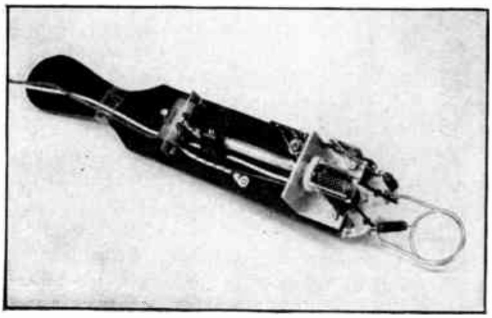


Fig. 16-20 — Probe-type grid-dip meter for v.h.f. use.

V.H.F. GRID-DIP METER

The oscillator circuit in the instrument just described is useful only to about 56 Mc. In the v.h.f. region it is necessary to use tubes and circuits especially designed for the purpose. Fig. 16-20 shows a "probe"-type v.h.f. grid-dip meter using an acorn tube in the ultraudion oscillator circuit. The instrument is built in a form that makes it easy to get into tight corners to check the resonant frequency of circuits in v.h.f. receivers and transmitters. The circuit diagram is given in Fig. 16-21.

The oscillator circuit is mounted on the end of a paddle made from Masonite, with the coil extending beyond the end so it can be coupled closely to the circuit to be measured. A push-to-operate switch is mounted at the handle end so the power is on only when the unit is being used. Small dry cells furnish the "A" and "B" supply. The instrument will work well with a 0–1 d.c. milliammeter.

The two-turn coil specified in Fig. 16-21 gives a range of 128 to 160 Mc., approximately. The frequency can be extended upward by using a one-turn coil. In either case the circuit may be calibrated by using a Lecher wire system of appropriate length.

Measuring Inductance and Capacitance

The ability to measure the inductance of coils, the capacitance of condensers, or the resonant frequency of a tuned circuit frequently saves time that might otherwise be spent in cut-and-try. A convenient instrument for this purpose is the grid-dip oscillator, described earlier in this chapter.

For measuring inductance, the coil to be measured is connected to a condenser of known capacitance as shown at A in Fig. 16-22. A mica condenser may be used as a standard; a 100-$\mu\mu$fd. 5-per-cent tolerance unit will serve for most purposes. With the unknown coil connected to the standard condenser, the pick-up loop is coupled to the coil and the oscillator frequency adjusted for the grid-current dip, using the loosest coupling that gives a detectable indication. The inductance is then given by the formula

$$L_{\mu h.} = \frac{25{,}330}{C_{\mu\mu fd.}\, f_{Mc.}^2}$$

A calibrated variable condenser is required for measuring capacitance. The circuit used is shown at B in Fig. 16-22. The frequency of the circuit, using any convenient coil, is first measured with the unknown capacitance disconnected and the calibrated condenser set near maximum. The unknown is then connected and the calibrated condenser readjusted to resonance. The unknown capacitance is then equal to the difference between the capacitances at the two settings of the calibrated condenser. Obviously only capacitances smaller than the maximum capacitance of the calibrated condenser can be measured by this method. Since high accuracy in capacitance measurement is not ordinarily required, a satisfactory standard is any condenser of the straight-line capacitance type, for which a sufficiently good calibration curve can be constructed by noting the dial divisions at which the plates just start to mesh and are completely meshed, and assuming that the capacitance change is linear within those limits. The minimum and maximum capacitance (corresponding closely enough to these condenser settings) can be obtained from the manufacturer's data on the particular variable condenser used.

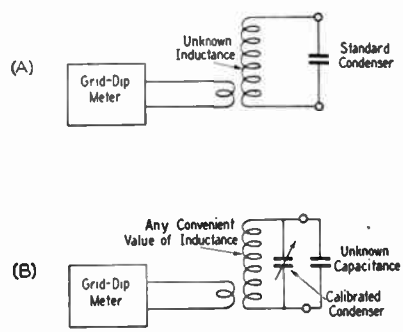


Fig. 16-22 — Set-ups for measuring inductance and capacitance with the grid-dip meter.

The Oscilloscope

The cathode-ray oscilloscope is an instrument of great versatility, and in conjunction with the instruments herein described, should be a valuable addition to the practical amateur station. The oscilloscope is useful on d.c., and audio and radio frequencies, and is particularly suited to a.f. and r.f. measurements because, compared to other types of measuring equipment, it introduces relatively little error at such frequencies.

CATHODE-RAY TUBES

The heart of the oscilloscope is the **cathode-ray tube**, 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, and can trol circuits, are connected directly to the rotor arms of their respective potentiometers, R_1, R_8 and R_9.

The socket for the cathode-ray tube is not fastened to any of the structural members of the unit, but is used as a plug, with the socket terminals enclosed in a tubular aluminum shield made by cutting down a National type T-78 tube shield. The base plate of this assembly is used as the support for a two-terminal tie point that holds isolating resistors R_{10} and R_{11}. These resistors are mounted inside the socket shield, as close to the tube base as possible. A ½-inch hole is drilled through the side of the shield to pass the cabled and shielded d.c. leads that run from the tube socket into the divider network in the aluminum shield box. A ceramic feed-through bushing requiring a ⅜-inch clearance hole passes through the opposite side of the socket shield to serve as the vertical input terminal. C_6 is connected between this bushing and the vertical deflection-plate pin on the tube socket. C_5, the coupling condenser for the horizontal plates, is mounted inside the larger shield box, near the horizontal-amplitude control, R_{12}.

The horizontal input terminals of the 'scope are mounted on the rear of the shield box, alongside of the audio transformer. The transformer secondary is connected to produce a turns ratio of approximately 1-to-1, which is sufficient to produce more than enough sweep voltage. A double-pole toggle switch is used to open the primary circuit of the audio transformer and to connect the external terminal to the amplitude control when the 'scope is used for transmitter monitoring. In this case sweep voltage is obtained from the audio system of the transmitter. R_{12} is connected on the tube side of the sweep switch, so that it remains in the circuit at all times to give control of voltage applied to the horizontal plates.

Details for using this oscilloscope to monitor a 'phone transmitter and to check both linearity and percentage modulation are contained in Chapter Nine. It should be remembered that an external resistor, R_E in Fig. 16-26, must be used in series with the lead to the horizontal input terminals to reduce the audio voltage to the desired level. Instructions for selection of this resistor are given in Chapter Nine.

● LINEAR SWEEPS AND AMPLIFIERS

Probably the chief use of the oscilloscope in amateur work is in measuring the percentage modulation in 'phone transmitters and in serving as a continuous monitor of modulation percentage. An oscilloscope for this purpose may be quite simple and inexpensive, consisting only of a small cathode-ray tube and an appropriate power supply as described above. However, by providing amplifiers for the deflection plates and furnishing a linear sweep circuit, the possibilities of the instrument are greatly extended. It then becomes possible, for example, to examine audio-frequency waveforms and to check and locate the causes of distortion in a.f. amplifiers.

Gas-Tube Sweep Generator

A typical circuit for a linear sweep generator and amplifier is shown in Fig. 16-28. The tube is a gas triode or grid-control rectifier. The

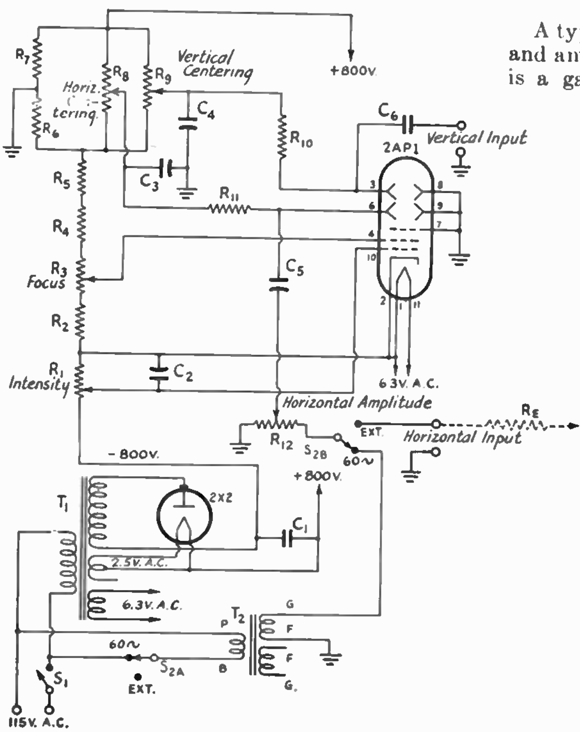


Fig. 16-26 — Circuit diagram of the simple oscilloscope for modulation monitoring.

C_1 — 1 μfd., 1000 volts, oil-filled.
C_2, C_3, C_4 — 0.01-μfd. 600-volt paper.
C_5 — 0.1 μfd., 1000 volts, paper.
C_6 — 0.001 μfd., 600 volts, mica.
R_1 — 20,000-ohm potentiometer, linear taper.
R_2 — 4700 ohms, ½ watt.
R_3 — 50,000-ohm potentiometer, linear taper.
R_4, R_5 — 33,000 ohms, 1 watt.
R_6, R_7 — 47,000 ohms, 1 watt.
R_8, R_9 — 50,000-ohm potentiometer, linear taper.
R_{10}, R_{11} — 1 megohm, ½ watt.
R_{12} — 0.25-megohm potentiometer, linear taper.
S_1 — S.p.s.t. toggle switch.
S_2 — D.p.d.t. toggle switch.
T_1 — Replacement-type receiver transformer, 350 v. each side of c.t., 70 ma. (Stancor P-6011.)
T_2 — Interstage audio transformer. (UTC S-2, with half of secondary unused, to produce approx. 1:1 turns ratio.)

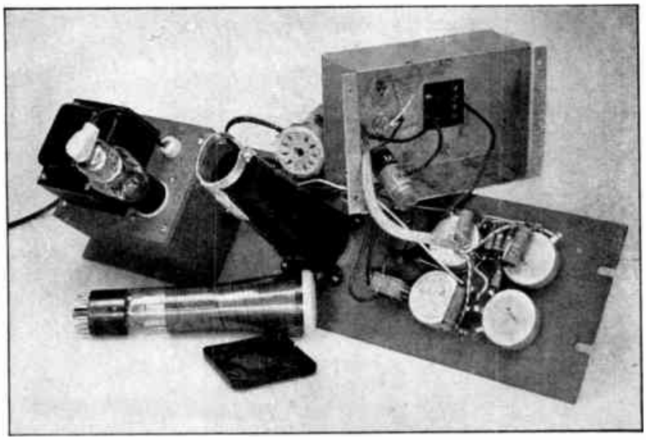


Fig. 16-27 — Rear view of the rack-mounting oscilloscope. The shield covering the voltage-divider components has been removed to show wiring. Mounted on the shield are the audio transformer and the horizontal input terminals. The 'scope tube and its socket have been removed.

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 in Fig. 16-29, is applied, the condenser between plate and cathode acquires a charge through R_6R_7. 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 the condenser once more charges through R_6R_7. If they are large enough, the voltage across the condenser rises linearly with time up to the breakdown point. This linear voltage change is used for the sweep. The fly-back time is the time required for discharge through the tube; to keep this time small, the resistance during discharge must be low.

The "sawtooth" rate is controlled by varying the capacitance between plate and cathode and the resistance of R_6R_7. To obtain a stationary pattern, the sweep is 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 self-generated sweep frequency.

The pentode amplifier in Fig. 16-28 can be used either to amplify the sweep-voltage output of the 884 oscillator, or to amplify any external voltage that it may be desired to use as a horizontal sweep. The gain control, R_{11}, provides a means for adjusting the width of the pattern on the cathode-ray tube screen.

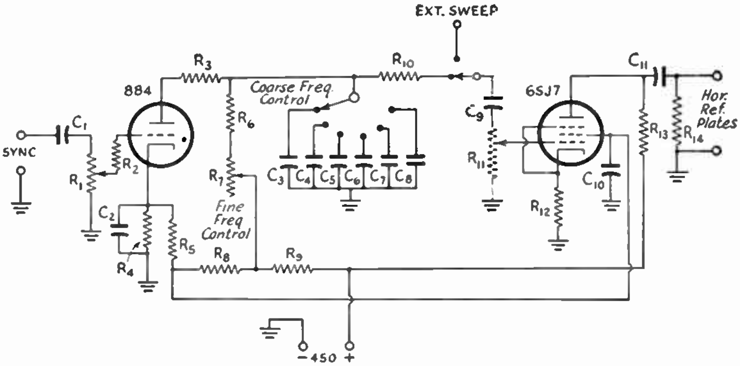


Fig. 16-28 — Linear sweep generator and horizontal amplifier.

C_1 — 0.1-μfd. paper.
C_2 — 25-μfd. 25-volt electrolytic.
C_3 — 0.25-μfd. paper, 600 volts.
C_4 — 0.1-μfd. paper, 600 volts.
C_5 — 0.04-μfd. paper, 600 volts.
C_6 — 0.015-μfd. paper, 600 volts.
C_7 — 0.005-μfd. paper or mica, 600 volts.
C_8 — 0.0022-μfd. mica.
C_9, C_{11} — 0.5-μfd. paper, 600 volts.
C_{10} — 8-μfd. electrolytic, 450 volts.
R_1 — 0.25-megohm potentiometer.
R_2 — 22,000 ohms, ½ watt.
R_3 — 470 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
R_5 — 22,000 ohms, 1 watt.
R_6 — 0.33 megohm, ½ watt.
R_7 — 1-megohm potentiometer.
R_8, R_9 — 62,000 ohms, 1 watt.
R_{10} — 1 megohm, ½ watt.
R_{11} — 0.5-megohm potentiometer.
R_{12} — 820 ohms, ½ watt.
R_{13} — 0.1 megohm, 1 watt.
R_{14} — Bleed for horizontal deflection plates.

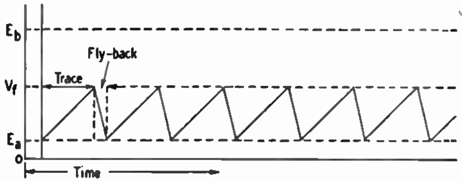


Fig. 16-29 — Condenser charging curves showing how a sawtooth wave is produced by a gaseous-tube linear sweep oscillator.

The output of the amplifier should be connected to the horizontal deflection plates of the tube. If this circuit is to be used with the oscilloscope previously described, the output terminals may be connected directly to Terminals 6 and 9 on the 2AP1 socket. In such case C_5 in Fig. 16-26 should be disconnected, but all other connections should be left unchanged.

Vertical Amplifiers

When using an oscilloscope for checking audio-frequency waveforms a "vertical" amplifier is a practical necessity. For most purposes the amplifier will be satisfactory if its frequency-response characteristic is flat over the a.f. range and if it has a gain of 100 or so. A typical circuit is shown in Fig. 16-30. It will be recognized as being practically similar to the "horizontal" amplifier of Fig. 16-28. A high-resistance gain control is desirable, to avoid loading the audio circuits to which the amplifier is connected.

When such an amplifier is used with the oscilloscope of Fig. 16-26, the output terminals should be connected between Terminals 3 and 8 on the 2AP1 socket. It is advisable to connect Terminal 3 to the arm of a 2-position ceramic switch, one contact going to the vertical amplifier and the other to C_6 in Fig. 16-26. This permits using either r.f. or a.f. input to the vertical deflection plates, disconnecting the a.f. amplifier circuit when r.f. is to be applied.

Constructional Considerations

In building an oscilloscope, care should be taken to see that the tube is shielded from stray electric and magnetic fields that might deflect the beam, and means should be provided to protect the operator from accidental shock, since the voltages employed with the larger tubes are quite high. In general, the preferable form of construction is to enclose the instrument completely in a metal cabinet. It is good practice to provide an interlock switch that automatically disconnects the high-voltage supply when the cabinet is opened for servicing or other reasons.

In laying out the unit, the cathode-ray tube must be placed so that the alternating magnetic field from the power 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.

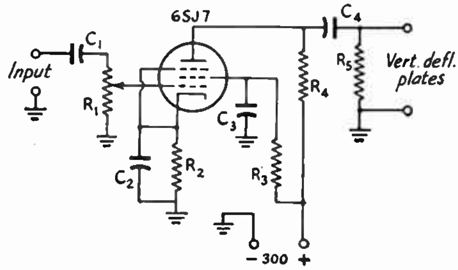


Fig. 16-30 — Circuit diagram of vertical amplifier.
C_1, C_3, C_4 — 0.1-μfd. paper, 400 volts.
C_2 — 25-μfd. 25-volt electrolytic.
R_1 — 1-megohm potentiometer.
R_2 — 1500 ohms, ½ watt.
R_3 — 2.2 megohms, 1 watt.
R_4 — 0.47 megohm, 1 watt.
R_5 — Bleed resistor for vertical deflection plates.

It is important that provision be included either for switching off the electron beam or reducing the spot intensity when no signal voltage is being applied. A thin, bright line or a spot of high intensity will "burn" the tube screen.

If trouble is experienced in obtaining a clean pattern from a high-power transmitter because of r.f. voltage introduced by the 115-volt line, by-pass condensers (0.01 or 0.1 μfd.) should be connected in series across the primary of the power transformer, the common connection between the two being grounded to the case.

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 of the antenna system include the measurement of transmission-line performance.

● 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 consists of a pick-up antenna and an indicating device such as a rectifier and microammeter, or 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 at least several wavelengths distant from the antenna and at heights corresponding with the desired angle of radiation.

The absorption frequency meter shown in Fig. 16-5 may be used as a field-strength meter if provided with a pick-up antenna. It is convenient to have the indicating device separate from the actual pick-up. This arrangement allows the pick-up unit to be set up out in the field to pick up radiation from the antenna under test, while the meter unit is near where adjustments are to be made. Antenna adjustment thus becomes a one-man job.

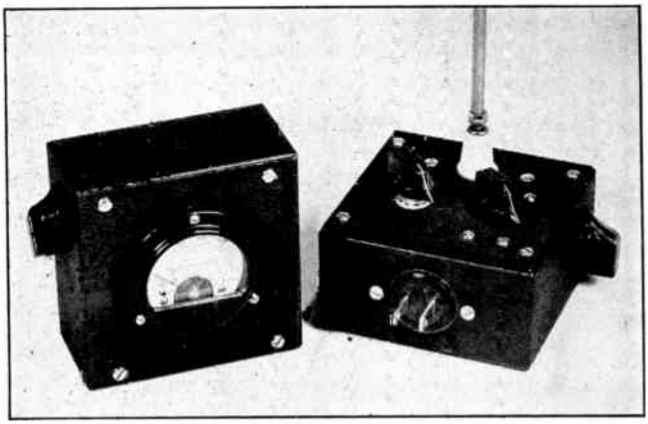


Fig. 16-31 — Remote-indicating field-strength meter, consisting of an r.f. pick-up and rectifier unit, and a meter unit. The knob on the left side of the meter unit is the switch for the shunt. On the pick-up unit the two controls are the bandswitch (left) and tuning. The knob at the right is for the resistor-shorting switch.

The unit shown in Figs. 16-31 to 16-33, inclusive, is particularly suitable for measurements in the v.h.f. range. It is constructed in two sections, one containing a tuned circuit, crystal rectifier, and antenna connection, and the other housing a microammeter for registering the rectified current from the crystal. The two units are fitted with matching plug and socket, permitting them to be used together, or they may be interconnected by means of a cable which can be any length up to several hundred feet. Three coils are used, so that measurements may be made on 28, 50 and 144 Mc. A resistor is inserted in series with the crystal and meter, to lessen the loading effect on the tuned circuit and to make the response of the crystal more linear with variations in radiated power. As the resistor reduces the sensitivity somewhat, a switch is provided to short it out in case measurements are to be made with extremely low power or at large distances from the transmitting antenna. A 100-microampere meter is used to give high sensitivity, and a shunt is available to multiply the range of the meter by three. This shunt is also provided with a switch so that low or high readings can be taken without making a trip to the pick-up unit. The crystal is the 1N21 type. Germanium crystals (1N34) also may be used with good results.

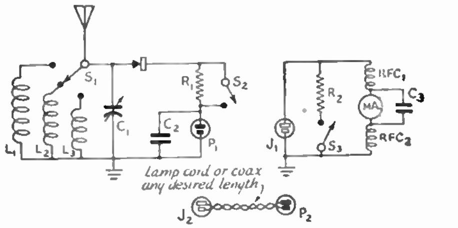


Fig. 16-32 — Wiring diagram of the remote-indicating field-strength meter.

C_1 — 25-μμfd. midget variable.
C_2, C_3 — 0.001-μfd. mica.
R_1 — 1000 ohms, ½ watt.
R_2 — 220 ohms, ½ watt.
L_1 — 28-Mc. coil — 7 turns No. 22 enamel, ¼ inch long, on ¾-inch dia. form (National PRF-1).
L_2 — 50-Mc. coil — 6 turns No. 22 enamel, ¼ inch long, on 9/16-inch dia. form (National PRF-1).
L_3 — 144-Mc. coil — 3 turns No. 18 enamel, ¼ inch long, ⅜-inch dia., self-supporting.
J_1, J_2 — Universal receptacle, two-pole retainer-ring type (Amphenol 61-F).
MA — 0-100 microammeter (0-500 microammeter or 0-1 milliammeter may be used, with reduced sensitivity).
P_1, P_2 — Polarized plug, two-pole retainer-ring type (Amphenol 61-MP).
S_1 — 3-position wafer-type switch.
S_2, S_3 — S.p.s.t. snap switch.
RFC_1, RFC_2 — 2.5 mh. choke (National R-100).

The two units are housed in 2 × 4 × 4-inch steel boxes with front and back removable. In the pick-up unit all parts except the resistor-shorting switch and connecting plug are mounted on the top panel, permitting easy wiring of the assembly. The interconnecting plug and socket are the polarized type, with one prong on the plug slightly larger than the other. The plug will fit a standard a.c. outlet, so the interconnecting cable (ordinary rubber-covered lamp cord) can double as a long a.c. extension cord.

The antenna connection is a steatite feed-through bushing fitted with a "banana-plug" socket. A convenient pick-up antenna is made by drilling and tapping a ¼-inch rod for 6/32 thread to take the threaded end of a banana plug. The length of the antenna will vary the sensitivity of the unit. If measurements are to be made with high power levels, a rod a few inches in length will suffice, but for ordinary work a 24-inch length will be suitable.

CHECKING STANDING WAVES

Standing waves on a transmission line can be measured if it is possible to measure the current at every point along the line, or the voltage between the two conductors at every point along the line. Rough checks can be made by going along the line with an absorption wavemeter having a crystal rectifier, taking care to keep the pick-up coil (or pick-up antenna) at the same distance from the line at

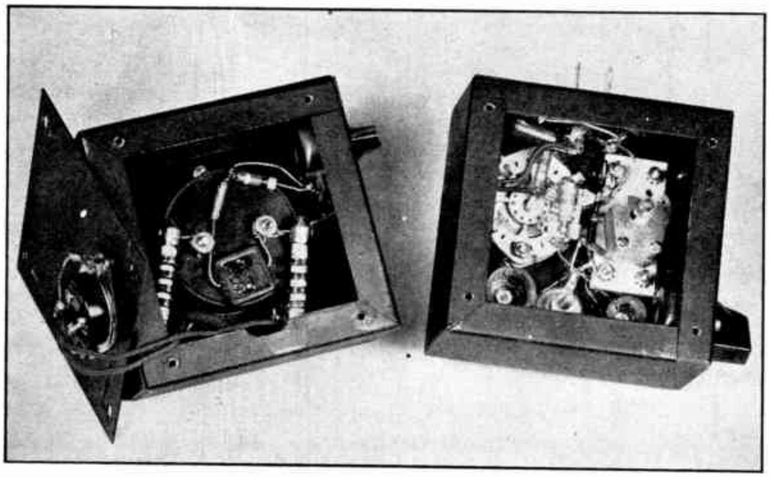


Fig. 16-33 — Inside view of the two units of the remote-indicating field-strength meter.

every measurement. With such a device the milliammeter usually will indicate current loops if a small pick-up coil is used, and voltage loops if a short pick-up antenna is used.

On two-conductor lines the current can be checked by means of the device shown in Fig. 16-34. The hooks, which should be sharp enough to cut through the insulation (if any)

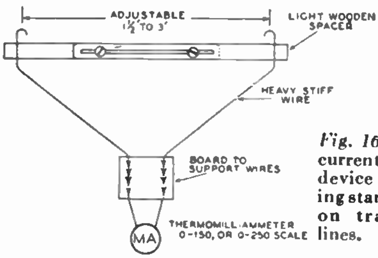


Fig. 16-34 — Line-current measuring device for checking standing waves on transmission lines.

on the wires, are placed on one of the wires. The spacing between the hooks should be adjusted to give a suitable reading on the meter. The standing-wave ratio can be determined by taking readings to determine the maximum and minimum currents along the line. At any one position along the line the currents in the two wires should be identical. If they are not, the line is carrying current either induced by the field of the antenna or coupled into the line through stray capacitance at the transmitter end.

BRIDGE-TYPE STANDING-WAVE INDICATORS

The standing-wave ratio on a transmission line can be found without actually going along the line and measuring the current or voltage. The basis of such measurement is the separation of the power traveling outward along the transmission line to the load from the power reflected back from the load toward the source. At any point along the line the following relationship is true:

$$S.W.R. = \frac{V_o + V_r}{V_o - V_r}$$

where $S.W.R.$ = Standing-wave ratio

V_o = Outgoing component of voltage

V_r = Reflected component of voltage

V_o and V_r are both taken at the point where the s.w.r. measurement is made.

An ordinary voltage measurement on the line simply shows the resultant of the two voltage components, but by using special bridge-type circuits it is possible to obtain voltage readings that are proportional to each component. Two circuits of this type are shown in Fig. 16-35. The one at A is a resistance-capacitance bridge and that at B a Maxwell-type bridge. Both bridges are theoretically independent of the applied frequency, and are practically so up to the frequency where stray inductance, capacitance, and coupling between circuit elements and wiring become of importance. In both circuits the radio-frequency voltmeter, V, must be a high-impedance device. The conditions for "balance" — that is, for the voltmeter to read zero regardless of the

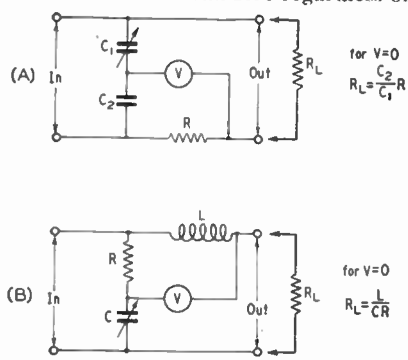


Fig. 16-35 — Fundamental circuits of two bridge-type standing-wave indicators. The upper circuit is used in the "Micro-Match" unit; the lower is a Maxwell bridge.

voltage applied to the input terminals — are given in the equations to the right of each diagram. C_1 in Fig. 16-35A, and C in the circuit at B, are made adjustable so that the ratio of the bridge can be varied for various load resistances, R_L.

If the load, R_L, is a transmission line, it will look like a pure resistance equal to its characteristic impedance when the line is correctly terminated. Consequently, the voltmeter will read zero when the line is perfectly matched, if the bridge has previously been balanced for

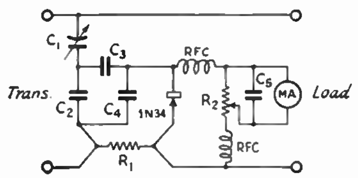


Fig. 16-36—Circuit diagram of the "Micro-Match" standing-wave indicator.

C_1 — 3-15-μμfd. midget variable.
C_2, C_4 — 220-μμfd. mica.
C_3 — 82-μμfd. mica.
C_5 — 0.0047-μfd. mica.
R_1 — 1.1-ohm resistor (9 10-ohm 1-watt carbon resistors in parallel).
R_2 — 5000-ohm potentiometer.
MA — 0-1 d.c. milliammeter.
RFC — 2.5-mh. r.f. choke.

that same value of resistance. If the line is not matched, the reflected power will cause the voltmeter to register a reading that is proportional to the *reflected* voltage. If the load is connected to the "input" terminals and the source of power to the "output" side, the voltmeter reading will be proportional to the *outgoing* component of voltage. Substituting the two voltmeter readings in the formula above will then give the standing-wave ratio.

Practical circuits corresponding to the two in Fig. 16-35 are given in Figs. 16-36 and 16-37. The r.f. voltmeter is a crystal rectifier and 0-1 d.c. milliammeter (or microammeter) with chokes and resistors for keeping the r.f. out of the meter circuit. In order to keep the voltmeter impedance high and to improve the linearity, it is advisable to use as much resistance in series with the meter as possible while still obtaining full-scale indications at the r.f. power level used.

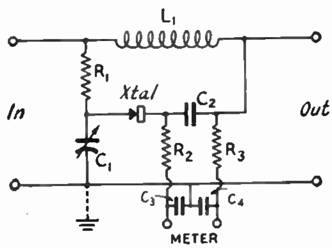


Fig. 16-37 — Circuit diagram of the Maxwell-bridge standing-wave indicator. The meter should have a full-scale range of 1 milliampere or less.

C_1 — 10-100-μμfd. Ceramicon variable.
C_2 — 470-μμfd. mica.
C_3, C_4 — (Optional) 100-μμfd. mica.
R_1 — 500 ohms, nonreactive.
R_2, R_3 — 10,000 ohms, ½-watt carbon.
L_1 — Approx. 29 turns No. 18, diameter 0.6 inch, 2.5 inches long.
XTAL — 1N34 or equivalent.

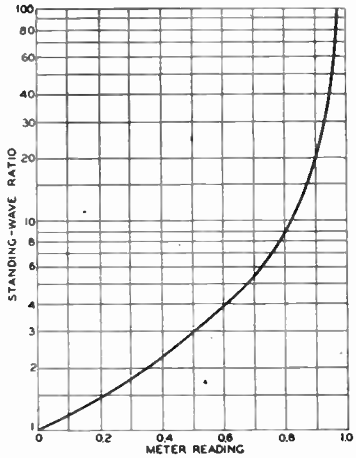


Fig. 16-38 — Standing-wave ratio in terms of meter reading (relative to full scale) after setting outgoing voltage to full scale. This graph is a plot of the formula

$$S.W.R. = \frac{V_o + V_r}{V_o - V_r}$$

Several precautions must be observed in constructing and using such instruments. The leads must be kept short, to avoid introducing reactance that would prevent obtaining proper balance. The rectifier-circuit wiring should be kept out of the fields of the other components insofar as possible, since stray pick-up in this wiring will give a "residual" voltmeter reading that will not balance out. It is absolutely essen-

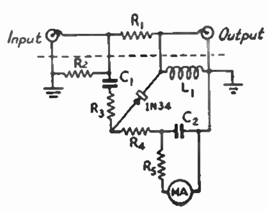


Fig. 16-39 — Circuit diagram of resistance bridge for measuring s.w.r., as adapted for coaxial lines. This circuit operates at very low power level and provision must be made for reducing the transmitter power to a low value when using it. The part of the circuit shown above the dotted line should be shielded from the remainder of the circuit.

C_1, C_2 — 470-μμfd. mica.
R_1 — 1-watt composition resistor, value equal to impedance of line being measured.
R_2 — 10 ohms, 1 watt.
R_3, R_4 — 56-ohm 1-watt composition. Exact value not important but the two resistors must have the same value.
R_5 — 470 ohms, ½ watt.
L_1 — Good r.f. choke at operating frequency. Not required if antenna system is closed type that offers d.c. return. At 28 Mc., 40 turns of No. 36 d.c.c. wound on a 1-watt 0.1-megohm resistor is satisfactory, or a 2.5-mh. choke may be used.
MA — 0-1 milliammeter.

tial that the resistors have negligible capacitance and inductance; wire-wound resistors cannot be used with any success.

To calibrate such a bridge, connect a noninductive resistor equal to the characteristic impedance of the line to be used to the output terminals, apply an r.f. voltage to the input terminals, and adjust the variable condenser for minimum reading. Then reverse the bridge so that the power source is connected to the output terminals and the resistor load to the input. Adjust the r.f. voltage (by changing the coupling to the transmitter) to make the meter read full scale. Then reverse the bridge connections and check the reading. If it is more than one or two per-cent of the full-scale reading it will be necessary to try different arrangements of the wiring until the null reading can be brought as close to zero as possible. The variable condenser can be calibrated in terms of various line impedances by substituting load resistances of the appropriate values, noting the setting for balance at each resistance value. Both circuits can be used over the range of 50 to 300 ohms, approximately.

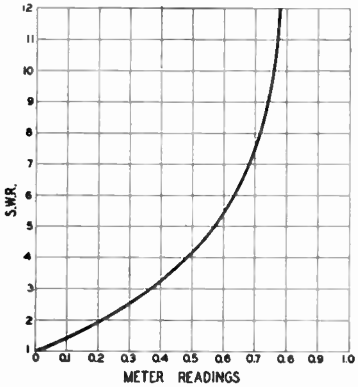


Fig. 16-40 — Calibration curve of the bridge-type s.w.r. indicator of Fig. 16-39. This curve will not apply if other circuit constants are used.

In using standing-wave indicators, the readings will be reliable only when there are no "antenna" currents on the transmission line. (See Chapter Ten.) If there is no stray pick-up on the line, it will not matter which line conductor is connected to which output terminal; the meter reading will be the same both ways. The same is true of the connections to the power source. If the way the line or source is connected does make a difference in the readings the results can be considerably in error. Coaxial lines usually are less troublesome in this respect than parallel-conductor lines.

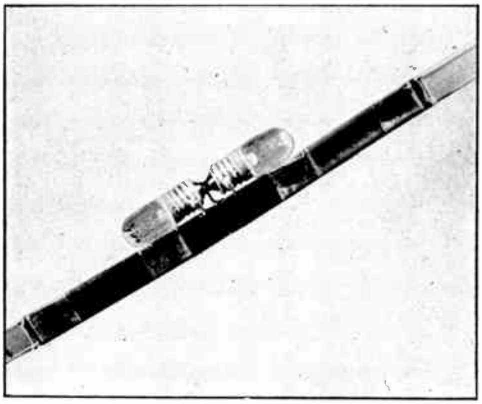


Fig. 16-41 — The "twin-lamp" standing-wave indicator.

Resistance Bridge

The bridge circuit in Fig. 16-39 uses equal resistance arms and can be used with only one value of line impedance. However, it is not necessary to reverse the line and power source to obtain a standing-wave reading. When properly constructed (noninductive resistors must be used) the meter reading will be zero when a noninductive resistance equal to the line impedance is connected to the output terminals.

To use the bridge, the output terminals are first open-circuited and the input voltage adjusted to give a full-scale reading on the milliammeter. When the line is connected to the output terminals, the meter reading (relative to the full-scale reading) indicates the s.w.r. as shown in Fig. 16-40. This curve was obtained by calibration of an instrument using the circuit constants given in Fig. 16-39. An actual calibration may be made in any case by using a series of noninductive resistors of different values as the load. The s.w.r. is equal to the ratio of the actual resistance used as a load to the value of resistance for which the bridge is designed.

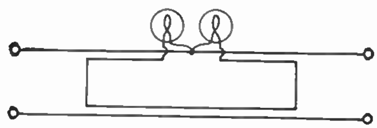


Fig. 16-42 — Wiring diagram of the "twin-lamp" standing-wave indicator.

The "Twin-Lamp"

A simple and inexpensive standing-wave indicator for 300-ohm line is shown in Fig. 16-41. It consists only of two flashlight lamps and a short piece of 300-ohm line. When laid flat against the line to be checked, the combination of inductive and capacitive coupling is such that outgoing power on the line causes the lamp nearest to the transmitter to light, while reflected power lights the lamp nearest the load. When the line is matched and no power is reflected, the lamp toward the antenna will be dark. The power input to the line should be adjusted to make the lamp nearest the transmitter light to full brilliance. When the lamp

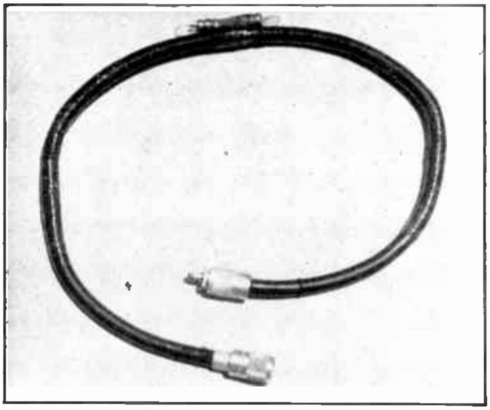


Fig. 16-43 — A "twin-lamp" standing-wave indicator for coaxial line.

nearest the load just begins to glow, the s.w.r. is about 1.5 to 1.

To construct the "twin-lamp," take a short length (a foot or two) of 300-ohm Twin-Lead and remove about ¼ inch of insulation from one wire at the center of the piece. Then take a second piece, 4 to 10 inches long (depending on the frequency and the transmitter power) and short-circuit both ends. Cut one wire in the exact center of the piece and peel the ends back on either side just far enough to provide leads to the flashlight lamps. Use the lowest-current flashlight bulbs or dial lamps available. Solder the tips of the bulbs together and connect them to the bare point in the long section of line, then solder the ends of the cut portion of the short piece to the shells of the bulbs. Figs. 16-41 and 16-42 should make the construction clear. The whole unit forms a "test section" that can be inserted in series with the line to be measured.

Figs. 16-43 and 16-44 show how the "twin-lamp" can be adapted for use with RG-8/U or RG-11/U coaxial cable. The circuit is the same as given in Fig. 16-42, but uses a loop formed from 75-ohm Twin-Lead fitted into the coax cable. A loop length of about 12 inches is satisfactory. As shown in Fig. 16-44, a length of the vinyl outer covering is removed from the cable, then the outer-braid conductor is carefully cut and removed by "bunching" the braid to increase its diameter sufficiently to slide it over the cable. The polyethylene insulation is then slotted to take the loop and a slot is cut in the piece of braid to permit the lamp connections to come through. When reassembled the whole section is covered with Scotch Tape.

While the "twin-lamp" cannot readily be used for quantitative measurement of s.w.r., it is very useful in checking the effect of load adjustments on the standing-wave ratio.

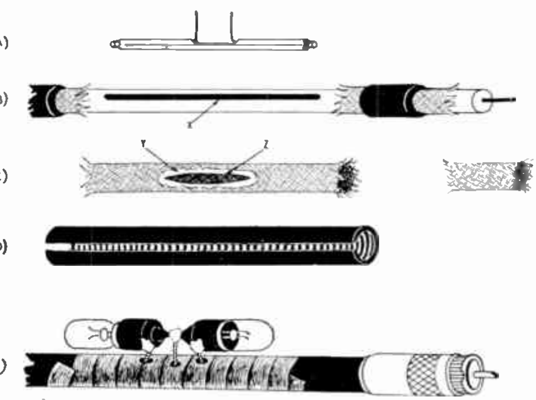


Fig. 16-44 — Steps in the construction of the coax "twin-lamp." The pick-up loop (A) is a length (approximately 12 inches) of 75-ohm Twin-Lead with the ends soldered together and covered with insulating material. There is no connection between the lamp circuit and the line except at the common connection between the bulbs; this point may be joined to the outer braid.

References

The construction of a bridge-type standing-wave indicator is critical both as to layout and materials. Additional information will be found in the following *QST* articles:

Jones and Sontheimer, "The 'Micromatch,'" April, 1947; "Additional Notes on the 'Micromatch,'" July, 1947.

Pattison, Morris and Smith, "A Standing-Wave Meter for Coaxial Lines," July, 1947.

Wright, "The 'Twin-Lamp,'" October, 1947.

Tiffany, "A Universal Transmission Bridge," December, 1947.

Keay, "The 'Coax Twin-Lamp,'" November, 1948.

CHAPTER 17

Assembling a Station

An amateur station is generally far better known by its signal and good operation than by its physical appearance. Good operating and a clean signal will build a reputation faster than thousands of dollars invested in special equipment and an elaborate "shack," and it is this very fact that makes amateur radio the democratic hobby that it is. However, most amateurs take pride in the arrangement of their stations, in the same way that they are careful of the appearance and arrangement of anything else which is part of the household. An antenna installation is the only external indication of the amateur station, and the degree of neatness required is generally determined by the district where the amateur lives and the attitude of the neighbors.

The actual location inside the house of the "shack" — the room where the transmitter and receiver are located — depends, of course, on the free space available for amateur activities. Fortunate indeed is the amateur with a separate room that he can devote to his amateur station, or the few who can have a special small building separate from the main house. However, most amateurs must share a room with other domestic activities, and amateur stations will be found tucked away in a corner of the living room, a bedroom, a large closet, or even under the kitchen stove! A spot in the cellar or the attic can almost be classed as a separate room, although it may lack the "finish" of a normal room.

This station is tucked away in a corner of the attic, and everything is accessible to the operator without his leaving the chair. The second-hand furniture was not an attempt to be fancy, but simply the only way the operator could find any lumber during a shortage. (W4JIZ/W4KFC, Annandale, Va.)

Regardless of the location of the station, however, it should be designed for maximum operating convenience and safety. It is foolish to have the station arranged so that the throwing of several switches is required to go from "receive" to "transmit," just as it is silly to have the equipment arranged so that the operator is in an uncomfortable and cramped position during his operating hours. The reasons for building the station as safe as possible are obvious, if you are interested in spending a number of years with your hobby!

● CONVENIENCE

The first consideration in any amateur station is the operating position, which includes the operator's table and chair and the pieces of equipment that are in constant use (the receiver, send-receive switch, and key or microphone). The table should be as large as possible, to allow sufficient room for the receiver or receivers, frequency-measuring equipment, monitoring equipment, control switches, and keys and microphones, with enough space left over for the logbook, a pad and pencil, and perhaps a *large* ash tray. Suitable space should be included for radiogram blanks and a call book, if these accessories are in frequent use. If the table is small, or the number of pieces of equipment is large, it is often necessary to build a shelf or rack for the auxiliary equipment, or to mount it in some less convenient location in or under the table. If one has the facilities, a semicircular "console" can be built of wood, or a simpler solution is to use two small wooden cabinets to support a table top of wood or Masonite. Home-built tables or consoles can be finished in any of the available oil stains, varnishes, paints or lacquers. Many operators use a large piece of plate glass over part of their table, since it furnishes a good writing surface and can cover miscellaneous charts and tables, prefix lists, operating aids, calendar, and similar accessories.

If the major interests never require frequent band changing, or frequency changing within a band, the transmitter can be located some distance from the operator, in a location where the meters can be observed from time to time (and the color of the tube plates noted!). If frequent band or frequency changes are a part of the usual operating procedure, the transmitter should be mounted close to the operator, either along one side or above the receiver, so that the controls are easily accessible

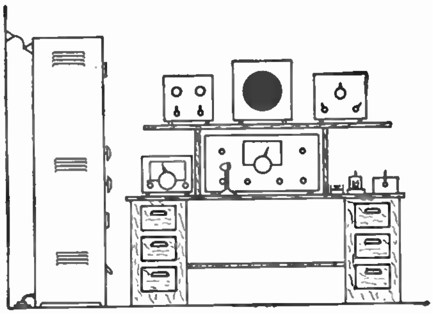


Fig. 17-1 — In a station assembled for maximum ease in frequency or band changing, the transmitter should be located next to the operating position, as shown above. On the operating table, the receiver is in front of the operator and VFO or crystal-switching oscillator on the left. (The VFO or crystal oscillator could be part of the transmitter proper, but most operators seem to prefer a separate VFO.)

The frequency standard and other auxiliary equipment can be mounted on a shelf above the receiver. The operating table can be an old desk, or a top supported by two small wooden cabinets. The "send-receive" switch is to the right of the telegraph keys — other switches are on the transmitter or the individual units.

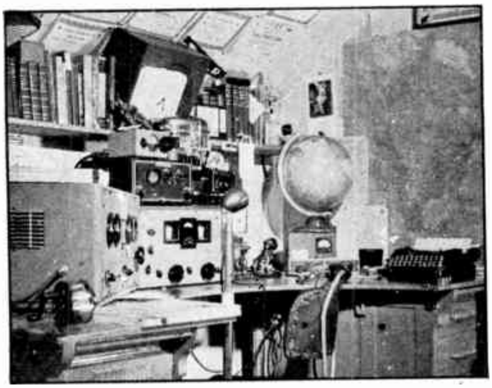
The above arrangement can be made to look cleaner by arranging all of the equipment on the table behind a single panel or a set of panels. In this case, provision must be made for getting behind the panel for servicing the units.

without the need for leaving the operating position.

A compromise arrangement would place the VFO or crystal-switched oscillator at the operating position and the transmitter in some convenient location not adjacent to the operator. Since it is usually possible to operate over a portion of a band without retuning the transmitter stages, an operating position of this type is an advantage over one in which the operator must leave his position to make a change in frequency.

Controls

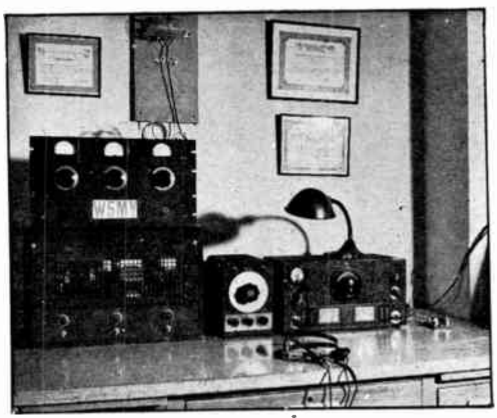
The operator has an excellent chance to exercise his ingenuity in the location of the operating controls. The most important controls in the station are the receiver tuning dial and the send-receive switch. The receiver tuning dial should be located four to eight inches



A convenient operating position can be obtained by building a "horseshoe-type" operating desk as shown here. Considerably more equipment can be placed on the desk around the operator than if an ordinary desk is used. (W9AND, Dixon, Ill.)

above the operating table, and if this requires mounting the receiver off the table, a small shelf or bracket will do the trick. With the single exception of the amateur whose work is almost entirely in traffic or rag-chew nets, which require little or no attention to the receiver, it will be found that the operator's hand is on the receiver tuning dial most of the time. If the tuning knob is too high or too low, the hand gets cramped after an extended period of operating, hence the importance of a properly-located receiver. The majority of c.w. operators tune with the left hand, preferring to leave the right hand free for copying messages and handling the key, and so the receiver should be mounted where the knob can be reached by the left hand. 'Phone operators aren't tied down this way, and tune the communications receiver with the hand that is more convenient.

The hand key should be fastened securely to the table, in a line just outside the right



When one specializes in clean-cut c.w. operation on all bands, he is likely to come up with a neat arrangement like this. The transmitter runs 400 watts, despite its small size. The small unit between transmitter and receiver is the VFO. (W5MY, San Antonio, Texas.)

shoulder and far enough back from the front edge of the table so that the elbow can rest on the table. A good location for the semiautomatic or "bug" key is right next to the handkey, although some operators prefer to mount the automatic key in front of them on the left, so that the right forearm rests on the table parallel to the front edge.

The best location of the microphone is directly in front of the operator, so that he doesn't have to shout across the table into it, or run up the speech-amplifier gain so high that all manner of external sounds are picked up.

In any amateur station worthy of the name, it should be necessary to throw no more than one switch to go from the "receive" to the "transmit" condition. In 'phone stations, this switch should be located where it can be easily reached by the hand that isn't on the receiver. In the case of c.w. operation, this switch is most conveniently located to the right or left of the key, although some operators prefer to have it mounted on the left-hand side of the operating position and work it with the left hand while the right hand is on the key. Either location is satisfactory, of course, and the choice depends upon personal preference. Some operators use a foot-controlled switch, which is a convenience but doesn't allow too much freedom of position during long operating periods.

If the microphone is hand-held during 'phone operation, a "push-to-talk" switch on the microphone is convenient, but hand-held microphones tie up the use of one hand and

A high-powered station in a room with enough space to locate the transmitters along one wall and the operating table along another. Note the convenient location of the control switches, and the receiver raised above the table. (W1CH, Worcester, Mass.)

are not too desirable, although they are widely used in mobile and portable work. A breast, chin or throat microphone is safer for mobile work, if the operator is also the driver of the vehicle.

The location of other switches, such as those used to control power supplies, filaments, 'phone/c.w. change-over and the like, is of no particular importance, and they can be located on the unit with which they are associated. This is not strictly true in the case of the 'phone/c.w. DX man, who sometimes has need to change in a hurry from c.w. to 'phone. In this case, the change-over switch should be at the operating table, although the actual change-over should be done by a relay that the switch controls.

If a rotary beam is used the control of the beam should be convenient to the operator. The beam-direction indicator, however, can be located anywhere within sight of the operator, and does not have to be located on the operating table unless it is small, or included with the beam control.

When several fixed beams are used, the selection of any one should be possible from the operating position, to minimize the time required to select the proper one. This generally means using a series of antenna relays or a stepping switch.

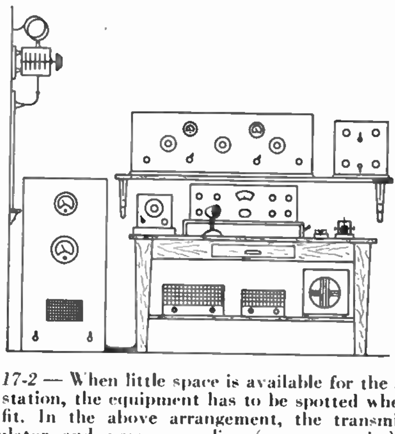


Fig. 17-2 — When little space is available for the amateur station, the equipment has to be spotted where it will fit. In the above arrangement, the transmitter, modulator and power supplies (separate units) are sandwiched in alongside the operating table and on a shelf above the table. The antenna tuning unit is mounted over the feed-through insulators that bring the antenna line into the "shack," and loudspeaker and small power supplies are mounted under the table. The operating position is clean, however, with the VFO, receiver and keys at table level. The tuning knob of this receiver would be uncomfortably low if the receiver weren't raised by the wooden arch, and the "send-receive" switch is mounted on the right-hand side of this arch, next to the hand key. Interconnecting leads should be cabled along the back of the table and table legs, to keep them inconspicuous.

Frequency Spotting

In a station where a VFO is used, or where a number of crystals is available, the operator should be able to turn on only the oscillator of his transmitter, so that he can spot accurately his location in the band with respect to other stations. This allows him to see if he has anything like a clear channel (if such a thing exists in the amateur bands!), or to see what his frequency is with respect to another station. Such a provision can be part of the "send-receive" switch. Switches are available with a center "off" position, a "hold" position on one side,

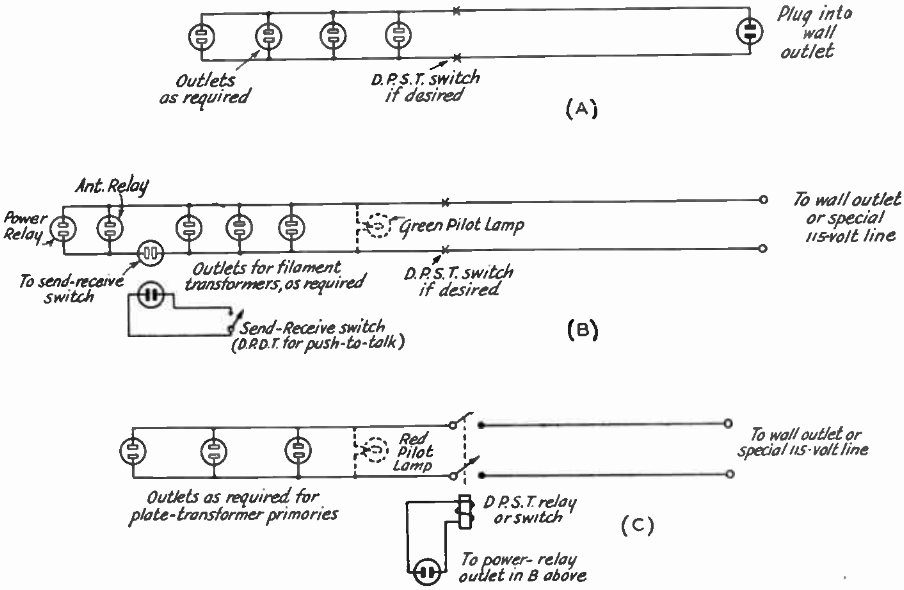


Fig. 17-3 — Power circuits for a high-power station. A shows the outlets for the receiver, monitoring equipment, speech amplifier and the like. The outlets should be mounted inconspicuously on the operating table. B shows the transmitter filament circuits and control-relay circuits, if the latter are used. C shows the plate-transformer primary circuits, controlled by the power relay. A heavy-duty switch can be used instead of the relay, in which case the antenna relay would be connected in circuit C.

If 115-volt pilot lamps are used, they can be connected as shown. Lower-voltage lamps must be connected across suitable windings on transformers.

With "push-to-talk" operation, the "send-receive" switch can be a d.p.d.t. affair, with the second pole controlling the "on-off" circuit of the receiver.

for turning on the oscillator only, and a "lock" position on the other side for turning on the transmitter and antenna relays. If oscillator keying is used, the key serves the same purpose, provided a "send-receive" switch is available to turn off the high-voltage supplies and prevent a signal going out on the air during adjustment of the oscillator frequency.

For 'phone operation, the telegraph key or an auxiliary switch can control the transmitter oscillator, and the "send-receive" switch can then be wired into the control system so as to control the oscillator as well as the other circuits.

An example of the compact station, complete on the operating table. The receiver is mounted on the left side of the table, for left-hand tuning. The beam-direction indicator and switches are housed in a small box sitting on the VFO. (W2NFU, Forest Hills, N. Y.)

Comfort

Of prime importance is the comfort of the operator. If you find yourself getting tired after a short period of operating, examine your station to find what causes the fatigue. It may be that the chair is too soft or hasn't a straight back or is the wrong height for you. The key or receiver may be located so that you assume an uncomfortable position while using them. If you get sleepy fast, the ventilation may be at fault. (Or you may need sleep!)

POWER CONNECTIONS AND CONTROL

Following a few simple rules in wiring your power supplies and control circuits will make it an easy job to change units in the station. If the station is planned in this way from the start, or if the rules are recalled when you are rebuilding, you will find it a simple matter to revise your station from time to time without a major rewiring job.

The regular wall outlets in a home are generally rated at 15 amperes at 115 volts, and so will furnish sufficient power for receivers, monitoring equipment, speech amplifiers, and anything that doesn't draw too high an intermittent load (such as a keyed transmitter or Class B modulator). A low-powered transmitter, under one or two hundred watts, can be supplied by an ordinary wall outlet. To make a neat installation, it is better to run a single pair of wires from the outlet over to the operating table or some central point, rather than to use a number of adapters at the wall outlet.

In a high-powered station, the receiver and auxiliary equipment can get their power from the wall outlet, but it is advisable to run in a special, heavy three-wire line from the meter box for the transmitter. This three-wire line will, of course, be 115 volts either side of neutral (ground), or 230 volts across the outside. In many cases it is possible to run the filaments and constant loads from one side of this three-wire line and the intermittent loads (plate transformers) from the other side. In this case the filament voltages will *rise* slightly with the application of load, because of the reduced *net* current in the neutral. However, this procedure often unbalances the system too much, resulting in considerable "blinking" of the lights, and the load must be distributed equally across the 230-volt circuit. This can be done by using plate transformers with 230-volt primaries, by dividing the load as equally as possible across both 115-volt circuits, or by using autotransformers that step down the 230 volts to 115 volts and connecting the plate-transformer primaries across the autotransformer secondaries. Obviously balancing the load is the cheapest "out" and the first one to try.

If the lights blink with keying or modulation of a low-powered transmitter that gets its power from a regular wall outlet, taking some of the power from another outlet may help to improve the regulation and is always worth a try.

When a special heavy line is run into the shack for a high-powered transmitter, it will generally be done by a licensed electrician who can advise you on the various types of outlets that are available. Some amateurs terminate their special lines in switch boxes, while others end the line in an electric-stove receptacle. In case you do the work yourself, it is wise to find out if there are any special regulations in your area covering the type of wire, insulation and outlet which must be used. The power companies are always willing to advise you if it looks as though you will be using more power!

Interconnections

The wiring of any station will entail two or three common circuits. The circuit for the receiver, monitoring equipment and the like, assuming it to be taken from a wall outlet, should be run from the wall to an inconspicuous point on the operating table, where it terminates in a multiple outlet large enough to handle the required number of plugs. A single switch between the wall outlet and the receptacle will then turn on all of this equipment at one time, or the plug can simply be pulled out at the wall when leaving the shack.

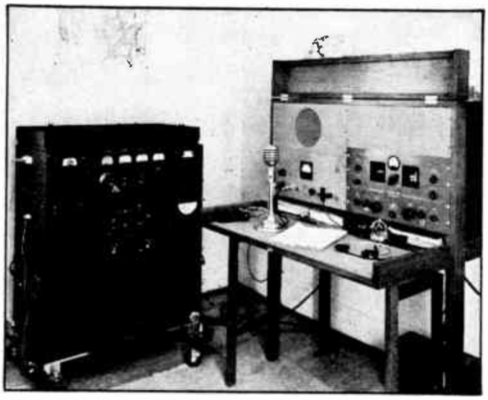
The second common circuit in the station is that supplying voltage to rectifier- and transmitter-tube filaments, bias supplies, and anything else that is not switched on and off during tranmsit and receive periods. The coil power for control relays should also be obtained from this circuit. The power for this circuit can come from a wall outlet or from the transmitter line, if a special one is used.

The third circuit is the one that furnishes power to the plate-supply transformers for the r.f. stages and for the modulator. When it is opened, the transmitter is disabled except for the filaments, and the transmitter should be safe to work on. However, one always feels safer when working on the transmitter if he has turned off every power supply pertaining to the transmitter.

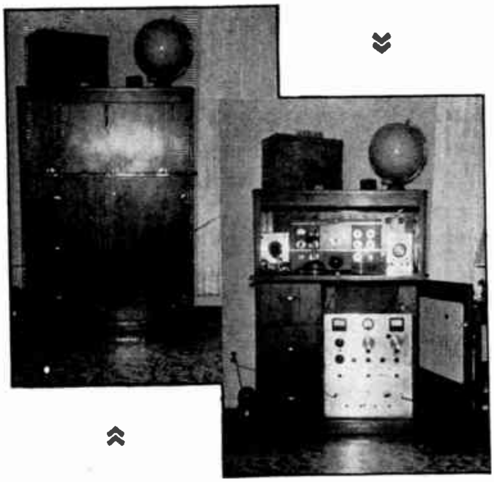
With these three circuits established, it becomes a simple matter to arrange the station for different conditions and with new units. Anything on the operating table (which runs all the time) ties into the first circuit. Any new power supply or r.f. unit gets its filament power from the second circuit. Since the third circuit is controlled by the send-receive switch (or relay), any power-supply primary that is to be switched on and off for send and receive connects to circuit No. 3.

Break-In and Push-To-Talk

In c.w. operation, "break-in" is any system that allows the transmitting operator to hear the other fellow's signal during the "key-up" periods between characters and letters. This allows the sending station to be "broken" by the receiving station at any time, to shorten calls, ask for "fills" in messages, and speed up operation in general. With present techniques, it requires the use of a separate receiving antenna and, with high power, some means for protecting the receiver from the transmitter when the key is "down." Several methods, applicable to high-power stations, are described in Chapter Eight. If the transmitter is low-powered (50 watts or so), no special



In this example of a compact high-power station, the operating table folds up when not in use and covers the receiver and speech amplifier. Special furniture, like this homemade operating table, goes a long way toward solving the space problem for many amateurs. (W4HAV, Fort Thomas, Ky.)



This station goes all the way in concealment by housing the entire station in a special cabinet. When the cabinet is opened, the operating table is formed and all pieces of gear are accessible. (W6YNX, Mountain View, Calif.)

equipment is required except the separate receiving antenna and a receiver that "recovers" fast. Where break-in operation is used, there should be a switch on the operating table to turn off the plate supplies when adjusting the oscillator to a new frequency, although during all break-in work this switch will be closed.

"Push-to-talk" is an expression derived from the "push" switch on some microphones, and it means a 'phone station with a single control for all change-over functions. Strictly speaking, it should apply only to a station where this single send-receive switch must be held in place during transmission periods, but any fast-acting switch will give practically the same effect. A control switch with a center "off" position, and one "hold" and one "lock" position, will give more flexibility than a straight "push" switch. The one switch must control the antenna change-over relay, the transmitter power supplies, and the receiver "on-off" circuit. This latter is necessary to disable the receiver during transmit periods, to avoid acoustic feed-back.

Switches and Relays

It is dangerous to use an overloaded switch in the power circuits. After it has been used for some time, it may fail, leaving the power on the circuit even after the switch is thrown to the "off" position. For this reason, large switches, or relays with adequate ratings, should be used to control the plate power. Relays are rated by coil voltages (for their control circuits) and by their contact ratings (the current they will carry safely).

When relays are used, the send-receive switch closes the circuit to their coils, thus closing the relay contacts. The relay contacts are in the power circuit being controlled, and thus the switch handles only the relay-coil current.

SAFETY

Of prime importance in the layout of the station is the personal safety of the operator and of visitors, invited or otherwise, during normal operating practice. If there are small children in the house, every step must be taken to prevent their accidental contact with power leads of any voltage. A locked room is a fine idea, if it is possible, otherwise housing the transmitter and power supplies in metal cabinets is an excellent, although expensive, solution. Lacking a metal cabinet, a wooden cabinet or a wooden framework covered with wire screen is the next-best solution. Many stations have the power supplies housed in metal cabinets in the operating room or in a closet or basement, and this cabinet or entry is kept locked — with the key out of reach of everyone but the operator. The power leads are run through conduit to the transmitter, using ignition cable for the high-voltage leads. If the power supplies and transmitter are in the same cabinet, a lock-type main switch for the incoming line power is a good precaution.

A simple substitute for a lock-type main switch is an ordinary line plug with a short connecting wire between the two pins. By wiring a female receptacle in series with the main power line in the transmitter, the shorting plug will act as the main safety lock. When the plug is removed and hidden, it will be impossible to energize the transmitter, and a stranger or child isn't likely to spot or suspect the open receptacle.

An essential adjunct to any station is a shorting stick for discharging any high voltage to ground before any work or coil changing is done in the transmitter. Even if interlocks and power-supply bleeders are used, the failure of one or more of these components may leave the transmitter in a dangerous condition. The shorting stick is made by mounting a small metal hook, of wire or rod, on one end of a dry stick or bakelite rod. A piece of ignition cable or other well-insulated wire is then run from the hook on the stick to the chassis or common ground of the transmitter, and the stick is hung alongside the transmitter. Whenever the power is turned off in the transmitter to work on the rig, or to change coils, the shorting stick is first used to touch the several high-voltage leads (tank condenser, filter condenser, tube plate connection, etc.) to insure that there is no high voltage at any of these points. Most commercial installations require the use of this simple device, and it has saved many a life. Use it!

Fusing

A minor hazard in the amateur station is the possibility of fire through the failure of a component. If the failure is complete and the component is large, the house fuses will generally blow. However, it is unwise and inconvenient to depend upon the house fuses to protect the lines running to the radio equipment, and every power supply should have its own set of fuses, with the fuse ratings selected at about 150 or 200 per cent of the maximum rating of the supply. If, for example, a power transformer is rated at 600 watts, it would draw about 5 amperes from the a.c. line ($600 \div 115 = 5.2$), and a 10-ampere fuse should be used in the primary circuit of the transformer. Circuit breakers can be used instead of fuses if desired.

Wiring

Control-circuit wires running between the operating position and a transmitter in another part of the room should be hidden, if possible. This can be done by running the wires under the floor or behind the base molding, bringing the wires out to terminal boxes or regular wall fixtures. Such construction, however, is generally only possible in elaborate installations, and the average amateur must content himself with trying to make the wires as inconspicuous as possible. If several pairs of leads must be run from the operating table to the transmitter, as is generally the case, a single piece of rubber- or vinyl-covered multiconductor cable will always look neater than several pieces of rubber-covered lamp cord.

The antenna wires always present a problem, unless coaxial-line feed is used. Open-wire line from the point of entry of the antenna line should always be arranged neatly, and it is generally best to support it at several points. Many operators prefer to mount their antenna-tuning assemblies right at the point of entry of the feedline, together with an antenna change-over relay (if one is used), and then the link from the tuning assembly to the transmitter can be made of inconspicuous coaxial line or Twin-Lead. If the transmitter is mounted near the point of entry of the antenna line, it simplifies the problem of "What to do with the feeders?"

There was enough room at this station to build the transmitter into the wall, and to protect it with glass doors. In an installation like this, it is convenient to have access to the rear of the transmitter units, for making connection to them and for testing. If the rear cannot be reached, all power leads should be cabled up along the side walls, at the rear. (W6NY, Whittier, Calif.)

General

You can check your station arrangement by asking yourself the following questions. If all of your answers are an honest "Yes," your station will be one of which you can be proud.

1) Is your station safe, under normal operating conditions, both for the operator and the visitor?

2) Is the operating position comfortable, even after several hours of operating?

3) Do you throw not more than one switch to go from "receive" to "transmit"?

4) Does it take only a short time to explain to another amateur how to work your station?

5) Do you show your station to visiting amateurs or laymen without apologizing for its appearance?

The Amateur's Workshop

TOOLS AND MATERIALS

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 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. It is an excellent idea for the amateur who does constructional work to add to his supply of tools from time to time as finances permit.

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 should you be in a position to acquire them.

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.
Hack saw, 12-inch blades.
Center punch for marking hole centers.
Hammer, ball-peen, 1-lb. head.
Heavy knife.
Yardstick or other straightedge.
Carpenter's brace with adjustable hole cutter or socket-hole punches (see text).
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 ⅛-inch and Nos. 18, 28, 33, 42 and 50.
Combination oil stone for sharpening tools.
Solder and soldering paste (noncorroding).
Medium-weight machine oil.

ADDITIONAL TOOLS

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, crosscut.
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.

Twist Drills

Twist drills are made of either high-speed steel or carbon steel. The latter type is 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 in Table 18-I will be most commonly used in construction of amateur 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.

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 a full kit of well-kept sharp-edged tools.

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 oilstoning of the cutting edges of a drill or reamer will extend the time between grindings.

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.

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 inches. (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.

TABLE 18-I
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	—	2-56
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.

CHASSIS WORKING

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

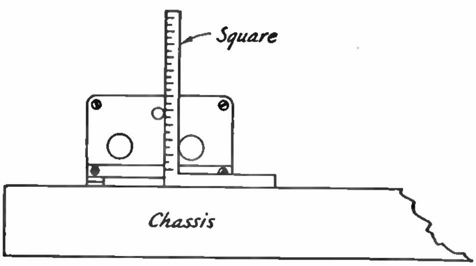


Fig. 18-1 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

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 are off center and 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.

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

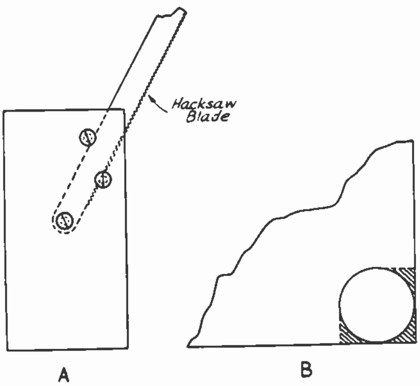


Fig. 18-2 — To cut rectangular holes in a chassis corner, holes may be filed out as shown in the shaded portion of B, making it possible to start the hack-saw blade along the cutting line. A shows how a single-ended handle may be constructed for a back-saw blade.

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. 18-1. 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.

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. When 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 counterclockwise.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. 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.

Rectangular Holes

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. 18-2, and using these holes for starting and turning the hack saw. 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.

● CONSTRUCTION NOTES

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 use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hack saw, 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 too far until 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.

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 Table 18-I) 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. Machine oil applied to the tap usually makes cutting easier and sticking less troublesome.

Wiring

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 are 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, Chapter Twenty-Four). 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.

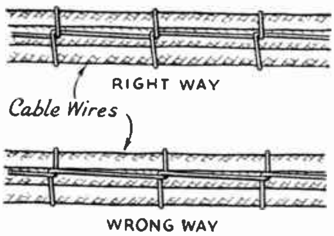


Fig. 18-3 — Right and wrong methods of lacing cable. With the right way the leading line is pinched under each turn and will not loosen if a break occurs in the lacing.

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.

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. 18-3 shows the correct way of lacing cabled wires.

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 fiber **terminal strips** as anchorages. These strips also form handy mountings for wire-terminal resistors, etc.

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.

Soldering paste, if of the noncorroding 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.

COMPONENT VALUES

Values of composition resistors and small condensers (mica and ceramic) are specified throughout this *Handbook*, in terms of "preferred values." In the preferred-number system, all values represent (approximately) a constant-percentage increase over the next lower value. The base of the system is the number 10. Only two significant figures are used. Table 18-II shows the preferred values based on tolerance steps of 20, 10 and 5 per cent. All other values are expressed by multiplying or dividing the base figures given in the table by the appropriate power of 10. (For example, resistor values of 33,000 ohms, 6800 ohms, and 150 ohms are obtained by multiplying the base figures by 1000, 100, and 10, respectively.)

"Tolerance" means that a variation of plus or minus the percentage given is considered satisfactory. For example, the actual resistance of a "4700-ohm" 20-per-cent resistor can lie anywhere between 3700 and 5600 ohms, approximately. The permissible variation in the same resistance value with 5-per-cent tolerance would be in the range from 4500 to 4900 ohms, approximately.

Only those values shown in the first column of Table 18-II are available in 20-per-cent tolerance. Additional values, as shown in the second column, are available in 10-per-cent tolerance; still more values can be obtained in 5-per-cent tolerance.

In the component specifications in this *Handbook*, it is to be understood that when no tolerance is specified the *largest* tolerance available in that value will be satisfactory.

TABLE 18-II
Standard Component Values

20% Tolerance	*10% Tolerance*	*5% Tolerance*
10	10	10
		11
	12	12
		13
15	15	15
		16
	18	18
		20
22	22	22
		24
	27	27
		30
33	33	33
		36
	39	39
		43
47	47	47
		51
	56	56
		62
68	68	68
		75
	82	82
		91
100	100	100

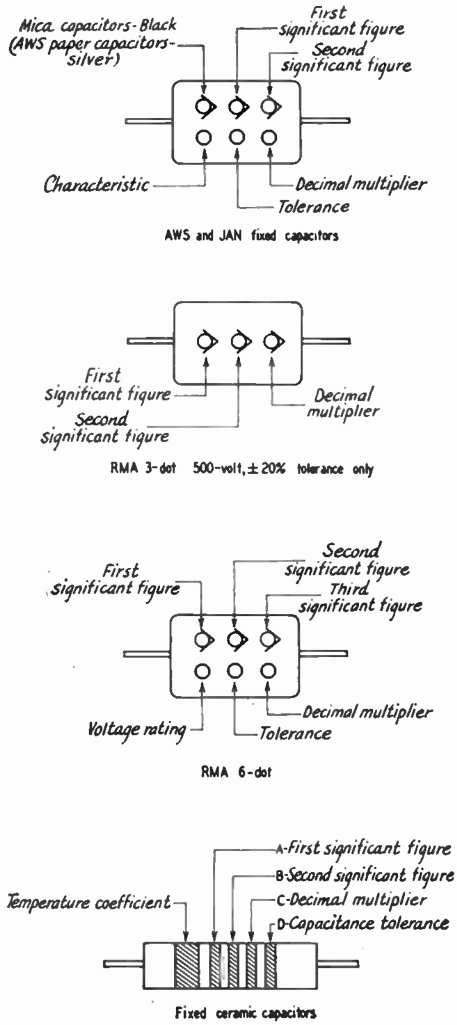


Fig. 18-4 — Color coding of fixed mica, molded paper, and tubular ceramic condensers. The color code for mica and molded paper condensers is given in Table 18-III. Table 18-IV gives the color code for tubular ceramic condensers.

Values that do not fit into the preferred-number system (such as 500, 25,000, etc.) easily can be substituted. It is obvious, for example, that a 5000-ohm resistor falls well within the tolerance range of the 4700-ohm 20-per-cent resistor used in the example above. It would not, however, be usable if the tolerance were specified as 5 per cent.

COLOR CODES

Standardized color codes are used to mark values on small components such as composition resistors and mica condensers, and to identify leads from transformers, etc. The resistor-condenser number color code is given in Table 18-III.

Fixed Condensers

The methods of marking "postage-stamp" mica condensers, molded paper condensers, and tubular ceramic condensers are shown in Fig. 18-4. Condensers made to American War Standards or Joint Army-Navy specifications are marked with the 6-dot code shown at the top. Practically all surplus condensers are in this category. The 3-dot RMA code is used for condensers having a rating of 500 volts and ±20% tolerance only; other ratings and tolerances are covered by the 6-dot RMA code.

Examples: A condenser with a 6-dot code has the following markings: Top row, left to right, black, yellow, violet; bottom row, right to left, brown, silver, red. Since the first color in the top row is black (significant figure zero) this is the AWS code and the condenser has mica dielectric. The significant figures are 4 and 7, the decimal multiplier 10 (brown, at right of second row), so the capacitance is 470 μμfd. The tolerance is ± 10%. The final color, the characteristic, deals with temperature coefficients and methods of testing, and may be ignored.

A condenser with a 3-dot code has the following colors, left to right: brown, black, red. The significant figures are 1, 0 (10) and the multiplier is 100. The capacitance is therefore 1000 μμfd.

A condenser with a 6-dot code has the following markings: Top row, left to right, brown, black, black; bottom row, right to left, black, gold, blue. Since the first color in the top row is neither black nor silver, this is the RMA code. The significant figures are 1, 0, 0 (100) and the decimal multiplier is 1 (black). The capacitance is therefore 100 μμfd. The gold dot shows that the tolerance is ± 5% and the blue dot indicates 600-volt rating.

Ceramic Condensers

Conventional markings for ceramic condensers are shown in the lower drawing of Fig. 18-4. The colors have the meanings indicated in Table 18-IV. In practice, dots may be used instead of the *narrow* bands indicated in Fig. 18-4.

Example: A ceramic condenser has the following markings: Broad band, violet; narrow bands or dots, green, brown, black, green. The significant figures are 5, 1 (51) and the decimal multiplier is 1, so the capacitance is 51 μμfd.

TABLE 18-IV

Color Code for Ceramic Condensers

Color	*Significant Figure*	*Decimal Multiplier*	*Capacitance Tolerance* More than 10 μμfd. (in %)	*Capacitance Tolerance* Less than 10 μμfd. (in μμfd.)	*Temp. Coeff. p.p.m. deg. C.*
Black	0	1	± 20	2.0	0
Brown	1	10	± 1		— 30
Red	2	100	± 2		— 80
Orange	3	1000			— 150
Yellow	4				— 220
Green	5		± 5	0.5	— 330
Blue	6				— 470
Violet	7				— 750
Gray	8	0.01		0.25	30
White	9	0.1	± 10	1.0	500

TABLE 18-III

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.

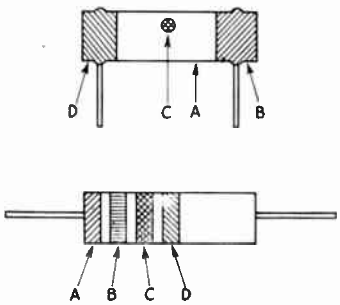
The temperature coefficient is — 750 parts per million per degree C., as given by the broad band, and the capacitance tolerance is ± 5%.

Fixed Composition Resistors

Composition resistors (including small wire-wound units molded in cases identical with the composition type) are color-coded as shown in Fig. 18-5. Colored bands are used on resistors having axial leads; on radial-lead resistors the colors are placed as shown in the drawing. When bands are used for color coding the body color has no significance.

Examples: A resistor of the type shown in the lower drawing of Fig. 18-5 has the following color bands: A, red; B, red; C, orange; D, no color. The significant figures are 2, 2 (22) and the decimal multiplier is 1000. The value of resistance is therefore 22,000 ohms and the tolerance is ± 20%.

A resistor of the type shown in the upper drawing has the following colors: body (A), blue; end (B), gray; dot, red; end (D), gold. The significant figures are 6, 8 (68) and the decimal multiplier is 100, so the resistance is 6800 ohms. The tolerance is ± 5%.



Fixed composition resistors

Fig. 18-5 — Color coding of fixed composition resistors. The color code is given in Table 18-III. The colored areas have the following significance:

A — First significant figure of resistance in ohms.
B — Second significant figure.
C — Decimal multiplier.
D — Resistance tolerance in per cent. If no color is shown, the tolerance is ± 20%.

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.

Loudspeaker 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*

Eliminating Broadcast Interference

It is your duty as an amateur to make sure that the operation of your station does not interfere with broadcasting or other radio services because of any shortcomings in your equipment. Failure to observe this rule may lead to curtailed operating privileges — a situation that is easily avoidable if you build and adjust your transmitter according to good practice.

However, there is a larger obligation — to eliminate broadcast interference to the greatest possible extent even when your own transmitter is not at fault. The institution of amateur radio cannot continue to flourish in the face of ill feeling on the part of a large segment of the general public — ill feeling that is only too readily generated if the public's favorite radio programs are broken up by amateur transmissions. It is no exaggeration to say that the future of amateur radio depends in large part on the efforts you exert now to make it possible for your neighbors to continue to enjoy their radio reception while you pursue your transmitting activities. It is unfortunately true that most interference to broadcasting is directly the fault of present-day broadcast-receiver construction. Nevertheless, the amateur can and should help to alleviate interference even though the responsibility for it does not lie with him.

The regulation of the Federal Communications Commission covering interference to broadcasting is quoted below:

§12.152. *Restricted operation.* (a) If the operation of an amateur station causes general interference to the reception of transmissions from stations operating in the domestic broadcast service when receivers of good engineering design including adequate selectivity characteristics are used to receive such transmissions and this fact is made known to the amateur station licensee, the amateur station shall not be operated during the hours from 8 o'clock P.M. to 10:30 P.M., local time, and on Sunday for the additional period from 10:30 A.M. until 1 P.M., local time, upon the frequency or frequencies used when the interference is created. (b) In general, such steps as may be necessary to minimize interference to stations operating in other services may be required after investigation by the Commission.

FCC recognizes the fact that much of the interference that occurs is because receivers are not capable of rejecting signals far outside the frequency band to which the receiver is tuned. That is why the phrases "general interference" and "receivers of good engineering design including adequate selectivity characteristics" are used in Rule 12.152. "Quiet hours" are not imposed unless it is shown that the interference is actually the fault of the transmitter.

Once you have determined that your transmitter is free from parasitic oscillations, spurious radiations, key clicks and modulation splatter, you can tackle the BCI problem with a clear conscience and the firm conviction that the answer is to be found in the b.c. receiver. Be sure your transmitter is clean *first.* From then on you have a twofold job: convincing the owner of the receiver that his set is at fault (not always the easiest thing in the world, especially if the receiver is fairly new), and finding out just why the interference occurs. The first is almost wholly a matter of using the right approach; you may need all the tact at your command to convince him that you know what you're talking about and are sincerely trying to help. His natural tendency, as one with no technical knowledge of radio at all, will be to blame you because you're coming in on the broadcast band where you obviously don't belong. You may have to overcome the suspicion that everything you say about his receiver is just so much camouflage to cover up something wrong with your transmitter.

In brief, to be successful in eliminating BCI you have got to win the listener's coöperation.

GETTING LISTENER COÖPERATION

The battle is 75 per cent won when you've earned the listener's confidence in your technical ability and your sincerity in wanting to clear up interference. Here are a few pointers on how to go about it.

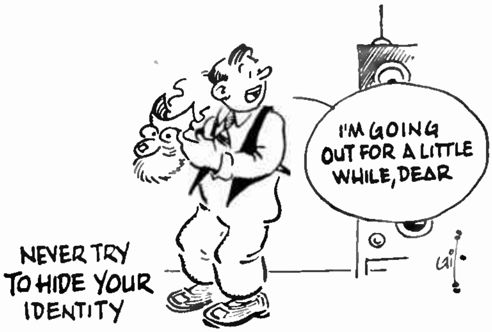
Clean House First

We've said above that the first obligation of every amateur is to clean up his transmitter so it has no radiations outside the bands assigned for amateur use. Even then, you'll probably find that you have a BCI problem in your own house.

So clean up your household BCI first! It is always convincing if you can say — and demonstrate — that you do not interfere with broadcast reception in your own home.

Don't Hide Your Identity

If a listener thinks that you are "trying to get away with something" he will not only be unwilling to coöperate, but may be actively hostile. As a general rule, whenever you change location, or mode of transmission, or increase power, or put up a new antenna, check with your neighbors to make sure that they are not



experiencing interference. Announce your presence and conduct occasional tests on the air, requesting anyone whose reception is being spoiled to let you know about it so that you may take steps to eliminate the trouble.

Act Promptly

Do something to show the listener that you are concerned for his welfare as soon as a complaint is received. The average person will tolerate a limited amount of interference, but no one can be expected to put up with frequent and extended interruption of his listening pleasure. The sooner you take steps to eliminate the interference, the more agreeable the listener will be; the longer he has to wait for you, the less willing he will be to coöperate.

Present Your Story Tactfully

Put yourself in the listener's place. He has a right, he believes, to interference-free reception of the broadcast programs he likes. When you interfere, his natural reaction is to assume that you are the one at fault. When you call on him, explain that you do not operate on the frequencies to which he wants to listen, and the real trouble is that you and he happen to be located so close to each other. Explain to him that there are thousands of stations operating simultaneously, all the time, and that the problem of rejecting all but the one he happens to want to hear is one of receiver design. Point out that the average broadcast receiver is made to sell as cheaply as possible, and that features that would prevent interference from near-by stations are left out.

It should be explained to the listener that if it is simply the presence of your strong signal on his receiving antenna that causes the difficulty, the situation can be cleared up by a wavetrap. In other cases the wiring of the receiver itself is picking up your signal, and such cases can be cured only by suppressing this unwanted pick-up in the receiver itself; in other words, some modifications will have to be made in the receiver if he is to expect interference-free reception.

Arrange for Tests

Most listeners are not very competent observers of the various aspects of interference. If at all possible, enlist the help of another amateur and have him operate your transmitter while you see what happens at the affected broadcast set. You can then determine for yourself where the trouble is most likely to be.

It is a good idea to take along a wavetrap when you arrange such a test. If the receiver is one having an external antenna, it may be possible to cure the interference then and there.

Avoid Working on the Receiver

If your tests show that the fault has to be remedied in the receiver itself, *do not offer to work on the receiver.* It is not your fault that the receiver design is defective. Recommend that the work be done by a reliable serviceman, and offer to advise the latter as to the cause and cure if necessary.

It is inadvisable to tackle broadcast receivers, particularly the midget varieties, unless you have had experience working on them. In any event, if you do work on the receiver yourself the chances are that if anything goes wrong later on you'll be blamed for it. Explain that, while you may be technically competent to make the necessary modifications, radio servicing is best left to those who specialize in it, and that you are sure he, the owner, will prefer to have the work done by someone whom he can hold responsible.

If the owner of the receiver obviously prefers to have you make the modifications, do so only with the understanding that it is purely as a favor and because you are anxious to coöperate. Make him understand, with as much tact as possible, that the responsibility for the interference does not lie with you (your transmitter having previously been checked and found OK); if the receiver responds to frequencies to which it is not tuned that is a defect in its design. You also have no obligation to pay for having the receiver modified. If you do the work yourself you should not make any charge, of course. In that event, insist that you must take the receiver to your own shop in order to work on it properly; you will be able to tell immediately whether the changes you make effect an improvement and therefore can work more rapidly and conveniently — and without turning the owner's

living room into a repair shop. If it is necessary to do some work in the listener's home, *be neat* in the work you do. Remember, the listener's living room cannot be treated in the same manner you would treat your own ham shack!

In General

In this "public relations" phase of the problem a great deal depends on your own attitude. Most people will be willing to meet you half way, particularly when the interference is not of long standing, if you as a person make a good impression. Your personal appearance is important. So is what you say about the receiver. A display of lofty technical superiority is more likely to generate resentment than coöperation. Above all, don't make remarks on the air about "bum broadcast receivers" and "cheap midgets." No one takes kindly to hearing his possessions publicly derided. If you discuss your BCI problems on the air, do it in a constructive way — one calculated to increase listener coöperation, not destroy it.

RADIO-CLUB BCI COMMITTEES

Organized amateur radio clubs can do a lot to pave the way toward coöperation between individual amateurs and the broadcast listeners. Most clubs maintain interference committees charged with handling both the public relations and the technical aspects of BCI. Through such committees, technical assistance is made available to all members of the club so that those less qualified can have the benefit of the experience of others. The committee should also maintain contact with the local radio servicemen, supplying them with information and technical assistance whenever possible. The committee can maintain valuable contacts with the local newspapers, broadcast stations and other authorities to provide the right kind of publicity for the efforts of individuals or groups who are trying to clear up BCI problems.

League Aids

The Communications Department of ARRL, as one of its services to affiliated clubs, has prepared material suggesting various ways in which local clubs can form interference committees, and methods by which such groups can function efficiently for the good of all concerned. This material is available to affiliated clubs on request, addressed to ARRL headquarters.

Causes and Cure of BCI

There are no magic cures for all cases of interference to standard AM broadcasting. The great number of different types of broadcast receivers makes it necessary to tailor the remedy to the specific set. However, interference does usually fall into one or more rather well-defined categories. A knowledge of the general types of interference and the methods required to eliminate it will lead to a rapid appraisal of the situation and will avoid much cut-and-try in finding a cure.

Transmitter Defects

Out-of-band radiation is something that must be cured at the transmitter. Parasitic oscillations are a frequently unsuspected source of such radiations, and no transmitter can be considered satisfactory until it has been thoroughly checked for both low- and high-frequency parasitics. Very often parasitics show up only as transients, causing key clicks in c.w. transmitters and "splashes" or "burps" on modulation peaks in AM transmitters. Methods for detecting and eliminating parasitics are discussed in Chapter Six.

In c.w. transmitters the sharp make and break that occurs with unfiltered keying causes transients that, in theory, contain frequency components through the entire radio spectrum. Practically, these transients do not have very much amplitude at frequencies very far away from the transmitting frequency. Nevertheless they are often strong enough in the immediate vicinity of the transmitter to cause serious interference to broadcast reception. Key clicks can be eliminated by the methods detailed in Chapter Eight.

A distinction must be made between clicks generated in the transmitter itself and those set up by the mere opening and closing of the key contacts when current is flowing. The latter are of the same nature as the clicks heard in a receiver when a wall switch is thrown to turn a light on or off, and may be more troublesome nearby than the clicks that actually go out on the signal. A filter for eliminating them usually has to be installed as close as possible to the key contacts.

Overmodulation in AM 'phone transmitters generates transients similar to key clicks. It can be prevented either by using automatic systems for limiting the modulation to 100 per cent, or by continuously monitoring the modulation. Methods for both are described in Chapter Nine. In this connection, the term "overmodulation" means any type of nonlinear modulation that results from overloading or inadequate design. This can occur even though the actual modulation percentage is less than 100.

BCI is frequently made worse by radiation from the transmitter, power wiring, or the r.f. transmission line. This is because the signal causing the interference, in such cases, is radiated from wiring that is nearer the broadcast receiver than the antenna itself. In such cases much depends on the method used to couple the transmitter to the antenna, a subject that is discussed in Chapter Ten. If it is at all possible, too, the antenna itself should be placed so that it is not in close proximity to house wiring, telephone and power lines, and similar conductors.

Oscillator Harmonic Responses

When the transmitter is operating on a low-frequency band — 3.5 or 7 Mc., usually — a near-by superheterodyne receiver tuned to the broadcast band frequently will give a response to the amateur signal when a harmonic of the receiver oscillator falls on a frequency equal to the amateur frequency plus or minus the receiver's intermediate frequency. These harmonic responses tune in and out on the receiver dial just like a broadcast signal, although the tuning rate is more rapid.

These spurious signals would not occur if the receiver oscillator had no harmonics, but there is usually nothing that can be done about the oscillator circuit design. The problem is to reduce the amplitude of the amateur signal in the front end of the b.c. receiver. If the receiver uses an external antenna a wavetrap at the receiver antenna terminals may help. It may also be helpful to reduce the length of the receiving antenna — and particularly to avoid a length that might be near resonance at the transmitter frequency — or to change its direction with respect to the transmitting antenna. If the signal is being picked up by the antenna it will disappear when the antenna is disconnected. If it is still present under these circumstances the pick-up is in the set wiring or the power circuits. A line filter may be tried for the latter. Pick-up on the set wiring can only be cured by installing some shielding around the r.f. circuits. Copper window screening cut and fitted to size will usually do the trick.

Since harmonic responses occur at definite frequencies on the receiver dial, it is always possible to choose an operating frequency that will not give such a response on top of the broadcast stations that are favored in the vicinity. While your signal may still be heard when the receiver is tuned off the local stations, it will at least not interfere with program reception.

Cross-Talk

With some of the older receivers, particularly of the nonsuperheterodyne type, interference occurs only when the receiver is tuned to a strong broadcast signal and disappears between stations. This is cross-modulation, a result of rectification in one of the early stages of the receiver. It is not so likely to occur in more modern sets using a remote-cut-off tube in the antenna stage.

One remedy is to install remote-cut-off tubes in the r.f. stages and put in an a.v.c. circuit. However, this is a major operation and frequently is not practicable. The remaining thing is to reduce the strength of the amateur signal at the grid of the first tube in the receiver. Wavetraps, a smaller antenna, and a different antenna position should be tried. Additional shielding about the r.f. circuits also will sometimes effect an improvement.

Blanketing

"Blanketing" is a form of interference that partially or completely masks reception, no matter where the broadcast receiver is tuned. Each time the carrier is thrown on, whether by keying or for modulation, the program disappears or is greatly reduced in amplitude. Amplitude modulation in such a case is usually distorted rather severely.

When the transmitter is operated on the lower frequencies this type of interference occurs only when the receiver and transmitter are very close together. It is the result of simple overloading of the receiver by the very strong field in the vicinity of the transmitting antenna. It occurs principally on receivers using external antennas (as contrasted with a built-in loop), and can be reduced by the steps recommended above; i.e., using a short receiving antenna, repositioning the antenna with respect to the transmitting antenna so the pick-up is reduced, or using wavetraps and line filters.

When the transmitter is operated on 28 Mc. or v.h.f. "blanketing" occurs rather rarely, and then only when the transmitting and receiving installations are located exceptionally close together.

Audio-Circuit Rectification

The most frequent cause of interference from operation at the higher frequencies is from rectification of a signal that by one means or another gets into the audio system of the receiver. In the milder cases an amplitude-modulated signal will be heard with reasonably good quality, but is not tunable — that is, it is present no matter what the frequency to which the receiver dial is set. An unmodulated carrier may have no observable effect in such cases beyond causing a little hum. However, if the signal is very strong there will be a reduction of the audio output level of the receiver whenever the carrier is thrown on. This causes an annoying "jumping" of the program when the interfering signal is keyed. With 'phone transmission the change in audio level is not so objectionable because it occurs at less frequent intervals. Also, ordinary rectification gives no audio output from a frequency-modulated signal, so the interference can be made almost completely unnoticeable if FM or PM is used instead of AM.

Interference of this type is most prevalent in a.c.-d.c. receivers. The pick-up may occur in the audio-circuit wiring or the interfering signal may get into the audio circuits by way of the line cord. Power-line pick-up can be treated by means of line filters, but pick-up in the receiver wiring requires individual attention. Remedies that have been found successful are described in the sections following.

CHECKING AND CURING BCI

When a case of broadcast interference comes to your attention, set a definite time to conduct tests and then prepare to do the job as expeditiously as possible. Provide yourself with one or two wavetraps and line filters, since they can be tried immediately without getting into the receiver. As suggested before, get another

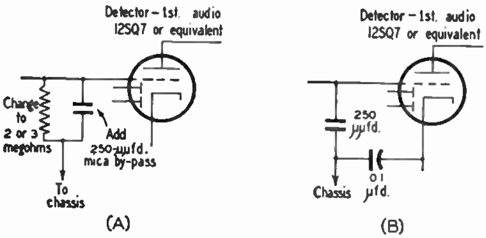


Fig. 19-1 — Two methods of eliminating r.f. from the grid of a combined detector/first-audio stage. At A, the value of the grid leak is reduced to 2 or 3 megohms, and a mica by-pass condenser is added. At B, both grid and cathode are by-passed.

amateur to operate your transmitter while you do the actual observing and testing at the listener's receiver. The procedure outlined below will save time in getting at the source of the trouble and in satisfactorily eliminating it.

1) Determine whether the interference is tunable or not. This will usually indicate the methods required for elimination of the trouble, as it will show which of the general types of interference discussed above is present. In severe cases it is possible that two or more types will be present at the same time, and steps will be necessary to eliminate each type.

2) If the set has an external antenna, disconnect it and turn the volume control up full. If the interference is no longer present, it is merely necessary to prevent the r.f. appearing on the antenna from entering the set. If wavetraps reduce the amplitude of the interfering signal but do not eliminate it entirely, try a short piece of wire as a receiving antenna.

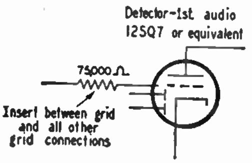


Fig. 19-2 — Using a 75,000-ohm resistor to form a low-pass filter with the tube capacitance. The resistor must be mounted at the tube pin, between the grid and all other grid connections.

Alternatively, the antenna may be relocated. It should be placed as far as possible from the transmitting antenna, and should run at right angles to it to minimize coupling.

If the interference persists after the antenna is disconnected, the search is narrowed to an investigation of whether the signal is coming in on the power lines, or is being picked up directly on the receiver wiring.

3) Check for power-line interference by using a sensitive wavemeter such as that described in Chapter Sixteen of this *Handbook* to probe along the a.c. cord that connects the set to the power source. Checks should be made at the transmitter frequency, and also at harmonic frequencies. If r.f. is detected in the line, by-pass both sides of the a.c. line to ground with 0.005-μfd. mica condensers at the point where the line cord enters the set. (A simple plug-and-socket adapter can be made up for this purpose before visiting the listener.) If this does not completely eliminate the interference, try a line filter designed for the operating frequency.

4) If it is evident that the interference is being picked up on the receiver wiring, explain the situation to the owner and tell him that the exact cause cannot be determined without removing the chassis from the cabinet, and that, in any event, the receiver will have to be modified somewhat if the interference is to be eliminated. As suggested before, recommend that the actual work be done by a radio serviceman. Offer to check into the cause yourself, if he wishes and will allow you to take the set to your shop (with the understanding that

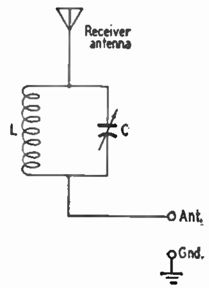


Fig. 19-3 — A simple wavetrap circuit. *L* and *C* must resonate at the frequency of the interfering signal. Suitable constants are tabulated below.

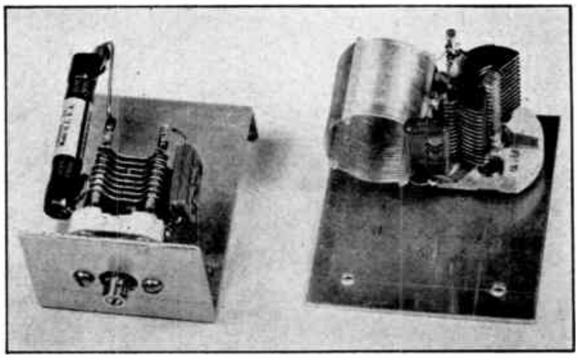
Band	*C*	*L*				
3.5	140 μμfd.	16 μh.,	32 turns	#22,	1″ diam.,	1″ long
7	100 μμfd.	6	19	#22,	1″	1″
14	50 μμfd.	3.5	14	#18,	1″	1″
21	35 μμfd.	2.2	12	#18,	1″	1″
28	25 μμfd.	1.5	9	#18,	1″	1″

you will not make any changes in the receiver without his express permission) so the serviceman can be told what needs to be done.

5) In the event that the owner allows you to take the receiver, set it up near your transmitter and check to see if the amplitude of the interfering signal is changed by various settings of the receiver volume control. If the volume of the interference changes with changes in the volume control, the r.f. is entering the set *ahead* of the volume control. If it is unaffected by the volume control, it is getting into the audio stages at a point following the volume control.

6) Pin the source down, if it is ahead of the volume control, by removing one tube at a time until one is found that kills the interfer-

Fig. 19-4 — Two examples of simple wavetraps. In the unit on the left, a 2-meter r.f. choke (Ohmite Z-0) is used with a 50-μμfd. variable condenser shunted by a 22-μμfd. mica condenser to form a trap for the 14-Mc. band. The larger unit on the right uses a 32-turn "Miniductor" (B & W) with a 100-μμfd. variable condenser shunted by a 67-μμfd. mica condenser to cover the 3.5- to 4-Mc. range. Both units are bracket-mounted with provision for mounting within the cabinet of a broadcast receiver. The circuit is shown in Fig. 19-3.



ence when it is removed. In sets using series-connected filaments, this will be possible only if a tube of equal heater rating, and with all but the heater pins clipped off, is *substituted* for the tube.

7) Determine which element (or elements) of the tube is picking up the interference by touching each tube pin with a test lead about three feet long. The lead, acting as an antenna, will cause the interference to increase when it is placed on a tube pin that is contributing to the interference. Once the sensitive points have been determined, the trouble can be eliminated by shielding the leads connected to the tube element that is affected, and by shielding

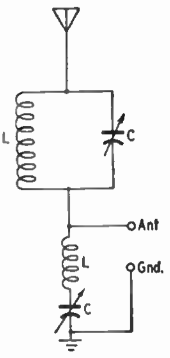


Fig. 19-5 — Combination of parallel-tuned and series-tuned wavetraps to obtain greater attenuation of the interfering signal. Values for *LC* are the same as those listed below Fig. 19-3.

the tube itself. Grid leads are the principal offenders, especially the long leads that run from a tube cap to a tuning condenser, and it may be necessary to shield several parts of the set before the interference is eliminated.

8) If the pick-up is found to be in the audio system — as is the case in many sets, especially when the transmitter is operating at 28 Mc. or higher — it can be eliminated by one or another of the methods shown in Figs. 19-1 and 19-2. Fig. 19-1A is a method that has proved successful with many a.c.-d.c. receivers. The value of the grid leak in the combined detector/first-audio tube (usually a 12SQ7 or its equivalent) is reduced to 2 or 3 megohms. The grid is then by-passed for r.f. with a 250-μμfd. mica condenser. Fig. 19-1B is a similar method. A third method that has worked in a.c.-d.c. receivers requires only that the heater of the detector/first-audio stage be by-passed to ground with a 0.001-μfd. condenser. The method shown in Fig. 19-2 uses a 75,000-ohm ½-watt resistor to form, with the tube capacitance, a low-pass filter. The resistor is connected between the grid pin of the audio stage and *all other* wires connected to the grid. In all cases, both sides of the a.c. line should be by-passed to chassis with 0.001- to 0.01-μfd. condensers.

Wavetraps and A.C. Line Filters

In its simplest form, a wavetrap consists of a parallel-tuned circuit that is connected in series with the broadcast antenna and the antenna post of the receiver. It should be designed to resonate at the frequency of the interfering signal. The circuit of a simple trap is shown in Fig. 19-3. If interference results from operation in more than one amateur band several traps may be connected in series, each tuned to the center of one of the bands in which operation is contemplated. A more elaborate form is illustrated in Fig. 19-5, in which a combination of a parallel-tuned circuit and a series-tuned circuit is used. To adjust the wavetrap, have another licensed amateur operate the transmitter while you tune

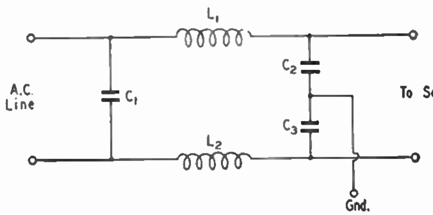


Fig. 19-6 — A.c. line filter for receivers. The values of C_1, C_2 and C_3 are not generally critical; capacitances from 0.001 to 0.01 μfd. can be used. L_1 and L_2 can be a 2-inch winding of No. 18 enamelled wire on a half-inch diameter form.

the trap for maximum attenuation of the interference. The trap should be connected to the broadcast receiver and the normal receiving antenna should be connected in series with the trap, as shown in the figures.

A common form of a.c. line filter is shown in

Fig. 19-6. This type of filter will usually do some good if the signal is being picked up on the house wiring and transferred to the set by way of the line cord. The values used for the coils and condensers are in general not critical. The effectiveness of the filter will depend considerably on the ground connection used, and it may be necessary to try grounding to several different possible ground connections to secure the best results. A filter of this type will usually not be very helpful if the signal is being picked up on the line cord itself, which may be the case when the transmitter is on v.h.f. In such a case it should be installed inside the receiver chassis and grounded to the chassis at the point where the line cord enters.

The tuned filter shown in Fig. 19-7 is often more effective than the untuned type when only one frequency needs to be eliminated. After installation, the condenser is simply adjusted to reduce the interference to the greatest possible extent.

It is advisable to mount either type of filter in a small shield, both to prevent pick-up in the filter itself and to make it less conspicuous when it has to be installed in a listener's home.

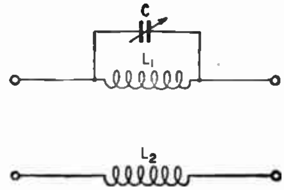


Fig. 19-7 — Resonant filter for the a.c. line. A single condenser tunes both L_1 and L_2, which are unity-coupled, one wound on top of the other. Constants for amateur bands are tabulated below.

Band	*C*	L_1 - L_2
3.5	140 + 150 (fixed)	25 t. No. 18, 1¼″ dia. × 2⅜″ long.
7	140 μμfd.	18 t. No. 18, 1¼″ dia. × 2⅜″ long.
14	100 μμfd.	12 t. No. 18. 1¼″ dia. × 2⅜″ long.
21	50 μμfd.	10 t. No. 18, 1¼″ dia. × 2⅜″ long.
10	25 μμfd.	9 t. No. 18. 1½″ dia. × 2⅜″ long.

D.c.c. wire is recommended for all coils.

Interference with Television

Interference with reception of television signals presents a more difficult problem than interference with ordinary AM broadcasting. In the latter case it is comparatively easy to clean up a transmitter so that it will have no spurious radiations in the broadcast band. Clearing up interference difficulties then becomes a matter of overcoming deficiencies in the selectivity of the broadcast receiver.

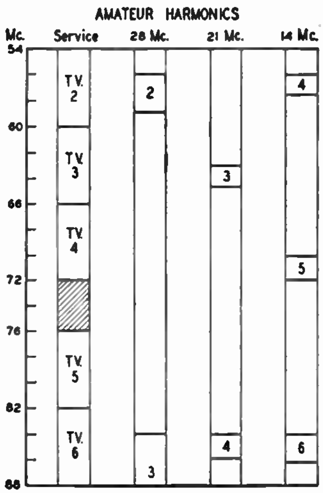
In the case of television reception similar receiver deficiencies exist, and must be treated by methods similar to those used for low-frequency broadcasting. However, a more serious situation for the amateur arises because harmonics of his transmitting frequency fall in many of the television channels. The relationship between television channels and harmonics of amateur bands from 14 through 28 Mc. is shown in Fig. 19-8. Harmonics of the 7- and 3.5-Mc. bands are not shown because they fall in every television channel. Also, the harmonics above 54 Mc. from these bands are of such high order that they are usually rather low in amplitude. They are not, however, too weak to interfere if the television receiver is quite close to the amateur transmitter.

Low-order harmonics — up to about the fourth or fifth — are usually the most difficult to eliminate. The degree of harmonic suppression required is very great, particularly when the television receiver is nearby and the signals from TV stations are weak. Effective harmonic suppression has three separate phases:

1) Reducing the amplitude of harmonics generated in the transmitter. This is a matter of circuit design and operating conditions.

2) Preventing stray radiation from the transmitter and from associated wiring. This requires shielding and filtering of all circuits and leads from which radiation can take place.

3) Preventing harmonics from being fed into the antenna.



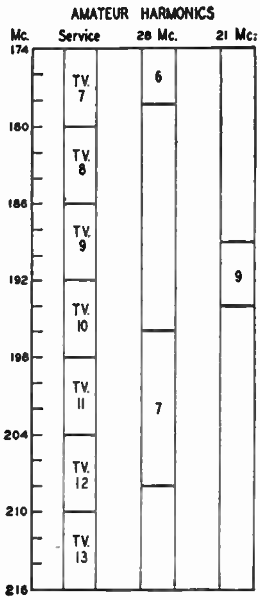


Fig. 19-8 — Relationship of amateur-band harmonics to TV channels. Harmonic interference is most likely to be serious in the low-channel group (54 to 88 Mc.)

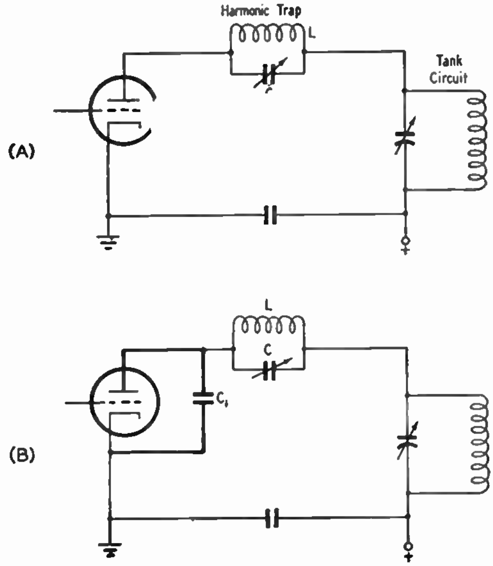


Fig. 19-9 — Using trap circuits in the amplifier plate lead to prevent a given harmonic from appearing in the tank circuit. The *LC* circuit should be tuned to the harmonic to be trapped out. A high *C/L* ratio will give best results and have least effect on the normal operation of the amplifier. The capacitance in use should be at least 25 to 50 μμfd. at harmonic frequencies in the 50–88 Mc. range. The trap condenser, *C*, should have sufficient plate spacing to stand the resonant harmonic voltage that builds up in the circuit; midget condensers are adequate for moderate power.

The circuit at B uses an additional condenser, C_1, to provide a low-impedance path between plate and cathode for harmonics.

Transmitter Design

Every stage in the transmitter that can generate harmonics can be the cause of TVI. The following methods have been found to be effective in reducing harmonic output:

1) Do all your frequency multiplying in stages operating at a very low power level and build up the power on the output frequency only. The lower the power of the intermediate stages in the transmitter the less the chance that they will cause harmonics of appreciable amplitude to be radiated.

2) Use a high ratio of capacitance to inductance in tank circuits, and keep the r.f. path from plate to cathode as short as possible. Use low-inductance connections, such as copper strip, in preference to wire in the plate-cathode circuit. A high-*C* grid circuit also is helpful.

3) Use link coupling between the driver and the final r.f. amplifier. This prevents harmonics generated in the driver stage from being amplified in the last stage.

4) Don't overbias and overdrive any stage in the transmitter. In the last stage, don't go any farther beyond cut-off bias than is necessary for proper modulation, if you're using amplitude modulation, and use just enough driving power to give linear modulation. In c.w. and FM transmitters it is advisable to use no more than cut-off bias and to keep the excitation down. Tubes operated in this way will work at 60- to 70- per-cent efficiency, and if the tube ratings are observed the difference in output as compared with overdriving is small.

5) In transmitters operating at the higher frequencies it may be necessary to use a trap, tuned to the frequency of the harmonic that causes interference, in the plate lead of the final amplifier. In some cases this may also be necessary in the driver stage. The connections are shown in Fig. 19-9A. A still better arrangement is shown at B in Fig. 19-9. In this case a low-impedance path from plate to cathode, for the harmonic frequencies, is provided by C_1 and the shortest practicable leads as indicated by the heavy lines. C_1 should have a capacitance of 25 to 50 μμfd. and should have inherently low inductance. The vacuum-type condenser is excellent for this purpose. Alternatively, a suitable condenser can be constructed of concentric tubes of aluminum or copper, or flat strips of the same materials. Such condensers can be built so that practically the entire path from plate to cathode is in the condenser and not in external leads. The capacitance of the normal tank tuning condenser can be reduced correspondingly, since C_1 becomes part of the tank circuit.

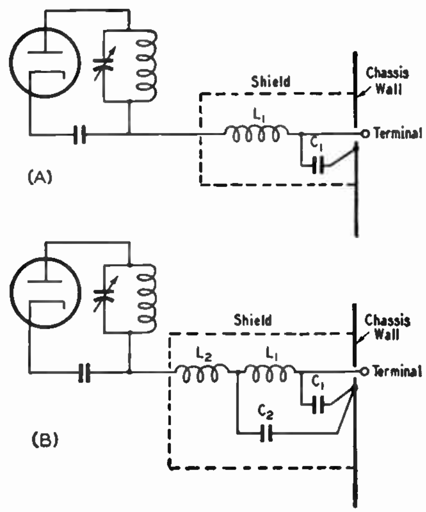


Fig. 19-10 — R.f. filters in power-supply leads to prevent harmonic radiation. Values for L_1 and C_1 are best determined by experiment. Capacitances from 100 μμfd. to 0.001 μfd. usually will work well. L_1 should have an inductance of 1 mh. or less, and can simply be a 1-watt resistor (high values) used as a winding form and wound full of No. 26 to No. 30 wire, depending on the current to be carried.

A double filter, as shown at B, may be required to do a good job of filtering out a number of different harmonics that are rather widely separated in frequency. In such a case L_2 should be the larger of the two coils. If different capacitance values are required at C_1 and C_2, the larger should be placed at C_2.

The effectiveness of such filters, and the proper constants for them, can be determined with the help of a crystal-detector wavemeter of the type described in Chapter Sixteen.

Preventing Stray Radiation

Even a minute amount of harmonic current on power leads can cause destructive interference in near-by television receivers. These currents invariably exist if no attempt is made to eliminate them. Radiation also can take place from the r.f. circuits of the transmitter. However, the latter type of radiation seems to be less serious than radiation from power leads, no doubt because the transmitter circuits usually are small compared with the lengths of supply leads. It can be reduced to negligible proportions by a moderate amount of shielding around the transmitter. An ordinary metal cabinet usually provides sufficient shielding. Alternatively, a shield to enclose the transmitter can be constructed easily and inexpensively from copper window screening.

Supply leads can be filtered by by-passing each lead at the point where it leaves the chassis and by inserting chokes in the leads at the same point, as indicated in Fig. 19-10. Double filters sometimes may be required. Shielding around the transmitter helps the filters to do a better job, since it prevents coupling between the transmitter circuits and the power leads *outside* the transmitter.

The presence of harmonic currents on power leads can be detected with the aid of a crystal-detector wavemeter of the type described in Chapter Sixteen. Such a wavemeter is an absolute essential for effective adjustment of filters, because it will show what progress is being made in each lead as individual filters are tried. If any reading at all is obtained around leads, on harmonics falling in television channels, interference with television reception in the immediate neighborhood is certain to result.

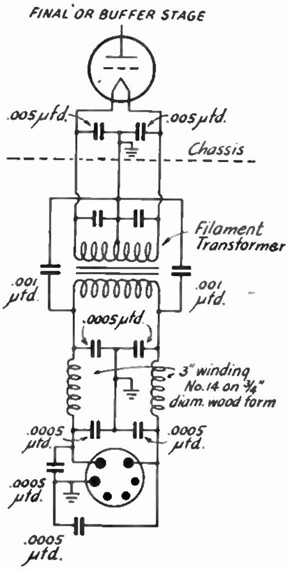


Fig. 19-11 — **Filament-supply filter. When the filament transformer is on the amplifier chassis the filter may be installed in the transformer-primary circuit, as shown. If the filament transformer is remotely located the filter should be in the filament leads themselves, before leaving the amplifier chassis.**

The wavemeter also will show which parts of the transmitter itself are "hot" at the harmonic frequencies.

In push-pull amplifiers the even harmonics are particularly strong in the r.f. return circuit to the cathodes, and special care must therefore be used to filter the filament-supply leads. If the filament transformer is mounted on the same chassis as the amplifier, the primary leads may be filtered by the arrangement shown in Fig. 19-11. In cases where the filament transformer is not on the amplifier chassis the same type of filter should be installed in the filament leads before leaving the r.f. chassis. The wire in the chokes must be heavy enough to carry the filament current without excessive voltage drop.

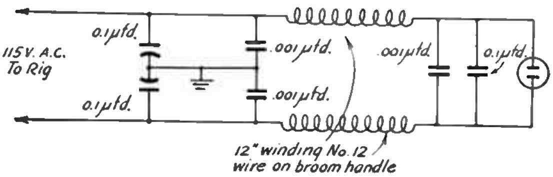


Fig. 19-12 — **Power-line filter for installation on the line side of the complete transmitter.**

It is also advisable to filter the a.c. line carrying the entire power for the transmitter. A suitable circuit is shown in Fig. 19-12.

Preventing Harmonic Radiation from the Antenna

Harmonics that get into the antenna system should be low in amplitude after the measures described above have been taken. Nevertheless, they may be strong enough to interfere with television reception in the immediate vicinity. Harmonics may appear at the antenna either because of insufficient selectivity in the coupling system between antenna and transmission line or because of coupling through stray capacitance between the final tank circuit and the transmission line.

Stray capacitive coupling is probably the more difficult to eliminate. Methods for reducing it are discussed in Chapter Ten in the section on coupling the transmitter to the line.

The additional selectivity provided by a tuned antenna coupler of the type discussed in Chapter Ten is very helpful in reducing harmonic radiation from the antenna. Such a coupler always should be used in preference to going directly from an output link to the line, particularly on frequencies from 14 to 28 Mc. where the low-order harmonics fall in television channels. The remaining harmonic output, when the harmonic current flows as a true transmission-line current, can be further reduced by using linear traps across the line. A shorted quarter-wave line has a very high impedance at the fundamental frequency and can be connected across a transmission line without affecting the energy transfer. However, such a "trap" is one-half wave long at the second harmonic, and therefore appears as a short-circuit across the line for the second harmonic. (This is also true at all *even* harmonics.) If the linear trap is connected ¼ wavelength from the antenna-coupler terminals, as indicated in the drawing, the terminals also will be short-circuited at the second harmonic.

A third harmonic can be reduced by using a linear trap one-half wavelength long at the third harmonic (⅙ wavelength at the fundamental). However, such a trap would upset the operation of the transmission line at the fundamental frequency, so its length must be extended an additional 1/12 wavelength to make

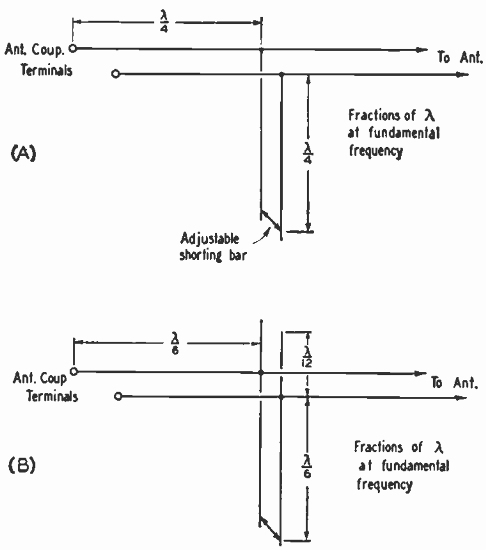


Fig. 19-13 — Linear trap circuits for second harmonic (A) and third harmonic (B). These will suppress harmonic currents of the transmission-line type but will not suppress parallel currents. The traps may, however, be helpful in detuning the line for parallel currents if attached at a distance half that given in the drawings from the antenna-coupler terminals.

the total length ¼ wave at the fundamental. This is shown in Fig. 9-13B. A short-circuit for the third harmonic will be reflected at the antenna-coupler terminals when the trap is attached ⅙ wavelength from the coupler.

Linear traps are suitable only when the antenna is used for single-band operation. If the antenna is a multiband affair traps may be cut for each band and changed when operation is shifted to another band. Since this becomes a cumbersome operation, it is best to reduce the harmonic as much as possible *before* it gets into the antenna system.

In making up a linear trap, remember that its physical length will be determined by the velocity factor of the line from which it is made. If solid-dielectric line is used for the trap it will be considerably shorter than the same fraction of a wavelength in space.

RECEIVER DEFICIENCIES

If your transmitter generates harmonics that fall in a television channel there will be interference in that channel regardless of the characteristics of the receiver. But even if the transmitter harmonics are satisfactorily low in amplitude there may still be interference if the receiver's shielding and selectivity are inadequate. Spurious responses because of receiver inadequacies are particularly likely to occur when the receiver and transmitter are quite close. They usually result from the fact that the strong fundamental-frequency signal from the transmitter overloads some circuit in the receiver.

Many television receivers have "front ends" that are inherently unselective and not well balanced — that is, they will give strong response to parallel currents on the receiving transmission line. Usually, the transmission line picks up a great deal more energy from a near-by transmitter than the television receiving antenna itself, causing parallel currents that should be, but are not, rejected by the receiver's input circuit. A strong signal that overloads the first or second stages in the receiver will cause the receiver itself to generate harmonics that fall in the television channels. This situation can be cured by using shielded transmission line ("twinax") on the receiving installation and by connecting wave-traps (tuned to the fundamental frequency of the transmitter) in both transmission line leads right at the receiver antenna terminals.

The use of twinax transmission line also will be helpful in reducing response to harmonics actually being radiated from the transmitter or transmitting antenna. In most receiving installations the transmission line is very much longer than the antenna itself, and is consequently far more exposed to the harmonic fields from the transmitter. Much of the harmonic pick-up, therefore, is on the receiving transmission line when the transmitter and receiver are quite close together. Shielded line, plus relocation of either the transmitting or receiving antenna to take advantage of directive effects, often will result in reducing the harmonic pick-up to a level that does not interfere with reception.

Many television receivers do not have enough isolation between the antenna and intermediate-frequency circuits. As a result, signals that fall in or near the intermediate-frequency passband (roughly 22 to 27 Mc. in most current receivers) will cause interference either to the picture or to the sound. Wave-traps tuned to the fundamental frequency of the interfering signal and installed right at the receiver antenna terminals will reduce interference of this type. If the receiver and transmitter are very close a complete cure may not be possible without shielding the receiver's i.f. circuits. I.f. interference is particularly likely from the 21-Mc. band when the receiver has its sound i.f. channel centered at or near 21.25 Mc. Realigning the receiver to a somewhat higher frequency (sound channel at 21.9 Mc.) usually will cure this type of interference.

Another type of interference, wholly attributable to lack of receiver selectivity, occurs from operation in the 50-Mc. band. A strong 50-Mc. signal on the receiving antenna will overload the receiver, particularly when the receiver is tuned to Channel 2. Wavetraps tuned to the frequency of the interfering signal, installed at the antenna input terminals of the receiver, will help reduce this type of interference. It is also helpful to work at the low-frequency end of the 50-Mc. band, since this frequency is farthest removed from Channel 2 Shielding of the receiver's r.f. circuits also may be necessary.

CHAPTER 20

Operating a Station

The enjoyment of our hobby usually comes from the operation of our station once we have finished its construction. Upon the *station* and its *operation* depend the communication records that are made. We have taken every bit of care that was possible in constructing our transmitter, receiver, frequency-measuring and monitoring equipment, and in erecting a suitable antenna system. Unless we use good judgment and care in operating our stations, we shall fall far short of realizing the utmost in results achieved. More than this, unless we do the right thing, we may interfere with other stations or delay their work, thus acquiring a bad reputation. Occasionally you will pick up an amateur whose method of operating is so clean-cut, so devoid of useless effort, so snappy and systematic, that your respect is gained and it is a pleasure to listen and work with him. One benefits proportionately from what one puts into amateur radio, so make the most of your hobby in the time and opportunities given you.

For best results, the transmitter should be adjusted for stable, satisfactory operation on several amateur frequencies. Known settings for definite frequencies will enable the operator to change frequency quickly at any time. Whenever such a change is made, be sure to check the transmitter frequency accurately. There is absolutely *no excuse* for a station operating off frequency. Any frequency calibrations should be checked often to guard against variations.

The operator and his methods have much to do with limiting the range of the station. The operator must have a good "fist," or good voice procedure. He must have patience and judgment. Some of these qualities in operating will make more station records than many kilowatts of power. Engineering or applied common sense is as essential to the radio operator as to the experimenter. Do not make several changes in the transmitter or receiver hoping for better results. Make one change at a time until the basic trouble or the best adjustment is found.

An operator with a slow, steady, clean-cut method of sending has a big advantage over the poor operator. Good sending is partly a matter of practice but patience and judgment are just as important qualities of an operator as a good "fist." The technique of speaking in connected thoughts and phrases is equally important for the operator who uses voice.

TOLERANCE AND COURTESY

None of us particularly enjoys working through interfering signals. As amateurs we have always had the "interference problem." It's nothing new. We accept it as a part of operation. We have eased the situation to a considerable extent by using VFOs and crystal filters in our receivers. That's part of the solution.

Except in emergencies, when an FCC declaration may require certain stations to remain silent or shift frequency (see Chapter Twenty-Two), each amateur ordinarily may operate on any frequency he chooses in any band, provided he abides by existing regulations. No amateur or any group of amateurs has any *exclusive* claim to any frequency in any band. We must work together, each respecting the rights of others. QRM is often unavoidable, and we occasionally find ourselves on the frequency used by another station. There are recommended ways in which conflicts of operating times and frequencies may be resolved. Network operators coördinate frequencies through ARRL headquarters registrations and individual amateurs and groups work out time sharing and frequency use to avoid difficulties in mutually agreed ways.

Each amateur should be tolerant of the other fellow's interests. He should exercise operating courtesy and common sense. Our normal operating interests in amateur radio vary considerably. Some prefer to rag-chew, others handle traffic, others work DX, others concentrate on working certain areas, countries or states, still others get on for an occasional contact only to check a new rig or antenna. "Why do they have to pick *my* frequency for a traffic net?" "What's the idea of chewing the rag on a traffic-net frequency?" "Why do these eggs have to use *my* frequency for their contest QSOs?" We have heard such expressions, and more. But QRM is one of the things we amateurs have to live with. Grin and bear it; it could be worse. Conduct your operating in a way designed to alleviate it as much as possible.

Before putting the transmitter on the air, listen on your own frequency. If you hear a trunk line or traffic net in operation on the frequency you intend to use, or if you hear any stations rag-chewing or conducting any form of communication on that frequency, stand by until you are sure no QRM will be caused by your operations, or shift to another frequency. Remember, those other chaps can cause you as much QRM as you cause them, sometimes more! The majority of amateurs own more than one crystal, many have VFO. It is not always necessary to stick to a single operating frequency.

Spend some time listening on all the frequencies you use for transmitting. You soon will learn what uses other amateurs are making of those spots. If you find, for example, that a net meets on one of the frequencies you use, or someone uses it for a regular schedule, you will soon learn the time that the net or schedule operates, and will be able to coöperate to avoid a conflict. It has become quite general operating procedure these days to work stations on or near your own frequency. This practice will automatically assist in reducing interference.

If we will each do our part to operate with tolerance and consideration, avoiding an attitude of running roughshod over other operators, we will do much to make ham operating more productive and enjoyable for ourselves as well as the other fellow!

● C.W. PROCEDURE

Official ARRL stations, *both* those using voice and c.w., observe the rules regarded as "standard practice" carefully. Any actively-operating c.w. stations will do well to copy these rules, and follow them when operating.

Calling Stations

1) *Calls.* The calling station should make the call by transmitting the call signal of the station called three times, the letters DE, followed by his own station call sent three times. Short calls with frequent "breaks" to listen have proved to be the best method. Repeating the call of the station called five times and signing not more than twice (repeating not more than three times) has proved excellent practice, thus: W6EY W6EY W6EY W6EY W6EY DE W1AW W1AW (etc.) $\overline{\text{AR}}$. Stations equipped with break-in are ideally suited for calling, for they can hear the station being called the instant he comes on the air, either indicating an answer to his call or obviating further calling.

CQ. The general-inquiry call (CQ) should be sent not more than five times without interspersing one's station identification. The length of repeated calls is carefully limited in intelligent amateur operating. CQ is not to be used when testing or when the sender is not expecting or looking for an answer. Never send a CQ "blind." Always listen on the frequency first.

The directional CQ: To reduce the number of useless answers and lessen QRM, every CQ call should be made informative when possible. Stations desiring communication should follow each CQ by an indication of direction, district, state, or the like.

Examples: A United States station looking for any Hawaiian amateur calls: CQ KH6 CQ KH6 CQ KH6 DE W4IA W4IA W4IA K. A Western station with traffic for the East Coast when looking for an intermediate relay station calls: CQ EAST CQ EAST CQ EAST DE W5IGW W5IGW W5IGW K. A station with messages for points in Massachusetts calls: CQ MASS CQ MASS CQ MASS DE W7CZY W7CZY W7CZY K. In each example indicated it is understood that the combination used is repeated three times.

Hams who do not raise stations readily may find that their sending is poor, their calls ill-timed or judgment in error. When conditions are right to bring in signals from the desired locality, the way to raise stations is to use the appropriate frequency and to call these stations. Reasonably short calls, with appropriate and brief breaks to listen, will raise stations with minimum time and trouble.

2) *Answering a Call:* Call three times (or less); send DE; sign three times (or less); after contact is established decrease the use of the call signals of both stations to *once or twice.* When a station receives a call without being certain that the call is intended for it, QRZ? may be used. It means "By whom am I being called?" QRZ should not be used in place of CQ.

3) *Ending Signals and Sign-Off:* The proper use of AR, K, $\overline{\text{KN}}$, $\overline{\text{SK}}$ and CL ending signals is as follows:

$\overline{\text{AR}}$ — End of transmission. Recommended after call to a specific station before contact has been established.

Example: W6ABC W6ABC W6ABC DE W9LMN W9LMN W9LMN $\overline{\text{AR}}$. Also at the end of transmission of a radiogram, immediately following the signature, preceding identification.

K — Go ahead (any station). Recommended after CQ and at the end of each transmission during QSO when there is no objection to others breaking in.

Example: CQ CQ CQ DE W1ABC W1ABC W1ABC K or W9XYZ DE W1ABC K.

$\overline{\text{KN}}$ — Go ahead (specific station), all others keep out. Recommended at the end of each transmission during a QSO, or after a call, when calls from other stations are not desired and will not be answered.

Example: W4FGH DE XU6GRL $\overline{\text{KN}}$.

$\overline{\text{SK}}$ — End of QSO. Recommended before signing *last* transmission at end of a QSO.

Example: . . . $\overline{\text{SK}}$ W8LMN DE W5BCD.

CL — I am closing station. Recommended when a station is going off the air, to indicate that it will not listen for any further calls.

Example: . . . $\overline{\text{SK}}$ W7HIJ DE W2JKL CL.

4) *Test signals* used to adjust a transmitter or at the request of another station to permit the latter to adjust receiving equipment usually consist of a series of Vs with the call signal of the transmitting station at frequent intervals. Remember that a test signal can be a totally unwarranted cause of QRM, and *always listen first,* find a clear spot if possible, or desist momentarily if it appears that your test signal will cause QRM.

5) *Receipting* for conversation or traffic:

Never send acknowledgment until the transmission has been entirely received. "R" means "All right, OK, I understand *completely*." Use R *only* when *all* is received correctly.

6) *Repeats.* When most of a transmission is lost, a call should be followed by correct abbreviations to ask for repeats. When a few words on the end of a transmission are lost, the last word received correctly is given after ?AA, meaning "all after" this should be repeated. When a few words on the beginning of a transmission are lost, ?AB for "all before" a stated word should be used. The quickest way to ask for a fill in the middle of a transmission is to send the last word received correctly, a question mark, then the next word received correctly. Another way is to send "?BN (word) ? (word)."

Do not send words twice (QSZ) unless it is requested. Send single. Do not fall into the bad habit of sending double *without a request* from fellows you work. Message-handling practices and procedure are discussed in another chapter.

General Practices

When a station has receiving trouble, the operator asks the transmitting station to "QSV." The letter "R" is often used in place of a decimal point (e.g., "3R5 Mc.") or the colon in time designation (e.g., "2R30 P.M."). A long dash for "zero" is in common use.

The law concerning superfluous signals should be noted carefully by every amateur. Do not hold the key down for long periods when testing. If you must test, disconnect the antenna system and use an equivalent "dummy" antenna. Send your call frequently when operating. Pick a time for adjusting the station apparatus when few stations will be bothered.

Long calls after communication has been established are unnecessary and inexcusable. The up-to-date amateur station uses "break-in." The best sending speed is a medium speed with the letters quickly formed and sent evenly with proper spacing. The standard-type telegraph key is best for all-round use. Before any freak keys are used, a few months should be spent listening-in and practicing. Regular daily practice periods, two or three half-hour periods a day, are best to acquire real familiarity and proficiency with code.

No excuse can be made for "garbled" copy. Operators should copy what is sent and refuse to acknowledge a whole transmission until every word has been received correctly. *Good operators never guess.* "Swing" in a fist is *not* the mark of a good operator, is undesirable. Unusual words are sent twice, the word repeated following the transmission of "?" If not *sure*, a good operator systematically asks for a fill or repeat.

Don't say, "QRM" or "QRN" when you mean "QRS."

Don't CQ unless there is definite reason for so doing. When sending CQ, use judgment. Sign your call frequently, interspersed with calls, and at the end of all transmissions.

On Good Sending

Although the primary facets of good sending, key adjustment, etc., have been enumerated in Chapter One, something need be said for the benefit of those operators who have followed the principles outlined therein and are embarking on their amateur careers. The more one operates, the more susceptible he is to pitfalls of unfortunate and ill-advised operating practices relating directly to his sending habits.

Every amateur should retain an acute consciousness of what his sending sounds like on the air. Complacency in operating is one of the leading causes of degeneration. Be your own worst critic, and cultivate the ability to shrug off the compliments of your friends until you are sure *in your own mind* that your sending deserves them.

Assuming that an operator has learned sending properly, and comes up with a precision "fist" — not fast, but clean, steady, making well-formed rhythmical characters and spacing beautiful to listen to — he then becomes subject to outside pressures to his own possible detriment in everyday operating. He will want to "speed it up" because the operator at the other end is going faster, and so he begins, unconsciously, to run his words together. Unless he keeps a close check on himself, he will begin to develop a "swing." This will become evident when friends recognize his "fist" even when he is operating at another station. In times of stress he may become nervous and start to "foul up" his characters, a fault which can easily become a habit if he is not careful.

Perhaps one of the easiest ways to get into bad habits is to do too much playing around with special keys. Too many operators spend only enough time with a straight key to acquire "passable" sending, then subject their newly-developed "fists" to the entirely different movements of bugs, side-swipers, electronic keys, or what-have-you. All too often, this results in the ruination of what may have become a very good "fist," and the operator finds that he can no longer send passably either with straight key or bug — or any other kind of key. Stick with the straight key a while. The results will justify it.

While it is true that nothing is so important in learning to send properly as correct learning methods and basic practices, there is considerable value in retaining a keen awareness of influences to which you become subject in your daily operating, and to become a good judge of which of these influences are good and which are bad. One must retain the lessons of the former and resist those of the latter. The good sending you hear on the air comes from operators who have not only learned correctly, but who have also continued to practise the principles of good sending.

Think about your sending a little. Are you satisfied with it? You should not be — ever. Nobody's sending is perfect, and therefore *every* operator should continually strive for improvement. Do you ever run words together — like Q for MA, or P for AN — especially when you are in a hurry? Practically everybody does at one time or another. Do you have a "swing"? Any recognizable "swing" is a deviation from perfection.

Check your spacing in characters, between characters and between words occasionally by making a recording of your fist on an inked tape recorder. This will show up your faults as nothing else will, and regardless of how good you thought your sending was, it will show that you *do* have faults. From there on, you can practise their correction.

Speed-Key Adjustment

Manual skill can be acquired only by practice, but no amount of practice will produce accurate sending if the key itself is improperly adjusted. The adjustment of a straight key is discussed in Chapter One and the requirement for proper adjustment is likewise important in using a "bug" or speed key. In using a bug every effort should be made to achieve *good control* rather than speed. In early practice the dot rate should be adjusted to not more than six per second.

Make sure that the movable and fixed dot contact points are parallel and have good contact over their entire surface. The pivot bearings should be adjusted so that no play can be felt when finger pressure is applied vertically to the shaft at its outer end.

For the preliminary adjustments, the weights should be at least halfway down the shaft. For a given speed, the exact position will vary considerably with the stiffness of the flat spring.

Back off the horizontal adjusting screw until the end of the shaft is resting against the damper weight. Apply pressure to the thumb paddle moving the shaft slowly toward the dot side, without allowing it to vibrate. If the adjustment is correct, the entire shaft will remain straight as it leaves the stop screw and the damper weight. The stop screw should be backed out as far as possible to allow good damping action by the damper weight without bending the flat spring when the thumb paddle is pressed slowly to the dot side.

Again press the shaft to the dot side without allowing it to vibrate. Vary the left-side stop screw until there is about 1/8 inch between the side of the shaft and the damper weight. The swing should not be much greater or less than this figure.

The dash adjustment is made by varying the contact screw until the operating paddle moves the same distance to the *left* of center to make a dash as it moves to the *right* to make dots.

The coil springs should have about the same tension. The spring should return the shaft quickly and positively from the dot side to the rest position against the damper. The dash spring is then carefully adjusted to a corresponding tension.

The length of a dot should be equal to the space between dots. If the paddle is held to the dot side, a series of at least 15 to 20 dots will be made with most "bugs" before there is any noticeable reduction in spacing between dots. If the dot adjustment is screwed in too far, a short series of heavy dots with very little separation will result. If the adjustment is screwed out too far, very light dots with excessive spacing will result. This adjustment is critical.

Many "bugs" are set to make excessively-fast dots. It will be found that most keys having a normal stiffness in the flat spring can be operated at a speed of about 30 w.p.m. with both weights toward the outer end of the shaft. Most operators cannot properly control a "bug" if the dot speed exceeds 11 per second. The rate at which your "bug" is adjusted can be determined by making a string of dots on recorder tape for 3 to 5 seconds, timing with a stop watch, and counting the dots.

The milliammeter method of adjustment involves connecting a battery (or any suitable source of d.c.), rheostat and milliammeter in series with the bug contacts. A typical set-up might use a 22½-volt battery, a 1000-ohm rheostat and a 0–100 milliammeter. With the key contacts closed, adjust the rheostat at the start with all the resistance in the circuit (to avoid burning out the meter!) until the meter reads 100 ma. A string of dots is then produced with the "bug" and the average-current reading on the meter is noted. If the dots are too light, the reading will be less than 50 ma.; if too heavy, more than 50 ma.

● USING A BREAK-IN SYSTEM

If you aim to have the best, and every ham does, you will have break-in, whether of the push-to-talk or open-the-key variety. If you haven't the ideal installation yet, by all means take every advantage of the other fellow's facilities when break-in is offered! Break-in avoids unnecessarily long calls, prevents QRM, gives you more communication per hour of operating. Brief calls with frequent short pauses for reply can approach (but not equal) break-in efficiency.

A separate receiving antenna makes it possible to listen to most stations while the transmitting tubes are lighted. It is only necessary with break-in to pause just a moment occasionally when the key is up (or to cut the carrier momentarily and pause in a 'phone conversation) to listen for the other station. The click when the carrier is cut off is as effective as the word "break."

C.w. telegraph break-in is usually simple to arrange. With break-in, ideas and messages to be transmitted can be pulled right through the holes in the QRM. Snappy, effective, efficient, enjoyable amateur work really requires but a simple switching arrangement in your station to cut off the power and switch 'phones from monitor to receiver. If trouble occurs the sending station can "stand by" (QRX), or it can take traffic until the reception conditions at the distant point are again good.

In calling, the transmitting operator sends the letters "BK" at frequent intervals during his call so that stations hearing the call may know that break-in is in use and take advantage of the fact. *He pauses at intervals* during his call, to listen for a moment for a reply from the station being called. If the station being called does not answer, the call can be continued. If the station called answers someone else, he will be heard and the calling can be broken off. With *full* break-in, the transmitter may be remotely-controlled so no receiver switching is necessary. A tap of the key, and the man on the receiving end can interrupt (if a word is missed) since the receiver is monitoring, awaiting just such directions constantly. But it is not necessary that *you* have such complete perfect facilities to take advantage of break-in when the stations you work are break-in-equipped. It is not intelligent handling of a station or coöperation with an operator advertising that he has "break-in" with his calls, to sit idly by minute after minute of a properly-sent call. After the first invitation to break is given and at each subsequent pause, turn on your transmitter and tap your key — and you will find that conversation or business can start immediately.

VOICE OPERATING

Voice work has become increasingly popular over the years. With the availability of new NFM equipment it will make additional progress. The use of proper procedure to get best results is just as important as in using code. In telegraphy words must be spelled out letter by letter. It is therefore but natural that abbreviations and short-cuts should have come into widespread use; they make it possible to convey intelligence faster. In voice work, however, abbreviations are not necessary, and should have no part in our operating procedure when using voice.

The letter "K" has been agreed to in telegraphic practice so that the operator will not have to pound out the separate letters that spell the words "go ahead." The voice operator can more readily and understandably say the words "go ahead" or "over," or "come in please."

One laughs on c.w. by spelling out the letters HI. Strangely enough, there are some voice operators who have never thought much about procedure who say HI or "Aitch eye" instead of actually laughing. Use a laugh when one is called for. Be natural as you would with your family and friends.

The matter of reporting *readability* and *strength* is as important to 'phone operators as to those using code. With telegraph nomenclature, it is necessary to spell out words to describe signals or use the abbreviated signal reporting system (RST . . . see Chapter Twenty-Four). Using voice, we have the ability to "say it with words." "Readability four, Strength eight" is the best way to give a quantitative report. Reporting can be done so much more meaningfully with ordinary words: "You are weak but you are in the clear and I can understand you, so go ahead," or "Your signal is strong but you are buried under local interference." Why not say it with words?

Voice Equivalents to Code Procedure

Voice	*Code*	*Meaning*
Go ahead; over	K	Self-explanatory.
Wait; stand by	$\overline{\text{AS}}$, QRX	Self-explanatory.
Okay	R	Receipt for a correctly-transcribed message or for "solid" transmission with no missing portions.
Repeat each word twice	QSZ	Self-explanatory.
All after	AA	Repeat everything after ..... (word).
I will repeat; I say again	$\overline{\text{IMI}}$	Repeating a difficult word, phrase or expression.

'Phone-Operating Practice

Efficient voice communication, like good c.w. communication, demands good operating. Adherence to certain points "on getting results" will go a long way toward improving our 'phone-band operating conditions.

Use push-to-talk technique. Where possible arrange on-off switches or controls for fast

Voice-Operating Hints

1) Listen before calling.

2) Make short calls with breaks to listen. Avoid long CQs; do not answer any.

3) Use push-to-talk. Give essential data concisely in first transmission.

4) Make reports honest. Use definitions of strength and readability for reference. Make your reports informative and useful. Honest reports and *full* word description of signals save amateur operators from FCC trouble.

5) Limit QSO length. Two minutes or less will convey much information. When three or more stations converse in round tables, brevity is essential.

6) Display sportsmanship and courtesy. Bands are congested . . . make transmissions meaningful . . . give others a break.

7) Check transmitter adjustment . . . avoid AM overmodulation and splatter. Do not radiate when moving VFO frequency or checking NFM swing. Use receiver b.f.o. to check stability of signal. Complete testing before busy hours!

back-and-forth exchanges that emulate the practicality of the wire telephone. This will help reduce the length of transmissions and keep brother amateurs from calling you a "monologuist" — a guy who likes to hear himself talk!

Listen with care. Keep noise and "backgrounds" out of your operating room to facilitate good listening. It is natural to answer the strongest signal, but take time to listen and give some consideration to the *best* signals, regardless of strength. Every amateur cannot run a kilowatt, but there is no reason why every amateur cannot have a signal of good quality, and utilize uniform operating practices to aid in the understandability and ease of his own communications.

Interpose your call regularly and at frequent intervals. Three short calls are better than one long one. In calling CQ, one's call should certainly appear at least once for every five or six CQs. Calls with frequent breaks to listen will save time and be most productive of results. In identifying, always transmit your *own* call *last*. *Don't* say "This is W1ABC standing by for W2DEF"; say "W2DEF, this is W1ABC, over." FCC regulations *require* that the call of the transmitting station be sent *last*.

Monitor your own frequency. This helps in timing calls and transmissions. Send when there is a chance of being copied successfully — not when you are merely "more QRM." Timing transmissions is an art to cultivate.

Keep modulation constant. By turning the gain "wide open" you are subjecting anyone listening to the diversion of whatever noises are present in or near your operating room, to say nothing of the possibility of feed-back, echo due to poor acoustics and modulation excesses due to sudden loud noises. Speak near the microphone, and don't let your gaze wander all over the station causing sharply-varying input to your speech amplifier; at the same time, keep far enough from the microphone so your signal is not modulated by your breathing. Change distance or gain only as necessary to insure uniform transmitter performance without overmodulation, splatter or distortion.

Make connected thoughts and phrases. Don't mix disconnected subjects. Ask questions consistently. Pause and get answers.

Have a pad of paper handy. It is convenient and desirable to jot down questions as they come in the course of discussion in order not to miss any. It will help you to make intelligent to-the-point replies.

Steer clear of inanities and soap-opera stuff. Our amateur radio and also our personal reputation as a serious communications worker depend on us.

Avoid repetition. Don't repeat back what the other fellow has just said. Too often we hear a conversation like this: "Okay on your new antenna there, okay on the trouble you're having with your receiver, okay on the company who just came in with some ice cream, okay . . . [etc.]." Just *say* you received everything OK. Don't try to prove it.

Use phonetics only as required. When clarifying genuinely doubtful expressions and in getting your call identified positively we suggest use of the ARRL Phonetic List. Limit such use to really-necessary clarification.

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. All voice-operated stations should use a *standard* list as needed to identify call signals or unfamiliar expressions.

ARRL Word List for Radiotelephony

ADAM	JOHN	SUSAN
BAKER	KING	THOMAS
CHARLIE	LEWIS	UNION
DAVID	MARY	VICTOR
EDWARD	NANCY	WILLIAM
FRANK	OTTO	X-RAY
GEORGE	PETER	YOUNG
HENRY	QUEEN	ZEBRA
IDA	ROBERT	

Example: W1AW . . . W 1 ADAM WILLIAM.

Round Tables. One of the most popular kinds of contact on the 'phone bands is the round table, in which several amateurs establish contact with each other and pass the conversation from one to another. The effect is somewhat the same as personal conversation in a group, with one important difference: the person talking cannot be interrupted. The round table has many advantages if run properly. It clears frequencies of interference, especially if all stations involved are on the same frequency, while the enjoyment value remains the same, if not greater. By use of push-to-talk, the conversation can be kept lively and interesting, giving each station operator ample opportunity to participate without waiting overlong for his turn.

Round tables can become very unpopular if they are not conducted properly. The monologuist, off on a long spiel about nothing in particular, cannot be interrupted: *make your transmissions short and to the point.* "Butting in" is discourteous and unsportsmanlike; *don't enter a round table, or any contact between two other amateurs, unless you are invited.* It is bad enough trying to understand voice through prevailing interference without the added difficulty of poor quality: *check your transmitter adjustments frequently.* In general, follow the precepts as hereinbefore outlined for the most enjoyment in round tables as well as any other form of radiotelephone communication.

WORKING DX

Most amateurs at one time or another make "working DX" a major aim. As in every other phase of amateur work, there are right and wrong ways to go about getting best results in working foreign stations, and it is the intention of this section to outline a few of them.

The ham who has trouble raising DX stations readily may find that poor transmitter efficiency is not the reason. He may find that his sending is poor, or his calls ill-timed, or his judgment in error. When conditions are right to bring in the DX, and the receiver sensitive enough to bring in several stations from the desired locality, the way to work DX is to use the appropriate frequency and timing and *call these stations*, as against the common practice of calling "CQ DX."

The call CQ DX means slightly different things to amateurs in different bands:

a) On v.h.f., CQ DX is a general call ordinarily used only when the band is open, under favorable "skip" conditions. For v.h.f. work such a call is used for looking for new states and countries, also for distances beyond the customary "line-of-sight" range on most v.h.f. bands.

b) CQ DX on our 7-, 14- and 28-Mc. bands may be taken to mean "General call to any foreign station." The term "foreign station" usually refers to any station in a foreign continent. (*Experienced* amateurs in the U. S. A. and Canada do *not* use this call, but *answer* such calls made by foreign stations.)

c) CQ DX used on 3.5 Mc. under winter-night conditions may be used in this same manner. At other times, under average 3.5-Mc. propagation conditions, the call may be used in domestic work when looking for new states or countries in one's own continent, usually applying to stations located over 1000 miles distant from your own.

The way to work DX is not to use a CQ call at *all* (in our continent). Instead, use your best tuning skill — and listen — and listen — and listen. *You have to hear them before you can work them.* Hear the desired stations first; time your calls well. Use your utmost skill. A sensitive receiver is often more important than the power input in working foreign stations. Before you can expect to be successful in working any particular foreign country or area, you should be able to hear ten or a dozen stations from that area.

One of the most effective ways to work DX is to know the operating habits of the DX stations sought. Doing too much transmitting on the DX bands is not the way to do this. Again, *listening* is effective. Once you know the operating habits of the DX station you are after you will know when and where to call, and when to remain silent waiting your chance.

Many DX stations use the signals HM, MH, LM and ML to indicate where they are tuning for replies. The meanings of these signals are as follows:

HM — Will start to listen at *high*-frequency end of band and tune toward *middle* of band.
MH — Will start to listen in the *middle* of the band and tune toward the *high*-frequency end.
LM — Will start to listen at *low*-frequency end of band and tune toward *middle* of band.
ML — Will start to listen in the *middle* of the band and

DX OPERATING CODE
(For W/VE Amateurs)

Some amateurs interested in DX work have caused considerable confusion and QRM in their efforts to work DX stations. The points below, if observed by all W/VE amateurs, will go a long way toward making DX more enjoyable for everybody.

1. Call DX only after he calls CQ, QRZ?, signs $\overline{\text{SK}}$, or 'phone equivalents thereof.
2. Do *not* call a DX station:
   a. On the frequency of the station he is working until you are *sure* the QSO is over. This is indicated by the ending signal $\overline{\text{SK}}$ on c.w. and any indication that the operator is listening, on 'phone.
   b. Because you hear someone else calling him.
   c. When he signs $\overline{\text{KN}}$, $\overline{\text{AR}}$, CL, or 'phone equivalents.
   d. Exactly on his frequency.
   e. After he calls a directional CQ, unless of course you are in the right direction or area.
3. Keep within frequency-band limits. Some DX stations operate outside. Perhaps they can get away with it, but you cannot.
4. Observe calling instructions of DX stations. "10U" means call ten kc. *up* from his frequency, "15D" means 15 kc. *down*, etc.
5. Give honest reports. Many foreign stations *depend* on W and VE reports for adjustment of station and equipment.
6. Keep your signal clean. Key clicks, chirps, hum or splatter give you a bad reputation and may get you a citation from FCC.
7. *Listen* for and *call* the station you want. Calling CQ DX is not the best assurance that the *rare* DX will reply.
8. When there are several W or VE stations waiting to work a DX station, avoid asking him to "listen for a friend." Let your friend take his chances with the rest. Also avoid engaging DX stations in rag-chews against their wishes.

DATE TIME	STATION CALLED	CALLED BY	HIS FREQ OR DIAL	HIS SIGNALS RST	MY SIGNALS RST	FREQ. MC.	EMISSION TYPE	POWER INPUT WATTS	TIME OF ENDING QSO	OTHER DATA
10-20-47										
6:15 PM	WØTQD	x	3.65	589x	569x	3.5	A-1	250	6:43	Lots of tfc! Rec'd 6, sent 10.
7:20	CQ	x				7				
7:21	x	W4TWI	7.24	369	579X				7:32	Too much QRM! Gave it up.
9:32	W3UA	x				3.95	A-3	100		Guess I was snowed under.
10-21-47										
7:05AM	VK4DV	x	14.03			14	A-1	250		Answered a W6
7:07	AC4YN	x	14.02							ND
7:09	VK2ADW	x	14.07	339	559X				7:20	Sydney, Australia First VK!!
7:31	CQ	x								No luck
7:42	W6RBQ	x	14.05	589	579				8:02	Had to QRT for breakfast. Nice chat.
8:02	— off									

KEEP AN ACCURATE AND COMPLETE STATION LOG AT ALL TIMES! F.C.C. REQUIRES IT.

A page from the official ARRL log is shown above, answering every Government requirement in respect to station records. Bound logs made up in accord with the above form can be obtained from Headquarters for a nominal sum or you can prepare your own, in which case we offer this form as a suggestion. The ARRL log has a special wire binding and lies perfectly flat on the table.

tune toward the *low*-frequency end.

Example: If the procedure will be to tune from the middle of the band to the high end, a CQ call goes: CQ DE G5BY MH K.

ARRL has recommended some operating procedures to DX stations aimed at controlling some of the thoughtless operating practices sometimes used by W/VE amateurs. A copy of these recommendations (Operating Aid No. 5) can be obtained free of charge from ARRL headquarters.

In any band, particularly at line-of-sight frequencies, when directional antennas are used, the directional CQ such as CQ W5, CQ north, etc., is the preferable type of call. Mature amateurs agree that CQ DX is a wishful rather than a practical type of call for most stations in the North Americas looking for contacts in foreign countries. Ordinarily, it is a cause of unnecessary QRM.

Conditions in the transmission medium make all field strengths from a given region more nearly equal at a distance, irrespective of power used. In general, the higher the frequency band, the less important power considerations become.

KEEPING AN AMATEUR STATION LOG

The FCC requires every amateur to keep a complete station operating record. It may also contain records of experimental tests and adjustment data. A stenographer's notebook can be ruled with vertical lines in any form to suit the user. The Federal Communications Commission requirements are that a log be maintained that shows (1) the date and time of *each* transmission, (2) *all calls* and transmissions made (whether two-way contacts resulted or not), (3) the input power to the last stage of the transmitter, (4) the frequency band used, (5) the time of *ending* each QSO and the operator's identifying signature for responsibility for each session of operating. Messages may be written in the log or separate records kept — but record must be made for one year as required by the FCC. For the convenience of amateur station operators ARRL stocks both logbooks and message blanks, and if one uses the official log he is sure to comply fully with the Government requirements if the precautions and suggestions included in the log are followed.

Message Handling

It is a common belief among non-traffic-handling amateurs that the amateur service is permitted, in normal times, to handle only unimportant "noncommercial" messages, and that the important and sometimes so-called "commercial" traffic handled during time of emergency is permitted to be handled by special dispensation of the Federal Communications Commission. Nothing could be farther from the truth. Amateurs may handle *any* domestic traffic, whether "pertaining to commerce" or not, at any time, provided only that they receive no compensation, direct or indirect, for such handling. Therefore it will be seen that the handling of third-party traffic by amateurs during periods of communications emergency is a normal extension of the third-party traffic handling carried on by amateurs from day to day.

Amateur traffic handling is highly developed and effective, *if one knows how to use it.* Don't expect that you can get on the air with the message you have written and give it to the first station that comes along and expect miracles to happen. You fellows who get your fun principally from DX, rag-chewing and building equipment should appreciate that you must place the occasional message *you* start and wish to have reach its destination, not in the hands of others like yourselves, but in the hands of one of the many operators who *specializes* in keeping schedules and handling messages, one who gets his fun mainly out of this branch of our hobby, who knows the best current routes and is in a position to use them.

Station owners may originate traffic *of any kind* going to any part of the United States, Hawaii, Puerto Rico and Alaska. Messages with radio amateurs in Canada, Chile and Peru may be handled under certain restrictions. Important traffic in emergencies or messages from expeditions for delivery in Canada must be put on a landwire by the U. S. amateur station handling. International regulations prohibit the handling of third-party messages to the majority of foreign countries. Messages relating to experiments and personal remarks of such unimportance that recourse to the public telegraph service would be out of the question may be handled freely with the amateurs of any country, but third-party messages only under special arrangements between U. S. A. and other governments, and only to the extent agreed upon by the contracting governments.

In handling messages, we are doing something really worth while. We want to start only good worth-while messages from our stations. Our efforts should be directed to making the quality of our message service high. The number of messages we handle is of secondary importance. The *kind of messages* we originate or start from our station, the *speed* with which the messages pass through our station, and the *reliability and accuracy* with which the messages are handled are the things of paramount importance.

Just as the ultimate aim of amateur radio on all frequency bands is *communication,* so is the relaying of word by radiogram a "natural" when one has something to say to a party beyond immediate reach. Not all hams perhaps appreciate the utility that results from using amateur message service in our ham correspondence. However, no ham, not even a new member of the brotherhood, but feels the satisfaction of having really accomplished something tangible in exchanging a message (recorded communication) with another amateur. Of course, not all beginners develop the advanced operating technique of the finished message handler, but it is within the reach of all who will try. In this chapter we shall discuss basic points to follow in message-handling activities.

The amateur who handles traffic is automatically training himself to do the kind of a job official agencies desire in emergencies, and he becomes a valuable exponent of the whole amateur service.

Messages should be put in complete form before transmitting them. *Incomplete messages should not be accepted.* As messages are often relayed through several stations before arriving at their destination, *no abbreviations should be used in the text* as mistakes are bound to happen when the text is shortened in this manner. To people not acquainted with radio abbreviations, messages written in shortened form are meaningless. Delivering stations must be careful to see that messages are written out fully.

Message Form

Each message originated and handled should contain the following component parts in the order given:

(a) Number
(b) Station of Origin
(c) Check
(d) Place of Origin
(e) Time Filed

(f) Date
(g) Address
(h) Text
(i) Signature

A standard form enables one to know just what is coming next, and makes accuracy possible with speed. Start some messages to familiarize yourself with the proper way to write and send traffic in good form. Just as you would be ashamed to admit it if you could not qualify as an experienced amateur by at least "15-w.p.m." code capability, be equally proud of your basic knowledge of how properly to form and send record communications.

a) Every message transmitted should bear a "number." On the first day of each calendar year, each transmitting station establishes a new series of numbers, beginning at Nr. 1. Keep a sheet with a consecutive list of numbers handy. File all messages without numbers. When you send the messages, assign numbers to them from the "number sheet," scratching off the numbers on that list as you do so, making a notation on the number sheet of the station to which the message was sent and the date. Such a system is convenient for reference to the number of messages originated each month.

b) The "station of origin" refers to the call of the station at which the message was filed. This should always be included so that a "service" message may be sent back to the originating station if something interferes with the prompt handling or delivery of a message. In the example in (d) below, W1INF is the station of origin.

c) Every word and numeral *in the text* of a message counts in the check. Full information on checking messages is given later in this chapter.

d) The "place of origin" refers to the name of the city from which the message was started. If a message is filed at League headquarters by someone in West Hartford, Conn., the preamble reads *Nr 457 W1INF ck 21 West Hartford Conn 8R57 p June 11, etc.*

If a message is sent to your station by mail the preamble shows the place of origin as the town from which the message came. If a message was filed at ARRL headquarters and if it came by mail from Wiscasset, Maine, the preamble would run like this to avoid confusion: *Hr msg nr 457 W1INF ck 21 WISCASSET Maine via West Hartford Conn 8R57 p June 11, etc.*

e) The time filed is the time at which the message is received at the station for transmission. "NFT" in a preamble means no filing time.

f) Every message shall bear a "date" and this date is transmitted by each station handling the message. The date is the "day filed" at the originating station unless otherwise specified by the sender.

g) The "address" refers to the name, street and number, city, state and telephone number of the party to whom the message is being sent. A *very complete address* should always be given to *insure* delivery. When accepting messages this point should be stressed. In transmitting the message the address is followed by a double dash or break sign (— . . . —) and it always precedes the text.

h) The "text" consists of the words in the body of the message. No abbreviations should ever be substituted for the words in the text of the message. The text follows the address and is set off from the signature by another break (— . . . —).

i) The "signature" is usually the name of the person sending the message. When no signature is given it is customary to include the words "no sig" at the end of the message to avoid confusion and misunderstanding. When there *is* a signature, it follows the break; *the abbreviation "sig" is not transmitted.*

The presence of unnecessary capital letters, periods, commas or other marks of punctuation may alter the meaning of a text. For this reason commercial communication companies use a shiftless typewriter (capitals only). The texts of messages are typed in block letters (all capitals) *devoid of punctuation*, underlining and paragraphing, *except where expressed in words.* In all communication work accuracy is of first importance. Spell out figures and punctuation.

Numbering Messages

Use of a "number sheet" or consecutive list of numbers enables any operator to tell quickly just what number is "next." Numbers may be crossed off as the messages are filed for origination. Another method of use consists of filing messages in complete form *except for the number.* Then the list of numbers is consulted and numbers assigned as each message is sent. As the operator you work acknowledges (QSLs) each message, cross off the number used and note the call of the station and the date opposite this number.

The original number supplied each message by the operator at the originating station is transmitted by each station handling the message. No new numbers are given the message by intermediate stations.

Checking Traffic — The Landline Check

The ARRL check is the landline or "text-only" count, consisting of the count of only the words in the body or text of the message. It is quicker and easier to count in this fashion than to use the cable count of words in address, text and signature check which is followed in marine-operating work, this simplification being the reason for its adoption. When in the case of a few exceptions to the basic rule in landline checking, certain words in 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.

Counting Words in Messages

The check includes: (1) all words, figures and letters in the body, and (2) the following extra words:

a) Signatures 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) Alternative names and/or street addresses, and such extras as "personal" or "attention ———————."

Examples: "Mother, Father, James and Henry" is a family signature, no names counted extra. "John Brown, Second Lieutenant" and "Richard Johnson, Secretary Albany Auto Club" are each one signature with no words counted as extra. An official title or connection is part of one signature, not extra. "Technical Department, Grammer *and Mix*" as a signature would count two extra words, those italicized after the first name counting as extras. The check of a message with ten words text and two such extras in the signature would be "CK 12 2 extra."

Dictionary words in most languages count as one word irrespective of length of the word. Figures, decimal points, fraction bars, etc., count as one word *each*. 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 in another.

Examples:

Emergency (English dictionary)	1 word
Nous arriverons dimanche (French dictionary)	3 words
DeWitt (surname)	1 word
E.L.B.D. (initials)	4 words
United States (country)	1 word
Prince William Sound	3 words
M.S. *City of Belgrade* (motor ship)	2 words
EXCEPTIONS	
A.M., P.M.	1 word
F.O.B. (or fob)	1 word
O.K.	1 word
Per cent (or percent)	1 word

Figures, punctuation marks, bars of division, decimal points—count each separately as one word. The best practice is to spell out all such when it is desired to send them in messages. In groups consisting of letters and figures *each* letter and figure will count as one word. Abbreviations of weights and measures in common use count as one word each.

Examples:

10 000 000 (figures)	8 words
Ten millions (dictionary words)	2 words
5348 (figures)	4 words
67.98 (figures)	5 words
64A2	4 words
45¼ (figures and bar of division)	5 words

Groups of letters which are not dictionary words of one of the languages enumerated, or combinations of such words, will count at the rate of five letters or fraction thereof to a word. In the case of combinations each dictionary word so combined will count as a word. In addition, USS, USCG, etc., written and sent as compact letter groups, count as one word.

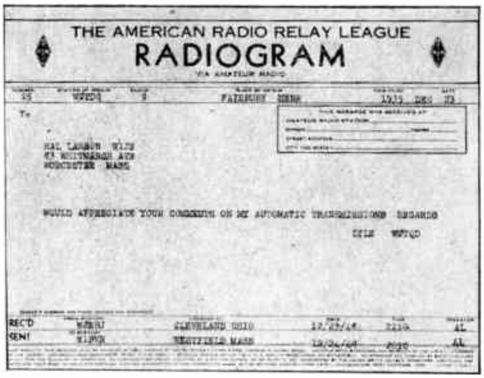
Examples:

Dothe (improperly combined)	2 words
allright, alright (improperly combined)	2 words
ARRL	1 word

At the request of sender the words "report back delivery," asking for a service showing success or failure in delivering at the terminal station, may be inserted after the check, or "rush" or "get answer" similarly, such words counting as extras in the group or check designation as just covered by example. "Phone" or "don't phone" or other sender's instructions in the address are not counted as extra words. In transmitting street addresses where the words east, west, north or south are part of the address, spell out the words in full. Suffixes "th," "nd," "st," etc., should not be transmitted. Example: Transmit "19 W 9th St" as "19 West 9 St." "F St NE" should be sent "F St Northeast."

Isolated characters each count as one word. Words joined by a hyphen count as separate words. Such words are sent as two words, without the hyphen. A hyphen or apostrophe each counts as one word; however, they are seldom transmitted. Two pair of quotation marks or two parentheses signs count as one word. Punctuation is *never* sent in radio messages except at the express command of the sender. *Even then it is spelled out.*

Here is an example of a plain-language message in correct ARRL form, carrying the landline check:



All messages should be checked carefully to insure accuracy. Request originators *to spell out all punctuation marks that must appear in delivered copies. Likewise, never abbreviate in texts,* nor use ham abbreviations except in conversations.

Originating Traffic

Messages to other amateurs are a natural means of exchanging comment and maintaining friendships. The simplest additional way to get messages is to offer to send a few for friends, reminding them that the message service is free and no one can be held responsible for delay or nondelivery. Wide-awake amateurs have distributed message blanks to tourist camps. Lots of good traffic has been collected through a system of message-collection boxes placed in public buildings and hospitals. A neatly-typed card should be displayed nearby explaining the workings of our ARRL traffic organization, and *listing the points to which the best possible service can be given.*

Messages that are not complete in every respect *should not be accepted* for relaying. *Complete address* on every message is very important.

To represent amateur radio properly, placards, when used, should avoid any possible confusion with telegraph and cable services. Any posters should refer to *amateur radiograms,* and explain that messages are sent through *amateur radio stations,* as a *hobby, free,* without cost (since amateurs can't and will not accept compensation). The exact conditions of the service should be stated or explained as completely as possible, including the fact that there is *no guarantee of delivery*. The individual in charge of the station has full powers to refuse any traffic unsuitable for radio transmission, or addressed to points where deliveries cannot be made. Relaying is subject to radio conditions and favorable opportunity for contacting. Better service can be expected on 15-word texts *of apparent importance* than on extremely long messages. Amateur radio traffic should *not* be accepted for "all over the world."

Careful planning and organized schedules are necessary if a *real* job of handling traffic is to be done. Advance schedules are essential to assist in the distribution of messages. It may be possible to schedule stations in cities to which you know quantities of messages will be filed. Distribute messages, in the proper directions, widely enough so that a few outside stations do not become seriously overburdened. Operators must route traffic properly — not merely aim to "clear the hook."

It is better to handle a small or moderate volume of traffic *well* than to attempt to break records in a manner that results in delayed messages, nondeliveries, and the like which certainly cannot help in creating any public good will for amateur radio.

Relay Procedure

Messages shall be relayed to the station nearest the location of the addressee and over the greatest distance permitting reliable communication.

No abbreviations shall be substituted for the words in the text of a message with the exception of "service messages," to be explained. Delivering stations must be careful that no confusing abbreviations are written into delivered messages.

Sending words twice is a practice to avoid. Use it only when expressly called for by the receiving operator, when receiving conditions are poor.

Messages shall be transmitted as many as three times at the request of the receiving operator. Failing to make a complete copy after three attempts, the receiving operator shall cancel the message (QTA).

Agreement to handle (relay or deliver) a message properly and promptly is always tacitly implied in accepting traffic. When temporarily *not* in a position to handle so, it is a service to amateur radio and your fellow ham to *refuse* (courteously) a message.

An operator with California traffic does not hear any Western stations so he decides to call a directional "CQ" as per ARRL practice. He calls, *CQ CALIF CQ CALIF DE W1INF W1INF W1INF*, repeating the combination three times.

He listens and hears W9BRD in Chicago calling him, *W1INF W1INF W1INF DE W9BRD W9BRD W9BRD AR.*

Then he answers W9BRD, indicating that he wishes him to take the message. W1INF says, *W9BRD W9BRD DE W1INF R QSP MILL VALLEY CALIF NEAR SF? K.*

After W9BRD has given him the signal to go ahead, the message is transmitted, thus:

HR MSG NR 78 W1INF CK15 WEST HARTFORD CONN NFT [for "no filing time"] NOV 18

ALAN D WHITTAKER JR W6SG
79 ELINOR AVE
MILL VALLEY CALIF

SUGGEST YOU USE ARRL TRUNK LINE K THROUGH W5NW TO HANDLE PROPOSED VOLUME TRAFFIC REGARDS

BUBB W1JTD

W9BRD acknowledges the message like this: *W1INF DE W9BRD NR 78 R K.* Not a single *R* should be sent unless the whole message has been correctly received.

Full handling data are placed on the message for permanent record at W1INF. The operator at W9BRD has now taken full responsibility for doing his best in forwarding the message.

Abbreviated Procedure

Abbreviated procedure deserves a word in the interest of brevity on the air. Abbreviated practices help to cut down unnecessary transmission. However, make it a rule not to abbreviate unnecessarily when working an operator of unknown experience.

NIL is shorter than *QRU CU NEXT SKED.* Instead of using the completely spelled-out preamble *HR MSG NR 287 W1ORP CK 18 PUTNAM CONN OCTOBER 28 TO,* etc., transmission can be saved by using *287 W1ORP 18 PUTNAM CT OCT 28 TO,* etc. One more thing that conserves operating time is the cultivation of the operating practice of writing down "W1JE 615P 1/13/49" with the free hand during the sending of the next message.

"Handling" a message always includes the transmission and receipt of radio acknowledgment (QSL) of same, and entry of date, time and station call *on the traffic*, as handled, for purposes of record.

Getting Fills

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 AL* or *RPT MSG* should not be sent unless nearly all of the message is lost.

Each abbreviation used after a question mark (. . — — . .) asks for a repetition of that particular part of a message.

When a few word groups in conversation or message handling have been missed, a selection of one or more of the following abbreviations will enable you to ask for a repeat on the parts in doubt. 'Phone stations, of course, request fills by using the full wording specified, without attempt at abbreviation.

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 for only 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. *RPT FROM . . . . TO . . . .* is a long way of asking for message fills.

The figure four (. . . . —) is a time-saving abbreviation which deserves popularity with traffic men. It is another of those hybrid abbreviations whose original meaning, "Please start me, where?" has come to us from Morse practice.

Delivering Messages

When it is possible to deliver messages in person, that is usually the most effective way. When the telephone does not prove instrumental in locating the party addressed in the message, it is usually quickest to mail the message.

ARRL delivery rules:

Messages received by stations shall be delivered immediately.

Every domestic message shall be relayed within forty-eight (48) hours after receipt, or if it cannot be relayed within this time shall be mailed to the addressee.

Messages for points outside North America must not be held longer than half the length of time required for them to reach their destination by mail.

When a message cannot be delivered, or if it is unduly delayed, a "service" message should be written and started back to the "office of origin."

Each operator who reads these pages is asked to assume *personal responsibility* for accuracy, speed of each message handled, and *delivery*, that we may approach a 100% delivery figure.

Fixed-Text Messages—"ARL" Check

ARL? means, "Do you have the list of ARRL-Numbered Radiograms, and are you ready for such a message," ARL (reply) then means, "I have the ARRL-Numbered Radiogram list. I am ready for such a message." A list of the texts applicable to possible relief-emergency uses follows:

ONE	All safe. Do not be concerned about disaster reports.
TWO	Coming home as soon as possible.
THREE	Am perfectly all right. Don't worry.
FOUR	Everyone safe here. Only slight property damage.
FIVE	All well here. Love to folks.
SIX	Everyone safe, writing soon.
*SEVEN	Reply by amateur radio.
EIGHT	All safe, writing soon, love.
NINE	Come home at once.
TEN	Will be home as soon as conditions permit.
ELEVEN	Cannot get home. Am perfectly all right. Will be home as soon as conditions permit.
*TWELVE	Are you safe? Anxious to hear from you.
*THIRTEEN	Is ...... safe? Anxious to hear.
*FOURTEEN	Anxious to know if everything is OK. Please advise.
*FIFTEEN	Advise at once if you need help.
*SIXTEEN	Please advise your condition.
*SEVENTEEN	Kindly get in touch with us.
*EIGHTEEN	Please contact me as soon as possible (at ....).

*Not to be solicited in emergency.

The list of fixed texts was prepared mainly with possible emergency needs and utility in mind; it is a special tool for special occasions. It may be used only when stations at each end of a QSO are equipped with identical lists. Never forget to put "ARL" in the check — or the delivering station might deliver a "number" instead of the words it stands for.

Example: NR1 W1AW CK ARL 1 Newington Conn March 2 [Address] BT THREE BT John AR.

A list of additional texts, including holiday-greeting types, will be sent free of charge to anyone requesting it. It is included also with each ARRL logbook purchased.

The Service Message

A service message is a message sent by one station to another station, relating to the service which we are or are not able to give in message handling. The service message may refer to nondeliveries, to delayed transmission, to errors, or to any phase of message-handling activity. It is not proper to abbreviate words in the texts of regular messages, but it is quite desirable and correct to use abbreviations in these station-to-station messages relating to traffic-handling work. Example:

HR SVC NR 291 W4IA CK XX ARLINGTON VA NFT AUG 19
L C MAYBEE W7GE
110 SOUTH SEVENTH AVE
PASCO WASHN —...—
UR NR 87 AUG 17 TO CUSHING SIG BOB HELD HR UNDLD PSE GBA —...—
BATTEY W4IA

Secrecy of Communications

Provisions of the Communications Act make it a misdemeanor to give out information of any sort to any person except the addressee of a message. It is in no manner unethical to deliver an unofficial copy of a radiogram, if you carefully mark it *duplicate* or *unofficial copy* and do it to improve the speed of handling a message or to insure certain and prompt delivery. Do not forget that there are heavy fines prescribed by Federal laws for divulging the contents of messages to anyone *except* the person addressed in a message.

Counting Messages

To provide a method of comparison between stations handling traffic, the ARRL counting system is presented below. Each time a message is *handled by radio* it counts one in the traffic total.

A message received in person, by telephone, by telegraph, or by mail, *filed at the station and transmitted by radio* in the proper form, counts as *one originated.*

A message *received by radio and delivered in person,* by telephone, by telegraph, or by mail, counts as *one delivered.*

A message *received by radio and transmitted by radio* counts as two messages *relayed* (one when received and one when transmitted).

All messages counted under the above classes must be handled within a 48-hour delay period in order to count as "messages handled." A "service" message counts the same as any other type of message.

In addition to the basic count of *one* for each time a message is handled by radio, an *extra credit* of one point for each delivery made by mail, telephone, in person, by messenger or other external means *other than use of radio* (which would count as a "relay" of course) will also be allowed. A message received by an operator for himself or his station or a party on the immediate premises counts *only* "one delivered." A message for a third party delivered by *additional means or effort* will receive a point under "extra delivery credits."

The message total shall be the *sum* of the messages originated, delivered and relayed, and the "extra" delivery credits. Each station's message file and log shall be used to determine the report submitted by that particular station. Messages with identical texts (so-called rubber-stamp messages) shall count once only for *each* time the complete text, preamble and signature are sent by radio.

In whatever volunteer work it is engaged, a station has an amateur status, and the total is a strictly "amateur" total if handled on amateur frequencies.

Examples of Counting

Let us assume that at the end of the month one operator of a large amateur station receives several messages from another station. (a) Some of these messages are for relaying by radio. (b) Some of them are for local delivery. (c) There are still other messages, the disposal of which cannot be accurately predicted. They are for the immediate neighborhood but either can be mailed or forwarded to another amateur by radio. A short-haul telephone toll call will deliver them but the chances of landing them nearer the destination by radio are pretty good. This operator's "trick" ends at midnight. He must make the report with some messages "on the hook," to be carried over for the next month's report.

a) The messages on the hook that are to be relayed have been received and are to be sent. They count as "1 relayed" in the report that is made out now, and they will also count as "1 relayed" in the next month's report (the month during which they were forwarded by radio).

b) By mailing or 'phoning the messages at once, they count as "1 delivered" for the current report. By holding them until next day they will count in the *next* report as "1 delivered." Also, they will each have a count of one *extra* delivery credit since they had to be telephoned, mailed, etc.

c) The messages in this class may be carried forward into the next month. If they *have* to be mailed then they will count in the *next* report as "1 delivered." If they are relayed, we count them as "1 relayed," "1 received" in the preceding month (already reported), and "1 relayed" for the next month, the month in which it was sent forward by radio. If the operator wishes to count this message *at once* as delivered it must be mailed promptly and counted at once.

Some examples of counting:

The operator of Station A gets a message by radio from Station B addressed to himself. This counts as "1 delivered" by himself and by Station A. There is no extra delivery credit possible for no additional delivery effort was needed.

The operator of Station A takes a verbal message from a friend for relaying. He gives it to Station B over the telephone. Operator A does not handle the message by radio. Station B and Operator B count the message as "1 originated." A cannot count the message in *any* manner.

The operator and owner of Station A visits Station B and *while operating* there takes a message for relaying. The operator and owner of B cannot operate for a day or two so the message is carried back to Station A by Operator A who relays it along within a few hours. The traffic report of both Station A and Station B shows "1 relayed" for this work.

Please note that "handling" a message always includes the transmission and receipt of radio acknowledgment (QSL) of same, and the entry of date, time, and station call on the traffic, as handled, for purposes of record. Only messages promptly handled and with information so recorded shall be counted in ARRL totals.

Reporting

Whether the principal accomplishments of the station are in traffic handling or other lines, *what you are doing* is always of interest. One part of *QST* is devoted to Station Activities, this written up by your elected section communications manager. His address is given on page 6 of each *QST*. Reports from *all active hams,* sent the SCM at the end of each month and covering the 30 days just previous, are welcomed.

Amateur Stations at Exhibits and Fairs

Whatever type of exhibit is planned, write ARRL in advance, in order to receive sample material to make your amateur booth more complete. A *portable* station can be installed and operated by an already-licensed amateur, subject to FCC notification of location, etc., as provided by regulations. No license coverage is needed if no station is operated, of course.

If the time is short and there is no opportunity for special organization of schedules to insure reliable routing and delivery, quite likely exhibit work, to be most productive of good-will results, had best *not* include message-handling plans — at least not from the booth-station itself where subject to noise, electrical interference and other handicaps. To handle such traffic as offered with *real efficiency*, it should be distributed for origination with existing schedules of the *several most reliable local amateur stations. By dividing the traffic filed with other stations it may be sent more speedily on its way.*

"Show stations" must avoid origination of "poor traffic" by rigid supervision and elimination of meaningless messages with guessed-at, inaccurate and incomplete addresses *at the source.*

Foreign Traffic Restrictions

Any and all kinds of traffic may be handled between amateur stations in different parts of the United States, Hawaii, Alaska and Puerto Rico. There is no qualification or restriction except that amateur status must be observed and no compensation, direct or indirect, be accepted for station operations or services.

Internationally the general regulations attached to the international communications treaty state the limitations to which work between amateur stations in different foreign countries is subject. In practically every nation outside our own country and its possessions, the government owns or controls the public communications systems. Since these systems are maintained as a state monopoly, foreign amateurs have been prohibited by their governments from exchanging traffic which might be regarded as in "competition" with state-owned telegraphs. The international treaty regulations reflect this condition and the domestic traffic restrictions (internal policy) of the majority of foreign countries. Any country ratifying the Madrid (1932), Cairo (1938) and Atlantic City (1947) conventions can make its domestic regulations as liberal as it likes; in addition it may conclude special agreements with other governments for amateur communications that are *more liberal* than the quoted terms of the treaty itself. If no specific formal negotiations have been concluded, however, amateurs must observe the following (treaty) regulations in conducting international amateur work:

§ 1. Radiocommunications between amateur stations of different countries shall be forbidden if the administration of one of the countries concerned has notified that it objects to such radiocommunications.

§ 2. (1) When transmissions between amateur stations of different countries are permitted they must be made in plain language and must be limited to messages of a technical nature relating to tests and to remarks of a personal character for which, by reason of their unimportance, recourse to the public telecommunications service is not justified. It is absolutely forbidden for amateur stations to be used for transmitting international communications on behalf of third parties.

(2) The preceding provisions may be modified by special arrangements between the countries concerned.

Referring to the first paragraph above, in the years since the Washington convention (1927) *no* prohibition on amateur communication (international QSOs) has been filed by *any* country with the Berne Bureau. In some countries, principally European, amateurs are restricted by regulation to privileges much *less* than made available by international agreement. In the U. S. A. it is the policy, and of course necessary to take care of our greater number of amateurs, to give amateurs the fullest frequency allocations and rights possible under international treaty provisions, and to permit free exchange of domestic noncommercial traffic in addition. This policy has justified itself, giving the public amateur radio traffic service and developing highly-skilled operators and technicians who have the ability to keep the U. S. A. in the lead in radio matters.

The second paragraph quoted prohibits *international* handling of third-party traffic, except where two governments have a special arrangement for such exchange. In any event, traffic relating to experimental work, and personal remarks which would not be sent by commercial communications channels, may be sent, when in communication with foreign amateurs.

Previous special arrangements, extending the basic international telecommunications treaty arrangements, have also been effected through ARRL and U. S. A. representations. The special U. S. A.-Canadian agreement will be explained later. Similar arrangements with Chile and Peru permit the handling to those countries of certain types of traffic.

The Canadian Agreement

The special reciprocal agreement concluded between our country and the Dominion of Canada, at the behest of the ARRL, permits Canadian and U. S. amateurs to exchange messages of importance under certain restrictions. This agreement is an expansion of the international regulations to permit the handling of important traffic.

The authorized traffic is described as follows:

"1. Messages that would not normally be sent by any existing means of electrical communication and on which no tolls must be charged.

"2. Messages from other radio stations in isolated points not connected by any regular means of electrical communications; such messages to be handed to the local office of the telegraph company by the amateur receiving station for transmission to final destination, e.g., messages from expeditions in remote points such as the Arctic, etc.

"3. Messages handled by amateur stations in cases of emergency, e.g., floods, etc., where the regular electrical communication systems become interrupted; such messages to be handed to the nearest point on the established commercial telegraph system remaining in operation."

The arrangement applies to the United States and its territories and possessions including Alaska, the Hawaiian Islands, Puerto Rico, the Virgin Islands and the Panama Canal Zone. The agreements with Chile and Peru are similar to the above.

General

Message handling is one of the major activities that lies in our power as amateurs to do to show our amateur radio in a respected light, rather than from a novelty standpoint. Regardless of experimental, QSL-collecting, friendly rag-chew and DX objectives, we doubt if the amateur exists who does not want to know how to phrase a message, how to put the preamble in order, how to communicate wisely and well when called upon to do so. Scarcely a month passes but what some of us in some section of our ARRL are called upon to add to the service record of the amateur.

It is important that deliveries be made in businesslike fashion to give the best impression, so that in each instance a new friend and booster for amateur radio may be won. Messages should be typed or neatly copied, preferably on a standard blank, retaining original for the FCC station file. The designation and address of the delivering station should be plainly given so a reply can be made by the same route, if desired.

For those who would disparage some message texts as unimportant, perhaps a reminder is in order that in the last analysis it is not the importance to the ham that handles it that counts, but the importance to the party that sends and the party that receives a message.

The individual handling of traffic in small quantities as well as large is to a very great extent the material that we amateurs use for developing our operating ability, for organizing our relay lines, and for making ourselves such a very valuable asset to the public and our country in every communications emergency that comes along.

For those "breaking-in" may we say that any ORS, trunk-liner or experienced ARRL traffic handler will be only too glad to answer your questions and give additional pointers regarding procedure and your station set-up, to help you make your station a really effective communications set-up. Since experience is the only real teacher, newcomers are reminded that becoming highly proficient in any branch of the game is partly just a matter of practice. Start a few messages, to get accustomed to the form. Check some messages to become familiar with the official ARRL (landline) check. You will find increased enjoyment in this side of amateur radio by adding to your ability to perform; by your familiarity with these things the chance of being able to serve your community or country in emergency will be greater. Credit will be reflected on amateur radio as a whole thereby.

Traffic Handling Develops Skill

The dispatch of messages makes operators keen and alert. The better the individual operator, the better the whole organization. Proper form in handling traffic, in getting fills, and in general operating procedure develops operators who excel in "getting results." Station performance depends 90% on operating ability and 10% on the equipment involved, granting of course that station and operator are always interdependent. Experience in message handling develops a high degree of operating "intelligence."

Message handling leads to organization naturally, through the need for schedules and coöperation between operators. It offers systematic training in "real" operating. It leads to planned, useful, unselfish, constructive work for others at the same time it represents the highest form of operating "skill" and enjoyment to its devotees. Emphasis should be placed on the importance of traffic handling in training operators in the use of procedure — and in general operating reliability. The value of the amateur (as a group), in cases of local or national emergency, depends to a great extent on the *operating ability* of individual operators. This ability is largely developed by message handling.

Practice in handling traffic familiarizes one with detailed time-saving procedure, and develops general skill and accuracy to a higher extent than obtains from "just rag-chewing" or haphazard work.

Emergency Operation

The radio amateur best justifies his existence by the service he renders his community in times of disaster and distress when all other media of summoning assistance have failed. The pleasure he derives from the pursuit of his hobby during normal times establishes a debit that he can offset only by his steadfast determination to be adequately prepared and willing to be of service when disaster strikes his community.

In the event of a communications emergency all amateurs are dedicated to serve in the public interest, within the limitations of their ability, to provide temporary communications for a stricken area until normal facilities are restored.

Amateur Emergency Communication

When customary communications circuits are interrupted or overloaded a communications emergency is said to exist. If such a condition of interruption or overload is accompanied by general suffering on the part of the inhabitants of the affected region, the pressure for the restoration of communications is increased proportionally to the degree of suffering. Under such circumstances the amateur service is frequently called upon to carry a portion of the traffic load until such time as normal facilities are restored.

COÖPERATION WITH THE AMERICAN NATIONAL RED CROSS

An official "understanding" between the Red Cross and the amateur service, in effect for many years, has been reviewed and reaffirmed. This appears in the *Disaster Preparedness and Relief Manual* (ARC 209) published by the Red Cross. This manual, for the guidance of the Red Cross chapters, assigns the function of providing and maintaining communication services to a subcommittee on transportation and communication. In preparing to meet the needs for communications incident to a disaster relief operation, the subcommittee is charged with surveying all communications resources within a chapter jurisdiction to obtain coöperation and to plan necessary coördination and mobilization of appropriate facilities in any emergency situation.

Here are pertinent extracts of the ARC 209 information:

. . . Test drills should be held from time to time in coöperation with the Emergency Corps of the American Radio Relay League. . . .

The American Radio Relay League and the American Red Cross have developed an understanding to obtain the maximum coöperation in emergency communications in time of disaster. The American Radio Relay League recognizes the American Red Cross as the primary disaster relief agency. The American Red Cross recognizes that the American Radio Relay League through its emergency coördinators or other designated representatives of the American Radio Relay League Emergency Corps can render valuable aid in handling emergency communications when other facilities have been disrupted. The American Red Cross welcomes the coöperation of the American Radio Relay League with local Red Cross chapters to assure the best possible communications by such facilities. This coöperation with the American Red Cross chapters may be furthered by individual chapter's designation of the American Radio Relay League Emergency Coördinator or other designated representative of the American Radio Relay League Emergency Corps to serve as a member of the American Red Cross chapter disaster subcommittee on transportation and communication.

Amateurs may serve many agencies but it is to be noted that ARRL recognizes the Red Cross as the primary disaster relief agency, and it is therefore entitled to our best possible arrangements. The Communications Department policy is to create and extend effective emergency radio coverage to every possible community and Red Cross chapter jurisdiction, where advance preparedness may pay dividends in the form of service. All amateurs are urged to study the section of this manual describing the organization of the ARRL Emergency Corps in order to fit themselves best for this important service.

The ARRL Emergency Corps

The ARRL Emergency Corps is composed of licensed amateurs who have voluntarily registered their qualifications and equipment for communication duty in the public service when disaster strikes.

Every licensed amateur, whether or not a member of ARRL, is eligible for membership in the Emergency Corps. The only other qualification is a sincere desire to serve. There are two grades of membership in the Corps: (a) full membership, under which the applicant pledges active participation in periodic tests, and (b) supporting membership, requiring only limited participation as time permits. The possession of emergency-powered equipment is desirable, but is not a requirement for either grade.

Emergency Corps activities in each ARRL section are under the direction of the Section Emergency Coördinator, appointed by the Section Communications Manager as his representative in such matters. The amateurs in each community within the section register their facilities with the local emergency coordinator, who, in turn, represents the amateur service in its dealings with local civic and relief agencies. The following paragraphs detail the duties and obligations of these officials.

THE SECTION EMERGENCY COÖRDINATOR

The Section Emergency Coördinator (SEC) is appointed by the SCM to take charge of the promotion of the ARRL Emergency Corps organization throughout the section. He reports progress and plans to the SCM, and acts as his executive in furthering provisions for emergency amateur radio communications in every community likely to suffer in case of a natural disaster or other emergency. He recommends ECs to the SCM, and determines the jurisdictional areas of ECs as required.

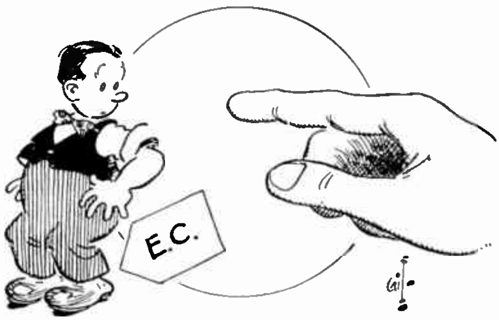
The SEC's functions are:

1. The coördination and implementation of a section program for the promotion of organization for emergency amateur radio work for all section communities.
2. Policy and planning recommendations to the section communications manager concerning all emergency amateur radio activities.
3. Recommendations for EC appointments and endorsements for community emergency coördinators, throughout the section.
4. The determination of the areas of jurisdiction of the several community emergency coördinators.
5. The promulgation of Emergency Corps membership drives, meetings, activities, tests, procedures, etc. at section levels. (In this the SEC will work through local ECs. ECs sign the local Emergency Corps members' cards; SECs may sign those not in an area normally covered by a local EC, while trying to secure an EC appointee for the area.)
6. The consolidation of the monthly reports obtained from the community coördinators, and the submission of monthly progress summaries, to his SCM and to ARRL headquarters, covering section planning, drills, and the current status of the AEC.
7. The supervision, at section levels, of community emergency-radio provisions supplied by the amateur service.

Inasmuch as his responsibilities require him to act as the assistant SCM for emergency organization matters pertinent to the amateur service, the SEC post is one of top importance. In view of this the section emergency coordinator must be an appointee who will devote his full energy and effort to this one important emergency organizing program for amateur radio. The purpose of this post is to provide full-time thought and backing for the ARRL Emergency Corps organization.

THE EMERGENCY COÖRDINATOR

In every community where qualified amateurs can be found the appointment of an emergency coördinator will be made by the SCM. For specific agencies a regional emergency coördinator to undertake organization covering a watershed, railway line or other area may be appointed.



DOES YOUR COMMUNITY HAVE ONE ?

Coördinators designated for a community should head local amateur committees which they organize and on which each amateur frequency-band-mode group is represented. The duties of the emergency coördinator include:

(1) Organizing meetings of all available amateurs and emergency workers.

(2) The designation of assistants for an emergency planning committee.

(3) The calling of committee meetings and the making of appointments to handle particular responsibilities.

(4) The initiation of code-class programs for those working for new amateur licenses.

(5) Liaison and general planning for assumed community emergency contingencies.

(6) The establishment of regular drill periods, and simulated emergency tests of equipment and operators.

(7) Monthly assessment of progress and reports for ARRL-*QST*.

In carrying out the above seven-point program, the EC issues AEC membership cards to each amateur licensee joining in his area, indicating "full" or "supporting" class of membership. His work includes the registration of local emergency equipment and operators to include the locations of stations, equipment, operating frequencies, emergency power sources; the normal availability, experience, telegraph operating speeds, occupation, address, working hours, and the 'phone numbers of operators. He furthers the adoption of a preparedness program for local amateurs designed to promote highly-skilled operating ability and the building of emergency equipment.

An important phase of his work includes contacting agencies to be served (such as Red Cross, national, state, civic and military authorities, utilities, meteorological agency, etc.), to make available *to them* complete knowledge of the extent of amateur facilities, with the addresses and the telephone availability of himself and his alternate committee members, and to determine *from them* their probable traffic load and estimated emergency needs, and the important outside points to or from which they must receive information. Liaison must be maintained with many other services which the amateur service must assist, or supplement (such as broadcasting, police, aviation, military, etc.).

In emergency the EC should advise the most effective disposition of operators and stations, to promote efficiency by:

(a) Designating main stations for key points.
(b) Manning stations suitable for 16- or 24-hour needs (8-hour shifts recommended).
(c) Reducing the interference levels by creating planned operator reserves. (Keeping a reserve of operators fresh for needs, asking voluntary coöperation in staying off the air to render maximum service while using a *minimum* number of nets and *stations* and causing minimum interference.)

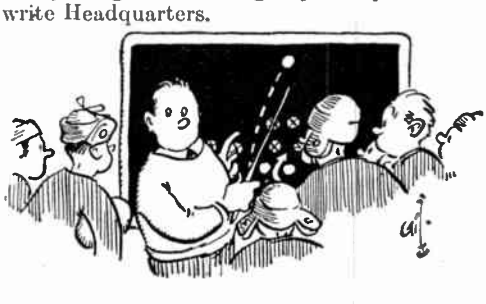
He utilizes frequencies and modes which are suited to make best use of *all frequencies*, and to minimize interference. The analysis of station frequencies and equipment, from registrations, has an important bearing on these duties. Plans must be predicated on local situations and equipment: (a) Local v.h.f. links. (b) Low-frequency nets and schedules for point-to-point work at greater distances. (c) Selecting the most skilled operators for the circuits having heavier loads. (d) Designating telegraph channels for accurate record communications and properly to load telegraph bands, without causing undue radio telephone-band congestion by overloading. (e) Designating telephone channels where speed for discussions is helpful and secrecy provisions, by reason of trained and properly cautioned operators — are believed adequate for a particular service use.

In the performance of his function, coördination and coöperation between different amateur service groups must be arranged and maintained. Friendly rivalry in competitive tests may be encouraged. In actual emergency (and plans for emergency) responsible amateur leaders in any and all groups must put aside petty rivalry in favor of working together closely. Facilities should be pooled for best results while minimizing interference. If the single "best" net cannot be utilized, we suggest careful assignment or division of nets between (a) different agencies served or (b) between certain groups of distant cities or points covered for a specific purpose.

Advance Planning Is Necessary

Individual preparedness and organization of amateurs, community by community, are the most essential elements of a successfully-functioning amateur emergency service. Annual ARRL Field Days supply a tremendous incentive to clubs and individuals to perfect and test emergency-powered equipment. Training in message handling and the principles of working together under unusual circumstances combines in this outing an annual "shakedown" of communications gear afield, to challenge individual and group efforts. But unless each individual in addition to working casually in annual events attaches himself to community organization and the support of local plans, our group effort may sometimes fall short of its organized capabilities of performing a public service.

The chief lesson, from past emergencies, calls for support and membership of every active operator in the ARRL Emergency Corps. By joining, one is supporting organized amateur radio for any and every emergency! The AEC member can enjoy and benefit from *advance* discussions, literature, and participation in exercises covering such matters as procedure, priorities, station dispositions, and other factors. By such alignment with AEC, one justifies his license as "in the public interest." Such support contributes to our emergency-readiness. Emergency-readiness contributions to public welfare as well as our self-training and experimenting characteristics permit FCC to grant amateur licenses as of public value. Once in the AEC you are on the "inside" in any amateur service plans worked out by your emergency coördinator. Your local ARRL coördinator can provide appropriate blanks for joining the Emergency Corps . . . or write Headquarters.



ADVANCE PLANNING IS NECESSARY!

THREE PHASES OF EMERGENCY PLANNING

1. There are two types of provisions to be made for emergency radio coverage: (a) that for local short-haul work, (b) that for necessary out-of-the-area or distant coverage. Emergency coördinators have to make plans to cover both cases for implementation, separately or together, depending on the seriousness of the situation.

2. In using radio in the simplest form of emergency work, there is but one objective: to establish contact with an outside point to report difficulties, summon aid, or communicate intelligence of a limited character, when other means fail or are unavailable. *Coverage to some outside point is the thing!*

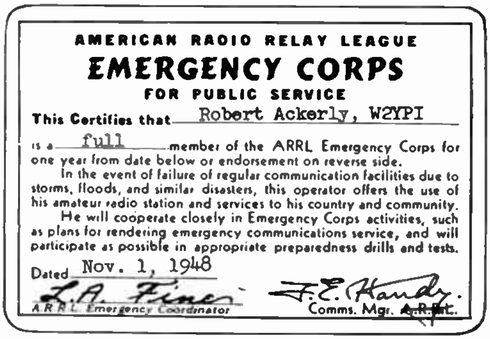
3. When a populated center has its communications crippled by causes that also disrupt transportation, food and water supply, and other organized public services, the *real emergency* is basically the same, but the effects are greatly magnified by the increased requirements. In addition to communicating considerable distances to get help from outside the paralyzed area, there is a problem of maintaining necessary communications between workers *within the area.*

Short-Haul Work — Local Nets

Local emergency networks should be established on a regular basis, wherever facilities permit, or at the very least on a stand-by basis with monthly tests and simulated drills in all the larger cities. Some mobiles and portables usually are assigned to cover points deemed strategically important by city officials for expected emergency use. Ordinarily local nets will use v.h.f. bands. Whether two meters or six meters or ten meters is used will depend on the community and the number of voluntary amateur workers equipped on a common frequency. In very large cities it may be justifiable to have local emergency nets using more than one amateur frequency band, with proper provision for traffic exchanged through common points. It is suggested that all regularly-operating v.h.f. nets dedicate themselves to emergency operation in the event of any public need. The responsible net control stations will be glad to assist the appropriate ARRL ECs to permit occasional exceptional tests as well as participation in all major Emergency Corps exercises.

H.F. Stations for Outside Contact

Local stations working on lower amateur frequencies should be included in every community plan, for contact to outside points via ARRL trunk lines, section nets, and other h.f. regional nets. Amateur service planners should attempt to determine from agencies they expect to serve from what cities or points *outside* the immediate local area supplies, assistance or exchanges of communication would be required


AMERICAN RADIO RELAY LEAGUE
EMERGENCY CORPS
FOR PUBLIC SERVICE
This Certifies that Robert Ackerly, W2YPI
is a full member of the ARRL Emergency Corps for one year from date below or endorsement on reverse side.
In the event of failure of regular communication facilities due to storms, floods, and similar disasters, this operator offers the use of his amateur radio station and services to his country and community.
He will coöperate closely in Emergency Corps activities, such as plans for rendering emergency communications service, and will participate as possible in appropriate preparedness drills and tests.
Dated Nov. 1, 1948
L.A. Finer — A.R.R.L. Emergency Coordinator
F.E. Handy — Comms. Mgr. A.R.R.L.

EMERGENCY CORPS MEMBERSHIP CARD
Have You Got Yours?

in event of possible emergencies. The needs of some agencies may warrant advance designation of a station. Or the setting up of traffic schedules or stand-by c.w. telegraph arrangements between neighboring cities for accurate handling of record communications can be tried on a weekly or other recurrent basis. It is *not* desirable to have too many stations operating on h.f. in haphazard fashion when there is an emergency. In any major catastrophe designated stations should be manned by several shifts of amateurs on a 24-hour basis, to avoid overworking one-man set-ups. Create the least band congestion and give the best service!

THE RÔLE OF AMATEUR COMMUNICATION

Conditions in emergency are such that *every available channel* must be used effectively. Naturally the telephone and telegraph wires are going to be used first when they are available for use. In the interest of suppression of interference, *radio should* not *be used where wire service is available,* except to line up additional avenues for communication to fall back on as a situation gets worse. Broadcast stations are best only for transmitting information to large groups of people. Some amateurs have actually written us to ask how they arrange to connect with the audio channels to the local broadcaster "that they may be ready for emergency." The chance that an amateur station will have to transmit over the local broadcast station, or will have his station used in place of that station in emergency, is extremely remote! Amateurs have important emergency functions, *not* in reaching the general public, but in handling necessary point-to-point communications effectively — as nearly like the service given by the local telegraph office and usual amateur traffic work as possible. The agencies served in past emergencies tell us that they want message handling to be accurate, secret and fast, and the messages reliably recorded in writing whenever practicable. Amateur networks will be depended on to form an invaluable communications system, secondary

OUR DUTIES ARE:

Primarily, the handling of strictly *emergency* traffic, that is, agency traffic between the scene of a disaster and a source of relief;

Secondarily, the prompt and efficient handling of third-party personal inquiry traffic which flows between a disaster area and all parts of the country.

to wire services whenever they conform to these specifications.

Strive for Efficiency

Whatever happens in emergency, you will find hysteria; also some amateurs who are activated by the thought that they must be "sleepless heroes." So activated, a situation can result that approaches chaos. In some past general emergencies we have had too many stations and operators on the air — too much meaningless communication! Instead of operating almost all the local stations full time or more in emergencies, how much better it would be to man the best-located and best-equipped stations, suitable for the work in hand, and man these by relief shifts of the best-qualified amateurs. This is the way to reduce interference. This is the way to secure well-operated stations, too. Instead of harassed, overworked operators who are inefficient and make mistakes, we need to put our operating on a 6- or 8-hour basis for each operator, and make it possible to keep the key stations on the air two or even three shifts per day (by relief operators) if and when required. Each station ought to have spare personnel to telephone messages, keep logs, prepare traffic for transmission, etc., during intensive operation in emergencies.

Select the Channel To Suit the Need

It is a characteristic of operators, both those using voice and those using telegraph, that they believe their mode of communication superior. For certain specified purposes and distances one or the other may indeed be preferable. Accuracy, speed, secrecy (avoiding public rumors) are all desirable and important considerations in emergencies. Accurate point-to-point work with messages receipted properly in two-way handling is preferable to any work patterned after broadcasting in emergencies, to prevent undesired duplication of messages. Duplications cause uncalled-for interference and irritate those sending and receiving messages. The consideration of the number of channels available and the need for efficiency calls for *increased utilization of telegraph* both by reasons of secrecy and minimized interference. C.w. "copy" minimizes garbling. Voice work is speedy for fast local liaison, where accuracy and recorded data are not absolutely needed and secrecy is not required. Many agencies (ARC, WU, WXSvc) that we expect to serve explain that secrecy from the public is important, that rumors may not start. For contacting the public, as from time to time is necessary, newspapers and broadcasting are recognized as the preferred media.

● WE ARE KNOWN BY OUR CONDUCT

Credit for the amateur service depends on teamwork; coördinators will assign a few stations to circuits for agencies that have a volume of traffic or important-enough needs for much consultation with a few scheduled points. To facilitate movement of *any* traffic between focal points in and near the "area," *most* stations of the few that represent a community (let us assume in each band) should work on hourly or half-hourly schedules with nets or individuals covering important points — or points from which wires to all the world are available. In an emergency, nets are molded from the available stations best to fit immediate needs which can only then be visualized.

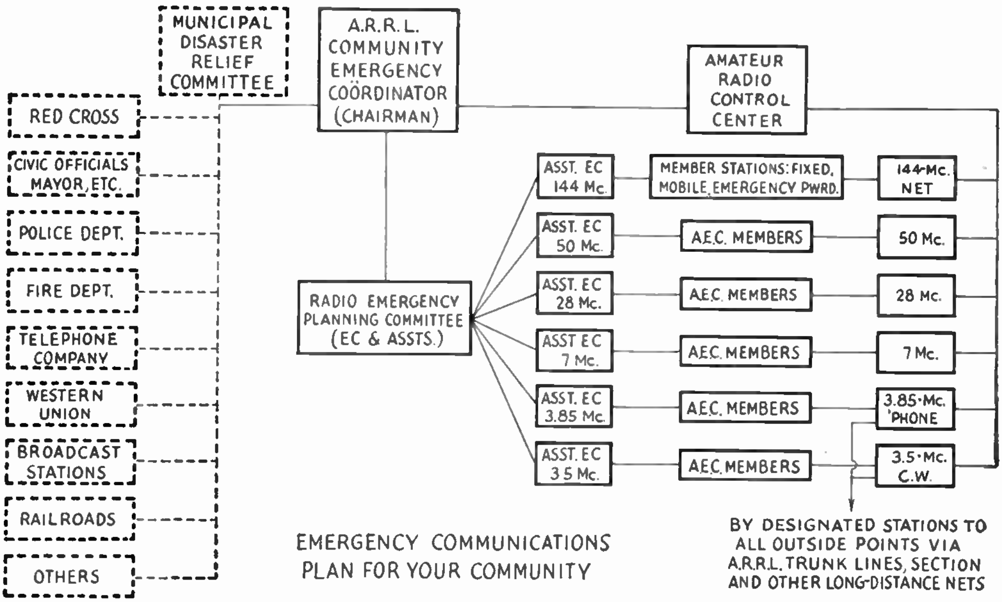
In the Local Emergency

As many amateurs as possible should be familiar with plans of the emergency coördinator and his committee. Local advices from three or four active local stations, and contact with relief operators, as well as the agencies meeting a critical situation, enable a coördinator to multiply the value of the amateur service to a community *many* times. By having *early* reports from several amateurs, those most active can be reported as on the job to agencies contacted. Reporters with stations will get any outgoing traffic to the points in their normal communication range. From early reports the first picture of stations and frequencies, and the points scheduled, develops into something tangible for all officials who need amateur radio service.

Tentative plans, based on studies of possible local contingencies, usually require modification to make the most effective disposition of local amateur resources possible to meet this "exact" communications emergency situation. Some stations become unavailable. A first set-up must be modified from hour to hour in serious local emergencies to meet changes in the situation, reach new points, reduce local interference, cover the important circuits with the more-skilled and efficient operators. A telephone, messenger contact, every manner of keeping up a local flow of information on station coverage, schedules, needs, etc., helps a coördinator meet the varying needs of a situation, and do credit to the amateur service.

In the Remote Emergency

Absolute silence, for those *not* in the area itself, is most often the very best form of coöperation — *unless and until* one definitely logs



EMERGENCY COMMUNICATIONS PLAN FOR YOUR COMMUNITY

an emergency station giving a directive call for his very city.

On a repeated QRRR *that goes unanswered*, remote stations should of course try to establish contact, to relay or assist as seems best after ascertaining the nature of the problem. QRRR is a call for operators *in an emergency situation only*, and operators misusing this call should be specifically referred to the precise definition of that call as given in amateur literature.

GENERAL CONSIDERATIONS IN EMERGENCY

It is impossible to state exact rules to cover every possible situation that will arise, but these principles supported by discussion and advice that will lead amateur operators to use their heads and their receivers to a greater extent than microphone or key, should add creditably to future performance of radio amateurs in emergencies.

Use All Channels Intelligently

A hurricane or flood brings conditions like war in some respects. The principle of effective military communication is to use all channels as intelligently and as efficiently as possible, without fear or favoritism. As long as wire circuits function and no overload or disruption occurs, the prudent coördinator will merely martial his forces and set in motion the plan of establishing schedules and nets in such fashion as to be ready to act constructively as need arises. Test messages may be in order. Agencies served may send inquiries or information by radio to check delivery and actual contact. When need requires, the communication load may be diverted to radio and delays in establishment of the radio facility will have been avoided.

There are always more amateurs outside an isolated area where communications by radio are desperately desired than there will be *in* such an area. Interference from outside stations is one of the main handicaps of the fellow in the emergency area who has self-power and low power — but if this fellow uses judgment and places himself in the center of a telegraph band or ties in with his ARRL section or state net on its frequency he often has no interference at all.

Keep the QRM Level Down

The operator working under emergency conditions should establish reliable contacts with stations as near to him as possible or at points logical for the handling of his traffic. Amateurs in localities remote from any emergency should be asked to *stand by;* to QRX without calls or interruptions until it is seen they can assist *better* than anyone else. DO NOT make an excuse for getting into a situation which someone else is really better fitted to handle, by reason of either his location or his signal. As a rule it is *not* necessary for stations many hundreds of miles distant to make extended efforts for each emergency occasion, since the very nature of most such traffic makes it better to handle it as a 50- to 250-mile proposition. If these distant stations will merely be silent until called upon, and defer their discussion about a rumored emergency until afterward, substituting a "learning-by-listening" attitude, the actual situation may be much more constructively handled. Interference levels can be kept much lower. One or two stations guarding each emergency-station frequency to pipe down unthinking newcomers (who start to interfere) by asking their coöperation will

QRRR

QRRR is the "official ARRL land SOS." It is a distress call for emergency use only, to be used only by a station definitely asking assistance!

Operators *outside* emergency zones, deprived of power in major floods and other disasters, have in the past failed to appreciate adequately the problem of battery- and low-powered stations trying to clear traffic and establish contact to "outside" points. It is *all wrong* for operators having picked up an inquiry message addressed to a point in a devastated area to work themselves up to the point of sending "CQ flood area" or sometimes even "QRRR." *Listening* and coöperating by keeping quiet are practically always best for "outside stations" until they are called for. It is *also a service* for amateurs to forward inquiry traffic, during the secondary phases of an emergency after all-important first messages have been sent and answers delivered, but the practice of *solicitation* of public traffic of the inquiry type through broadcasting stations or otherwise is rather to be discouraged. All questions of priority and procedure should be decided by evaluation of each particular matter in the light of the degree of public interest involved.

usually be quite sufficient to cover each telephone station or c.w. station actually engaged with emergency traffic.

The problem of reducing interference levels so the important communications in the area concerned can go through is one that most of all requires the enlistment and registration and coöperation of every licensee. The glory seeker who itches to be at his microphone or key is like the novelty seeker who intercepts police calls and tries to be in on the arrest, or who congests the streets and increases public losses by trying to beat the fire engines to the fire! Practice in operating behavior, procedure, and in the knack of properly framing messages, is needed. Actual practice "on the air" and over telephone circuits in transcribing *messages* is heartily recommended. Taking part in the ARRL Field Day and in net traffic operating is excellent for experience.

Never forget that *emergency* traffic is still *traffic*. An operator trained in handling record traffic is an operator who can be depended upon.

No Authority to Repeat Bulletins; Policing

Policing and observer stations recommended by ARRL or appointed by FCC in communications emergencies may report any failure to observe the FCC's emergency orders or rules, for disciplinary action. Addressed transmissions to amateurs (bulletins) should include their source. They should be repeated exactly, if at all, and *may not be repeated* without specific authority for so doing. Rumors are started by expansion, or contraction, and subsequent repetition of broadcast dispatches. To delete qualifying words, expand, exaggerate or alter meanings is criminal.

Authentication and Routing

The *signed* message is the best identification for the person who receives a message. We recommend that amateurs link emergency stations direct to civil and military authorities, so that all messages derive from reliable sources. Broadcasters and others are often criticized for transmitting hearsay and rumors, even for sending traffic requiring delivery "broadcast" and the like. The more nearly our amateur service can make results conform to the pattern of commercial *communications* and *not* to the field of broadcasting, the higher are our achievements acclaimed by agencies served in emergencies!

Red Cross Traffic Routing

The Office of Telecommunications Services of the American National Red Cross requests that all official Red Cross traffic during emergencies be routed via amateur stations which have a direct connection with the nation-wide Red Cross teletype system. Three stations are presently provided with such facilities: K3NRW in Washington, D. C., W9DUA in Evanston, Ill., and W6CXO in San Francisco. These stations will monitor 3550 kc., 7100 kc., and 14,050 kc. during periods of communications emergency, and will feed all Red Cross traffic onto the teletype wires.

In the event that it is impractical to contact any one of these three stations it is recommended that stations of the ARRL National Emergency Net, alerted during period of emergency, be used for such routing. National Emergency Net stations monitor the National Emergency Frequencies of 3550 kc., 3875 kc., and 7100 kc. to expedite the handling of long-haul traffic.

Watch the latest issues of *QST* for additional information on emergency operating work, possible changes in above calls or National Emergency Frequencies. During emergencies, follow W1AW bulletins for the latest emergency orders and then-current operating information.

League Operating Organization

Amateur operation must have point and constructive purpose to win public respect. Each individual amateur is the ambassador of the entire fraternity in his public relations and attitude toward his hobby. ARRL field organizations adds point and purpose to amateur operating.

The Communications Department of the League is concerned with the practical operation of stations in all branches of amateur activity. Appointment or awards are available for rag-chewer, traffic enthusiast, 'phone operator, DX man and experimenter.

There are seventy-two ARRL Sections in the League's field organization, which embraces the United States, Canada and certain other territory. Operating affairs in each Section are supervised by a Section Communications Manager elected by members in that section for a two-year term of office. Organization appointments are made by the section managers. The election of officials is covered in detail in the League's Constitution and By-Laws. Section communications managers' addresses for all sections are given in full in each issue of *QST*. SCMs welcome monthly activity reports from all amateur stations in their jurisdiction. Full information on appointments may be obtained from SCMs and is also contained in a League booklet, *Operating an Amateur Radio Station*, which will be sent from Headquarters on request (10¢ to nonmembers).

Whether your activity embraces 'phone or telegraphy, or both, there is a place for you in League organization.

LEADERSHIP POSTS

To advance each type of station work and group interest in amateur radio, and to develop practical communications plans with the greatest success, appointments of leaders and organizers in particular single-interest fields are made by SCMs. Each leadership post is important. Each provides activities and assistance for appointee groups and individual members along the lines of natural interest. While some posts further the general ability of amateurs to communicate efficiently at all times, by pointing activity toward networks and round tables, others are aimed specifically at establishment of provisions for organizing the amateur service as a stand-by communications group to serve the public in disaster or emergency of any sort.

Section Communications Manager

The Section Manager is the section executive or administrator in operating matters. He is the only elected official for the section alone, and the office is open to election each two years, or oftener if a vacancy occurs. Requirements for nomination and the system of mail balloting are covered in the operating booklet. Section managers report on all forms of amateur activity (for *QST*) monthly. Every active amateur licensee is invited to report his station activity to his SCM at the end of each calendar month for the preceding 30-days. He appoints all the following in accordance with section needs and individual qualifications.

Emergency Coördinator and Section Coördinator

A Section Emergency Coördinator is appointed to promote and develop the ARRL Emergency Corps in each section. The SCM's chief assistant on this subject customarily is fully responsible for every detail of the section emergency program. At the community level each town should have a full Emergency Coordinator, and he designates his own assistants in different amateur groups, geographical sections of larger cities and the like. See the chapter on Emergency Operation for detailed statement of the functions and responsibilities of SECs and ECs.

Route Manager

The Route Manager is the authority on schedules and routes, and his station must be active in traffic and organization work. The route manager's duties include coöperation with all radio amateurs in his territory in organizing and maintaining traffic routes, nets and schedules. RMs also test candidates for ORS appointment as directed by the SCM, Advice to amateurs wanting schedules or traffic routings via trunk lines, section nets, etc., will be given by RMs on request.

'Phone Activities Manager

The 'Phone Activities Manager may sponsor 'phone-operating activities in his territory, in the name of the League. The PAM appointment, while paralleling that of RM in some respects, has to do with the upbuilding of ARRL-section 'phone organization. The 'phone activities manager also tests candidates for OPS when referred by the SCM. 'Phone nets may develop ability to handle traffic or follow special objectives as worked out with net members by the PAM.

STATION APPOINTMENTS

ARRL's field organization has a place for every active amateur who has a station. The Communications Department organization exists to increase individual enjoyment in amateur radio work, and we extend a cordial invitation to every amateur to participate fully in the activities and to apply to the SCM for ORS, OPS, OES, OBS or OO appointment as soon as sufficiently experienced in amateur radio work.

The section manager makes appointments for specific work in accordance with the qualifications and rules for such appointments. He makes cancellations, likewise, for inactivity, inaptitude or failure to perform adequately the actions contemplated in appointment. All appointment certificates must be returned to SCMs annually for endorsement to keep them in effect — no trouble to this if there is continuing activity. The object is to keep field-organization standards high, and to insure a live-functioning organization in each amateur group at all times.

Official Relay Station Appointment

Every radio-telegraphing amateur interested in traffic work and operating activities, who can meet the qualifications, is eligible for appointment of his station as Official Relay Station. Brass pounders handle traffic because they enjoy such work. The potential value to his community and country of the operator who handles traffic is enhanced by his ability, as well as by the readiness of his station and schedules to function in emergency.

The appointment identifies the holder with high standards of amateur operating, and indicates personal keenness and responsibility. The holder voluntarily agrees to keep station equipment operative at all times, to report each month, and with absolute reliability to forward and deliver messages regularly through his station. Secure application forms from your local SCM.

Official 'Phone Station Appointment

This appointment is for every qualified ham who uses his microphone more than his key in his amateur station, who takes pride in the manner of signal he puts on the air and who aims to have his station really accomplish worth-while communication work. Official 'Phone Station appointees endeavor to live up to the Amateur's Code. OPS appointment aids 'phone-operating enjoyment by helping to promote good voice-operating practices and readiness for meeting the demands of emergency work.

Cultivation of operating ability that is essential to assure accuracy, conciseness and speed for point-to-point work is encouraged. Communications (two-way) techniques are the aim. Section 'phone-net members are encouraged to become OPS. Official 'Phone Station appointees, like ORSs agree voluntarily in accepting appointment that they will keep stations active and report on activities to the SCM monthly. Application forms are available from your SCM.

Official Experimental Station

The Official Experimental Station appointment is designed to promote operating progress from 50 Mc. through the microwaves. A broad group aim is production of data to aid in discussion and knowledge of transmission phenomena peculiar to each of our higher-frequency bands. The correlation of reports and results on the broadest possible scale will assist us in knowing how to use antenna structures, from notation of the pattern of these radiations in different terrain and circumstances, as regards transmitted-wave polarization, absorption, refraction and reflection.

OESs are concerned with new v.h.f. communications systems and equipment adaptations, such as amateur radio teletypewriter development. From time to time OESs contribute to a group bulletin; they also report propagation data in the form of monthly operating notes. The OES appointment is available only to members with operative equipment and a continuing interest "above 50 Mc."

This newest *station* appointment is popular with v.h.f. workers; write SCM for blanks.

Official Bulletin Station

OBS appointees send specifically-addressed information to *all radio amateurs* constituting the latest bulletins on amateur radio subjects, on all frequency bands using both code and voice transmission. Section managers give preference, in making appointment, to amateurs with stations of considerable power, whose operators are located where needed geographically to give v.h.f. local coverage, to get full-section coverage on h.f. bands, and as quotas permit. Where possible OBSs at distances from Headquarters are selected in line with their ability to copy data direct from W1AW.

Applicants for this appointment must submit their qualifications to the section manager with the proposed dates, times and frequencies for transmission of the bulletins. In deciding on the times of transmission schedules preference is given to those times when the largest number of amateurs is listening, that is, the hours between 6 P.M. and midnight. In establishing these schedules, a minimum of three regular transmissions per week are required to justify holding OBS appointment. OBSs are expected to send a monthly activity report to the SCM. OBSs who use c.w. are often copied by new amateurs for the code practice afforded. Nearly 600 appointees receive and send bulletin data and cover the country. OBS use the call "QST" (c.w.) or "Calling All Radio Amateurs" (voice) before each bulletin transmission.

Official Observer

ARRL's Coöperative Monitoring Service is supported by appointees known as Official Observers who have saved many a ham from "FCC trouble" by asking him to look into his signal and adjustments *before* technical maladjustments have come to FCC attention. Observers report "a.c." notes, unstable signals, overmodulation, parasitics, harmonics, off-frequency signals, illegal "broadcasting," or other abuses. Special mailing and report forms are provided by ARRL. Activity in using the notification forms is required for holding appointment.

Observers are examined and classified by SCMs for the type of work in which they engage. The top OO appointments require qualification at intervals in actual Frequency Measuring Tests. The OO certificate is thus a badge of technical proficiency in the measurement field. A Roman numeral designation, after the signature of the Observer indicates the class of Observer sending a report. The established classes are as follows:

I. Precise frequency checking.
II. General frequency checking (plus or minus 5 kc. at 14 Mc.)
III. Radiotelephone checks: modulation, stability, quality.
IV. Radiotelegraph checks: notes, clicks, chirps, stability.

Applications for observer appointment will be welcomed from the members who would like to help in bettering operating conditions in the different amateur bands. Observers must have good receiving equipment, and an accurate frequency meter, oscilloscope, etc., appropriate to the observing fields they expect to cover, as prerequisite to appointment. Class I and II observers must measure *within* their prescribed limits of accuracy on at least two, of four Frequency Measuring Tests ARRL runs each year, to retain their appointments. New observers must pass such a test to qualify. All initial appointments will be made as Class III or IV unless such measurement results are available to aid classification. All to SCMs.

Emblem Colors

Members wear the emblem with black-enamel background. A red background for an emblem will indicate that the wearer is SCM. SECs, ECs, RMs, PAMs may wear the emblem with green background. Observers and all station appointees are entitled to wear emblems with blue background.

● SECTION NETS AND TRUNK LINES

Amateurs can add much experience and pleasure to their own amateur lives, and substance and accomplishment to the credit of all of amateur radio, when organized into effective interconnection of cities and towns.

The successful operation of a net depends a lot on the Net Control Station. This station should be chosen carefully and be one that will not hesitate to enforce each and every net rule and set the example in his own operation.

A progressive net grows, obtaining new members both directly and through other net members. Bulletins may be issued at intervals to keep in direct contact with the members regarding general net business, to keep tab on net procedure and make suggestions for improvement, to keep track of active members and weed out inactive ones.

Official Relay Stations at key points are organized in trunk-line formation, covering fourteen east-west and north-south routes, connecting with numerous section and local networks and feeder systems for the purpose of efficient dispatch of traffic. Speedy and reliable work is carried on, the operation entirely on separate spot frequencies in the 3.5-Mc. amateur band. A station must hold ORS appointment to be considered for a trunk-line post.

Radio Club Affiliation

ARRL is pleased to grant affiliation to any amateur society having (1) 51% of the voting club membership made up of licensed United States or Canadian amateurs, and (2) 51% of its licensed amateurs also members of

ARRL. Where a society has common aims and wishes to add strength to that of other club groups to strengthen amateur radio by affiliation with the national amateur organization, a request addressed to the Communications Manager will bring the necessary forms and information to initiate the application for affiliation. Such clubs receive field-organization bulletins and special information at intervals for posting on club bulletin boards or for relay to their memberships. A travel plan providing communications, technical and secretarial contact from the Headquarters is worked out seasonally to give maximum benefits to as many as possible of the more than four hundred affiliated radio clubs. Papers on club work, suggestions for organizing, for constitutions, for radio courses of study, etc., are available on request.

Club Training Aids

One section of the ARRL Communications Department devotes its full time to the Training Aids Program. This program is a service to ARRL affiliated clubs. Material is supplied for club programs aimed at education, training and entertainment of club members, to make your club meetings more interesting and consequently better-attended.

Training Aids include such items as motion-picture films, film strips, slides, recordings, and lecture outlines. Also, code-proficiency training equipment such as recorders, tape transmitters and tapes will be loaned when such items are available.

All Training Aids materials are loaned free (except for shipping charges) to ARRL-affiliated clubs. Numerous groups use this ARRL service to good advantage. If your club is affiliated but has not yet taken advantage of this service, you are missing a good chance to add the available features to your meeting programs and general club activities. Watch club bulletins and *QST* or write the ARRL communications department for full details.

W1AW

The Maxim Memorial Station, W1AW, is dedicated to fraternity and service. Operated by the League headquarters, W1AW is located about four miles south of the Headquarters offices on a seven-acre site. The station is on the air daily, except holidays, and available time is divided between different bands and modes. Telegraph and 'phone transmitters are provided for all bands from 3.5 to 144 Mc. The normal frequencies in each band for c.w. and voice transmissions are as follows: 3555, 3950, 7215, 14,100, 14,280, 28,060, 29,000, 52,000 and 146,000 kc. Operating-visiting hours and the station schedule are listed every other month in *QST*.

All amateurs are invited to visit W1AW, as well as to work the station from their own shacks. The station was established to be a living memorial to Hiram Percy Maxim and to carry on the work and traditions of the amateur fraternity.

OPERATING ACTIVITIES

Within the ARRL field organization there are several special activities. The first Saturday night each month is set aside for all ARRL officials, officers and directors to get together over the air from their own stations. This activity is known to the gang as LO-NITE. For all appointees, quarterly tests called CD parties are scheduled to develop operating ability and a spirit of fraternalism. All League members may participate in the Annual ARRL Member Party held each January.

In addition to these special activities for appointees and members, ARRL sponsors various other activities open to all amateurs. The DX-minded amateur may participate in the Annual ARRL International DX Competition during February and March. This popular contest may bring you the thrill of working new countries. Then there is the ever-popular Sweepstakes in November. Of domestic scope, the SS affords the opportunity to work new states for that WAS award. The interests of v.h.f. enthusiasts are also provided for in special activities planned by ARRL.

As in all our operating, the idea of having a good time is combined in the Annual Field Day, with the more serious thought of preparing ourselves to render public service in times of emergency. A premium is placed on the use of equipment without connection to commercial power sources. Clubs and individual groups always have a good time in the "FD," learn much about the requirements for knockabout conditions afield.

ARRL contest activities are diversified to appeal to all operating interests, and will be found announced in detail in issues of *QST* preceding the different events.

AWARDS

The League-sponsored operating activities heretofore mentioned have useful objectives and provide much enjoyment for members of the fraternity. Achievement in amateur radio is recognized by various certificates offered through the League and detailed below.

WAS Award

WAS means "Worked All States." This award is available regardless of affiliation or nonaffiliation with any organization. Here are the few simple rules to follow in applying for a WAS Certificate:

1) Two-way communications must be established on the amateur bands with all forty-eight United States; any and all amateur bands may be used. A card from the District of Columbia may be submitted in lieu of one from Maryland.

2) Contacts with all forty-eight states must be made from the same location. Within a given community one location may be defined as from places no two of which are more than 25 miles apart.

3) Contacts may be made over any period of years, and may have been made any number of years ago, provided only that all contacts are from the same location.

4) Forty-eight QSL cards, or other written communications from stations worked confirming the necessary two-way contacts, must be submitted to ARRL headquarters.

5) Sufficient postage must be sent with the confirmations to finance their return. No correspondence will be returned unless sufficient postage is furnished.

6) The WAS award is available to all amateurs.

7) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford, Conn.

DX Century Club Award

Here are the rules under which the DX Century Club Award will be issued to amateurs who have worked and confirmed contact with 100 countries in the postwar period. If you worked fewer than 100 countries before the war and have since worked and confirmed a sufficient number to make the 100 mark, the DXCC is still available to you under the rules detailed on page 74 of June 1946 *QST*.

1) The Century Club Award Certificate for confirmed contacts with 100 or more countries is available to all amateurs everywhere in the world.

2) Confirmations must be submitted direct to ARRL headquarters for all countries claimed. Claims for a total of 100 countries must be included with first application. Confirmation from foreign contest logs may be requested in the case of the ARRL International DX Competition only, subject to the following conditions:

a) Sufficient confirmations of other types must be submitted so that these, plus the DX Contest confirmations, will total 100. In every case, Contest confirmations must not be requested for any countries from which the applicant has regular confirmations. That is, contest confirmations will be granted only in the case of countries from which applicants have no regular confirmations.

b) Look up the contest results as published in *QST* to see if your man is listed in the foreign scores. If he isn't, he did not send in a log and no confirmation is possible.

c) Give year of contest, date and time of QSO.

d) In future DX Contests, do not request confirmations until after the final results have been published, usually in one of the early fall issues. Requests before this time must be ignored.

3) The ARRL Countries List, printed periodically in *QST*, will be used in determining what constitutes a "country." The Miscellaneous Data chapter of this *Handbook* contains the Postwar Countries List.

4) Confirmations must be accompanied by a list of claimed countries and stations to aid in checking and for future reference.

5) Confirmations from additional countries may be submitted for credit each time ten additional confirmations are available. Endorsements for affixing to certificates and showing the new confirmed total (110, 120, 130, etc.) will be awarded as additional credits are granted. ARRL DX Competition logs from foreign stations may be utilized for these endorsements, subject to conditions stated under (2).

6) All contacts must be made with amateur stations working in the authorized amateur bands or with other stations licensed to work amateurs.

7) In cases of countries where amateurs are licensed in the normal manner, credit may be claimed only for stations using regular government-assigned call letters. No credit may be claimed for contacts with stations in any countries in which amateurs have been temporarily closed down by special government edict where amateur licenses were formerly issued in the normal manner.

8) All stations contacted must be "land stations" . . . contacts with ships, anchored or otherwise, and aircraft, cannot be counted.

9) All stations must be contacted from the same call area, where such areas exist, or from the same country in cases where there are no call areas. One exception is allowed to this rule: where a station is moved from one call area to another, or from one country to another, all contacts must be made from within a radius of 150 miles of the initial location.

10) Contacts may be made over any period of years from November 15, 1945, provided only that all contacts be made under the provisions of Rule 9, and by the same station licensee; contacts may have been made under different call letters in the same area (or country), if the licensee for all was the same.

11) All confirmations must be submitted exactly as received from the stations worked. Any altered or forged confirmations submitted for CC credit will result in disqualification of the applicant. The eligibility of any DXCC applicant who was ever barred from DXCC to reapply, and the conditions for such application, shall be determined by the Awards Committee. Any holder of the Century Club Award submitting forged or altered confirmations must forfeit his right to be considered for further endorsements.

12) OPERATING ETHICS: Fair play and good sportsmanship in operating are required of all amateurs working toward the DX Century Club Award. In the event of specific objections relative to continued poor operating ethics an individual may be disqualified from the DXCC by action of the ARRL Awards Committee.

13) Sufficient postage for the return of confirmations must be forwarded with the application. In order to insure the safe return of large batches of confirmations, it is suggested that enough postage be sent to make possible their return by first-class mail, registered.

14) Decisions of the ARRL Awards Committee regarding interpretation of the rules as here printed or later amended shall be final.

15) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford 7, Conn.

WAC Award

The International Amateur Radio Union issues WAC (Worked All Continents) certificates to all members of member-societies who submit proof of two-way communication with at least one station on each continent. Foreign amateurs submit their proof direct to member-societies of the IARU. Others may make application to ARRL, headquarters society of the Union. A c.w. and a telephony certificate are available. Also, special endorsement will be placed on certificates upon receipt of request accompanied by proof of having worked all continents on 50 Mc

Code Proficiency Award

Many hams can follow the general idea of a contact "by ear" but when pressed to "write it down" they "muff" the copy. The Code Proficiency Award invites every amateur to prove himself as a proficient operator, and sets up a system of awards for step-by-step gains in copying proficiency. It enables every amateur to check his code proficiency, to better that proficiency, and to receive a certification of his receiving speed.

This program is a whale of a lot of fun. The League will give a certificate to any licensed radio amateur who demonstrates that he can copy perfectly, for at least one minute, plain-language Continental code at 15, 20, 25, 30 or 35 words per minute, as transmitted during special monthly transmissions from W1AW, or from W6OWP, WØTQD and others mentioned in *QST*.

As part of the ARRL Code Proficiency program, W1AW transmits plain-language practice material each evening, Monday through Friday, at speeds from 9 to 35 w.p.m. All amateurs are invited to use these transmissions to increase their code copying ability. Non-amateurs are invited to utilize the lower speeds, 9, 12 and 15 w.p.m., which are transmitted for the benefit of persons studying the code in preparation for the amateur license examination. Refer to any issue of *QST* for details of the practice schedule.

Rag Chewers Club

The Rag Chewers Club is designed to encourage friendly contacts and discourage the "hello-good-by" type of QSO. Its purpose is to bond together operators interested in honest-to-goodness rag-chewing over the air. Membership certificates are available.

How To Get In: (1) Chew the rag with a member of the club for at least a solid half hour. This does not mean a half hour spent in trying to get a message over through bad QRM or QRN, but a solid half hour of conversation or message handling. (2) Report the conversation by card to The Rag Chewers Club, ARRL, Communications Department, West Hartford, Conn., and ask the member station you talk with to do the same. When both reports are received you will be sent a membership certificate entitling you to all the privileges of a Rag Chewer.

How To Stay In: (1) Be a conversationalist on the air instead of one of those tongue-tied infants who don't know any words except "cuagn" or "cul," or "QRU" or "nil." Talk to the fellows you work with and get to know them. (2) Operate your station in accordance with the radio laws and ARRL practice. (3) Observe rules of courtesy on the air. (4) Sign "RCC" after each call so that others may know you can talk as well as call.

A-1 Operator Club

The A-1 Operator Club should include in its ranks every good operator. To become a member, one must be nominated by at least two operators who already belong. General keying or voice technique, procedure, copying ability, judgment and courtesy all count in rating candidates under the club rules detailed at length in *Operating an Amateur Radio Station*. Aim to make yourself a fine operator, and one of these days you will be pleasantly surprised by an invitation to belong to the A-1 Operator Club, which carries a worth-while certificate in its own right.

Brass Pounders League

Every individual reporting more than a specified minimum in official monthly traffic totals is given an honor place in the *QST* listing known as the Brass Pounders League and a certificate to recognize his performance.

The value to amateurs in operator training, and the utility of amateur message handling to the members of the fraternity itself as well as to the general public, make message-handling work of prime importance to the fraternity. Fun, enjoyment, and the feeling of having done something really worth while for one's fellows is accentuated by pride in message files, records, and letters from those served.

Old Timers Club

The Old Timers Club is open to anyone who holds an amateur call at the present time, and who held an amateur license (operator or station) 20-or-more years ago. Lapses in activity during the intervening years are permitted.

If you can qualify as an "Old Timer," send us a brief chronology of your ham career, being sure to indicate the date of your first amateur license, and your present call. If the evidence submitted proves you eligible for the OTC, you will be added to the roster and will receive a membership certificate.

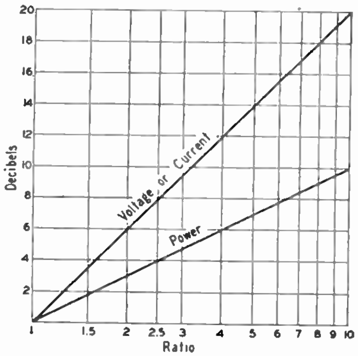
● INVITATION

Amateur radio is capable of giving enjoyment, self-training, social, and organization benefits in proportion to what the individual amateur puts into his hobby. All amateurs are invited to become ARRL members, to work toward awards, and to accept the challenge and invitation offered in field-organization appointments. Drop a line for the booklet *Operating an Amateur Radio Station*, which has detailed information on the field-organization appointments and awards. Accept today the invitation to take full part in all ARRL activities and organization work.

Miscellaneous Data

THE DECIBEL

In most radio communication the received signal is converted into sound. This being the case, it is useful to appraise signal strengths in terms of relative loudness as registered by the ear. A peculiarity of the ear is that an increase or decrease in loudness is responsive to the *ratio* of the amounts of power involved, and is practically independent of absolute *value* of the power. For example, if a person estimates that the signal is "twice as loud" when the transmitter power is increased from 10 watts to 40 watts, he will also estimate that a 400-watt signal is twice as loud as a 100-watt signal. In other words, the ear has a *logarithmic* response.



This fact is the basis for the use of the relative-power unit called the **decibel**. A change of one decibel (abbreviated **db.**) in the power level is just detectable as a change in loudness under ideal conditions. The power ratio and decibels are related by the following formula:

$$Db. = 10 \log \frac{P_2}{P_1}$$

Common logarithms (base 10) are used.

Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but *only when the impedance is the same for both values of voltage, or current.* The gain of an amplifier cannot be expressed correctly in db. if it is based on the ratio of the output voltage to the input voltage unless both voltages are measured across the same value of impedance. When the impedance at both points of measurement is the same, the following formula may be used for voltage or current ratios:

$$Db. = 20 \log \frac{V_2}{V_1} \text{ or } 20 \log \frac{I_2}{I_1}$$

The two formulas are shown graphically in the accompanying chart for ratios from 1 to 10. Gains (increases) expressed in decibels may be added arithmetically; losses (decreases) may be subtracted. A power decrease is indicated by prefixing the decibel figure with a minus sign. Thus +6 db. means that the power has been multiplied by 4, while −6 db. means that the power has been divided by 4. The chart may be used for other ratios by adding (or subtracting, if a loss) 10 db. each time the ratio scale is multiplied by 10, for power ratios; or by adding (or subtracting) 20 db. each time the scale is multiplied by 10 for voltage or current ratios.

Example: The power input to a transmitter is increased from 75 to 600 watts. Assuming that the efficiency is the same in both cases, the ratio of the new output power to the old is 600/75 = 8. From the chart, the signal will be increased 9 db. Note that increasing the power to 750 watts, a ratio of 10, would increase the signal to 10 db., a barely perceptible increase over 600 watts.

Example: A speech amplifier has an output of 10 watts when excited by 0.02 volt from a crystal microphone. The nominal impedance of the microphone is 50,000 ohms. In a 50,000-ohm load, the voltage developed by the 10 watts would be

$$E = \sqrt{PR} = \sqrt{10 \times 50{,}000} = \sqrt{500{,}000}$$
$$= 707 \text{ volts}$$

The voltage ratio of the amplifier therefore is 707/0.02 = 35,350. This is the same as 3.5 × 10,000. A voltage ratio of 10,000 (10^4) is equal to 4 × 20 = 80 db. From the chart, a voltage ratio of 3.5 = 11 db. Adding the two gives 11 + 80 = 91 db. as the gain of the amplifier.

Example: A transmission line is terminated in its characteristic impedance and operates without standing waves. The power put into the line is 150 watts, but the power measured at the output end is 100 watts. The ratio is 150/100 = 1.5. From the chart, this ratio is equal to 1.9 db. The loss in the line is therefore 1.9 db.

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

SYMBOLS FOR ELECTRICAL QUANTITIES

Admittance	Y, y
Angular velocity ($2\pi f$)	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, x
Reactance, Capacitive	X_C
Reactance, Inductive	X_L
Reluctivity	ν
Resistance	R, r
Resistivity	ρ
Susceptance	b
Speed of rotation	n
Voltage	E, e
Work	W

PILOT-LAMP DATA

Lamp No.	Bead Color	Base (Miniature)	Bulb Type	RATING Volts	Amp.
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[3]	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.

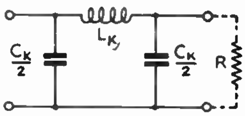
[4] Replace with No. 48.

[5] Use in 2.5-volt sets where regular bulb burns out too frequently.

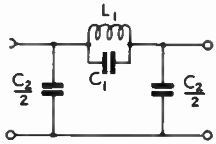
ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.
Ampere (amperes)	a.
Amplitude modulation	AM
Antenna	ant.
Audio frequency	a.f.
Centimeter	cm.
Continuous waves	c.w.
Cycles per second	c.p.s.
Decibel	db.
Direct current	d.c.
Electromotive force	e.m.f.
Frequency	f.
Frequency modulation	FM
Ground	gnd.
Henry	h.
High frequency	h.f.
Intermediate frequency	i.f.
Interrupted continuous waves	i.c.w.
Kilocycles (per second)	kc.
Kilovolt	kv.
Kilowatt	kw.
Magnetomotive force	m.m.f.
Medium frequency	m.f.
Megacycles (per second)	Mc.
Megohm	MΩ
Meter	m.
Microfarad	μfd.
Microhenry	μh.
Micromicrofarad	μμfd.
Microvolt	μv.
Microvolt per meter	μv.m.
Microwatt	μw.
Milliampere	ma.
Millivolt	mv.
Milliwatt	mw.
Modulated continuous waves	m.c.w.
Ohm	Ω
Power	P
Power factor	p.f.
Radio frequency	r.f.
Ultrahigh frequency	u.h.f.
Very-high frequency	v.h.f.
Volt (volts)	v.
Watt (watts)	w.

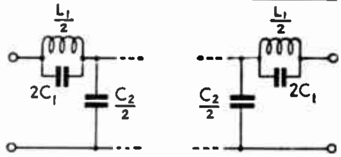
LOW-PASS FILTERS



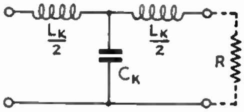
Constant-k π section



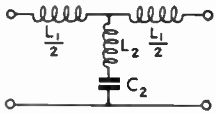
m-derived π section



m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section

m-derived end sections for use with intermediate T section

$L_K = \frac{R}{\pi f_c}$ $C_K = \frac{1}{\pi f_c R}$

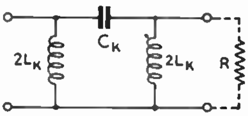
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m C_K$

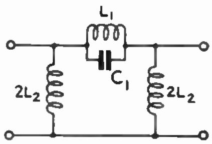
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m C_K$

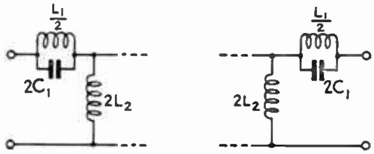
HIGH-PASS FILTERS



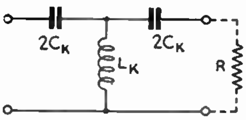
Constant-k π section



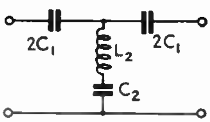
m-derived π section



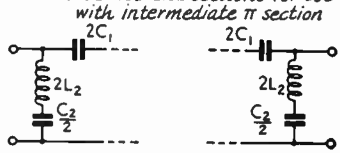
m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section



m-derived end section for use with intermediate T section

$L_K = \frac{R}{4\pi f_c}$ $C_K = \frac{1}{4\pi f_c R}$

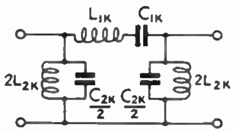
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

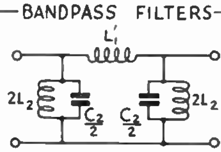
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

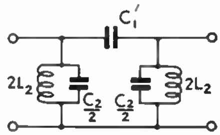
BANDPASS FILTERS



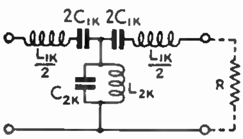
Constant-k π section



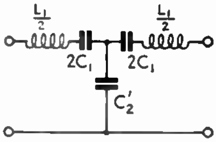
Three-element π section



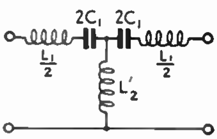
Three-element π section



Constant-k T section



Three-element T section



Three-element T section

$L_{1K} = \frac{R}{\pi(f_2-f_1)}$ $C_{1K} = \frac{f_2-f_1}{4\pi f_1 f_2 R}$

$L_{2K} = \frac{(f_2-f_1)R}{4\pi f_1 f_2}$ $C_{2K} = \frac{1}{\pi(f_2-f_1)R}$

$L_1 = L_{1K}$ $L_1' = \frac{R}{\pi(f_1+f_2)}$

$C_1 = \frac{f_2-f_1}{4\pi f_1^2 R}$ $L_2 = \frac{(f_2-f_1)R}{4\pi f_1^2}$

$C_2 = C_{2K}$ $C_2' = \frac{1}{\pi(f_1+f_2)R}$

$L_1 = \frac{f_1 R}{\pi f_2(f_2-f_1)}$ $C_1 = C_{1K}$

$C_1' = \frac{f_1+f_2}{4\pi f_1 f_2 R}$ $L_2 = L_{2K}$

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

FILTERS

The filter sections shown on the facing page can be used alone or, if greater attenuation and sharper cut-off are required, several sections can be connected in series. In the low- and high-pass filters, f_c represents the cut-off frequency, the highest (for the low-pass) or the lowest (for the high-pass) frequency transmitted without attenuation. In the bandpass-filter designs, f_1 is the low-frequency cut-off and f_2 the high-frequency cut-off.

If several low- (or high-) pass sections are to be used, it is advisable to use m-derived end sections on either side of a constant-k section, although an m-derived center section can be used. The factor m relates the ratio of the cut-off frequency and f_∞, a frequency of high attenuation. A value of 0.6 is usually used for m, although a deviation of 10 or 15 per cent from this value is not too serious in amateur work. For a value of m = 0.6, f_∞ will be $1.25f_c$ for the low-pass filter and $0.8f_c$ for the high-pass filter. Other values can be found from

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$ for the low-pass filter and

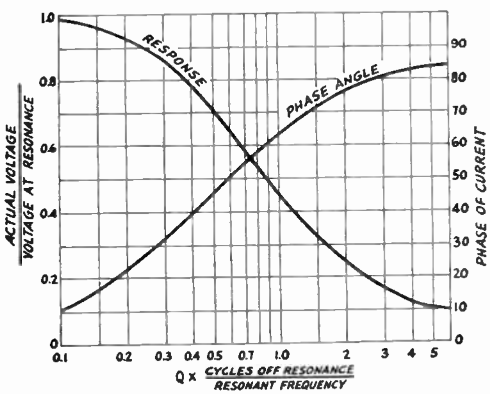
$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2}$$ for the high-pass filter.

The filters shown should be terminated in a resistance = R, and there should be little or no reactive component in the termination.

Simple audio filters can be made with powdered-iron-core chokes and paper condensers, but above this range mica condensers should be used. The values of the components can vary by ±5% with little or no reduction in performance. The more sections there are to performance. The more sections there are to the filter the greater is the need for accuracy in the components. High resistance in the coils or appreciable leakage in the condensers will also reduce the effectiveness of the filter.

TUNED-CIRCUIT RESPONSE

The graph below gives the response and phase angle of a high-Q parallel-tuned circuit.



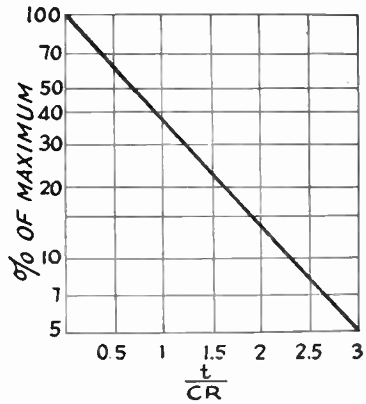
Circuit Q is equal to

$$2\pi fRC \quad \text{or} \quad \frac{R}{2\pi fL}$$

where L and C are the inductance and capacitance at the resonant frequency, f, and R is the parallel resistance across the circuit. The curves above become more accurate as the circuit Q is higher, but the error is not especially great for values as low as Q = 10.

VOLTAGE DECAY IN *RC* CIRCUITS

The accompanying chart enables calculation of the instantaneous voltage across the terminals of a condenser discharging through a resistance. The voltage is given in terms of percentage of the voltage to which the condenser is initially charged. To obtain the



voltage-decay time in seconds, multiply the factor (t/CR) by the time constant of the resistor-condenser circuit.

Example: A 0.01-μfd. condenser is charged to 150 volts and then allowed to discharge through a 0.1-megohm resistor. How long will it take the voltage to fall to 10 volts? In percentage, 10/150 = 6.7%. From the chart, the factor corresponding to 6.7% is 2.7. The time constant of the circuit is equal to CR = 0.01 × 0.1 = 0.001. The time is therefore 2.7 × 0.001 = 0.0027 second, or 2.7 milliseconds.

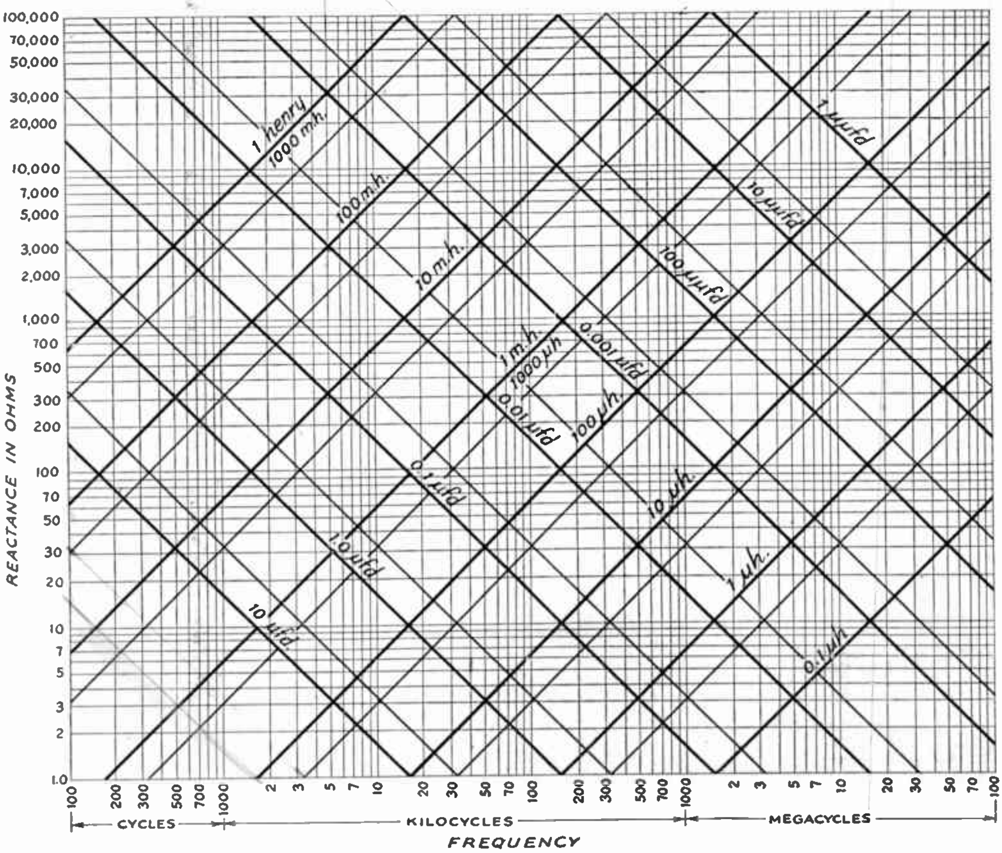
Example: An RC circuit is desired in which the voltage will fall to 50% of the initial value in 0.1 second. From the chart, t/CR = 0.7 at the 50%-voltage point. Therefore $CR = t/0.7 = 0.1/0.7 = 1.43$. Any combination of resistance and capacitance whose product (R in megohms and C in microfarads) is equal to 1.43 can be used; for example, C could be 1 μfd. and R 1.43 megohms.

TABLE OF DIELECTRIC CHARACTERISTICS

Dielectric material[1]	*Dielectric constant (K)*	*Power factor* 60 cycles	1 kc.	1 Mc.	10 Mc.	100 Mc.	*Dielectric strength (puncture voltage)*[2]	*Volume resistivity*[3] ( )
Air (normal pressure)	1.0						19.8–22.8	
AlSiMag A196	5.7–6.3	2.9		0.21	0.15		240	10^{14}
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	$2–30 \times 10^{10}$
Ceresin wax	2.5–2.6			0.12–0.21				
Cresol formaldehyde	6	10					400	
Dilectene	3.57					0.33		
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}
Formica MF-66	4.6–4.9		1.5	1.1			450	
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	$5 \times 10^{14}–10^{15}$
Lucite[7]	2.5–3	7	5	1.5–3	1.9		180–500	
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	
Mykroy	6.5–7			0.1–0.2			630	
Nylon	3.6			2.2				
Paper	2.0–2.6						1250	
Paraffin wax (solid)	1.9–2.6			0.1–0.3			300	$10^{15}–10^{19}$
Pemque	7.21			0.2				
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	$10^{10}–10^{13}$
Yellow	5.3–5.4			0.36–0.7			500	
Polyethylene	2.3–2.4	0.02	0.02	0.02–0.05			1000	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	$10^{14}–10^{18}$
Rubber (hard)[10]	2–3.5(3)			0.5–1			450	$10^{12}–10^{15}$
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	$10^{14}–10^{15}$
Titanium dioxide[12]	90–170		0.1	0.1				
Urea formaldehyde[13]	5–7	3–5	2–3	2–4	4		300–550	$10^{12}–10^{13}$
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				
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 apply to relatively thin sections and cannot be multiplied directly to give breakdown for thicker sections without added safety factor.
[3] In ohm-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.

INDUCTIVE AND CAPACITIVE REACTANCE VS. FREQUENCY CHART



By use of the chart above, the approximate reactance of any capacitance 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 capacitance 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 combination will tune. First locate the respective slanting lines for the capacitance 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

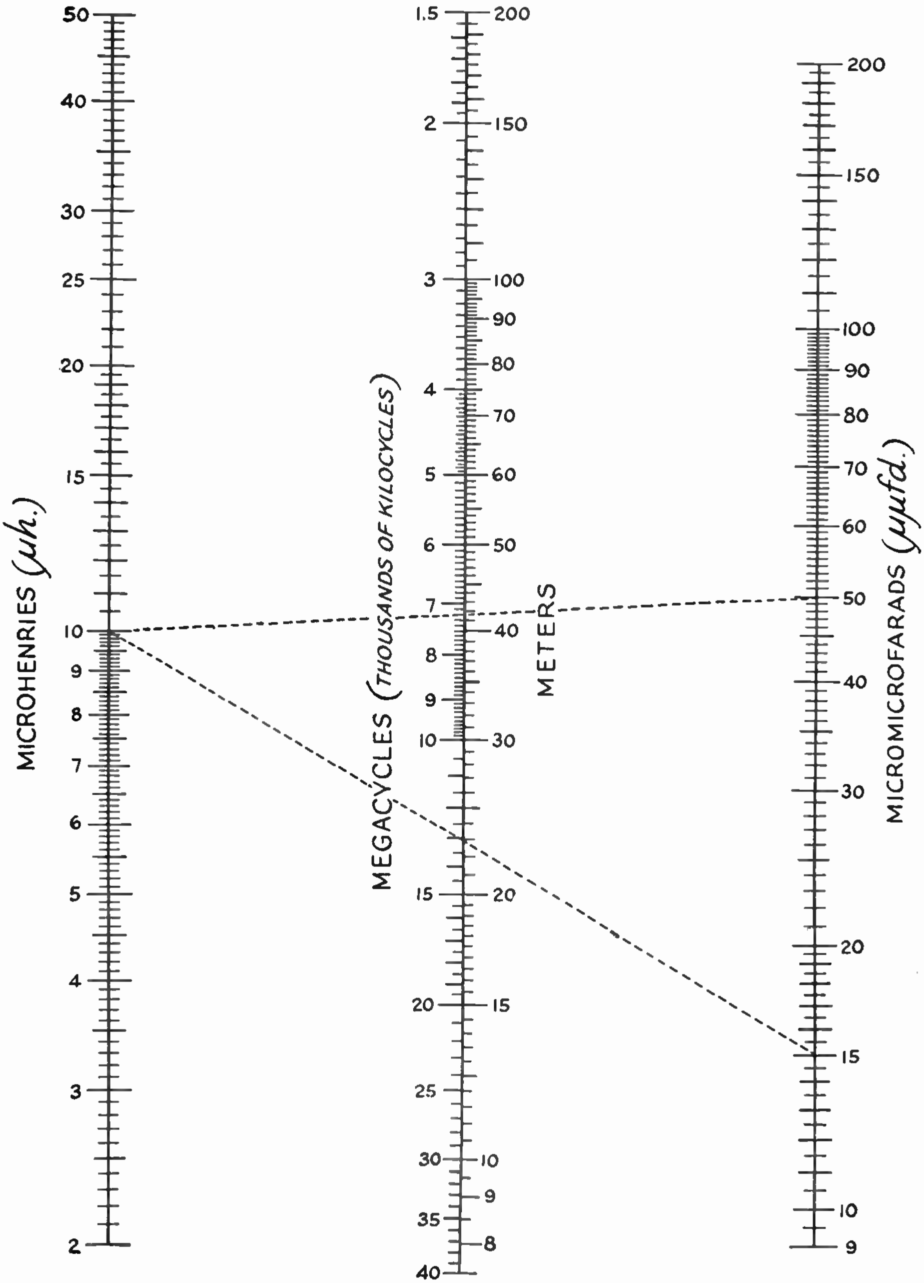
	Relative Conductivity [1]	*Temp. Coef.* [2] *of Resistance*
Aluminum (2S; pure)	59	0.0049
Aluminum (alloys):		
Soft-annealed	45–50	
Heat-treated	30–45	
Brass	28	0.002–0.007
Cadmium	19	
Chromium	55	
Climax	1.83	
Cobalt	16.3	
Constantin	3.24	0.00002
Copper (hard drawn)	89.5	0.004
Copper (annealed)	100	
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
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

[1] At 20° C., based on copper as 100. [2] Per °C. at 20° C.

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.

INDUCTANCE, CAPACITANCE AND FREQUENCY CHART — 1.5–40 MC.



This chart may be used to find the values of inductance and capacitance 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 capacitance of 15 μμfd. and a maximum capacitance of 50 μμfd. If it is to be used with a coil of 10-μh. inductance, what frequency range will be covered? The straightedge is connected between 10 on the left-hand scale and 15 on the right, giving 13 Mc. as the high-frequency limit. Keeping the straightedge 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 capacitances 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.

GREEK ALPHABET

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

THE R-S-T SYSTEM

READABILITY

1 — Unreadable.
2 — Barely readable, occasional words distinguishable.
3 — Readable with considerable difficulty.
4 — Readable with practically no difficulty.
5 — Perfectly readable.

SIGNAL STRENGTH

1 — Faint signals, barely perceptible.
2 — Very weak signals.
3 — Weak signals.
4 — Fair signals.
5 — Fairly good signals.
6 — Good signals.
7 — Moderately strong signals.
8 — Strong signals.
9 — Extremely strong signals.

TONE

1 — Extremely rough hissing note.
2 — Very rough a.c. note, no trace of musicality.
3 — Rough low-pitched a.c. note, slightly musical.
4 — Rather rough a.c. note, moderately musical.
5 — Musically-modulated note.
6 — Modulated note, slight trace of whistle.
7 — Near d.c. note, smooth ripple.
8 — Good d.c. note, just a trace of ripple.
9 — Purest d.c. note.

If the signal has the characteristic steadiness of crystal control, add the letter X to the RST report. If there is a chirp, the letter C may be added to so indicate. Similarly for a click, add K. The above reporting system is used on both c.w. and voice, leaving out the "tone" report on voice.

Q SIGNALS

Given below are a number of Q signals whose meanings most often need to be expressed with brevity and clearness in amateur work. (Q abbreviations take the form of questions only when each is sent followed by a question mark.)

QAV Are you calling me? I am calling. . . .
QRG Will you tell me my exact frequency in kilocycles? Your frequency is......kc.
QRH Does my frequency vary? Your frequency varies.
QRI How is the tone of my transmission? The tone of your transmission is....(1. good; 2. variable; 3. bad).
QRJ Are you receiving me badly? Are my signals weak? I cannot receive you. Your signals are too weak.
QRK What is the readability of my signals (1 to 5)? The readability of your signals is... (1 to 5).
QRL Are you busy? I am busy or busy with (........)
QRM Are you being interfered with? I am interfered with.
QRN Are you troubled by atmospherics? I am being troubled by atmospherics.
QRQ Shall I send faster? Send faster (......words per min.).
QRS Shall I send more slowly? Send more slowly (.... w.p.m.)
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? Please tell..........that I am calling him.
QRX When will you call again? I will call you again at .... hours (on ......kc.).
QRZ By whom am I being called? 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) at a time? Send........telegrams at a time.
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 sent me.
QSO Can you communicate with........direct (or through........)? I can communicate with ....direct (or through........).
QSP Will you relay to........? I will relay to........
QSV Shall I send a series of VVV........? Send a series of VVV.
QSW Will you send on......kc., etc.? I will send on ....kc., etc.
QSX Will you listen for......(call sign) on......kc? I am listening for....on....kc.
QSY Shall I change to........kilocycles without changing the type of wave? Change to........kc. without changing type of wave.
QSZ Shall I send each word or group twice? Send each word or group twice.
QTA Shall I cancel nr........as if it had not been sent? Cancel nr........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........
QTH What is your position (location)? My location is........(by any indication.)
QTR What is the exact time? The time is........

Special abbreviations adopted by ARRL:

QST General call preceding a message addressed to all amateurs and ARRL members. This is in effect "CQ ARRL."
QRRR Official ARRL "land SOS." A distress call for emergency use only by a station in an emergency situation.

ABBREVIATIONS FOR C.W. WORK

Abbreviations help to cut down unnecessary transmission. However, make it a rule not to abbreviate unnecessarily when working an operator of unknown experience.

AA	All after
AB	All before
ABT	About
ADR	Address
AGN	Again
ANT	Antenna
BCI	Broadcast interference
BCL	Broadcast listener
BK	Break; break me; break in
BN	All between; been
B4	Before
C	Yes
CFM	Confirm; I confirm
CK	Check
CL	I am closing my station; call
CLD-CLG	Called; calling
CUD	Could
CUL	See you later
CUM	Come
CW	Continuous wave
DLD-DLVD	Delivered
DX	Distance
ECO	Electron-coupled oscillator
FB	Fine business; excellent
GA	Go ahead (or resume sending)
GB	Good-by
GBA	Give better address
GE	Good evening
GG	Going
GM	Good morning
GN	Good night
GND	Ground
GUD	Good
HI	The telegraphic laugh; high
HR	Here; hear
HV	Have
HW	How
LID	A poor operator
MILS	Milliamperes
MSG	Message; prefix to radiogram
N	No
ND	Nothing doing
NIL	Nothing; I have nothing for you
NR	Number
NW	Now; I resume transmission
OB	Old boy
OM	Old man
OP-OPR	Operator
OSC	Oscillator
OT	Old timer; old top
PBL	Preamble
PSE-PLS	Please
PWR	Power
PX	Press
R	Received solid; all right; OK; are
RAC	Rectified alternating current
RCD	Received
REF	Refer to; referring to; reference
RPT	Repeat; I repeat
SED	Said
SEZ	Says
SIG	Signature; signal
SINE	Operator's personal initials or nickname
SKED	Schedule
SRI	Sorry
SVC	Service; prefix to service message
TFC	Traffic
TMW	Tomorrow
TNX-TKS	Thanks
TT	That
TU	Thank you
TXT	Text
UR-URS	Your; you're; yours
VFO	Variable-frequency oscillator
VY	Very
WA	Word after
WB	Word before
WD-WDS	Word; words
WKD-WKG	Worked; working
WL	Well; will
WUD	Would
WX	Weather
XMTR	Transmitter
XTAL	Crystal
YF (XYL)	Wife
YL	Young lady
73	Best regards
88	Love and kisses

W PREFIXES BY STATES

Alabama	W4
Arizona	W7
Arkansas	W5
California	W6
Colorado	WØ
Connecticut	W1
Delaware	W3
District of Columbia	W3
Florida	W4
Georgia	W4
Idaho	W7
Illinois	W9
Indiana	W9
Iowa	WØ
Kansas	WØ
Kentucky	W4
Louisiana	W5
Maine	W1
Maryland	W3
Massachusetts	W1
Michigan	W8
Minnesota	WØ
Mississippi	W5
Missouri	WØ
Montana	W7
Nebraska	WØ
Nevada	W7
New Hampshire	W1
New Jersey	W2
New Mexico	W5
New York	W2
North Carolina	W4
North Dakota	WØ
Ohio	W8
Oklahoma	W5
Oregon	W7
Pennsylvania	W3
Rhode Island	W1
South Carolina	W4
South Dakota	WØ
Tennessee	W4
Texas	W5
Utah	W7
Vermont	W1
Virginia	W4
Washington	W7
West Virginia	W8
Wisconsin	W9
Wyoming	W7

INTERNATIONAL PREFIXES

Below is the list of prefixes assigned to the countries of the world by the 1947 International Telecommunications Conference at Atlantic City. While changes of call swhere necessary to accord with these assignments, which were effective January 1, 1949, may not yet have been made in all cases, it is expected that all such changes will occur in the immediate future.

AAA-ALZ	United States of America
AMA-AOZ	(Not allocated)
APA-ASZ	Pakistan
ATA-AWZ	India
AXA-AXZ	Commonwealth of Australia
AYA-AZZ	Argentina Republic
BAA-BZZ	China
CAA-CEZ	Chile
CFA-CKZ	Canada
CLA-CMZ	Cuba
CNA-CNZ	Morocco
COA-COZ	Cuba
CPA-CPZ	Bolivia
CQA-CRZ	Portuguese Colonies
CSA-CUZ	Portugal
CVA-CXZ	Uruguay
CYA-CZZ	Canada
DAA-DMZ	Germany
DNA-DQZ	Belgian Congo
DRA-DTZ	Bielorussian Soviet Socialist Republic
DUA-DZZ	Republic of the Philippines
EAA-EHZ	Spain
EIA-EJZ	Ireland
EKA-EKZ	Union of Soviet Socialist Republics
ELA-ELZ	Republic of Liberia
EMA-EOZ	Union of Soviet Socialist Republics
EPA-EQZ	Iran
ERA-ERZ	Union of Soviet Socialist Republics
ESA-ESZ	Estonia
ETA-ETZ	Ethiopia
EUA-EZZ	Union of Soviet Socialist Republics
FAA-FZZ	France and Colonies and Protectorates
GAA-GZZ	Great Britain
HAA-HAZ	Hungary
HBA-HBZ	Switzerland
HCA-HDZ	Ecuador
HEA-HEZ	Switzerland
HFA-HFZ	Poland
HGA-HGZ	Hungary
HHA-HHZ	Republic of Haiti
HIA-HIZ	Dominican Republic
HJA-HKZ	Republic of Colombia
HLA-HMZ	Korea
HNA-HNZ	Iraq
HOA-HPZ	Republic of Panama
HQA-HRZ	Republic of Honduras
HSA-HSZ	Siam
HTA-HTZ	Nicaragua
HUA-HUZ	Republic of El Salvador
HVA-HVZ	Vatican City State
HWA-HYZ	France and Colonies and Protectorates
HZA-HZZ	Kingdom of Saudi Arabia
IAA-IZZ	Italy and Colonies
JAA-JSZ	Japan
JTA-JVZ	Mongolian People's Republic
JWA-JXZ	Norway
JYA-JZZ	(Not allocated)
KAA-KZZ	United States of America
LAA-LNZ	Norway
LOA-LWZ	Argentina Republic
LXA-LXZ	Luxemburg
LYA-LYZ	Lithuania
LZA-LZZ	Bulgaria
MAA-MZZ	Great Britain
NAA-NZZ	United States of America
OAA-OCZ	Peru
ODA-ODZ	Republic of Lebanon
OEA-OEZ	Austria
OFA-OJZ	Finland
OKA-OMZ	Czechoslovakia
ONA-OTZ	Belgium and Colonies
OUA-OZZ	Denmark
PAA-PIZ	Netherlands
PJA-PJZ	Curacao
PKA-POZ	Netherlands Indies
PPA-PYZ	Brazil
PZA-PZZ	Surinam
QAA-QZZ	(Service abbreviations)
RAA-RZZ	Union of Soviet Socialist Republics
SAA-SMZ	Sweden
SNA-SRZ	Poland
SSA-SUZ	Egypt
SVA-SZZ	Greece
TAA-TCZ	Turkey
TDA-TDZ	Guatemala
TEA-TEZ	Costa Rica
TFA-TFZ	Iceland
TGA-TGZ	Guatemala
THA-THZ	France and Colonies and Protectorates
TIA-TIZ	Costa Rica
TJA-TZZ	France and Colonies and Protectorates
UAA-UQZ	Union of Soviet Socialist Republics
URA-UTZ	Ukrainian Soviet Socialist Republic
UUA-UZZ	Union of Soviet Socialist Republics
VAA-VGZ	Canada
VHA-VNZ	Commonwealth of Australia
VOA-VOZ	Newfoundland
VPA-VSZ	British Colonies and Protectorates
VTA-VWZ	India
VXA-VYZ	Canada
VZA-VZZ	Commonwealth of Australia
WAA-WZZ	United States of America
XAA-XIZ	Mexico
XJA-XOZ	Canada
XPA-XPZ	Denmark
XQA-XRZ	Chile
XSA-XSZ	China
XTA-XWZ	France and Colonies and Protectorates
XXA-XXZ	Portuguese Colonies
XYA-XZZ	Burma
YAA-YAZ	Afghanistan
YBA-YHZ	Netherlands Indies
YIA-YIZ	Iraq
YJA-YJZ	New Hebrides
YKA-YKZ	Syria
YLA-YLZ	Latvia
YMA-YMZ	Turkey
YNA-YNZ	Nicaragua
YOA-YRZ	Rumania
YSA-YSZ	Republic of El Salvador
YTA-YUZ	Yugoslavia
YVA-YYZ	Venezuela
YZA-YZZ	Yugoslavia
ZAA-ZAZ	Albania
ZBA-ZJZ	British Colonies and Protectorates
ZKA-ZMZ	New Zealand
ZNA-ZOZ	British Colonies and Protectorates
ZPA-ZPZ	Paraguay
ZQA-ZQZ	British Colonies and Protectorates
ZRA-ZUZ	Union of South Africa
ZVA-ZZZ	Brazil
2AA-2ZZ	Great Britain
3AA-3AZ	Principality of Monaco
3BA-3FZ	Canada
3GA-3GZ	Chile
3HA-3UZ	China
3VA-3VZ	France and Colonies and Protectorates
3WA-3XZ	(Not allocated)
3YA-3YZ	Norway
3ZA-3ZZ	Poland
4AA-4CZ	Mexico
4DA-4IZ	Republic of the Philippines
4JA-4LZ	Union of Soviet Socialist Republics
4MA-4MZ	Venezuela
4NA-4OZ	Yugoslavia
4PA-4SZ	British Colonies and Protectorates
4TA-4TZ	Peru
4UA-4UZ	United Nations
4VA-4VZ	Republic of Haiti
4WA-4WZ	Yemen
4XA-4ZZ	(Not allocated)
5AA-5ZZ	(Not allocated)
6AA-6ZZ	(Not allocated)
7AA-7ZZ	(Not allocated)
8AA-8ZZ	(Not allocated)
9AA-9ZZ	(Not allocated)

A.R.R.L. COUNTRIES LIST

Official List for ARRL DX Contest and the Postwar DXCC

Country	Prefix
Aden and Socotra Island	VS9
Afghanistan	YA
Alaska	KL7
Albania	ZA
Aldabra Islands	
Algeria	FA
Andaman Ids. and Nicobar Ids.	VU
Andorra	PX
Anglo-Egyptian Sudan	ST
Angola	CR6
Antarctica	
Argentina	LU
Ascension Island	ZD8
Australia (including Tasmania)	VK
Austria	(MB9) OE
Azores Islands	CT2
Bahama Islands	VP7
Bahrein Island	VU7
Baker Island, Howland Island and Am. Phoenix Islands	KB6
Balearic Islands	EA6
Barbados	VP6
Basutoland	ZS8
Bechuanaland	ZS9
Belgian Congo	OQ
Belgium	ON
Bermuda Islands	VP9
Bhutan	
Bolivia	CP
Bonin Islands and Volcano Islands (e.g., Iwo Jima)	
Borneo, British North	VS3
Borneo, Netherlands	PK5
Brazil	PY
Brunei	VS5
Bulgaria	LZ
Burma	XZ
Cameroons, French	FE8
Canada	VE
Canal Zone	KZ5
Canary Islands	EA8
Cape Verde Islands	CR4
Caroline Islands	
Cayman Islands	VP5
Celebes and Molucca Islands	PK6
Ceylon	VS7
Chagos Islands	VQ8
Channel Islands	GC
Chile	CE
China	XU, C
Christmas Island	ZC3
Clipperton Island	
Cocos Island	TI
Cocos Islands	ZC2
Colombia	HK
Comoro Islands	
Cook Islands	ZK1
Corsica	FC
Costa Rica	TI
Crete	SV
Cuba	CM-CO
Cyprus	(MD7) ZC4
Czechoslovakia	OK
Denmark	OZ
Dodecanese Islands (e.g., Rhodes)	SV5
Dominican Republic	HI
Easter Island	
Ecuador	HC
Egypt	(MD5) SU
Eire (Irish Free State)	EI
England	G
Eritrea	I6
Ethiopia	ET
Faeroes, The	OY
Falkland Islands	VP8
Fanning Island (Christmas Island)	VR3
Fiji Islands	VR2
Finland	OH
Formosa (Taiwan)	C3
France	F
French Equatorial Africa	FQ8
French India	FN
French Indo-China	FI8
French Oceania (e.g., Tahiti)	FO8
French West Africa	FF8
Fridtjof Nansen Land (Franz Josef Land)	UA1
Galapagos Islands	
Gambia	ZD3
Germany	D
Gibraltar	ZB2
Gilbert & Ellice Islands and Ocean Island	VR1
Goa (Portuguese India)	CR8
Gold Coast (and British Togoland)	ZD4
Greece	SV
Greenland	OX
Guadeloupe	FG8
Guantanamo Bay	NY4
Guatemala	TG
Guiana, British	VP3
Guiana, Netherlands (Surinam)	PZ
Guiana, French, and Inini	FY8
Guinea, Portuguese	CR5
Guinea, Spanish	
Haiti	HH
Hawaiian Islands	KH6
Honduras	HR
Honduras, British	VP1
Hong Kong	VS6
Hungary	HA
Iceland	TF
Ifni	
India	VU
Iran	EP-EQ
Iraq	YI
Ireland, Northern	GI
Isle of Man	GD
Italy	I
Jamaica	VP5
Jan Mayen Island	
Japan	J
Jarvis Island, Palmyra group (Christmas Island)	KP6
Java	PK
Johnston Island	KJ6
Kenya	VQ4
Kerguelen Islands	
Korea	HL
Kuwait	
Laccadive Islands	VU4
Lebanon	AR8
Leeward Islands	VP2
Liberia	EL
Libya	(MD1-2) LI
Liechtenstein	HE1
Luxembourg	LX
Macau	CR9
Madagascar	FB8
Madeira Islands	CT3
Malaya	VS1, VS2
Maldive Islands	
Malta	ZB1
Manchuria	C9
Marianas Islands (Guam)	KG6
Marion Island (Prince Edward Island)	ZS
Marshall Islands	KX6
Martinique	FM8
Mauritius	VQ8
Mexico	XE
Midway Island	KM6
Miquelon and St. Pierre Islands	FP8
Monaco	CZ
Mongolian Republic (Outer)	
Morocco, French	CN
Morocco, Spanish	EA9
Mozambique	CR7
Nepal	
Netherlands	PA
Netherlands West Indies	PJ
New Caledonia	FK8
Newfoundland and Labrador	VO
New Guinea, Netherlands	PK6
New Guinea, Territory of	VK9
New Hebrides	FU8, YJ
New Zealand	ZL
Nicaragua	YN
Nigeria	ZD2
Niue	ZK2
Norfolk Island	VK9
Norway	LA
Nyasaland	ZD6
Oman	(MP4) VS9
Pakistan	AP
Palau (Pelew) Islands	
Palestine	ZC6
Panama	HP
Papua Territory	VK9
Paraguay	ZP
Peru	OA
Philippine Islands	KA
Phoenix Islands (British)	VR1
Pitcairn Island	VR6
Poland	SP
Portugal	CT
Principe and Sao Thome Islands	
Puerto Rico	KP4
Reunion Island	FR8
Rhodesia, Northern	VQ2
Rhodesia, Southern	ZE
Rio de Oro	
Roumania	YR
Ryukyu Islands (e.g., Okinawa)	J9
St. Helena	ZD7
Salvador	YS
Samoa, American	KS6
Samoa, Western	ZM
San Marino	M1
Sarawak	VS5
Sardinia	IS
Saudi Arabia (Hedjaz and Nejd)	HZ
Scotland	GM
Seychelles	VQ9
Siam	HS
Sierra Leone	ZD1
Sikkim	AC3
Solomon Islands	VR4
Somaliland, British	VQ6
Somaliland, French	FL8
Somaliland, Italian	(MD4)
South Georgia	VP8
South Orkney Islands	VP8
South Sandwich Islands	VP8
South Shetland Islands	VP8
Southwest Africa	ZS3
Soviet Union:	
European Russian Socialist Federated Soviet Republic	UA1-3-4-6
Asiatic Russian S.F.S.R.	UA9-0
Ukraine	UB5
White Russian Soviet Socialist Republic	UC
Azerbaijan	UD6
Georgia	UF6
Armenia	UG6
Turkoman	UH8
Uzbek	UI8
Tadzhik	UJ8
Kazakh	UL7
Kirghiz	UM8
Karelo-Finnish Republic	UN1
Moldavia	UO5
Lithuania	UP
Latvia	UQ
Estonia	UR
Spain	EA
Sumatra	PK4
Svalbard (Spitzbergen)	LA
Swan Island	KS4
Swaziland	ZS7
Sweden	SM
Switzerland	HB
Syria	YK
Tanganyika Territory	VQ3
Tangier Zone	EK
Tannu Tuva	
Tibet	AC4
Timor, Portuguese	CR10
Togoland, French	FD8
Tokelau (Union) Islands	
Tonga (Friendly) Islands	VR5
Trans-Jordan	ZC1
Trieste	
Trinidad and Tobago	VP4
Tristan da Cunha and Gough Island	ZD9
Tunisia	FT4
Turkey	TA
Turks and Caicos Islands	VP5
Uganda	VQ5
Union of South Africa	ZS
United States of America	W, K
Uruguay	CX
Vatican City	HV
Venezuela	YV
Virgin Islands	KV4
Wake Island	KW6
Wales	GW
Windward Islands	VP2
Wrangel Islands	
Yemen	
Yugoslavia	YT-YU
Zanzibar	VQ1

NOTE: Prefixes in parentheses are used by occupation forces.

INTERNATIONAL AMATEUR PREFIXES

To make possible identification of calls heard on the air, the international telecommunications conferences assign to each nation certain alphabetical blocks, from which *all* classes of stations are assigned prefixes. The following prefixes are used by amateurs:

AC3	Sikkim
AC4	Tibet
AP	Pakistan
AR8	Lebanon
C	China (unofficial)
C3	Formosa
C9	Manchuria
CE	Chile
CM, CO	Cuba
CN	Morocco, French
CP	Bolivia
CR4	Cape Verde Islands
CR5	Guinea, Portuguese
CR6	Angola
CR7	Mozambique
CR8	Goa (Portuguese India)
CR9	Macau
CR10	Timor, Portuguese
CT1	Portugal
CT2	Azores Islands
CT3	Madeira Islands
CX	Uruguay
CZ	Monaco
D	Germany
EA	Spain
EA6	Balearic Islands
EA8	Canary Islands
EA9	Morocco, Spanish
EI	Eire (Irish Free State)
EK	Tangier Zone
EL	Liberia
EP, EQ	Iran (Persia)
ET	Ethiopia
F	France
FA	Algeria
FB8	Madagascar
FC	Corsica
FD8	Togoland, French
FE8	Cameroons, French
FF8	French West Africa
FG8	Guadeloupe
FI8	French Indo-China
FK8	New Caledonia
FL8	Somaliland, French
FM8	Martinique
FN	French India
FO8	French Oceania (e.g., Tahiti)
FP8	St. Pierre & Miquelon Islands
FQ8	French Equatorial Africa
FR8	Reunion Island
FT4	Tunisia
FU8	New Hebrides
FY8	Guiana, French & Inini
G	England
GC	Channel Islands
GD	Isle of Man
GI	Ireland, Northern
GM	Scotland
GW	Wales
HA	Hungary
HB	Switzerland
HC	Ecuador
HE1	Liechtenstein
HH	Haiti
HI	Dominican Republic
HK	Colombia
HL	Korea
HP	Panama
HR	Honduras
HS	Siam
HV	Vatican City
HZ	Saudi Arabia (Hedjaz & Nejd)
I	Italy
I6	Eritrea
J	Japan
J9	Ryukyu Islands (e.g., Okinawa)
K	United States of America
KA	Philippine Islands
KB6	Baker, Howland & American Phoenix Islands
KC4	Little America
KG6	Guam, Saipan, Tinian
KH6	Hawaiian Islands
KJ6	Johnston Islands
KL7	Alaska
KM6	Midway Islands
KP	Puerto Rico
KP6	Palmyra Group, Jarvis Island
KS6	American Samoa
KS4	Swan Island
KV4	Virgin Islands
KW6	Wake Island
KX6 (J9)	Marshall Islands
KZ5	Canal Zone
LA	Norway
LI	Libya
LU	Argentina
LX	Luxembourg
LZ	Bulgaria
M1	San Marino
MB9	Austria
MD1	Cyrenaica
MD2	Tripolitania
MD3	Eritrea
MD4	Somalia
MD5	Suez Canal Zone
MD6	Iraq
MD7	Cyprus
MP4	Oman
NY4	Guantanamo Bay
OA	Peru
OE	Austria
OH	Finland
OK	Czechoslovakia
ON	Belgium
OQ	Belgian Congo
OX	Greenland
OY	Faeroes, The
OZ	Denmark
PA	Netherlands
PJ	Netherlands West Indies
PK1, 2, 3	Java
PK4	Sumatra
PK5	Borneo, Netherlands
PK6	Celebes & Molucca Islands
PK6	New Guinea, Netherlands
PX	Andorra
PY	Brazil
PZ	Guiana, Netherlands (e.g., Surinam)
SM	Sweden
SP	Poland
ST	Anglo-Egyptian Sudan
SU	Egypt
SV	Greece
SV	Crete
SV5	Dodecanese (e.g., Rhodes)
TA	Turkey
TF	Iceland
TG	Guatemala
TI	Costa Rica
TI	Cocos Island
UA1, 3, 4, 6	European Russian Socialist Federated Soviet Republic
UA9, Ø	Asiatic Russian S.F.S.R.
UB5	Ukraine
UC	White Russian Soviet Socialist Republic
UD6	Azerbaijan
UF6	Georgia
UG6	Armenia
UH8	Turkoman
UI8	Uzbek
UJ8	Tadzhik
UL7	Kazakh
UM8	Kirghiz
UN1	Karelo-Finnish Republic
UO5	Moldavia
UP	Lithuania
UQ	Latvia
UR	Estonia
VE	Canada
VK	Australia (including Tasmania)
VK9	Papua Territory
VK9	New Guinea, Territory of
VK9	Norfolk Island
VO	Newfoundland & Labrador
VP1	British Honduras
VP2	Leeward & Windward Islands
VP3	British Guiana
VP4	Trinidad & Tobago
VP5	Jamaica & Cayman Islands
VP5	Turks & Caicos Islands
VP6	Barbados
VP7	Bahama Islands
VP8	Falkland Islands
VP8	South Georgia
VP8	South Orkney Islands
VP8	South Sandwich Islands
VP8	South Shetland Islands
VP9	Bermuda Islands
VQ1	Zanzibar
VQ2	Northern Rhodesia
VQ3	Tanganyika Territory
VQ4	Kenya
VQ5	Uganda
VQ6	Somaliland, British
VQ8	Mauritius & Chagos Islands
VQ9	Seychelles
VR1	Gilbert & Ellice Islands & Ocean Island
VR2	Fiji Islands
VR3	Fanning Island (Christmas Island)
VR4	Solomon Islands
VR5	Tonga (Friendly) Islands
VR6	Pitcairn Island
VS1	Strait Settlements
VS2	Federated Malay States
VS3	British North Borneo
VS5	Sarawak, Brunei
VS6	Hong Kong
VS7	Ceylon
VS8 (VU7)	Bahrein Islands
VS9	Maldive Islands
VU	India
VU4	Laccadive Islands
VU7	Bahrein Islands
W	United States of America
XE	Mexico
XU	China
XZ	Burma
YA	Afghanistan
YI	Iraq
YK	Syria
YN	Nicaragua
YR	Roumania
YS	Salvador
YT, YU	Yugoslavia
YV	Venezuela
ZA	Albania
ZB1	Malta
ZB2	Gibraltar
ZC1	Trans-Jordan
ZC2	Cocos Islands
ZC3	Christmas Island
ZC4	Cyprus
ZC6	Palestine
ZD1	Sierra Leone
ZD2	Nigeria
ZD3	Gambia
ZD4	Togoland, Gold Coast
ZD6	Nyasaland
ZD7	St. Helena
ZD8	Ascension Island
ZD9	Tristan da Cunha & Gough Island
ZE1	Southern Rhodesia
ZK1	Cook Islands
ZK2	Niue
ZL	New Zealand
ZM	British Samoa
ZP	Paraguay
ZS1, 2, 4, 5, 6	Union of South Africa
ZS3	Southwest Africa
ZS7	Swaziland
ZS8	Basutoland
ZS9	Bechuanaland

Vacuum-Tube Data

For the convenience of the designer, the receiving-type tubes listed in this chapter are grouped by filament voltages and construction types (glass, metal, miniature, etc.). For example, all 6.3-volt metal tubes are listed in Table I, all lock-in base tubes are in Table III, all miniatures are in Table XI, and so on.

Transmitting tubes are divided into triodes and tetrodes-pentodes, then listed according to rated plate dissipation. This permits direct comparison of ratings of tubes in the same power classification.

For quick reference, all tubes are listed in numerical-alphabetical order in the index beginning on the following page.

Tube Ratings

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 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.

For certain air-cooled transmitting tubes, there are two sets of maximum values, one designated as CCS (Continuous Commercial Service) ratings, the other 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. 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. ICAS ratings are considerably higher than CCS ratings. They permit the handling of greater power, and 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.

Typical Operating Conditions

The typical operating conditions given for transmitting tubes represent, in general, maximum ICAS ratings where such ratings have been given by the manufacturer. They do not represent the *only* possible method of operation of a particular tube type. Other values of plate voltage, plate current, grid bias, etc., may be used so long as the maximum ratings for a particular voltage or current are not exceeded.

INDEX TO TUBE TABLES

BASE TYPE DESIGNATIONS

The type of base used on each tube listed in the tables is indicated in the base column by a letter.

The meaning of each letter is as follows:

A = Acorn
B = Glass button miniature
J = Jumbo
L = Lock-in
M = Medium
N = None or special type
O = Octal
S = Small
W = Wafer

INDEX TO VACUUM-TUBE TYPES

For convenience in locating data on specific tube types the index below lists all tubes in numerical-alphabetical order, showing the page number where individual tubes may be found in the classified-data section (pages 565-605) and the identifying base-diagram number in the base-diagram section (pages 558-564).

Type	Page	Base
00-A	576	4D
01-A	576	4D
0A2	582	5BO
0A3	583	4AJ
0A4G	582	4V
0B2	582	5BO
0B3	583	4AJ
0C3	583	4AJ
0D3	583	4AJ
0Y4	586	4BU
0Z4	586	4R
0Z4A	587	4R
1	586	4G
1A3	578	5AP
1A4P	571	4M
1A4T	571	4K
1A5GT	573	6X
1A6	571	6L
1A7GT	573	7Z
1AB5	573	5BF
1B3GT	586	3C
1B4P	571	4M
1B5	571	6M
1B7GT	573	7Z
1B8GT	573	8AW
1B47	582	—
1B48	586	—
1C5GT	573	6X
1C6	572	6L
1C7G	572	7Z
1C8	580	—
1C21	582	4V
1D5GP	572	5Y
1D5GT	572	5R
1D7G	572	7Z
1D8GT	573	8AJ
1E4G	573	5S
1E5GP	572	5Y
1E7G	572	8C
1F4	572	5K
1F5G	572	6X
1F6	572	6W
1F7G	572	7AD
1G4GT	573	5S
1G5G	572	6X
1G6GT	573	7AB
1H4G	572	5S
1H5G	573	5Z
1H6G	572	7AA
1J5G	572	6X
1J6G	572	7AB
1L4	578	6AR
1LA4	573	5AD
1LA6	573	7AK
1LB4	573	5AD
1LB6	573	8AX
1LC5	573	7AO
1LC6	573	7AK
1LD5	573	6AX
1LE3	573	4AA
1LG5	573	7AO
1LH4	573	5AG
1LN5	573	7AO
1N5GT	573	5Y
1N6G	573	7AM
1P5GT	573	5Y
1Q5GT	573	6AF
1R4	573	4AH
1R5	578	7AT
1S4	578	7AV
1S5	578	6AU
1SA6GT	573	6CA
1SB6GT	573	6CB
1T4	578	6AR
1T5GT	573	6AF
1U4	578	6AR
1U5	578	6BW
1-V	586	4G
1V5	580	—
1W5	580	—
1Z2	586	7CB
2A3	571	4D
2A4G	582	5S
2A5	571	6B
2A6	571	6G
2A7	571	7C
2AP1	583	11B
2B4	582	5A
2B6	571	7J
2B7	571	7D
2B25	586	3T
2BP1	583	12E
2C4	582	5AS
2C21	569	7BH
2C22	566	4AM
2C22	588	4AM
2C25	588	4D
2C26A	588	4BB
2C34	588	T-7DC
2C35	577	Fig. 38
2C39	595	—
2C40	588	Fig. 19
2C43	588	Fig. 19
2C44	588	Fig. 17
2C51	578	8CJ

Type	Page	Base
2D21	582	7BN
2E5	571	6R
2E22	601	5J
2E24	599	7CL
2E25	600	5BJ
2E26	599	7CK
2E30	578	7CQ
2E30	599	7CQ
2E31	580	—
2E32	580	—
2E35	580	—
2E36	580	—
2E41	580	—
2E42	580	—
2G5	571	6R
2G21	580	—
2G22	580	—
2K25	604	Fig. 60
2K26	604	Fig. 60
2K28	604	Fig. 61
2K33	604	Fig. 62
2K34	604	Fig. 58
2K35	604	Fig. 58
2K39	604	Fig. 59
2K41	604	Fig. 59
2K42	604	Fig. 59
2K43	604	Fig. 59
2K44	604	Fig. 59
2K46	604	Fig. 58
2K47	604	Fig. 58
2K56	604	Fig. 60
2S/4S	571	5D
2V3G	586	4Y
2W3	586	4X
2X2	586	4AB
2X2-A	586	4AB
2Y2	586	4AB
2Z2	586	4B
3A4	578	7BB
3A4	599	7BB
3A5	578	7BC
3A5	588	7BC
3A8GT	576	8AS
3AP1	583	7AN
3B5GT	576	7AP
3B7	573	7BE
3B7	588	7BE
3B24	586	T-4A
3B25	586	4P
3B26	586	Fig. 31
3B27	586	4B
3BP1	583	14A
3C5GT	576	7AQ
3C6	576	7BW
3C22	595	Fig. 30
3C23	582	3G
3C24	589	2D
3C28	589	Fig. 56
3C34	590	3G
3C37	596	—
3D6	573	6BB
3D6	599	6BB
3D23	601	Fig. 54
3D24	601	T-9J
3DP1	583	Fig. 49
3DX3	601	Fig. 40
3E6	573	7CJ
3E22	601	8BY
3E29	601	7BP
3EP1	583	11A
3GP1	583	11A
3JP1	583	14B
3K21	604	Fig. 58
3K22	604	Fig. 58
3K23	604	Fig. 59
3K27	604	Fig. 59
3K30	604	Fig. 58
3KP1	583	11M
3LE4	576	6BA
3LF4	576	6BB
3MP1	584	Fig. 2
3Q4	578	7BA
3Q5GT	576	7AQ
3S4	578	7BA
3V4	578	6BX
3-25A3	589	3G
3-25D3	589	2D
3-50A4	591	3G
3-50D4	591	2D
3-50G2	591	2D
3-75A2	593	2D
3-75A3	593	2D
3-100A2	594	2D
3-100A4	594	2D
3X-100A11	595	—
3-150A2	596	4BC
3-150A3	596	4BC
3X-150A3	596	—
3-250A2	597	2N
3-250A4	597	2N
3-300A2	598	4BC
3-300A3	598	4BC
4A6G	572	8L
4A6G	576	8L
4C32	597	2N

Type	Page	Base
4C34	597	2N
4C36	595	Fig. 56
4D22	602	Fig. 50
4D23	603	5BK
4D32	602	Fig. 51
4E27	602	7BM
4X150A	603	T-9J
4-65A	602	5BK
4-125A	603	5BK
4-250A	603	5BK
4-400A	603	5BK
5AP1	584	11A
5AZ4	586	5T
5BP1	584	11A
5CP1	584	14B
5FP1	584	5AN
5HP1	584	11A
5JP1	584	11E
5LP1	584	11F
5MP1	584	7AN
5R4GY	586	5T
5RP1	584	14F
5T4	586	5T
5T4P	584	12C
5U4G	586	5T
5UP1	584	12E
5V4G	586	5L
5W4	586	5T
5X3	586	4C
5X4G	586	5Q
5Y3G	586	5T
5Y4G	586	5Q
5Z3	586	4C
5Z4	586	5L
6A3	569	4D
6A4	569	5B
6A5G	566	6T
6A6	569	7B
6A7	569	7C
6A8	565	8A
6AB5	569	6R
6AB6G	566	7AU
6AB7	565	8N
6AC5G	566	6Q
6AC6G	566	7AU
6AC7	565	8N
6AD5G	566	6Q
6AD6G	566	7AG
6AD7G	566	8AY
6AE5G	566	6Q
6AE6GT	566	7AH
6AE7GT	567	7AX
6AF5G	567	6Q
6AF6G	569	7AG
6AF7G	567	8AG
6AG5	578	7BD
6AG6G	567	7S
6AG7	565	8Y
6AG7	599	8Y
6AH5G	567	6AP
6AH6	578	7CC
6AH7GT	567	8BE
6AJ5	578	7PM
6AJ7	565	8N
6AK5	585	7BD
6AK6	579	7BK
6AK6	599	7BK
6AK7	565	8Y
6AL5	579	6BT
6AL6G	567	6AM
6AL7GT	567	8CH
6AN5	579	7BD
6AN6	579	7BJ
6AQ5	579	7BZ
6AQ5	599	7BZ
6AQ6	579	7BT
6AQ7GT	567	8CK
6AR5	579	6CC
6AR6	567	6BQ
6AR7GT	567	8CG
6AS5	579	7CV
6AS6	579	7CM
6AS7GT	567	8BD
6AT6	579	7BT
6AU6	579	7BK
6AV6	579	7BT
6B4G	567	5S
6B5	569	6AS
6B6G	567	7V
6B7	570	7D
6B8	565	8E
6BA6	579	7CC
6BA7	579	8CT
6BD6	579	7CC
6BE6	579	7CH
6BF6	579	7BT
6BG6	567	5BT
6BH6	579	7CM
6BJ6	579	7CM
6C4	579	6BG
6C4	588	6BG
6C5	565	6Q
6C6	570	6F
6C7	570	7G
6C8G	567	8G

Type	Page	Base
6D4	582	5AY
6D6	570	6F
6D7	570	7H
6D8G	567	8A
6E5	570	6R
6E6	570	7B
6E7	570	7H
6E8G	567	8O
6F4	576	7BR
6F4	588	7BR
6F5	565	5M
6F6	599	7S
6F6	565	7S
6F7	570	7E
6F8G	567	8G
6G5	570	6R
6G6G	567	7S
6H4GT	567	5AF
6H5	570	6R
6H6	565	7Q
6H8G	567	8E
6J4	579	7BQ
6J5	565	6Q
6J6	579	7BF
6J6	588	7BF
6J7	565	7R
6J8G	567	8H
6K4	580	—
6K5GT	567	5U
6K6GT	567	7S
6K7	565	7R
6K8	565	8K
6L5G	567	6Q
6L6	565	7AC
6L6	600	7AC
6L6GX	600	7AC
6L7	565	7T
6M6G	567	7S
6M7G	567	7R
6M8GT	567	8AU
6N4	579	7CA
6N4	588	7CA
6N5	569	6R
6N6G	567	7AU
6N7	565	8B
6N7	588	8B
6P5GT	567	6Q
6P7G	567	7U
6P8G	567	8K
6Q5G	582	6Q
6Q6G	567	6Y
6Q7	565	7V
6R6G	567	6AW
6R7	565	7V
6S6GT	567	5AK
6S7	565	7R
6S8GT	567	8CB
6SA7	565	8R
6SB7Y	570	8R
6SC7	565	8S
6SD7GT	567	8N
6SE7GT	567	8N
6SF5	565	6AB
6SF7	565	7AZ
6SG7	565	8BK
6SH7	566	8BK
6SH7L	567	8BK
6SJ7	566	8N
6SJ7Y	566	8N
6SK7	566	8N
6SL7GT	567	8BD
6SN7GT	567	8BD
6SQ7	566	8Q
6SR7	566	8Q
6SS7	566	8N
6ST7	566	8Q
6SU7GTY	567	8BD
6SV7	566	7AZ
6SZ7	566	8Q
6T5	570	6R
6T6GM	568	6Z
6T7	566	7V
6T8	579	9E
6U5	570	6R
6U6GT	568	7AC
6U7G	568	7R
6V6	566	7AC
6V6GT	599	7AC
6V7G	568	7V
6W4GT	586	4CG
6W5G	586	6S
6W6GT	568	7AC
6W7G	568	7R
6X4	586	7CF
6X5	586	6S
6X6G	568	7AL
6Y3G	586	4AC
6Y5	586	6J
6Y6G	568	7AC
6Y7G	568	8B
6Z3	586	4G
6Z4	587	5D
6Z5	586	6K
6Z7G	568	8B
6ZY5G	586	6S

Type	Page	Base
7A4	568	5AC
7A5	568	6AA
A6	568	7AJ
7A7	568	8V
7A8	568	8U
7AB7	577	8BO
7AD7	568	8V
7AF7	568	8AC
7AG7	568	8V
7AH7	568	8V
7AP4	584	5AJ
7B4	568	5AC
7B5	568	6AE
7B6	568	8W
7B7	568	8V
7B8	568	8X
7BP1	584	5AN
7C4	577	4AH
7C5	568	6AA
7C6	568	8W
7C7	568	8V
7CP1	584	6AZ
7D7	568	8AR
7DP4	584	Fig. 46
7E5	577	8BN
7E6	569	8W
7E7	569	8AE
7EP4	584	11N
7F7	569	8AC
7F8	569	8BW
7G7	569	8V
7G8	569	8BV
7GP4	584	Fig. 47
7H7	569	8V
7J7	569	8AR
7JP4	584	14G
7K7	569	8BF
7L7	569	8V
7N7	569	8AC
7Q7	569	8AL
7R7	569	8AE
7S7	569	8BL
7T7	569	8V
7V7	569	8V
7W7	569	8BJ
7X7	569	8BZ
7Y4	586	5AB
7Z4	586	5AB
9AP4	584	6AL
9CP4	584	4AF
9JP1	584	8BR
10	589	4D
10	576	4D
10BP4	585	12D
10EP4	585	12D
10FP4	585	12D
10Y	588	4D
11/12	576	4F
12A5	574	7F
12A6	574	7AC
12A7	574	7K
12A7	586	7K
12A8GT	574	8A
12AH7GT	574	8BE
12AL5	579	6BT
12AP4	585	6AL
12AT6	579	7BT
12AT7	579	9A
12AU6	579	7CC
12AU7	579	9A
12AU7	588	9A
12AV6	579	7BT
12AW6	579	7CM
12AW7	579	7CM
12AX7	579	9A
12B6M	574	6Y
12B7	574	8V
12B7ML	574	8V
12B8GT	574	8T
12BA6	580	7CC
12BA7	580	8CT
12BD6	580	7CC
12BE6	580	7CH
12BF6	580	7BT
12C8	574	8E
12CP4	585	4AF
12DP4	585	5AN
12E5GT	574	6Q
12F5GT	574	5M
12G7G	574	7V
12H6	574	7Q
12J5GT	574	6Q
12J7GT	574	7R
12K7GT	574	7R
12K8	574	8K
12L8GT	574	8BU
12Q7GT	574	7V
12S8GT	574	Fig. 34
12SA7	574	8R
12SC7	574	8S
12SF5	574	6AB
12SF7	574	7AZ
12SG7	574	8BK
12SH7	574	8BK
12SJ7	574	8N

Type	Page	Base
12SK7	574	8N
12SL7GT	574	8BD
12SN7GT	574	8BD
12SQ7	574	8Q
12SR7	574	8Q
12SW7	574	8Q
12SX7	574	8BD
12SY7	574	8R
12Z3	586	4G
12Z5	586	7L
14A4	574	5AC
14A5	574	6AA
14A7	574	8V
14AF7	574	8AC
14AF7	576	8AC
14B6	574	8W
14B8	574	8X
14C5	574	6AA
14C7	575	8V
14E6	575	8W
14E7	575	8AE
14F7	575	8AC
14F8	575	8BW
14H7	575	8V
14J7	575	8BL
14N7	575	8AC
14Q7	575	8AL
14R7	575	8AE
14S7	575	8BL
14V7	575	8V
14W7	575	8BJ
14Y4	586	5AB
14Z3	586	4G
15	572	5F
15AP4	585	12D
15E	589	T-4AF
17	582	3G
18	575	6B
19	572	6C
19J6	580	7BF
19T8	580	9E
20BP4	585	12D
20	576	4D
20J8GM	575	8H
21A7	575	8AR
22	576	4K
24-A	571	5E
24-G	589	2D
24XH	585	Fig. 1
25A6	575	7S
25A7GT	575	8F
25A7GT	586	8F
25AC5G	575	6Q
25B5	575	6D
25B6G	575	7S
25B8GT	575	8T
25C6G	575	7AC
25D8GT	575	8AF
25L6	575	7AC
25N6G	575	7W
25S	571	6M
25T	589	3G
25X6GT	586	7Q
25Y4GT	586	5AA
25Y5	586	6E
25Z3	586	4G
25Z4	586	5AA
25Z5	586	6E
25Z6	586	7Q
26	576	4D
26A6	580	7BK
26A7GT	575	8BU
26C6	580	7BT
26D6	580	7CH
27	571	5A
28D7	576	8BS
28Z5	586	5AB
30	572	4D
31	572	4D
32	572	4K
32L7GT	575	8Z
32L7GT	586	8Z
33	572	5K
34	572	4M
35/51	571	5E
35A5	575	6AA
35B5	580	7BZ
35C5	580	7CV
35L6G	575	7AC
35T	591	3G
35TG	591	2D
35W4	586	5BQ
35Y4	586	5AL
35Z3	586	4Z
35Z4GT	586	5AA
35Z5G	586	6AD
35Z6G	586	7Q
36	570	5E
37	570	5A
38	570	5F
39/44	570	5F
40	576	4D
40Z5GT	587	6AD
41	570	6B
42	570	6B
43	575	6B
45	571	4D
45Z3	587	5M
45Z5GT	587	6AD
46	571	5C
47	571	5B
48	575	6A
49	572	5C
50	576	4D
50A5	575	6AA
50B5	580	7BZ

Type	Page	Base
50C5	580	7CV
50C6GT	575	7AC
50L6GT	575	7AC
50T	593	2D
50X6	587	7AJ
50Y6GT	587	7Q
50Y7GT	587	8AN
50Z6G	587	7Q
50Z7G	587	8AN
51	571	5E
52	570	5C
53	571	7B
53A	590	T-4B
55	571	6G
56	571	5A
56AS	570	5A
57	571	6F
57AS	570	6F
58	571	6F
58AS	570	6F
59	571	7A
70A7GT	575	8AB
70A7GT	587	8AB
70L7GT	575	8AA
70L7GT	587	8AA
71-A	577	4D
72	587	4P
73	587	4Y
75	570	6G
75TH	593	2D
75TL	593	2D
76	570	5A
77	570	6F
78	570	6F
79	570	6H
80	587	4C
81	587	4B
82	587	4C
83	587	4C
83-V	587	4AD
84/6Z4	587	5D
85	570	6G
85AS	570	6G
89	570	6F
99	577	4D
100TH	594	2D
100TL	594	2D
111H	593	2D
112-A	577	4D
117L7GT	575	8AO
117L7GT	587	8AO
117M7GT	575	8AO
117M7GT	587	8AO
117N7GT	575	8AV
117N7GT	587	8AV
117P7GT	575	8AV
117P7GT	587	8AV
117Z3	587	4BR
117Z4GT	587	5AA
117Z6GT	587	7Q
128AS	583	5A
150T	596	2N
152TH	596	4BC
152TL	596	4BC
182-B	577	4D
183	577	4D
203-A	594	4E
203-H	594	3N
204-A	598	T-1A
205D	588	4D
211	594	4E
212-E	598	T-2A
217A	587	4AT
217C	587	4AT
227A	594	T-4B
241B	598	T-2AA
242A	593	4E
242B	594	4E
242C	595	4E
249B	587	Fig. 53
250TH	597	2N
250TL	597	2N
254A	600	T-4C
254B	601	T-4C
261A	595	4E
270A	598	T-1A
276A	595	4E
282A	602	T-4C
284B	595	3N
284D	593	4E
295A	595	4E
300T	598	2N
303A	594	4E
304A	598	T-1A
304B	591	2D
304TH	598	4BC
304TL	598	4BC
305A	602	T-4CE
306A	600	T-5CB
307A	600	T-5C
308B	598	T-2A
310	589	4D
311	594	4E
311CH	595	Fig. 57
312A	602	T-6C
312E	598	T-2AA
316A	590	—
327A	594	T-4AD
327B	593	T-4AD
342B	594	4E
356A	592	T-4BD
361A	595	4E
376A	595	4E
410R	604	Fig. 58
482B	577	4D
483	577	4D

Type	Page	Base
527	598	T-4B
485	577	5A
559	577	Fig. 18
703A	589	—
705A	587	T-3AA
707B	604	Fig. 61
715B	601	—
717A	568	8BK
723AB	604	Fig. 60
756	591	4D
800	590	2D
801	589	4D
801A	589	4D
802	599	T-7C
803	603	5J
804	602	T-5C
805	596	3N
806	597	2N
807	600	5AW
808	592	2D
809	590	3G
810	596	2N
811	592	3G
812	592	3G
812A	592	3G
812H	593	3G
813	603	Fig. 28
814	602	T-5D
815	600	8BY
816	587	4P
822	597	3N
822S	597	2N
826	592	T-9A
828	602	5J
829	601	7BP
829A	601	7BP
829B	601	7BP
830	591	4D
830B	592	3G
831	598	T-1AA
832	600	7BP
832A	600	7BP
833A	598	T-1AB
834	592	2D
835	594	4E
836	587	4P
837	599	T-7C
838	595	4E
840	572	5J
841	589	4D
841A	592	3G
841SW	592	3G
843	589	5A
844	600	5AW
849	598	T-1A
850	603	T-3B
852	595	2D
860	603	T-4CB
861	603	T-1B
864	577	4D
865	600	T-4C
866	587	4P
866A	587	4P
866B	587	4P
866Jr	587	4B
871	587	4P
872	587	4AT
872A	587	4AT
874	582	4S
876	582	—
878	587	4P
879	587	4P
884	582	6Q
885	582	5A
886	582	—
902	585	Fig. 1
903	585	6AL
904	585	Fig. 3
905	585	Fig. 6
906P1	583	7AN
907	585	Fig. 6
908	585	7AN
908A	585	7CE
909	585	Fig. 6
910	585	7AN
911	585	7AN
912	585	Fig. 8
913	585	Fig. 1
914	585	Fig. 12
930B	592	3G
938	595	4E
950	572	5K
951	571	4M
954	577	5BB
955	577	5BC
955	588	5BC
956	577	5BB
957	577	5BD
958	577	5BD
958A	577	5BD
958A	588	5BD
959	577	5BE
967	582	3G
975A	587	4AT
991	582	—
1003	587	4R
1005	587	5AQ
1006	587	4C
1201	577	8BN
1203	577	4AH
1204	577	8BO
1206	569	8BV
1221	570	6F
1223	568	7R
1229	572	4K
1230	572	4D

Type	Page	Base
1231	569	8V
1232	569	8V
1247	581	—
1265	582	4AJ
1266	582	4AJ
1267	582	4V
1273	569	8V
1274	587	6S
1275	587	4C
1276	577	4D
1280	576	8V
1284	576	8V
1291	573	7BE
1293	573	4AA
1294	573	4AH
1299	573	6BB
1602	589	4D
1603	570	6F
1608	589	4D
1609	577	5B
1610	599	T-5CA
1611	566	7S
1612	566	7T
1613	599	7S
1614	600	7AC
1616	587	4P
1619	600	T-9H
1620	566	7R
1621	566	7S
1622	566	7AC
1623	590	3G
1624	601	T-5DC
1625	601	5AZ
1626	588	6Q
1627	596	2N
1628	590	T-4BB
1629	576	6RA
1631	576	7AC
1632	576	7AC
1633	576	8BD
1634	576	8S
1635	568	8B
1641	587	T-4AG
1642	569	7BH
1644	576	Fig. 7
1654	587	Fig. 41
1800	585	6AL
1801	585	Fig. 13
1802P1	584	11A
1803P4	585	6AL
1804P4	584	6AL
1805P1	584	11A
1806P1	583	11A
1809P1	584	8BR
1811P1	584	6AZ
1851	566	7R
1852	565	8N
1853	565	8N
2001	585	Fig. 2
2002	585	Fig. 1
2005	585	Fig. 1
2050	582	8BA
2051	583	8BA
24XH	585	Fig. 2
2523N/128AS	583	5A
5514	592	4BO
5516	600	7CL
5517	587	5BU
5556	588	4D
5562	601	Fig. 54
5590	580	7BD
5591	580	7BD
5618	599	7CU
5633	581	—
5634	581	—
5637	581	—
5638	581	—
5640	581	—
5641	582	—
5642	582	—
5645	582	—
5651	583	5BO
5679	569	7CX
5691	568	8BD
5692	568	8BD
5693	566	8N
7000	568	7R
7193	588	4AM
7700	570	6F
8000	597	2N
8001	602	T-7CB
8003	595	3N
8005	594	3G
8008	587	Fig. 11
8010-R	591	—
8012	590	T-4BB
8013-A	587	4P
8016	587	4AC
8020	587	4P
8025	590	4AQ
9001	580	7PM
9002	580	7TM
9003	580	7PM
9004	577	4BJ
9005	577	5BG
9006	580	6BH
AT-340	603	5BK
BA	586	4J
BH	586	4J
BR	586	4H
CE220	586	4P
CK501	581	—
CK502	581	—
CK503	581	—
CK504	581	—
CK505	581	—

Type	Page	Base
CK506	581	—
CK507	581	—
CK509	581	—
CK510	581	—
CK512	581	—
CK515BX	581	—
CK520AX	581	—
CK521AX	581	—
CK522AX	581	—
CK551AXA	581	—
CK553AXA	581	—
CK556AX	581	—
CK568AX	581	—
CK569AX	581	—
CK650AX	581	—
CK606BX	581	—
CK608CX	581	—
CK619CX	581	—
CK1005	587	5AQ
CK1006	587	4C
CK1007	587	T-9G
CK1009	587	—
DR-3B27	586	4B
DR123C	595	Fig. 26
DR200	596	2N
EF50	577	9C
F123A	595	Fig. 26
F127A	597	Fig. 26
G84	586	4B
GL2C44	577	Fig. 17
GL2C44	588	Fig. 17
GL5C24	597	Fig. 26
GL5D24	603	5BK
GL146	595	T-4BG
GL152	596	T-4BG
GL159	598	T-4BG
GL169	598	T-4BG
GL446A	577	Fig. 19
GL446A	588	Fig. 19
GL446B	588	Fig. 19
GL446B	577	Fig. 19
GL464A	588	Fig. 17
GL464A	577	Fig. 17
GL559	577	Fig. 18
GL592	597	Fig. 52
GL8012A	590	T-4BB
HD203A	596	3N
HF60	593	2D
HF75	593	2D
HF100	593	2D
HF120	594	4F
HF125	594	—
HF130	595	—
HF140	594	4F
HF150	595	—
HF175	595	T-3AC
HF200	596	2N
HF250	596	2N
HF300	597	2N
HK24	590	3G
HK54	591	2D
HK154	591	2D
HK158	591	2D
HK252L	596	4BC
HK253	587	4AT
HK254	594	2N
HK257	602	7BM
HK257B	602	7BM
HK304L	598	4BC
HK354	596	2N
HK354C	596	2N
HK354D	596	2N
HK354E	596	2N
HK354F	596	2N
HK454H	598	2N
HK454L	598	2N
HK654	598	2N
HV12	597	3N
HV18	596	2N
HV27	597	3N
HY6J5GTX	588	6Q
HY6L6GTX	600	7AC
HY6V6GTX	599	7AC
HY24	588	4D
HY25	590	3G
HY30Z	590	4BO
HY31Z	590	T-4D
HY40	590	3G
HY40Z	591	3G
HY51A	592	3G
HY51B	592	3G
HY51Z	592	4BO
HY57	591	3G
HY60	599	5AW
HY61	600	5AW
HY63	599	T-8DB
HY65	600	T-8DB
HY67	602	T-5DB
HY69	601	T-5D
HY75	589	2T
HY75A	589	2T
HY113	581	5K
HY114B	588	2T
HY115	581	5K
HY123	581	5K
HY125	581	5K
HY145	581	5K
HY155	581	5K
HY615	588	T-8AG
HY801A	589	4D
HY866Jr	587	4P
HY1231Z	590	T-4D
HY1269	601	T-5DB
HYE1148	588	T-8AG
KY21	583	—
KY866	583	Fig. 8

VACUUM-TUBE BASE DIAGRAMS

The diagrams on the following pages show standard socket connections corresponding to the base designations given in the column headed "Socket Connections" in the classified tube-data tables. Bottom views are shown throughout. Terminal designations are as follows:

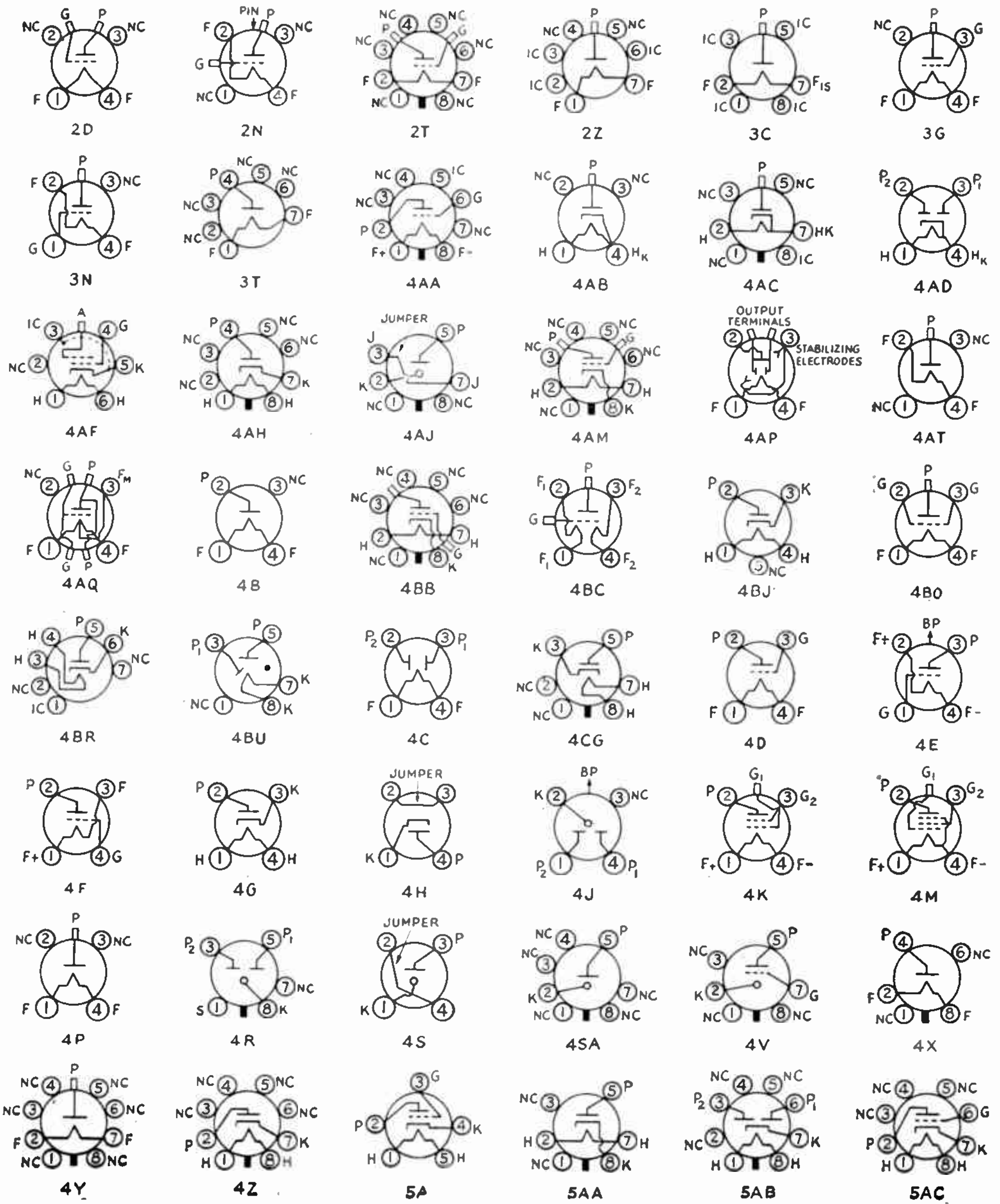
A = Anode	F = Filament	IS = Internal Shield	P_{BF} = Beam-Forming Plates
B = Beam	G = Grid	K = Cathode	RC = Ray-Control Electrode
BP = Bayonet Pin	H = Heater	NC = No Connection	Ref = Reflector or repeller
BS = Base sleeve	IC = Internal Connection	P = Plate (Anode)	S = Shell
D = Deflecting Plate		P_1 = Starter-Anode	TA = Target
			● = Gas-Type Tube
			U = Unit

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.

Generally when the No. 1 pin of a metal-type tube in Table I, with the exception of all triodes, is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

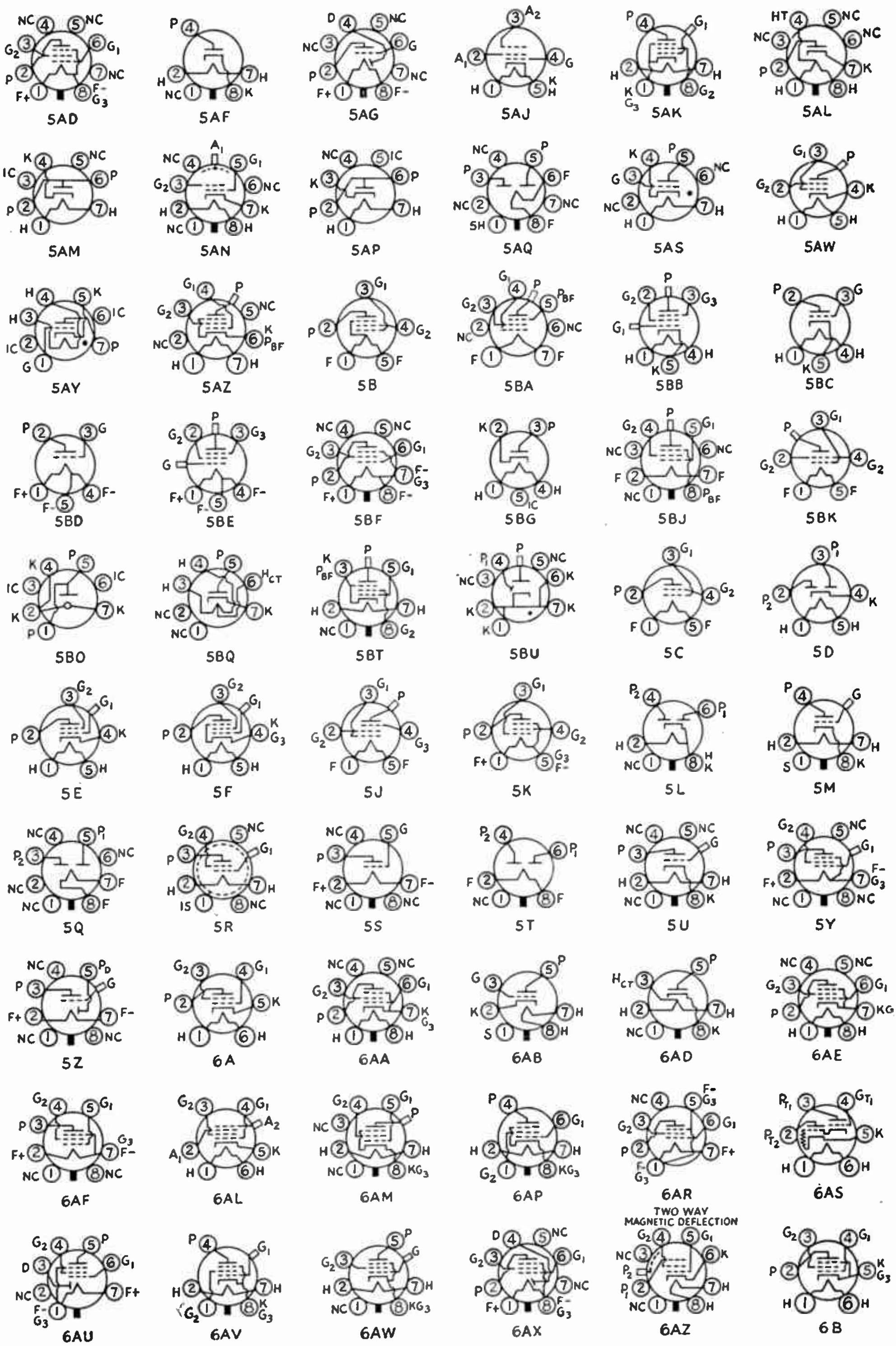
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are shown above.



R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



5AD 5AF 5AG 5AJ 5AK 5AL

5AM 5AN 5AP 5AQ 5AS 5AW

5AY 5AZ 5B 5BA 5BB 5BC

5BD 5BE 5BF 5BG 5BJ 5BK

5BO 5BQ 5BT 5BU 5C 5D

5E 5F 5J 5K 5L 5M

5Q 5R 5S 5T 5U 5Y

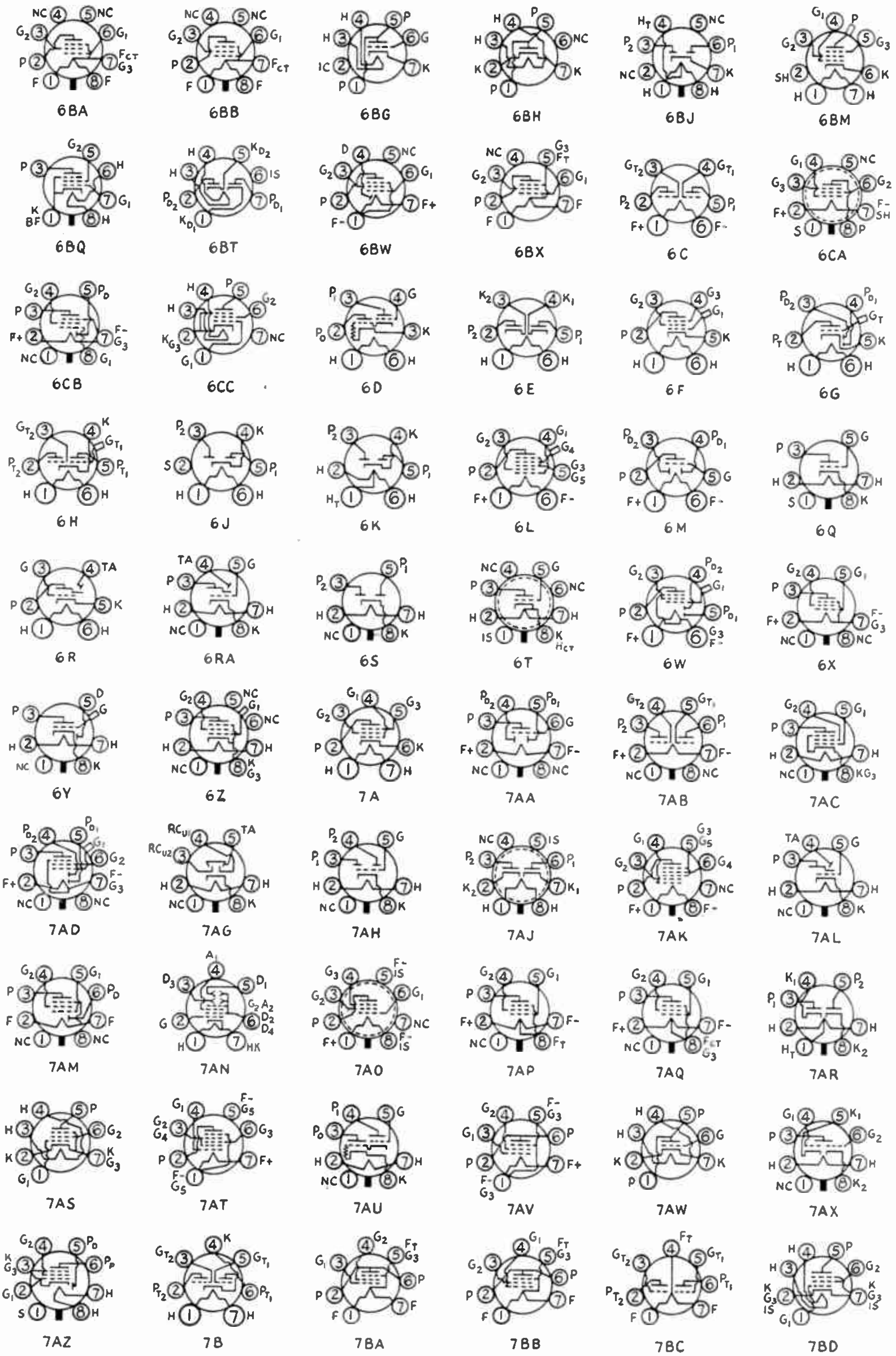
5Z 6A 6AA 6AB 6AD 6AE

6AF 6AL 6AM 6AP 6AR 6AS

6AU 6AV 6AW 6AX 6AZ (TWO WAY MAGNETIC DEFLECTION) 6B

R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



6BA 6BB 6BG 6BH 6BJ 6BM

6BQ 6BT 6BW 6BX 6C 6CA

6CB 6CC 6D 6E 6F 6G

6H 6J 6K 6L 6M 6Q

6R 6RA 6S 6T 6W 6X

6Y 6Z 7A 7AA 7AB 7AC

7AD 7AG 7AH 7AJ 7AK 7AL

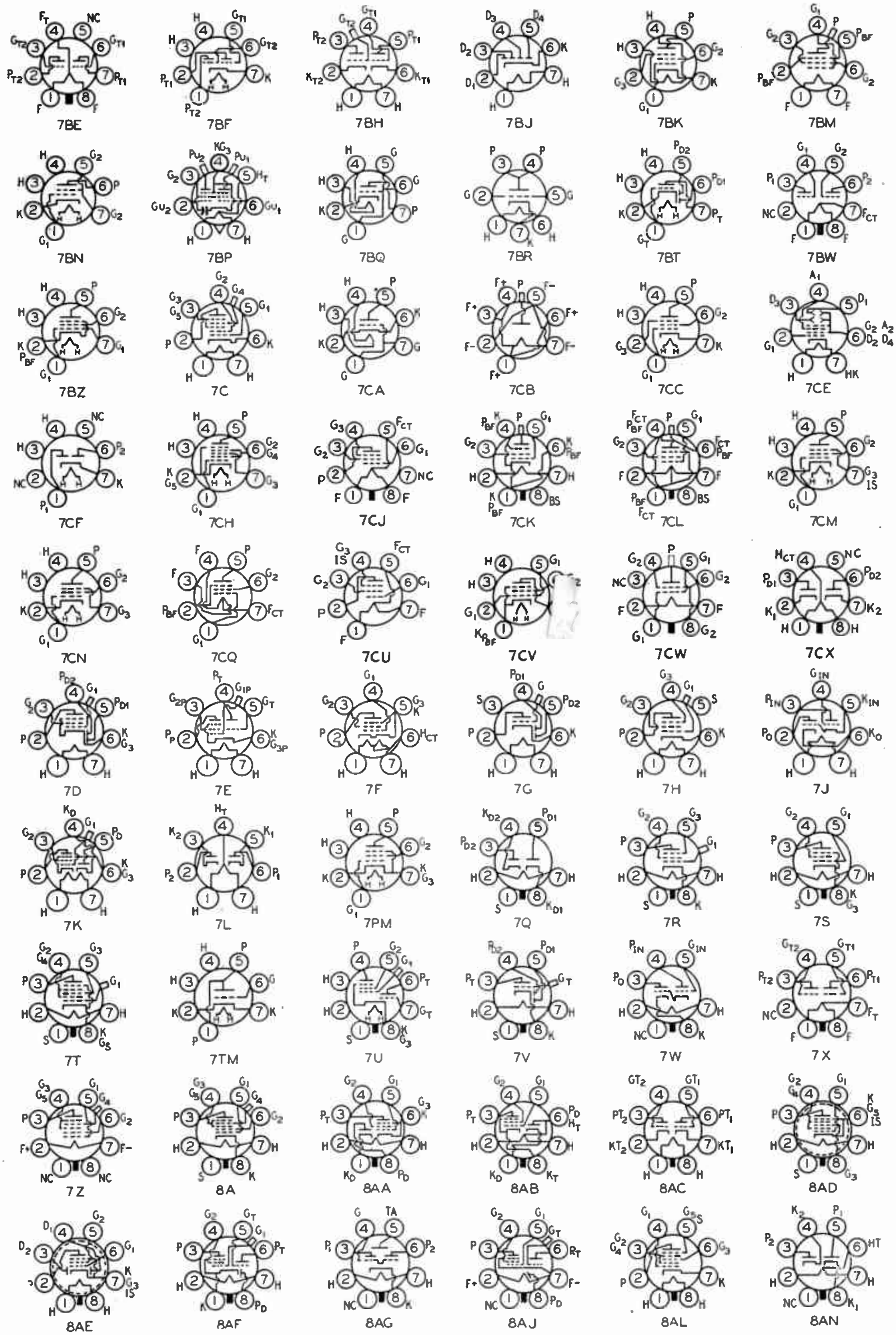
7AM 7AN 7AO 7AP 7AQ 7AR

7AS 7AT 7AU 7AV 7AW 7AX

7AZ 7B 7BA 7BB 7BC 7BD

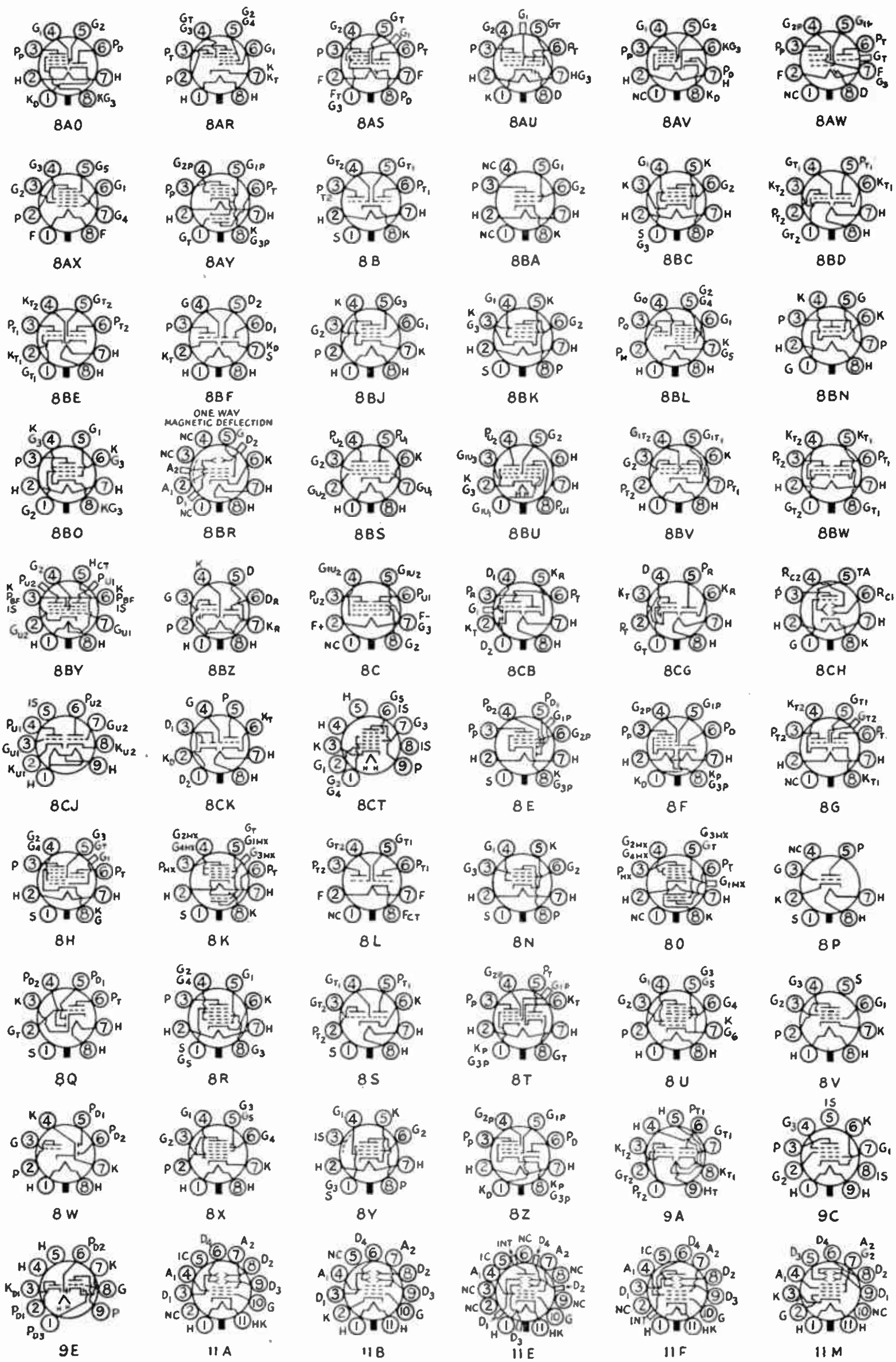
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



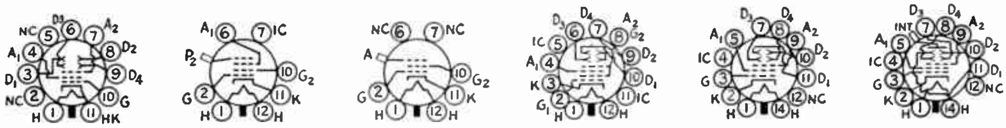
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.

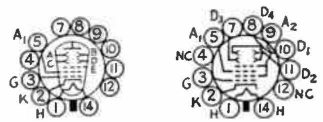


R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



11 N 12 C 12 D 12 E 14 A 14 B



14 F 14 G

SUPPLEMENTARY BASE DIAGRAMS

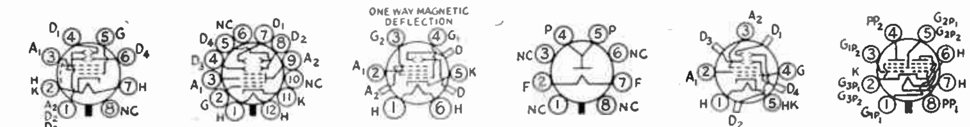


FIG. 1 FIG. 2 FIG. 3 FIG. 4 FIG. 6 FIG. 7

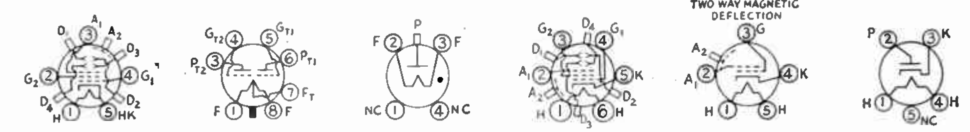


FIG. 8 FIG. 9 FIG. 11 FIG. 12 FIG. 13 FIG. 15

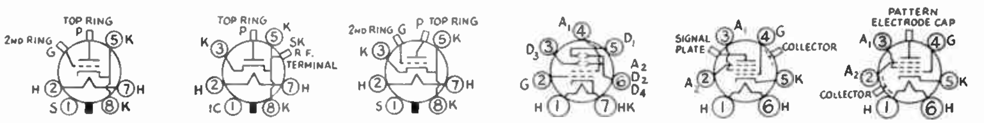


FIG. 17 FIG. 18 FIG. 19 FIG. 20 FIG. 21 FIG. 22

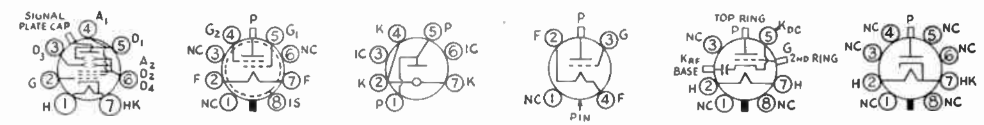


FIG. 23 FIG. 24 FIG. 25 FIG. 26 FIG. 30 FIG. 31

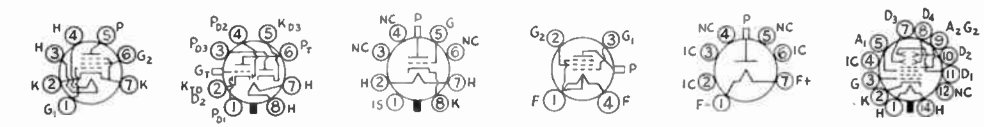


FIG. 32 FIG. 34 FIG. 38 FIG. 40 FIG. 41 FIG. 47

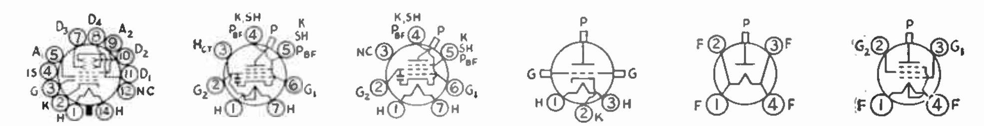


FIG. 49 FIG. 50 FIG. 51 FIG. 52 FIG. 53 FIG. 54

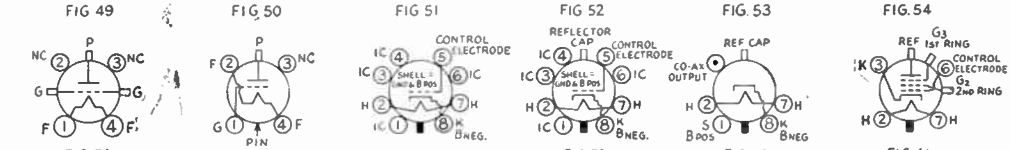


FIG. 56 FIG. 57 FIG. 58 FIG. 59 FIG. 60 FIG. 61

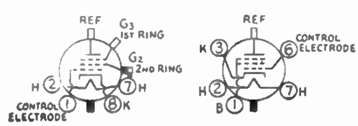


FIG. 62 FIG. 63

SUPPLEMENTARY "T"-GROUP BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.

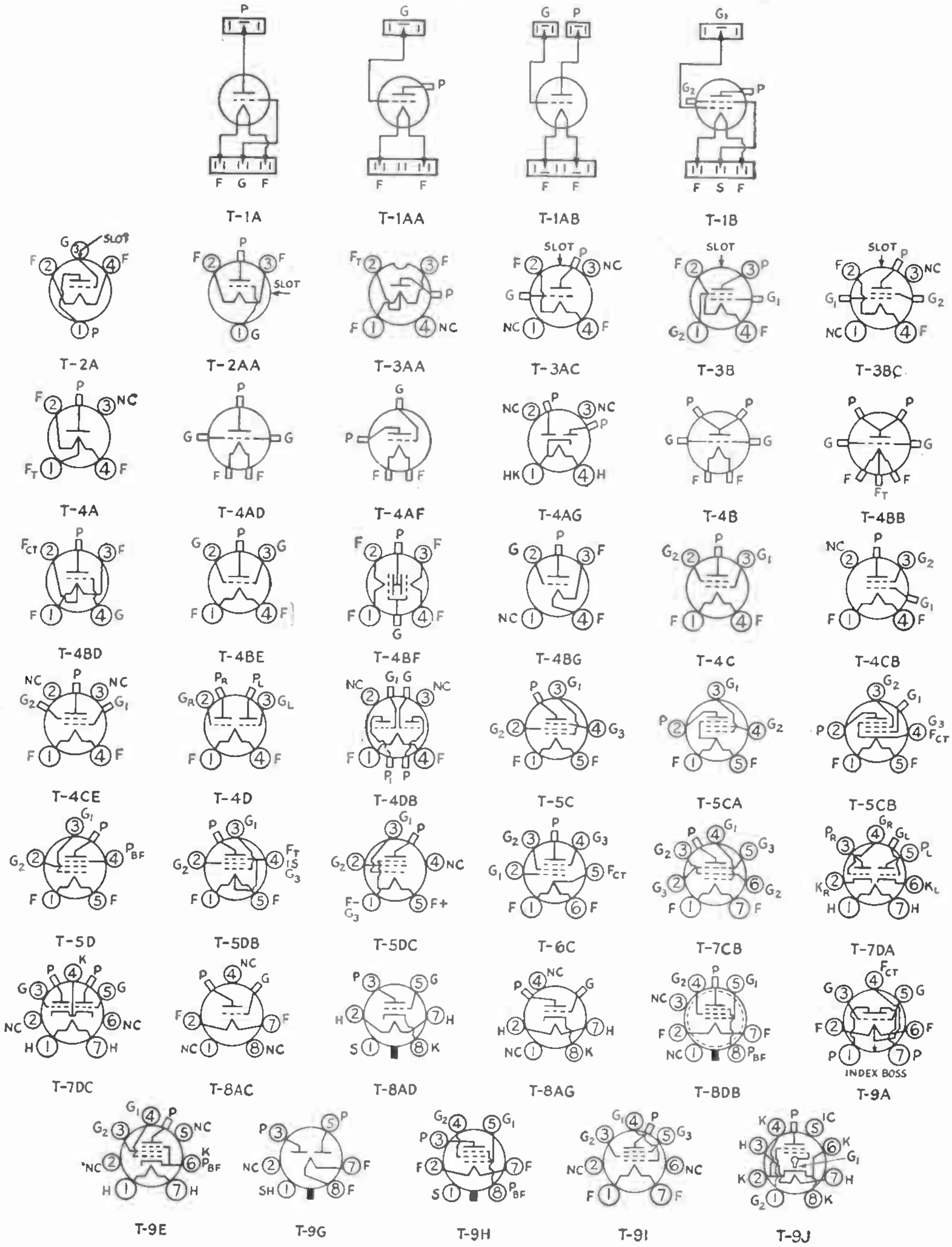


TABLE I—METAL RECEIVING TUBES

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	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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	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	6.3	0.45	8	5	0.015	Class-A Amp.	300	− 3.0	200	3.2	12.5	700000	5000	3500	—	—	6AB7 1853
6AC7 1852	Television Amp. Pentode	8N	6.3	0.45	11	5	0.015	Class-A Amp.	300	160*	150	2.5	10	1000000	9000	6750	—	—	6AC7 1852
6AG7	Sharp Cut-off Pentode	8Y	6.3	0.65	13	7.5	0.06	Class-A$_1$ Amp.	300	− 3.0	150	7/9	30/30.5	130000	11000	—	10000	3.0	6AG7
6AJ7	Sharp Cut-off Pentode	8N	6.3	0.45	—	—	—	Class-A Amp.	300	160*	300	2.5	10	1000000	9000	—	—	—	6AJ7
6AK7	Pentode Power Amp.	8Y	6.3	0.65	13	7.5	0.06	Class-A Amp.	300	− 3	150	7	30	130000	11000	—	10000	3.0	6AK7
6B8	Duplex-Diode Pentode	8E	6.3	0.3	6	9	0.005	Class-A Amp.	250	− 3.0	125	2.3	9.0	650000	1125	730	—	—	6B8
6C5	Triode	6Q	6.3	0.3	3	11	2	Class-A Amp.	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	6.3	0.3	5.5	4	2.3	Class-A Amp.	250	− 1.3	—	—	0.2	66000	1500	100	—	—	6F5
6F6	Pentode Power Amplifier	7S	6.3	0.7	6.5	13	0.2	Class-A$_1$ Pent.[5]	250 315	−16.5 −22.0	250 315	6.5 8.0	36 [7] 42	80000 75000	2500 2650	200 200	7000 7000	3.2 5.0	6F6
								Triode Amp.[1]	250	−20.0	—	—	34 [7]	2600	2600	6.8	4000	0.85	
								Class-AB$_2$ Amp.[5] Class-AB$_2$ Amp.[5]	375 350	340* −38.0	250 —	18 [7] —	77 [7] 22.5	Power output for 2 tubes at stated load, plate-to-plate			10000 [8] 6000 [8]	19.0 18.0	
								Class-AB$_2$ Amp.[1, 6]	350 350	730* −38	— —	— —	50/61 48/92	—	—	—	10000 [8] 6000 [8]	9 13	
6H6	Twin Diode	7Q	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	Triode	6Q	6.3	0.3	3.4	3.6	3.4	Class-A Amp.	250	− 8.0	—	—	9	7700	2600	20	—	—	6J5
6J7	Sharp Cut-off Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	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	Variable-μ Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	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	8K	6.3	0.3	—	—	—	Converter	250	− 3.0	100	6	2.5	Triode Plate (No. 2) 100 volts, 3.8 ma.					6K8
6L6	Beam Power Amplifier	7AC	6.3	0.9	10	12	0.4	Single Tube Class A$_1$	250 300	170* 220*	250 200	5.4/7.2 3.0/4.6	75/78 51/54.5	—	—	—	2500 4500	6.5 6.5	6L6
								Single Tube Class A$_1$	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	
								P.P. Class A$_1$[6]	270	125*	270	11/17	134/145	—	—	—	5000 [8]	18.5	
								P.P. Class A$_1$[6]	250 270	−16.0 −17.5	250 270	10/16 11/17	120/140 134/155	24500 23500	5500 5700	—	5000 [8] 5000 [8]	14.5 17.5	
								P.P. Class AB$_1$[5]	360	250*	270	5/17	88/100	Power output for 2 tubes. Load plate-to-plate			9000 [8]	24.5	
								P.P. Class AB$_1$[6]	360	−22.5	270	5/15	88/132				6600 [8]	26.5	
								P.P. Class AB$_2$[6]	360 360	−18.0 −22.5	225 270	3.5/11 5/16	78/142 88/205				6000 [8] 3800 [8]	31.0 47.0	
6L7	Pentagrid Mixer Amplifier	7T	6.3	0.3	—	—	—	R.F. Amp.	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				
6N7	Twin Triode	8B	6.3	0.8	—	—	—	Class-B Amp.	300	0	—	—	35/70	—	—	—	8000	10.0	6N7
6Q7	Duplex-Diode Triode	7V	6.3	0.3	5	3.8	1.4	Triode Amp.	250	− 3.0	—	—	1.1	58000	1200	70	—	—	6Q7
6R7	Duplex-Diode Triode	7V	6.3	0.3	4.8	3.8	2.4	Triode Amp.	250	− 9.0	—	—	9.5	8500	1900	16	10000	0.28	6R7
6S7	Remote Cut-off Pentode	7R	6.3	0.15	6.5	10.5	0.005	Class-A Amp.	250	− 3.0	100	2.0	8.5	1000000	1750	—	—	—	6S7
6SA7	Pentagrid Converter	8R[2]	6.3	0.3	—	—	—	Converter	250	0[3]	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				6SA7
6SC7	Twin-Triode	8S	6.3	0.3	—	—	—	Class-A Amp.	250	− 2.0	—	—	2.0	53000	1325	70	—	—	6SC7
6SF5	High-μ Triode	6AB	6.3	0.3	4	3.6	2.4	Class-A Amp.	250	− 2.0	—	—	0.9	66000	1500	100	—	—	6SF5
6SF7	Diode Variable-μ Pentode	7AZ	6.3	0.3	5.5	6	0.004	Class-A Amp.	250	− 1.0	100	3.3	12.4	700000	2050	—	—	—	6SF7
6SG7	Semivariable-μ Pentode	8BK	6.3	0.3	8.5	7	0.003	H.F. Amp.	250	− 2.5	150	3.4	9.2	Over 1 meg	4000	—	—	—	6SG7

TABLE I—METAL RECEIVING TUBES—*Continued*

Type	Name	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6SH7	Sharp Cut-off Pentode	8BK	6.3	0.3	8.5	7	0.003	Class-A Amp.	250	– 1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7
6SJ7[4]	Sharp Cut-off Pentode	8N	6.3	0.3	6	7	0.005	Class-A Amp.	250	– 3.0	100	0.8	3	1500000	1650	2500	—	—	6SJ7
6SK7	Variable-μ Pentode	8N	6.3	0.3	6	7	0.003	Class-A Amp.	250	– 3.0	100	2.4	9.2	800000	2000	1600	—	—	6SK7
6SQ7	Duplex-Diode Triode	8Q	6.3	0.3	3.2	3.0	1.6	Class-A Amp.	250	– 2.0	—	—	0.8	91000	1100	100	—	—	6SQ7
6SR7	Duplex-Diode Triode	8Q	6.3	0.3	3.6	2.8	2.40	Class-A Amp.	250	– 9.0	—	—	9.5	8500	1900	16	—	—	6SR7
6SS7	Variable-μ Pentode	8N	6.3	0.15	5.5	7.0	0.004	Class-A Amp.	250	– 3.0	100	2.0	9.0	1000000	1850	—	—	—	6SS7
6ST7	Duplex-Diode Triode	8Q	6.3	0.15	2.8	3	1.50	Class-A Amp.	250	– 9.0	—	—	9.5	8500	1900	16	—	—	6ST7
6SV7	Diode R.F. Pentode	7AZ	6.3	0.3	6.5	6	0.004	Class-A Amp.	250	– 1	150	2.8	7.5	800000	3400	—	—	—	6SV7
6SZ7	Duplex-Diode Triode	8Q	6.3	0.15	2.6	2.8	1.10	Class-A Amp.	250	– 3	—	—	1.0	58000	1200	70	—	—	6SZ7
6T7	Duplex-Diode Triode	7V	6.3	0.15	1.8	3.1	1.70	Class-A Amp.	250	– 3.0	—	—	1.2	62000	1050	65	—	—	6T7
6V6	Beam Power Amplifier	7AC	6.3	0.45	2.0	7.5	0.7	Class-A₁ Amp.[5]	250	–12.5	250	4.5/7.0	45/47	52000	4100	218	5000	4.5	6V6
								Class-AB₁ Amp.[6]	250	–15.0	250	5/13	70/79	60000	3750	—	10000[8]	10.0	
									285	–19.0	285	4/13.5	70/92	65000	3600	—	8000[8]	14.0	
1611	Pentode Power Amplifier	7S	6.3	0.7	—	—	—	Audio Amp.	Characteristics same as 6F6										1611
1612	Pentagrid Amplifier	7T	6.3	0.3	7.5	11	0.001	Class-A Amp.	250	– 3.0	100	6.5	5.3	600000	1100	880	—	—	1612
1620	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6J7										1620
1621	Power Amplifier Pentode	7S	6.3	0.7	—	—	—	Class-AB₂ Amp.[6]	300	–30.0	300	6.5/13	38/69	—	—	—	4000[8]	5.0	1621
								Class-A₁ Amp.[6 1]	330	500*	—	—	55/59	—	—	—	5000[8]	2.0	
1622	Beam Power Amplifier	7AC	6.3	0.9	—	—	—	Class-A₁ Amp.	300	–20.0	250	4/10.5	86/125	—	—	—	4000	10.0	1622
1851	Television Amp. Pentode	7R	6.3	0.45	11.5	5.2	0.02	Class-A Amp.	300	– 2.0	150	2.5	10	750000	9000	6750	—	—	1851
5693	Sharp Cut-off Pentode	8N	6.3	0.3	5.3	6.2	0.005	Class-A Amp.	250	– 3	100	0.85	3.0	1000000	1650	—	—	—	5693

* Cathode resistor—ohms.
[1] Screen tied to plate.
[2] For 6SA7GT use base diagram 8AD.
[3] Grid bias—2 volts if separate oscillator excitation is used.
[4] Also Type "6SJ7Y."
[5] Values are for single tube.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Plate-to-plate value.

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 Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2C22	Triode	4AM	6.3	0.3	2.2	0.7	3.60	Class-A Amp.	300	–10.5	—	—	11	6600	3000	20	—	—	2C22
6A5G	Triode Power Amplifier	6T	6.3	1.0	—	—	—	Class-A Amp.[1]	250	–45.0	—	—	60	800		4.2	2500	3.75	6A5G
								P.P. Class AB[5]	325	–68.0	—	—	80	—	5250	—	3000[6]	15.0	
								P.P. Class AB[5]	325	850*			80	—		—	5000[6]	10.0	
6AB6G	Direct-Coupled Amplifier	7AU	6.3	0.5	—	—	—	Class-A Amp.	250	0	Input		5.0	40000	1800	72	8000	3.5	6AB6G
									250	0	Output		34						
6AC5G	High-μ Power-Amplifier Triode	6Q	6.3	0.4	—	—	—	P.P. Class B[5]	250	0	—	—	5.0	36700	3400	125	10000[6]	8.0	6AC5G
								Dyn.-Coupled	250	—	—	—	32				7000	3.7	
6AC6G	Direct-Coupled Amplifier	7AU	6.3	1.1	—	—	—	Class-A Amp.	180	0	Input		7.0	—	3000	54	4000	3.8	6AC6G
									180	0	Output		45						
6AD5G	High-μ Triode	6Q	6.3	0.3	4.1	3.9	3.3	Class-A Amp.	250	– 2.0	—	—	0.9	—	1500	100	—	—	6AD5G
6AD6G[10]	Electron-Ray Tube	7AG	6.3	0.15	—	—	—	Indicator	100		0 for 90°; —23 for 135°; 45 for 0°. Target current 1.5 ma.								6AD6G
6AD7G	Triode-Pentode	8AY	6.3	0.85	—	—	—	Triode Amp.	250	–25.0	—	—	4.0	19000	325	6.0	—	—	6AD7G
								Pentode Amp.	250	–16.5	250	6.5	34	80000	2500	—	7000	3.2	
6AE5G	Triode Amplifier	6Q	6.3	0.3	—	—	—	Class-A Amp.	95	–15.0	—	—	7.0	3500	1200	4.2	—	—	6AE5G
6AE6GT[10]	Twin-Plate Triode with Single Grid	7AH	6.3	0.15	Remote cut-off			Class-A Amp.	250	– 1.5	—	—	6.5	25000	1000	25	—	—	6AE6GT
					Sharp cut-off			Class-A Amp.	250	– 1.5			4.5	35000	950	33			

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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[10]	Twin-Input Triode	7AX	6.3	0.5	—	—	—	Driver Amplifier	250	−13.5	—	—	5.0	9300	1500	14	—	—	6AE7GT
6AF5G	Triode	6Q	6.3	0.3	—	—	—	Class-A Amplifier	180	−18.0	—	—	7.0	—	1500	7.4	—	—	6AF5G
6AF7G	Twin Electron Ray	8AG	6.3	0.3	—	—	—	Indicator Tube											6AF7G
6AG6G	Power-Amplifier Pentode	7S	6.3	1.25	—	—	—	Class-A Amplifier	250	− 6.0	250	6.0	32	—	10000	—	8500	3.75	6AG6G
6AH5G	Beam Power Amplifier	6AP	6.3	0.9	—	—	—	Class-A Amplifier	350	−18	250		—	33000	5200	—	4200	10.8	6AH5G
6AH7GT	Twin Triode	8BE	6.3	0.3	—	—	—	Converter & Amp.	250	− 9.0	—	—	12[1]	6600	2400	16	—	—	6AH7GT
6AL6G	Beam Power Amplifier	6AM	6.3	0.9	—	—	—	Class-A Amplifier	250	−14.0	250	5.0	72	22500	6000	—	2500	6.5	6AL6G
6AL7GT	Electron-Ray Tube	8CH	6.3	0.15	—	—	—	Indicator	Outer edge of any of the three illuminated areas displaced 1/16 in. min. outward with +5 volts to its electrode. Similar inward disp. with —5 volts. No pattern with —6 volts grid.										6AL7GT
6AQ7GT	Duplex Diode Triode	8CK	6.3	0.3	2.3	1.5	2.8	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6AQ7GT
6AR6	Beam Power Amp.	6BQ	6.3	1.2	11	7	0.55	Class-A Amplifier	250	−22.5	250	5	77	21000	5400	95	—	—	6AR6
6AR7GT	Diode Triode	8CG	6.3	0.3	1.4	1	2	Class-A Amplifier	250	− 2	—	—	1.3	66500	1050	70	—	—	6AR7GT
6AS7G	Low-Mu Twin Triode	8BD	6.3	2.5	—	—	—	D.C. Amplifier	135	250*	—	—	125	280	7500	2.1	—	—	6AS7G
								Class-A₁ Amp. P.P.	250	2500*	—	—	100/106	280	225[9]	—	6000[6]	13	
6B4G	Triode Power Amplifier	5S	6.3	1.0	—	—	—	Power Amplifier		Characteristics same as Type 6A3—Table IV						—	—	—	6B4G
6B6G	Duplex-Diode High-μ Triode	7V	6.3	0.3	1.7	3.8	1.7	Detector-Amplifier		Characteristics same as Type 75—Table IV						—	—	—	6B6G
6BG6	Beam Power Amplifier	5BT	6.3	0.9	11	6.5	0.5	Deflection Amp.	400	−50	350	6.0	70	—	6000	—	—	—	6BG6
6C8G	Twin Triode	8G	6.3	0.3	—	—	—	Amp. 1 Section	250	− 4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6D8G	Pentagrid Converter	8A	6.3	0.15	—	—	—	Converter	250	− 3.0	100	Cathode current 13.0 Ma.			Anode grid (No. 2) Volts = 250[3]				6D8G
6E8G[10]	Triode-Hexode Converter	8O	6.3	0.3	—	—	—	Converter	250	− 2.0			Triode Plate 150 volts						6E8G
6F8G	Twin Triode	8G	6.3	0.6	—	—	—	Amplifier	250	− 8.0	—	—	9[1]	7700	2600	20	—	—	6F8G
6G6G	Pentode Power Amplifier	7S	6.3	0.15	—	—	—	Class-A Amplifier	180	− 9.0	180	2.5	15	175000	2300	400	10000	1.1	6G6G
								Class-A Amplifier[2]	180	−12.0	—	—	—	4750	2000	9.5	12000	0.25	
6H4GT	Diode Rectifier	5AF	6.3	0.15	—	—	—	Detector	100	—	—	—	4.0	—	—	—	—	—	6H4GT
6H8G	Duo-Diode High-μ Pentode	8E	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	100	—	8.5	650000	2400	—	—	—	6H8G
6J8G[10]	Triode Heptode	8H	6.3	0.3	—	—	—	Converter	250	− 3.0	100	2.8	1.2	Anode-grid (No. 2) 250 volts max.[3] 5 ma.					6J8G
6K5GT[10]	High-μ Triode	5U	6.3	0.3	2.4	3.6	2.0	Class-A Amplifier	250	− 3.0	—	—	1.1	50000	1400	70	—	—	6K5GT
6K6GT	Pentode Power Amplifier	7S	6.3	0.4	—	—	—	Class-A Amplifier	Characteristics same as Type 41—Table IV										6K6GT
6L5G	Triode Amplifier	6Q	6.3	0.15	2.8	5.0	2.8	Class-A Amplifier	250	− 9.0	—	—	8.0	—	1900	17	—	—	6L5G
6M6G	Power Amplifier Pentode	7S	6.3	1.2	—	—	—	Class-A Amplifier	250	− 6.0	250	4.0	36	—	9500	—	7000	4.4	6M6G
6M7G	Pentode Amplifier	7R	6.3	0.3	—	—	—	R.F. Amplifier	250	− 2.5	125	2.8	10.5	900000	3400	—	—	—	6M7G
6M8GT	Diode Triode Pentode	8AU	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	6.3	0.8	—	—	—	Power Amplifier	Characteristics same as Type 6B5—Table IV							—	—	—	6N6G
6P5GT[10]	Triode Amplifier	6Q	6.3	0.3	3.4	5.5	2.6	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	6P5GT
6P7G[10]	Triode-Pentode	7U	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as 6F7—Table IV										6P7G
6P8G	Triode-Hexode Converter	8K	6.3	0.8	—	—	—	Converter	250	− 2.0	75	1.4	1.5	Triode Plate 100 v. 2.2 ma.					6P8G
6Q6G	Diode-Triode	6Y	6.3	0.15	—	—	—	Class-A Amplifier	250	− 3.0	—	—	1.2	—	1050	65	—	—	6Q6G
6R6G	Pentode Amplifier	6AW	6.3	0.3	4.5	11	0.007	Class-A Amplifier	250	− 3.0	100	1.7	7.0	—	1450	1160	—	—	6R6G
6S6GT	Remote Cut-off Pentode	5AK	6.3	0.45	—	—	—	R.F. Amplifier	250	− 2.0	100	3.0	13	350000	4000	—	—	—	6S6GT
6S8GT	Triple Diode Triode	8CB	6.3	0.3	1.2	5	2	Class-A Amplifier	250	− 2.0	—	—	0.9	91000	1100	100	—	—	6S8GT
6SD7GT	Medium Cut-off Pentode	8M	6.3	0.3	9	7.5	.0035	R.F. Amplifier	250	− 2.0	100	1.9	6.0	1000000	3600	—	—	—	6SD7GT
6SE7GT	Sharp Cut-off Pentode	8N	6.3	0.3	8	7.5	.005	R.F. Amplifier	250	− 1.5	100	1.5	4.5	1100000	3400	3750	—	—	6SE7GT
6SH7L	Pentode R.F. Amp.	8BK	6.3	0.3	—	—	—	Class-A Amplifier	100	− 1.0	100	2.1	5.3	350000	4000	—	—	—	6SH7L
									250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	
6SL7GT	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3[1]	44000	1600	70	—	—	6SL7GT
6SN7GT	Twin Triode	8BD	6.3	0.6	—	—	—	Class-A Amplifier	250	− 8.0	—	—	9.0[1]	7700	2600	20	—	—	6SN7GT
6SU7GTY	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6SU7GTY

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6T6GM	Amplifier	6Z	6.3	0.45	—	—	—	Class-A Amplifier	250	− 1.0	100	2.0	10	1000000	5500	—	—	—	6T6GM
6U6GT	Beam Power Amplifier	7AC	6.3	0.75	—	—	—	Class-A Amplifier	200	−14.0	135	3.0	56	20000	6200	—	3000	5.5	6U6GT
6U7G	Variable-μ Pentode	7R	6.3	0.3	5	9	.007	Class-A Amplifier		Characteristics same as Type 6D6—Table III									6U7G
6V7G[10]	Duplex Diode-Triode	7V	6.3	0.3	2	3.5	1.7	Detector-Amplifier		Characteristics same as Type 85—Table III									6V7G
6W6GT	Beam Power Amplifier	7AC	6.3	1.25	—	—	—	Class-A Amplifier	135	− 9.5	135	12.0	61.0	—	9000	215	2000	3.3	6W6GT
6W7G	Pentode Det. Amplifier	7R	6.3	0.15	5	8.5	.007	Class-A Amplifier	250	− 3.0	100	2.0	0.5	1500000	1225	1850	—	—	6W7G
6X6G	Electron-Ray Tube	7AL	6.3	0.3	—	—	—	Indicator Tube	250	0 v. for 300°, 2 ma. —8 v. for 0°, 0 ma. Vane grid 125 v.									6X6G
6Y6G	Beam Power Amplifier	7AC	6.3	1.25	15	8	0.7	Class-A Amplifier	135	−13.5	135	3.0	60.0	9300	7000	—	2000	3.6	6Y6G
6Y7G[10]	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier		Characteristics same as Type 79—Table IV									6Y7G
6Z7G	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier	180 135	0 0	— —	— —	8.4 6.0	— —	— —	— —	12000 9000	4.2 2.5	6Z7G
717A	Sharp Cut-off Pentode	8BK	6.3	0.175	—	—	—	Class-A Amplifier	120	− 2.0	120	2.5	7.5	390000	4000	—	—	—	717A
1223	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amplifier		Characteristics some as 6C6—Table IV									1223
1635	Twin Triode Amplifier	8B	6.3	0.6	—	—	—	Class-B Amplifier	400	0	—	—	10/63	—	—	—	14000	17	1635
5691	Hi-Mu Twin Triode	8BD	6.3	0.6	2.4[7] 2.7[8]	2.3[7] 2.7[8]	3.6[7] 3.6[8]	Class-A Amp.	250	− 2	—	—	2.3[1]	44000	1600	70	—	—	5691
5692	Medium-Mu Twin Triode	8BD	6.3	0.6	2.3[7] 2.6[8]	2.5[7] 2.7[8]	3.5[7] 3.3[8]	Class-A Amp.	250	− 9	—	—	6.5[1]	9100	2200	18	—	—	5692
7000	Low-Noise Amplifier	7R	6.3	0.3	—	—	—	Class-A Amplifier		Characteristics same as Type 6J7—Table									7000

* Cathode resistor-ohms.
[1] Per plate.
[2] Screen tied to plate.
[3] Through 20,000-ohm dropping resistor.
[4] Values are for single tube.
[5] Values are for two tubes in push-pull.
[6] Plate-to-plate value.
[7] No. 1 triode.
[8] No. 2 triode.
[9] Peak a.f. volts G-G.
[10] Discontinued.

TABLE III—7-VOLT LOCK-IN-BASE TUBES

For other lock-in-base types see Tables VIII, IX, and X

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7A4	Triode Amplifier	5AC	7.0	0.32	3.4	3	4	Class-A Amplifier	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7A4
7A5	Beam Power Amplifier	6AA	7.0	0.75	13	7.2	0.44	Class-A_1 Amplifier	125	− 9.0	125	3.2/8	37.5/40	17000	6100	—	2700	1.9	7A5
7A6	Twin Diode	7AJ	7.0	0.16	—	—	—	Rectifier			Max. A.C. volts per plate—150. Max. Output current—10 ma.								7A6
7A7	Remote Cut-off Pentode	8V	7.0	0.32	6	7	.005	Class-A Amplifier	250	− 3.0	100	2.0	8.6	800000	2000	1600	—	—	7A7
7A8	Multigrid Converter	8U	7.0	0.16	7.5	9.0	0.15	Converter	250	− 3.0	100	3.1	3.0	50000	Anode-grid 250 volts max.[1]				7A8
7AD7	Pentode	8V	6.3	0.6	11.5	7.5	0.03	Class-A_1 Amp.	300	68*	150	7.0	28.0	300000	9500	—	—	—	7AD7
7AF7	Twin Triode	8AC	6.3	0.3	2.2	1.6	2.3	Class-A Amp.	250	−10	—	—	9.0	7600	2100	16	—	—	7AF7
7AG7	Sharp Cut-off Pentode	8V	7.0	0.16	7.0	6.0	0.005	Class-A_1 Amp.	250	250*	250	2.0	6.0	750000	4200	—	—	—	7AG7
7AH7	Pentode Amplifier	8V	6.3	0.15	7.0	6.5	0.005	Class-A_1 Amplifier	250	250*	250	1.9	6.8	1000000	3300	—	—	—	7AH7
7B4	High-μ Triode	5AC	7.0	0.32	3.6	3.4	1.6	Class-A Amplifier	250	− 2.0	—	—	0.9	66000	1500	100	—	—	7B4
7B5	Pentode Power Amplifier	6AE	7.0	0.43	3.2	3.2	1.6	Class-A_1 Amplifier	250	−18.0	250	5.5/10	32/33	68000	2300	—	7600	3.4	7B5
7B6	Duo-Diode Triode	8W	7.0	0.32	3.0	2.4	1.6	Class-A Amplifier	250	− 2.0	—	—	1.0	91000	1100	100	—	—	7B6
7B7	Remote Cut-off Pentode	8V	7.0	0.16	5	7	.005	Class-A Amplifier	250	− 3.0	100	2.0	8.5	700000	1700	1200	—	—	7B7
7B8	Pentagrid Converter	8X	7.0	0.32	10.0	9.0	0.2	Converter	250	− 3.0	100	2.7	3.5	360000	Anode-grid 250 volts max.[1]				7B8
7C5	Tetrode Power Amplifier	6AA	7.0	0.48	9.5	9.0	0.4	Class-A_1 Amplifier	250	−12.5	250	4.5/7	45/47	52000	4100	—	5000	4.5	7C5
7C6	Duo-Diode Triode	8W	7.0	0.16	2.4	3	1.4	Class-A Amplifier	250	− 1.0	—	—	1.3	100000	1000	100	—	—	7C6
7C7	Pentode Amplifier	8V	7.0	0.16	5.5	6.5	.007	Class-A Amplifier	250	− 3.0	100	0.5	2.0	2 meg.	1300	—	—	—	7C7
7D7	Triode-Hexode Converter	8AR	7.0	0.48	—	—	—	Converter	250	− 3.0	Triode Plate (No. 3) 150 v. 3.5 ma.								7D7

TABLE III—7-VOLT LOCK-IN-BASE TUBES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amp.	Capacitance µµfd. In	Capacitance µµfd. Out	Capacitance µµfd. Plate-Grid	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
7E6	Duo-Diode Triode	8W	7.0	0.32	—	—	—	Class-A Amplifier	250	− 9.0	—	—	9.5	8500	1900	16	—	—	7E6
7E7	Duo-Diode Pentode	8AE	7.0	0.32	4.6	4.6	.005	Class-A Amplifier	250	− 3.0	100	1.6	7.5	700000	1300	—	—	—	7E7
7F7	Twin Triode	8AC	7.0	0.32	—	—	—	Class-A Amplifier[2]	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7F7
7F8	Twin Triode	8BW	6.3	0.30	2.8	1.4	1.2	R.F. Amplifier	250 180	− 2.5 − 1.0	—	—	10.0 12.0	10400 8500	5000 7000	—	—	—	7F8
7G7/ 1232	Sharp Cut-off Pentode	8V	7.0	0.48	9	7	.007	Class-A Amplifier	250	− 2.0	100	2.0	6.0	800000	4500	—	—	—	7G7/ 1232
7G8/ 1206	Dual Tetrode	8BV	6.3	0.30	3.4	2.6	0.15	R.F. Amplifier[2]	250	− 2.5	100	0.8	4.5	225000	2100	—	—	—	7G8/ 1206
7H7	Semi-Variable-µ Pentode	8V	7.0	0.32	8	7	.007	R.F. Amplifier	250	− 2.5	150	2.5	9.0	1000000	3500	—	—	—	7H7
7J7	Triode-Heptode Converter	8AR	7.0	0.32	—	—	—	Converter	250	− 3.0	100	2.9	1.3	Triode Plate 250 v. Max.[1]					7J7
7K7	Duo-Diode High-µ Triode	8BF	7.0	0.32	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7K7
7L7	Sharp Cut-off Pentode	8V	7.0	0.32	8	6.5	.01	Class-A Amplifier	250	− 1.5	100	1.5	4.5	100000	3100	Cathode Resistor 250 ohms			7L7
7N7	Twin Triode	8AC	7.0	0.6	3.4[3] 2.9[4]	2.0[3] 2.4[4]	3.0[3] 3.0[4]	Class-A Amplifier[2]	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7N7
7Q7	Pentagrid Converter	8AL	7.0	0.32	—	—	—	Converter	250	0	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				7Q7
7R7	Duo-Diode Pentode	8AE	7.0	0.32	5.6	5.3	.004	Class-A Amplifier	250	− 1.0	100	1.7	5.7	1000000	3200	—	—	—	7R7
7S7	Triode Hexode Converter	8BL	7.0	0.32	—	—	—	Converter	250	− 2.0	100	2.2	1.7	2000000	Triode Plate 250 v. Max.[1]				7S7
7T7	Pentode Amplifier	8V	7.0	0.32	8	7	.005	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	7T7
7V7	Sharp Cut-off Pentode	8V	7.0	0.48	9.5	6.5	.004	Class-A Amplifier	300	160*	150	3.9	10	300000	5800	—	—	—	7V7
7W7	Sharp Cut-off Pentode	8BJ	7.0	0.48	9.5	7.0	.0025	Class-A Amplifier	300	− 2.2	150	3.9	10	300000	5800	—	—	—	7W7
7X7	Duo-Diode Triode	8BZ	6.3	0.3	—	—	—	Class-A Amplifier	250	− 1.0	—	—	1.9	67000	1500	100	—	—	7X7
1231	Pentode Amplifier	8V	6.3	0.45	8.5	6.5	.015	Class-A Amplifier	300	200*	150	2.5	10	700000	5500	3850	—	—	1231
1273	Nonmicrophonic Pentode	8V	7.0	0.32	6.0	6.5	.007	Class-A₁ Amplifier	250 100	− 3.0 − 1.0	100 100	0.7 1.8	2.2 5.7	1000000 400000	1575 2275	— —	— —	— —	1273
5679	Twin Diode	7CX	6.3	0.15	—	—	—	V.T.V.M. Rectifier	Same as 7A6										5679
XXL	Triode Oscillator	5AC	7.0	0.32	—	—	—	Oscillator	250	− 8.0	—	—	8.0	—	2300	20	—	—	XXL

* Cathode resistor—ohms. [1] Applied through 20000-ohm dropping resistor. [2] Each section. [3] Triode No. 1. [4] Triode No. 2.

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance µµfd. In	Capacitance µµfd. Out	Capacitance µµfd. Plate-Grid	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
2C21/ 1642	Twin-Triode Amplifier	M.	7BH	6.3	0.6	—	—	—	Class-A Amp.	250	−16.5	—	—	8.3	7600	1375	10.4	—	—	2C21/ 1642
6A3	Triode Power Amplifier	M.	4D	6.3	1.0	7.0	5.0	16.0	Class-A Amp.	250	−45	—	—	60	800	5250	4.2	2500	3.5	6A3
									Class AB₁ Amp.[10]	300 300	−62 850*	Fixed Bias Self Bias		80 80	—	—	—	3000[11] 5000[11]	15 10	
6A4	Pentode Power Amplifier	M.	5B	6.3	0.3	—	—	—	Class-A Amp.	180	−12.0	180	3.9	22	60000	2500	150	8000	1.5	6A4
6A6	Twin Triode Amplifier	M.	7B	6.3	0.8	—	—	—	Class-B Amp. P.P.	250 300	0 0	—	—	Power output is for one tube at stated load, plate-to-plate				8000 10000	8.0 10.0	6A6
6A7	Pentagrid Converter	S.	7C	6.3	0.3	8.5	9.0	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	S.	6R	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	S.	7AG	6.3	0.15	—	—	—	Indicator Tube	135 100		Ray Control Voltage = 81 for 0° Shadow Angle. Target current 1.5 ma. Ray Control Voltage = 60 for 0° Shadow Angle. Target current 0.9 ma.								6AF6G
6B5	Direct-Coupled Power Amplifier	M.	6AS	6.3	0.8	—	—	—	Class-A Amp.[9] Push-Pull Amp.[10]	300 400	0 −13.0	— —	6[1] 4.5[1]	45 40	241000 —	2400 —	58 —	7000 10000[11]	4.0 20	6B5

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance µµfd. In	Capacitance µµfd. Out	Capacitance µµfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6B7	Duplex-Diode Pentode	S.	7D	6.3	0.3	3.5	9.5	.007	Pentode R.F. Amp.	250	− 3.0	125	2.3	9.0	650000	1125	730	—	—	6B7
6C6	Sharp Cut-off Pentode	S.	6F	6.3	0.3	5	6.5	.007	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	6C6
6C7	Duplex Diode Triode	S.	7G	6.3	0.3	—	—	—	Class-A Amp.	250	− 9.0	—	—	4.5	—	20	1250	—	—	6C7
6D6	Variable-µ Pentode	S.	6F	6.3	0.3	4.7	6.5	.007	R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	6D6
6D7	Sharp Cut-off Pentode	S.	7H	6.3	0.3	5.2	6.8	.01	Class-A Amp.	250	− 3.0	100	0.5	2.0	—	1600	1280	—	—	6D7
6E5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	0	—	—	0.25	Target Current 4 ma.				—	6E5
6E6	Twin Triode Amplifier	M.	7B	6.3	0.6	—	—	—	Class-A Amp.	250	−27.5	Per plate—18.0			3500	1700	6.0	14000	1.6	6E6
6E7	Variable-µ Pentode	S.	7H	6.3	0.3	—	—	—	R.F. Amplifier			Characteristics same as 6U7G—Table II								6E7
6F7	Triode Pentode	S.	7E	6.3	0.3	—	—	—	Triode Unit Amp.	100	− 3.0	—	—	3.5	16000	500	8	—	—	6F7
									Pentode Unit Amplifier	250	− 3.0	100	1.5	6.5	850000	1100	900	—	—	
6U5/6G5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	Cut-off Grid Bias = −22 v.			0.24	Target Current 4 ma.				—	6U5/6G5
										100	Cut-off Grid Bias = −8 v.			0.19	Target Current 1 ma.					
6H5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube			Same characteristics as Type 6G5—Circular Pattern								6H5
6SB7Y	Pentagrid Converter	O.	8R	6.3	0.3	9.6	9.2	—	Converter	100	− 1	100	10.2	3.6	500000	900	—	—	—	6SB7Y
									Converter	250	− 1	100	10	3.8	1000000	950	—	—	—	
						Osc. Section in 88–108 Mc. Serv.				250	22000[6]	12000[7]	12.6/12.5	6.8/6.5	—	—	—	—	—	
6T5	Electron-Ray Tube	S.	6R	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	S.	5E	6.3	0.3	3.8	9	.007	R.F. Amplifier	250	− 3.0	90	1.7	3.2	550000	1080	595	—	—	36
37	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.9	2	Class-A Amp.	250	−18.0	—	—	7.5	8400	1100	9.2	—	—	37
38	Pentode Power Amplifier	S.	5F	6.3	0.3	3.5	7.5	0.3	Class-A Amp.	250	−25.0	250	3.8	22.0	100000	1200	120	10000	2.5	38
39/44	Remote Cut-off Pentode	S.	5F	6.3	0.3	3.8	10	007	R.F. Amplifier	250	− 3.0	90	1.4	5.8	1000000	1050	1050	—	—	39/44
41	Pentode Power Amplifier	S.	6B	6.3	0.4	—	—	—	Class-A Amp.	250	−18.0	250	5.5	32.0	68000	2200	150	7600	3.4	41
42	Pentode Power Amplifier	M.	6B	6.3	0.7	—	—	—	Class-A Amp.	250	−16.5	250	6.5	34.0	100000	2200	220	7000	3.0	42
52	Dual Grid Triode	M.	5C	6.3	0.3	—	—	—	Class-A Amp.[4]	110	0	—	—	43.0	1750	3000	5.2	2000	1.5	52
									Class-B, 2 tubes[5]	180	0	—	—	3.0[12]				10000	5.0	
56AS	Triode Amplifier	S.	5A	6.3	0.4	—	—	—	Class-A Amp.	Characteristics same as 56										56AS
57AS	Sharp Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 57										57AS
58AS	Remote Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 58										58AS
75	Duplex-Diode Triode	S.	6G	6.3	0.3	1.7	3.8	1.7	Triode Amplifier	250	− 1.35	—	—	0.4	91000	1100	100	—	—	75
76	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.5	2.8	Class-A Amp.	250	− 13.5	—	—	5.0	9500	1450	13.8	—	—	76
77	Sharp Cut-off Pentode	S.	6F	6.3	0.3	4.7	11	.007	R.F. Amplifier	250	− 3.0	100	0.5	2.3	1500000	1250	1500	—	—	77
78	Variable-µ Pentode	S.	6F	6.3	0.3	4.5	11	.007	R.F. Amplifier	250	− 3.0	100	1.7	7.0	800000	1450	1160	—	—	78
79	Twin Triode Amplifier	S.	6H	6.3	0.6	—	—	—	Class-B Amp.	250	0	—	—	10.6[12]	Power output is for one tube			14000	8.0	79
85	Duplex-Diode Triode	S.	6G	6.3	0.3	1.5	4.3	1.5	Class-A Amp.	250	−20.0	—	—	8.0	7500	1100	8.3	20000	0.35	85
85AS	Duplex-Diode Triode	S.	6G	6.3	0.3	—	—	—	Class-A Amp.	250	− 9.0	—	—	5.5	—	1250	20	—	—	85AS
89	Power Amplifier Pentode	S.	6F	6.3	0.4	—	—	—	Triode Amp.[2]	250	− 31.0	—	—	32.0	2600	1800	4.7	5500	0.9	89
									Pentode Amp.[8]	250	−25.0	250	5.5	32.0	70000	1800	125	6750	3.4	
1221	Pentode R.F. Amplifier	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Special non-microphonic. Characteristics same as 6C6										1221
1603[3]	Sharp Cut-off Pentode	M.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										1603
7700[3]	Sharp Cut-off Pentode	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										7700

* Cathode bias resistor—ohms.

1 Current to input plate (P_1).
2 Grids Nos. 2 and 3 connected to plate.
3 Low noise, nonmicrophonic tubes.
4 G_2 tied to plate.
5 G_1 tied to G_2.
6 Osc. grid leak ohms.
7 Screen dropping resistor ohms.
8 Grid No. 2, screen; grid No. 3, suppressor.
9 Values for single tube.
10 Values for two tubes in push-pull.
11 Plate-to-plate value.
12 No signal value.

TABLE V—2.5-VOLT RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance µµfd. In	Capacitance µµfd. Out	Capacitance µµfd. Plate-Grid	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	M.	5D	2.5	1.35	—	—	—	Detector	At 50 d.c. Volts per plate, cathode ma.=80										2S/4S
2A3	Triode Power Amplifier	M.	4D	2.5	2.5	7.5	5.5	16.5	Class-A Amp.	Characteristics same as Type 6A3, Table IV										2A3
2A5	Pentode Power Amplifier	M.	6B	2.5	1.75	—	—	—	Class-A Amp.	Characteristics same as Type 42, Table IV										2A5
2A6	Duplex-Diode Triode	S.	6G	2.5	0.8	1.7	3.8	1.7	Class-A Amp.	Characteristics same as Type 75, Table IV										2A6
2A7	Pentagrid Converter	S.	7C	2.5	0.8	—	—	—	Converter	Characteristics same as Type 6A7, Table IV										2A7
2B6	Direct-Coupled Amplifier	M.	7J	2.5	2.25	—	—	—	Amplifier	250	−24.0	—	—	40.0	5150	3500	18.0	5000	4.0	2B6
2B7	Duplex-Diode Pentode	S.	7D	2.5	0.8	3.5	9.5	.007	Pentode Amp.	Characteristics same as Type 6B7—Table IV										2B7
2E5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as Type 6E5—Table IV										2E5
2G5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as 6U5/6G5—Table IV										2G5
24-A	Tetrode R.F. Amplifier	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	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	Triode Detector-Amplifier	M.	5A	2.5	1.75	3.1	2.3	3.3	Class-A Amp.	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	Remote Cut-off Pentode	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	250	− 3.0	90	2.5	6.5	400000	1050	420	—	—	35/51
45	Triode Power Amplifier	M.	4D	2.5	1.5	4	3	7	Class-A Amp.	275	−56.0	—	—	36.0	1700	2050	3.5	4600	2.00	45
46	Dual-Grid Power Amp.	M.	5C	2.5	1.75	—	—	—	Class-A Amp.[2]	250	−33.0	—	—	22.0	2380	2350	5.6	6400	1.25	46
									Class-B Amp.[3]	400	0	—	—	12	Power output for 2 tubes			5800	20.0	
47	Pentode Power Amplifier	M.	5B	2.5	1.75	8.6	13	1.2	Class-A Amp.	250	−16.5	250	6.0	31.0	60000	2500	150	7000	2.7	47
53	Twin Triode Amplifier	M.	7B	2.5	2.0	—	—	—	Class-B Amp.	Characteristics same as Type 6A6, Table IV										53
55	Duplex-Diode Triode	S.	6G	2.5	1.0	1.5	4.3	1.5	Class-A Amp.	Characteristics same as Type 85, Table IV										55
56	Triode Amplifier, Detector	S.	5A	2.5	1.0	3.2	2.4	3.2	Class-A Amp.	Characteristics same as Type 76, Table IV										56
57	Sharp Cut-off Pentode	S.	6F	2.5	1.0	—	—	—	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	57
58	Remote Cut-off Pentode	S.	6F	2.5	1.0	4.7	6.3	.007	Screen-Grid R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	58
59	Pentode Power Amplifier	M.	7A	2.5	2.0	—	—	—	Class-A Triode[4]	250	−28.0	—	—	26.0	2300	2600	6.0	5000	1.25	59
									Class-A Pentode[5]	250	−18.0	250	9.0	35.0	40000	2500	100	6000	3.0	
RK15	Triode Power Amplifier	M.	4D[1]	2.5	1.75	—	—	—		Characteristics same as Type 46 with Class-B connections										RK15
RK16	Triode Power Amplifier	M.	5A	2.5	2.0	—	—	—		Characteristics same as Type 59 with Class-A triode connections										RK16
RK17	Pentode Power Amplifier	M.	5F	2.5	2.0	—	—	—		Characteristics same as Type 2A5										RK17

[1] Grid connection to cap; no connection to No. 3 pin. [2] Grid No. 2 tied to plate. [3] Grids Nos. 1 and 2 tied together. [4] Grids Nos. 2 and 3 connected to plate. [5] Grid No. 2, screen; grid No. 3, suppressor.

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance µµfd. In	Capacitance µµfd. Out	Capacitance µµfd. Plate-Grid	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	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.8	2.3	1000000	750	750	—	—	1A4P
1A4T	Variable-µ Tetrode	S.	4K	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.7	2.3	960000	750	720	—	—	1A4T
1A6	Pentagrid Converter	S.	6L	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	S.	4M	2.0	0.06	5	11	.007	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	S.	6M	2.0	0.06	1.6	1.9	3.6	Triode Class-A	135	− 3.0	—	—	0.8	35000	575	20	—	—	1B5/25S

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
1C6	Pentagrid Converter	S.	6L	2.0	0.12	10	10	—	Converter	180	−3.0	67.5	2.0	1.5	750000	Anode grid (No. 2) 135 max. volts				1C6
1F4	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	−4.5	135	2.6	8.0	200000	1700	340	16000	0.34	1F4
1F6	Duplex-Diode Pentode	S.	6W	2.0	0.06	4	9	.007	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	Sharp Cut-off Pentode	S.	5F	2.0	0.22	2.3	7.8	0.01	R.F. Amplifier	135	−1.5	67.5	0.3	1.85	800000	750	600	—	—	15
19	Twin-Triode Amplifier	S.	6C	2.0	0.26	—	—	—	Class-B Amp.	135	0	—	—	—	Load plate-to-plate			10000	2.1	19
30	Triode Detector Amplifier	S.	4D	2.0	0.06	—	—	—	Class-A Amp.	180	−13.5	—	—	3.1	10300	900	9.3	—	—	30
31	Triode Power Amplifier	S.	4D	2.0	0.13	3.5	2.7	5.7	Class-A Amp.	180	−30.0	—	—	12.3	3600	1050	3.8	5700	0.375	31
32	Sharp Cut-off Pentode	M.	4K	2.0	0.06	5.3	10.5	.015	R.F. Amplifier	180	−3.0	67.5	0.4	1.7	1200000	650	780	—	—	32
33	Pentode Power Amplifier	M.	5K	2.0	0.26	8	12	1	Class-A Amp.	180	−18.0	180	5.0	22.0	55000	1700	90	6000	1.4	33
34	Variable-μ Pentode	M.	4M	2.0	0.06	6	11	.015	R.F. Amplifier	180	−3.0	67.5	1.0	2.8	1000000	620	620	—	—	34
49	Dual-Grid Power Amp.	M.	5C	2.0	0.12	—	—	—	Class-A Amp.[1]	135	−20.0	—	—	6.0	4175	1125	4.7	11000	0.17	49
									Class-B Amp.[2]	180	0	—	—	Power output for 2 tubes				12000	3.5	
840	Pentode	S.	5J	2.0	0.13	—	—	—	Class-A Amp.	180	−3.0	67.5	0.7	1.0	1000000	400	400	—	—	840
950	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	−16.5	135	2.0	7.0	100000	1000	125	13500	0.575	950
RK24	Triode	M.	4D	2.0	0.12	—	—	—	Class-A Amp.	180	−13.5	—	—	8.0	5000	1600	8.0	12000	0.25	RK24
1229	Tetrode	M.	4K	2.0	0.06	—	—	—	Class-A Amp.	Special Type 32 for low grid-current applications										1229
1230	Triode	M.	4D	2.0	0.06	3.0	2.1	6.0		Special Type 30 for low grid-current applications										1230

[1] Grid No. 2 tied to plate.

[2] Grids Nos. 1 and 2 tied together.

TABLE VII—2.0-VOLT BATTERY TUBES WITH OCTAL BASES

Type	Name	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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	Heptode	7Z	2.0	0.06	10	14	0.26	Converter	Characteristics same as Type 1C6—Table VI										1C7G
1D5GP	Variable-μ Pentode	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1A4P—Table VI										1D5GP
1D5GT[4]	Variable-μ Tetrode	5R	2.0	0.06	—	—	—	R.F. Amplifier	180	−3.0	67.5	0.7	2.2	600000	650				1D5GT
1D7G	Pentagrid Converter	7Z	2.0	0.06	10.5	9.0	0.25	Converter	Characteristics same as Type 1A6—Table VI										1D7G
1E5GP	Pentode Amplifier	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1B4—Table VI										1E5GP
1E7G	Double Pentode Power Amp.	8C	2.0	0.24	—	—	—	Class-A Amplifier	135	−7.5	135	2.0[1]	6.5[1]	220000	1600	350	24000	0.65	1E7G
1F5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	Characteristics same as Type 1F4—Table VI										1F5G
1F7G[2]	Duplex-Diode Pentode	7AD	2.0	0.06	3.8	9.5	0.01	Detector-Amplifier	Characteristics same as Type 1F6—Table VI										1F7G
1G5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−13.5	135	2.5	8.7	160000	1550	250	9000	0.55	1G5G
1H4G	Triode Amplifier	5S	2.0	0.06	—	—	—	Detector-Amplifier	Characteristics same as Type 30—Table VI										1H4G
1H6G	Duplex-Diode Triode	7AA	2.0	0.06	1.6	1.9	3.6	Detector-Amplifier	Characteristics same as Type 1B5—Table VI										1H6G
1J5G	Pentode Power Amplifier	6X	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	2.0	0.24	—	—	—	Class-B Amplifier	Characteristics same as Type 19—Table VI										1J6G
4A6G	Twin Triode	8L	2.0	0.12	—	—	—	Class-A, 1 section	90	−1.5	—	—	1.1	26600	750	20			4A6G
			4.0	0.06				Class-B, 2 sections	90	−1.5	—	—	10.8[3]				8000	1.0	

[1] Total current for both sections; no signal.

[2] Type GV has 7AF base.

[3] Max. signal.

[4] Discontinued.

TABLE VIII—1.5-VOLT FILAMENT BATTERY TUBES

See also Table X for Special 1.4-volt Tubes

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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 M-watts	Type
1A5GT	Pentode Power Amplifier	O.	6X	1.4	0.05	—	—	—	Class-A$_1$ Amp.	90	−4.5	90	0.8	4.0	300000	850	240	25000	115	1A5GT
1A7GT	Pentagrid Converter	O.	7Z	1.4	0.05	—	—	—	Converter	90	0	45	0.6	0.55	600000	Anode-grid volts 90				1A7GT
1AB5	Pentode R.F. Amplifier	L.	5BF	1.2	0.05	2.8	4.2	0.25	R.F. Amplifier	90 150	0 −1.5	90 150	0.8 2.0	3.5 6.8	275000 125000	1100 1350	—	—	—	1AB5
1B7GT [4]	Heptode	O.	7Z	1.4	0.1	—	—	—	Converter	90	0	45	1.3	1.5	350000	Grid No. 1 resistor 200,000 ohms				1B7GT
1B8GT	Diode Triode Pentode	O.	8AW	1.4	0.1	—	—	—	Triode Amplifier Pentode Amp.	90 90	0 −6.0	— 90	— 1.4	0.15 6.3	240000 —	275 1150	— —	— 14000	— 210	1B8GT
1C5GT	Pentode Power Amplifier	O.	6X	1.4	0.1	—	—	—	Class-A$_1$ Amp.	90	−7.5	90	1.6	7.5	115000	1550	165	8000	240	1C5GT
1D8GT	Diode Triode Pentode	O.	8AJ	1.4	0.1	—	—	—	Triode Amp. Pentode Amp.	90 90	0 −9.0	— 90	— 1.0	1.1 5.0	43500 200000	575 925	25 —	— —	— —	1D8GT
1E4G	Triode Amplifier	O.	5S	1.4	0.05	2.4	6	2.40	Class-A Amp.	90 90	0 −3.0	—	—	4.5 1.5	11000 17000	1325 825	14.5 14	—	—	1E4G
1G4GT	Triode Amplifier	O.	5S	1.4	0.05	2.2	3.4	2.80	Class-A Amp.	90	−6.0	—	—	2.3	10700	825	8.8	—	—	1G4GT
1G6GT	Twin Triode	O.	7AB	1.4	0.1	—	—	—	Class-A Amp. Class-B Amp.	90 90	0 0	— —	— —	1.0 1/7	45000 34 volts input per grid	675	30	— 12000	— 675	1G6GT
1H5GT	Diode High-μ Triode	O.	5Z	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.14	240000	275	65	—	—	1H5GT
1LA4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	Characteristics same as 1A5GT									1LA4
1LA6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	45	0.6	0.55	Anode Grid Volts 90					1LA6
1LB4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	−9	90	1.0	5.0	200000	925	—	12000	200	1LB4
1LB6	Heptode Converter	L.	8AX	1.4	0.05	—	—	—	Converter	90	0	67.5	2.2	0.4	Grid No. 4—67.5 v., No. 5—0 v.					1LB6
1LC5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.2	7	.007	R.F. Amplifier	90	0	45	0.2	1.15	1500000	775	—	—	—	1LC5
1LC6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	35 [1]	0.7	0.75	Anode Grid Volts 45					1LC6
1LD5	Diode Pentode	L.	6AX	1.4	0.05	3.2	6	0.18	Class-A Amp.	90	0	45	0.1	0.6	950000	600	—	—	—	1LD5
1LE3	Triode Amplifier	L.	4AA	1.4	0.05	1.7	3	1.70	Class-A Amp.	90 90	0 −3	—	—	4.5 1.3	11200 19000	1300 760	14.5	—	—	1LE3
1LG5	Pentode R.F. Amp.	L.	7AO	1.4	0.05	—	—	—	Class-A Amp.	90	0	45	0.4	1.7	1000000	800	—	—	—	1LG5
1LH4	Diode High-μ Triode	L.	5AG	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.15	240000	275	65	—	—	1LH4
1LN5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.4	8	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	—	—	—	1LN5
1N5GT	Remote Cut-off Pentode	O.	5Y	1.4	0.05	3	10	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	1160	—	—	1N5GT
1N6G [4]	Diode-Power-Pentode	O.	7AM	1.4	0.05	—	—	—	Class-A Amp.	90	−4.5	90	0.6	3.1	300000	800	—	25000	100	1N6G
1P5GT	Pentode	O.	5Y	1.4	0.05	3	10	.007	R.F. Amplifier	90	0	90	0.7	2.3	800000	800	640	—	—	1P5GT
1Q5GT	Tetrode Power Amplifier	O.	6AF	1.4	0.1	—	—	—	Class-A Amp.	85 90	−5.0 −4.5	85 90	1.2 1.6	7.2 9.5	70000 75000	1950 2100	—	9000 8000	250 270	1Q5GT
1R4/1294	U.h.f. Diode	L.	4AH	1.4	0.15	—	—	—	Rectifier	Max. r.m.s. voltage per plate—30					Max. d.c. output current—340 μa.					1R4/1294
1SA6GT	Medium Cut-off Pentode	O.	6CA	1.4	0.05	6.2	8.6	0.01	R.F. Amplifier	90	0	67.5	0.68	2.45	800000	970	—	—	—	1SA6GT
1SB6GT	Diode Pentode	O.	6CB	1.4	0.05	3.2	3	0.25	Class-A Amp. R.C. Amplifier	90 90	0 0	67.5 90	0.38 Screen resistor 5 meg., grid 10 meg.	1.45	700000	665	—	— 1 meg.	— 110 [2]	1SB6GT
1T5GT	Beam Power Amplifier	O.	6AF	1.4	0.05	4.8	8	0.50	Class-A Amp.	90	−6.0	90	1.4	6.5	—	1150	—	14000	170	1T5GT
3B7/1291	U.h.f. Twin Triode	L.	7BE	2.8 [3]	0.11	1.4	2.6	2.6	Class-A Amp.	90	0	—	—	5.2	11350	1850	21	—	—	3B7/1291
1293	U.h.f. Triode	L.	4AA	1.4	0.11	1.7	3.0	1.7	Class-A Amp.	90	0	—	—	4.7	10750	1300	14	—	—	1293
3D6/1299	U.h.f. Tetrode	L.	6BB	2.8 [3]	0.11	7.5	6.5	0.30	Class-A Amp.	135	−6	90	0.7	5.7	—	2200	—	13000	500	3D6/1299
3E6	R.F. Pentode	L.	7CJ	1.4 2.8	0.10 0.05	5.5	7.5	0.007	Class-A Amp.	90	0	90	1.3	3.8	300000	2100	—	—	—	3E6
RK42	Triode Amplifier	S.	4D	1.5	0.6	—	—	—	Class-A Amp.		Characteristics same as Type 30—Table VI									RK42
RK43	Twin Triode Amplifier	S.	6C	1.5	0.12	—	—	—	Class-A Amp.	135	−3	—	—	4.5	14500	900	13	—	—	RK43

[1] Through series resistor. Screen voltage must be at least 10 volts lower than oscillator anode. [2] Voltage gain. [3] Center-tap filament permits 1.4-volt operation. [4] Discontinued.

TABLE IX—HIGH-VOLTAGE HEATER TUBES

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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 [8]	Pentode Power Amplifier	M.	7F	12.6 6.3	0.3 0.6	9.0	9.0	0.3	Class-A_1 Amp. [6]	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	O.	7AC	12.6	0.15	—	—	—	Class-A Amp.	250	−12.5	250	3.5	30	70000	3000	—	7500	3.4	12A6
12A7	Rectifier-Amplifier	M.	7K	12.6	0.3	—	—	—	Class-A Amp.	135	−13.5	135	2.5	9.0	102000	975	100	13500	0.55	12A7
12A8GT	Heptode	O.	8A	12.6	0.15	9.5	12	0.26	Converter	Characteristics same as 6A8—Table I										12A8GT
12AH7GT	Twin Triode	O.	8BE	12.6	0.15	Each Triode Sect.			Class-A Amp.	180	−6.5	—	—	7.6	8400	1900	16	—	—	12AH7GT
12B6M	Diode Triode	O.	6Y	12.6	0.15	—	—	—	Class-A Amp.	250	−2.0	—	—	0.9	91000	1100	100	—	—	12B6M
12B7ML	Pentode Amplifier	O.	8V	12.6	0.15	—	—	—	Class-A Amp.	250	−3.0	100	2.6	9.2	800000	2000	—	—	—	12B7ML
12B8GT [8]	Triode-Pentode	O.	8T	12.6	0.3	Triode Section Pentode Section			Class-A Amp. Class-A Amp.	100 100	−1 −3	— 100	— 2	0.6 8	73000 170000	1500 2100	110 360	— —	— —	12B8GT
12C8	Duplex-Diode Pentode	O.	8E	12.6	0.15	6	9	.005	Class-A Amp.	Characteristics same as 6B8—Table I										12C8
12E5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	5.5	2.60	Class-A Amp.	250	−13.5	—	—	50	—	1450	13.8	—	—	12E5GT
12F5GT	Triode Amplifier	O.	5M	12.6	0.15	1.9	3.4	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12F5GT
12G7G	Duplex-Diode Triode	O.	7V	12.6	0.15	—	—	—	Class-A Amp.	250	−3.0	—	—	—	58000	1200	70	—	—	12G7G
12H6	Twin Diode	O.	7Q	12.6	0.15	—	—	—	Rectifier	Characteristics same as 6H6—Table I										12H6
12J5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	3.6	3.40	Class-A Amp.	Characteristics same as 6J5—Table I										12J5GT
12J7GT	Sharp Cut-off Pentode	O.	7R	12.6	0.15	4.2	5.0	3.8	Class-A Amp.	Characteristics same as 6J7—Table I										12J7GT
12K7GT	Remote Cut-off Pentode	O.	7R	12.6	0.15	4.6	12	.005	R.F. Amplifier	Characteristics same as 6K7—Table I										12K7GT
12K8	Triode Hexode Converter	O.	8K	12.6	0.15	—	—	—	Converter	Characteristics same as 6K8—Table I										12K8
12L8GT	Twin Pentode	O.	8BU	12.6	0.15	5	6	0.70	Class-A_1 Amp.	180	−9.0	180	2.8	13.0	160000	2150	—	10000	1.0	12L8GT
12Q7GT	Duplex-Diode Triode	O.	7V	12.6	0.15	2.2	5	1.60	Class-A Amp.	Characteristics same as 6Q7—Table I										12Q7GT
12S8GT	Triple-Diode Triode	O.	Fig. 34	12.6	0.15	2.0	3.8	1.2	Class-A Amp.	250	−2.0	—	—	0.9	91000	1100	100	—	—	12S8GT
12SA7	Heptode	O.	8R	12.6	0.15	9.5	12	0.13	Converter	Characteristics same as 6SA7—Table I										12SA7
12SC7	Twin Triode	O.	8S	12.6	0.15	2.2	3.0	2.0	Class-A Amp.	Characteristics same as 6SC7—Table I										12SC7
12SF5	High-μ Triode	O.	6AB	12.6	0.15	4	3.6	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12SF5
12SF7	Diode Variable-μ Pentode	O.	7AZ	12.6	0.15	5.5	6.0	.004	Class-A Amp.	Characteristics same as 6SF7—Table I										12SF7
12SG7	Medium Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	Class-A Amp.	Characteristics same as 6SG7—Table I										12SG7
12SH7	Sharp Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	H-F Amplifier	Characteristics same as 6SH7—Table I										12SH7
12SJ7	Sharp Cut-off Pentode	O.	8N	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SJ7—Table I										12SJ7
12SK7	Remote Cut-off Pentode	O.	8N	12.6	0.15	6.0	7.0	.003	R.F. Amplifier	Characteristics same as 6SK7—Table I										12SK7
12SL7GT	Twin Triode	O.	8BD	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SL7GT—Table II										12SL7GT
12SN7GT	Twin Triode	O.	8BD	12.6	0.3	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table II										12SN7GT
12SQ7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.2	3.0	1.60	Class-A Amp.	Characteristics same as 6SQ7—Table I										12SQ7
12SR7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.6	2.8	2.40	Class-A Amp.	Characteristics same as 6R7—Table I										12SR7
12SW7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.0	2.8	2.4	Class-A_1 Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	12SW7
12SX7	Twin Triode	O.	8BD	12.6	0.3	3.0	0.8	3.6	Class-A_1 Amp. [5]	250	−8	—	—	9	7700	2600	20	—	—	12SX7
12SY7	Heptode Converter	O.	8R	12.6	0.15	Osc.-Grid leak 20000 ohms			Converter	250	−2	100	8.5	3.5	1000000	450	—	—	—	12SY7
14A4	Triode Amplifier	L.	5AC	14	0.16	3.4	3.0	4.00	Class-A Amp.	Characteristics same as 7A4—Table III										14A4
14A5	Beam Power Amplifier	L.	6AA	14	0.16	—	—	—	Class-A_1 Amp.	250	−12.5	250	3.5/5.5	30/32	70000	3000	—	7500	2.8	14A5
14A7/ 12B7	Remote Cut-off Pentode	L.	8V	14	0.16	6.0	7.0	.005	Class-A Amp.	250	−3.0	100	2.6	9.2	800000	2000	—	—	—	14A7/ 12B7
14AF7	Twin Triode	L.	8AC	14	0.16	2.2	1.6	2.30	Class-A Amp.	250	−10	—	—	9	7600	2100	16	—	—	14AF7
14B6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7B6—Table III										14B6
14B8	Pentagrid Converter	L.	8X	14	0.16	Ic2 = 4 Ma.			Converter	Characteristics same as 7B8—Table III										14B8
14C5	Beam Power Amplifier	L.	6AA	14	0.24	—	—	—	Class-A Amp.	Characteristics same as 6V6—Table I										14C5

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
14C7	R.F. Pentode	L.	8V	14	0.16	6.0	6.5	.007	Class-A Amp.	250	− 3.0	100	0.7	2.2	1000000	1575	—	—	—	14C7
14E6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7E6—Table III										14E6
14E7	Duplex-Diode Pentode	L.	8AE	14	0.16	4.6	5.3	.005	Class-A Amp.	Characteristics same as 7E7—Table III										14E7
14F7	Twin Triode	L.	8AC	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7F7—Table III										14F7
14F8	Twin Triode	L.	8BW	12.6	0.15	2.8	1.4	1.2	Class-A_1 Amp.	Characteristics same as 7F8										14F8
14H7	Semi-Variable-μ Pentode	L.	8V	14	0.16	8.0	7.0	.007	Class-A Amp.	250	− 2.5	150	3.5	9.5	800000	3800	—	—	—	14H7
14J7	Triode-Hexode Converter	L.	8BL	14	0.16	Ipt = 5 Ma.			Converter	Characteristics same as 7J7—Table III										14J7
14N7	Twin Triode	L.	8AC	14	0.32	—	—	—	Class-A Amp.	Characteristics same as 7N7—Table III										14N7
14Q7	Heptode Pentagrid Converter	L.	8AL	14	0.16	—	—	—	Converter	Characteristics same as 7Q7—Table III										14Q7
14R7	Duplex-Diode Pentode	L.	8AE	14	0.16	5.6	5.3	.004	Class-A Amp.	Characteristics same as 7R7—Table III										14R7
14S7	Triode Heptode	L.	8BL	14	0.16	Ipt = 5 Ma.			Converter	250	− 2.0	100	3	1.8	1250000	525	—	—	—	14S7
14V7	H.f. Pentode	L.	8V	14	0.24	—	—	—	Class-A Amp.	300	− 2.0	150	3.9	9.6	300000	5800	—	—	—	14V7
14W7	Pentode	L.	8BJ	14	0.24	Rk = 160 ohms			Class-A Amp.	300	− 2.2	150	3.9	10	300000	5800	—	—	—	14W7
18	Pentode	M.	6B	14	0.30	—	—	—	Class-A Amp.	Characteristics same as 6F6G										18
20J8GM	Triode Heptode Converter	O.	8H	20	0.15	—	—	—	Converter	250	− 3.0	100	3.4	1.5	Triode Plate (No. 6) 100 v. 1.5 ma.					20J8GM
21A7	Triode Hexode Converter	L.	8AR	21	0.16	—	—	—	Converter	250 150	− 3.0 − 3.0	100 Triode	2.8	1.3 3.5	— —	275 1900	— 32	— —	— —	21A7
25A6	Pentode Power Amplifier	O.	7S	25	0.3	8.5	12.5	0.20	Class-A Amp.	135	−20.0	135	8	37	35000	2450	85	4000	2.0	25A6
25A7GT [8]	Rectifier Power Pentode	O.	8F	25	0.3	—	—	—	Class-A Amp.	100	−15.0	100	4	20.5	50000	1800	90	4500	0.77	25A7GT
25AC5GT [8]	Triode Power Amplifier	O.	6Q	25	0.3	—	—	—	Class-A Amp.	110 165	+15.0 Used in dynamic-coupled circuit with 6AF5G driver	—	—	45	—	3800	58	2000 3500	2.0 3.3	25AC5GT
25B5 [8]	Direct-Coupled Triodes	S.	6D	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25B5
25B6G [8]	Pentode Power Amplifier	O.	7S	25	0.3	—	—	—	Class-A Amp.	95	−15.0	95	4	45	—	4000	—	2000	1.75	25B6G
25B8GT [8]	Triode Pentode	O.	8T	25	0.15	—	—	—	Class-A Amp.	Characteristics same as 12B8GT										25B8GT
25C6G [8]	Beam Power Amplifier	O.	7AC	25	0.3	—	—	—	Class-A_1 Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	25C6G
25D8GT	Diode Triode Pentode	O.	8AF	25	0.15	—	—	—	Triode Amp. Pentode Amp.	100 100	− 1.0 − 3.0	— 100	— 2.7	0.5 8.5	91000 200000	1100 1900	100 —	— —	— —	25D8GT
25L6	Beam Power Amplifier	O.	7AC	25	0.3	16	13.5	0.30	Class-A_1 Amp.	110	− 8.0	110	3.5/10.5	45/48	10000	8000	80	2000	2.2	25L6
25N6G [8]	Direct-Coupled Triodes	O.	7W	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25N6G
26A7GT	Twin Beam-Power Audio Amplifier	O.	8BU	26.5	0.6	Each Unit Push-Pull			Class-A Amp. Class-AB Amp. [3]	26.5 26.5	− 4.5 − 7.0	26.5 26.5	2/5.5 2/8.5	20/20.5 19/30	2500 —	5500 —	— —	1500 2500 [4]	0.2 0.5	26A7GT
32L7GT	Diode-Beam Tetrode	O.	8Z	32.5	0.3	—	—	—	Class-A Amp.	110	− 7.5	110	3	40	15000	6000	—	2500	1.5	32L7GT
35A5	Beam Power Amplifier	L.	6AA	35	0.15	—	—	—	Class-A_1 Amp.	110	− 7.5	110	3/7	40/41	14000	5800	—	2500	1.5	35A5
35L6G	Beam Power Amplifier	O.	7AC	35	0.15	13	9.5	0.80	Class-A_1 Amp.	110	− 7.5	110	3/7	40/41	13800	5800	—	2500	1.5	35L6G
43	Pentode Power Amplifier	M.	6B	25	0.3	8.5	12.5	0.20	Class-A Amp.	95	−15.0	95	4.0	20.0	45000	2000	90	4500	0.90	43
48 [8]	Tetrode Power Amplifier	M.	6A	30	0.4	—	—	—	Class-A Amp.	96	−19.0	96	9.0	52.0	—	3800	—	1500	2.0	48
50A5	Beam Power Amplifier	L.	6AA	50	0.15	—	—	—	Class-A_1 Amp.	110	− 7.5	110	4/11	49/50	10000	8200	—	2000	2.2	50A5
50C6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A_1 Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	50C6GT
50L6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A Amp.	110	− 7.5	110	4/11	49/50	—	8200	82	2000	2.2	50L6GT
70A7GT	Diode-Beam Tetrode	O.	8AB [1]	70	0.15	—	—	—	Class-A Amp.	110	− 7.5	110	3.0	40	—	5800	80	2500	1.5	70A7GT
70L7GT	Diode-Beam Tetrode	O.	8AA	70	0.15	—	—	—	Class-A_1 Amp.	110	− 7.5	110	3/6	40/43	15000	7500	—	2000	1.8	70L7GT
117L7GT/ 117M7GT	Rectifier-Amplifier	O.	8AO	117	0.09	—	—	—	Class-A Amp.	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117L7GT/ 117M7GT
117N7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	100	− 6.0	100	5.0	51	16000	7000	—	3000	1.2	117N7GT
117P7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117P7GT

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
1280	Pentode	L.	8V	12.6	0.15	6.0	6.5	0.007	Class-A_1 Amp.	Same as 14C7 (Special Non-microphonic)										1280
1284	U.h.f. Pentode	L.	8V	12.6	0.15	5.0	6.0	0.01	Class-A Amp.	250	− 3.0	100	2.5	9.0	800000	2000	—	—	—	1284
1629	Electron-Ray Tube	O.	6RA	12.6	0.15	—	—	—	Indicator Tube	Characteristics same as 6E5—Table IV										1629
1631	Beam Power Amplifier	O.	7AC	12.6	0.45	—	—	—	Class-A Amp.	Characteristics same as 6L6—Table I										1631
1632	Beam Power Amplifier	O.	7AC	12.6	0.6	—	—	—	Class-A Amp.	Characteristics same as 25L6										1632
1633	Twin Triode	O.	8BD	25	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table II										1633
1634	Twin Triode	O.	8S	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SC7—Table I										1634
1644	Twin Pentode	O.	Fig. 7	12.6	0.15	—	—	—	Class-A Amp.	180	− 9.0	180	2.8/4.6	13	160000	2150	—	10000	1.0	1644
XXD/ 14AF7	Twin Triode	L.	8AC	12.6	0.15	—	—	—	Class-A Amp.	250	−10	—	—	9.0	—	2100	16	—	—	XXD/ 14AF7
28D7	Double Beam Power Amplifier	L.	8BS	28.0	0.4	—	—	—	Class-A Amp.	28	390* 180*	28 [2] 28 [3]	0.7 [2] 1.2 [3]	9.0 [2] 18.5 [3]	— —	— —	— —	4000 [3] 6000 [4]	0.08 [2] 0.175 [3]	28D7

* Cathode resistor—ohms.

[1] 6.3-volt pilot lamp must be connected between Pins 6 and 7.
[2] Per section—resistance-coupled.
[3] P.p. operation—values for both sections..
[4] Plate to plate.
[5] Values are for each unit.
[6] Values are for single tube.
[7] Grids 2 and 3 connected to plate.
[8] Discontinued.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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 [7]	Triode Detector	M.	4D	5.0	0.25	3.2	2.0	8.50	Grid-Leak Det.	45	—	—	—	1.5	30000	666	20	—	—	00-A
01-A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	135	− 9.0	—	—	3.0	10000	800	8.0	—	—	01-A
3A8GT	Diode Triode Pentode	O.	8AS	1.4	0.1	2.6	1.2	2.0	Class-A Triode	90	0	—	—	0.15	240000	275	65	—	—	3A8GT
				2.8	0.05	3.0	10.0	0.012	Class-A Pentode	90	0	90	0.3	1.2	600000	750	—	—	—	
3B5GT	Beam Power Amplifier	O.	7AP	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	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	O.	7AQ	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.4	6.0	—	1550 1450	—	8000 10000	0.24 0.26	3C5GT
3C6	Twin Triode	L.	7BW	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	0	—	—	4.5	11200	1300	14.5	—	—	3C6
3LE4	Power Amplifier Pentode	L.	6BA	2.8	0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.8	9.0	110000	1600	—	6000	0.30	3LE4
3LF4	Power Amplifier Tetrode	L.	6BB	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 8.0	75000 80000	2200 2000	—	8000 7000	0.27 0.23	3LF4
3Q5GT	Beam Power Amplifier	O.	7AQ	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 7.5	—	2100 1800	—	8000	0.27 0.25	3Q5GT
4A6G	Twin Triode Amplifier	O.	8L	4	0.06	Triodes Parallel			Class-A Amp.	90	− 1.5	—	—	2.2	13300	1500	20	—	—	4A6G
				2	0.12	Both Sections			Class-B Amp.	90	0	—	—	4.6	—	—	—	8000	1.0	
6F4	Acorn Triode	A.	7BR	6.3	0.225	2.0	0.6	1.90	Class-A Amp.	80	150*	—	—	13.0	2900	5800	17	—	—	6F4
10	Triode Power Amplifier	M.	4D	7.5	1.25	4.0	3.0	7.00	Class-A Amp.	425	−39.0	—	—	18.0	5000	1600	8.0	10200	1.6	10
11/12 [7]	Triode Detector Amplifier	M.	4F/4D	1.1	0.25	—	—	—	Class-A Amp.	135	−10.5	—	—	3.0	15000	440	6.6	—	—	11/12
20 [7]	Triode Power Amplifier	S.	4D	3.3	0.132	2.0	2.3	4.10	Class-A Amp.	135	−22.5	—	—	6.5	6300	525	3.3	6500	0.11	20
22 [7]	Tetrode R.F. Amplifier	M.	4K	3.3	0.132	3.5	10	0.02	Class-A Amp.	135	− 1.5	67.5	1.3	3.7	325000	500	160	—	—	22
26	Triode Amplifier	M.	4D	1.5	1.05	2.8	2.5	8.10	Class-A Amp.	180	− 14.5	—	—	6.2	7300	1150	8.3	—	—	26
40 [7]	Triode Voltage Amplifier	M.	4D	5.0	0.25	2.8	2.2	2.00	Class-A Amp	180	− 3.0	—	—	0.2	150000	200	30	—	—	40
50	Triode Power Amplifier	M.	4D	7.5	1.25	4.2	3.4	7.10	Class-A Amp.	450	−84.0	—	—	55.0	1800	2100	3.8	4350	4.6	50

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
71-A	Triode Power Amplifier	M.	4D	5.0	0.25	3.2	2.9	7.50	Class-A Amp.	180	−43.0	—	—	20.0	1750	1700	3.0	4800	0.79	71-A
99 [6]	Triode Detector Amplifier	S.	4D	3.3	0.063	2.5	2.5	3.30	Class-A Amp.	90	−4.5	—	—	2.5	15500	425	6.6	—	—	99
112A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	180	−13.5	—	—	7.7	4700	1800	8.5	—	—	112A
182B/482B [*]	Triode Amplifier	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−35.0	—	—	18.0	—	1500	5.0	—	—	182B/482B
183/483 [7]	Power Triode	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−60.0	—	—	25.0	18000	1800	3.2	4500	2.0	183/483
485 [7]	Triode •	S.	5A	3.0	1.3	—	—	—	Class-A Amp.	180	−9.0	—	—	6.0	9300	1350	12.5	—	—	485
864	Triode Amplifier	S.	4D	1.1	0.25	—	—	—	Class-A Amp.	90	−4.5	—	—	2.9	13500	610	8.2	—	—	864
954	Pentode Detector, Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp.	250	−3.0	100	0.7	2.0	1.5 meg.	1400	2000	—	—	954
									Bias Detector	250	−6.0	100	—	Plate current to be adjusted to 0.1 ma. with no signal					—	
955	Triode Detector, • Amplifier, Oscillator	A.	5BC	6.3	0.15	1.0	0.6	1.40	Class-A Amp.	250	−7.0	—	—	6.3	11400	2200	25	—	—	955
										90	−2.5	—	—	2.5	14700	1700	25	—	—	
956	Variable-μ Pentode R.F. Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp.	250	−3.0	100	2.7	6.7	700000	1800	1440	—	—	956
									Mixer	250	−10.0	100	—	—	Oscillator peak volts—7 min.					
957	Triode Detector. Amplifier, Oscillator	A.	5BD	1.25	0.05	0.3	0.7	1.20	Class-A Amp.	135	−5.0	—	—	2.0	20800	650	13.5	—	—	957
958 958-A	Triode A.F. Amplifier, Oscillator	A.	5BD	1.25	0.1	0.6	0.8	2.60	Class-A Amp.	135	−7.5	—	—	3.0	10000	1200	12	—	—	958 958-A
959	Pentode Detector, Amplifier	A.	5BE	1.25	0.05	1.8	2.5	0.015	Class-A Amp.	145	−3.0	67.5	0.4	1.7	800000	600	480	—	—	959
7E5/1201	U.h.f. Triode	L.	8BN	6.3	0.15	3.6	2.8	1.50	Class-A Amp.	180	−3	—	—	5.5	12000	—	36	—	—	7E5/1201
7C4/1203	U.h.f. Diode	L.	4AH	6.3	0.15	—	—	—	Rectifier		Max. r.m.s. voltage—150				Max. d.c. output current—8 ma.					7C4/1203
7AB7/1204	Sharp Cut-off Pentode	L.	8BO	6.3	0.15	3.5	4.0	0.06	Class-A Amp.	250	−2	100	0.6	1.75	800000	1200	—	—	—	7AB7/1204
1276	Triode Power Amplifier	M.	4D	4.5	1.14	—	—	—	Class-A Amp.				Characteristics similar to 6A3							1276
1609	Pentode Amplifier	S.	5B	1.1	0.25	—	—	—	Class-A Amp.	135	−1.5	67.5	0.65	2.5	400000	725	300	—	—	1609
9004	U.h.f. Diode	A.	4BJ	6.3	0.15	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—5 ma.								9004
9005	U.h.f. Diode	A.	5BG	3.6	0.165	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—1 ma.								9005
EF-50	Sharp Cut-off Pentode	L.	9C	6.3	0.3	8	5	0.007	I.F.-R.F. Amp.	250	150*	250	3.1	10	600000	6300	—	—	—	EF-50
GL-2C44 GL-464A	U.h.f. Triode	O.	Fig. 17	6.3	0.75	—	—	—	Class-A Amp. and Modulator	250	100*	—	—	25.0	—	7000	—	—	—	GL-2C44 GL-464A
GL-446A GL-446B	U.h.f. Triode	O.	Fig. 19	6.3	0.75	—	—	—	Oscillator, Amp. or Converter	250	200*	—	—	15.0	—	4500	45	—	—	GL-446A GL-446B
559 GL-559	U.h.f. Diode	O.	Fig. 18	6.3	0.75	—	—	—	Detector or trans. line switch	5.0	—	—	—	24.0	—	—	—	—	—	559 GL-559
NU-2C35	Special Hi-Mu Triode	O.	Fig. 38	6.3	0.3	5.2	2.3	0.62	Shunt Voltage Regulator	8000	−200	—	—	5.0	525000	950	500	—	—	NU-2C35
VT52	Triode	M.	4D	7.0	1.18	5.0	3.0	7.7	Class-A_1 Amp.	220	−43.5	—	—	29.0	1650	2300	3.8	3800	1.0	VT52
X6030	Diode	L.	Fig. 4	3.0	0.6	—	—	—	Noise Diode	1400	—	—	—	0.535	—	—	—	—	—	X6030

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
XXB	Twin-Triode Frequency Converter	L.	Fig. 9	2.8/ 1.4 3.2[3]/ 1.6	0.05/ 0.10 — —	—	—	—	Converter[2]	90[1]	0	—	—	4.5[4] 4.5[5]	11200[4] 11200[5]	1300[4] 1300[5]	14.5[1]	—	—	XXB
											− 3	—	—	1.4[4] 1.4[5]	1900[4] 1900[5]	760[4] 760[5]	14.5[1]	—	—	
XXFM	Twin-Diode Triode	L.	8BZ	6.3	0.3	—	—	—	Class-A Amp.	250	− 1	—	—	1.9	6700	1500	100	—	—	XXFM
										100	0	—	—	1.2	85000	1000	85	—	—	

* Cathode resistor—ohms.
[1] Both sections.
[2] Section No. 2 recommended for h.f.o.
[3] Dry battery operation.
[4] Section No. 1.
[5] Section No. 2.
[6] Same as X99. Type V99 is same, but socket connections are 4E.
[7] Discontinued.

TABLE XI—MINIATURE RECEIVING TUBES

Other miniature types in Tables XIII and XV

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
1A3	H. F. Diode	B.	5AP	1.4	0.15	—	—	—	Detector F.M. Discrim.		Max. a.c. voltage per plate—117				Max. output current—0.5 ma.					—
1L4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	.008	Class-A Amp.	90	0	90	2.0	4.5	350000	1025	—	—	—	1N5GT
1R5	Pentagrid Converter	B.	7AT	1.4	0.05	—	—	—	Converter	90	0	67.5	3.0	1.7	500000	300	Grid No. 1 100000 ohms			1A7GT
1S4	Pentagrid Power Amp.	B.	7AV	1.4	0.1	—	—	—	Class-A Amp.	90	− 7.0	67.5	1.4	7.4	100000	1575	—	8000	0.270	1Q5GT
1S5	Diode Pentode	B.	6AU	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
									R-Coupled Amp.	90	0	90	Screen resistor 3 meg., grid 10 meg.					1 meg.	0.050	
1T4	Variable-μ Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	67.5	1.4	3.5	500000	900	—	—	—	1P5GT
1U4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	90	0.5	1.6	1500000	900	—	—	—	1N5GT
1U5	Diode Pentode	B.	6BW	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
2C51	Twin Triode	B.	8CJ	6.3	0.3	2.2	1.0	1.3	Class-A₁ Amp.	150	− 2	—	—	8.2[1]	—	5500	35	—	—	7F8
2E30	Beam Power Tetrode	B.	7CQ	6.0	0.7	10	4.5	0.5	Class-A₁ Single	250	450*	250	7.4[2]	44[2]	63000	3700	40[5]	4500	4.5	—
									Class-A₁ Amp.[3]	250	225*	250	14.8[2]	88[2]	—	—	80[5]	9000[6]	9	
									Class-AB₁ Amp.[3]	250	−25	250	13.5[2]	80[2]	—	—	48[5]	8000[6]	12.5	
									Class-AB₂ Amp.[3]	250	−30	250	20[2]	120[2]	—	—	40[5]	3800[6]	17	
3A4	Power Amplifier Pentode	B.	7BB	1.4 2.8	0.2 0.1	4.8	4.2	0.34	Class-A₁ Amp.	135 150	− 7.5 − 8.4	90 90	2.6 2.2	14.9[2] 14.1[2]	90000 100000	1900	—	8000	0.6 0.7	—
3A5	H.F. Twin Triode	B.	7BC	1.4 2.8	0.22 0.11	0.9	1.0	3.20	Class-A Amp.	90	− 2.5	—	—	3.7	8300	1800	15	—	—	—
3Q4	Power Amplifier Pentode	B.	7BA	1.4	0.1	Parallel Filaments			Class-A Amp.	90	− 4.5	90	2.1	9.5	100000	2150	—	10000	0.27	3Q5GT
				2.8	0.05	Series Filaments							1.7	7.7	120000	2000			0.24	
3S4	Power Amplifier Pentode	B.	7BA	1.4	0.1	Parallel Filaments			Class-A Amp.	90	− 7.0	67.5	1.4	7.4	100000	1575	—	8000	0.27	3Q5GT
				2.8	0.05	Series Filaments							1.1	6.1		1425			0.235	
3V4	Power Amplifier Pentode	B.	6BX	1.4	0.1	Parallel Filaments			Class-A Amp.	90	− 4.5	90	2.1	9.5	100000	2150	—	10000	0.27	3Q5GT
				2.8	0.05	Series Filaments			Class-A Amp.	90	− 4.5	90	1.7	7.7	120000	2000	—	10000	0.24	
6AG5	Sharp Cut-off Pentade	B.	7BD	6.3	0.3	—	—	—	Class-A Amp.	250 100	200* 100*	150 100	2.0 1.6	7.0 5.5	800000 300000	5000 4750	— —	— —	—	6SH7GT
6AH6	Sharp Cut-off Pentode	B.	7CC	6.3	0.45	10	2	0.03	Pentode Amp.	300	160*	150	2.5	10	500000	9000	—	—	—	6AC7
									Triode Amp.[7]	150	160*	—	—	12.5	3600	11000	40	—	—	
6AJ5	Sharp Cut-off Pentode	B.	7PM	6.3	0.175	—	—	—	R.F. Amplifier	28	200*	28	1.2	3.0	90000	2750	250	—	—	—
									Class-AB Amp.[3]	180	− 7.5	75	—	—	—	—	—	28000[6]	1.0	
6AK5	Sharp Cut-off Pentode	B.	7BD	6.3	0.175	4.3	2.1	0.03	R.F. Amplifier	180	200*	120	2.4	7.7	690000	5100	3500	—	—	—
										150	330*	140	2.2	7.0	420000	4300	1800	—	—	
										120	200*	120	2.5	7.5	340000	5000	1700			

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections[1]	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
6AK6	Power Amplifier Pentode	B.	7BK	6.3	0.15	3.6	4.2	0.12	Class-A Amp.	180	−9.0	180	2.5	15.0	200000	2300	—	10000	1.1	—
6AL5	U.h.f. Twin Diode	B.	6BT	6.3	0.3	—	—	—	Detector			Max. r.m.s. voltage—150. Max. d.c. output current—10 ma.[1]								6H6GT
6AN5	Power Amp. Pentode	B.	7BD	6.3	0.5	9.0	4.8	0.05	Class-A_1 Amp.	120	−6	120	12	35	12500	8000	—	—	—	6AG7
6AN6	Twin Diode	B.	7BJ	6.3	0.2	—	—	—	Detector	R.m.s. voltage per plate=75 volts; d.c. output=3.5 ma. with 25000 ohms and 8 μμfd. load; peak current per plate=10 ma.; peak inverse voltage=210.										—
6AQ5	Beam Power Tetrode	B.	7BZ	6.3	0.45	7.6	6.0	0.35	Class-A_1 Amp.	180	−8.5	180	4.0[2]	30[2]	58000	3700	29[5]	5500	2.0	6V6GT
										250	−12.5	250	7.0[2]	47[2]	52000	4100	45[5]	5000	4.5	
6AQ6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.15	1.7	1.5	1.80	Class-A Triode	250	−3.0	—	—	1.0	58000	1200	70	—	—	6T7G
										100	−1.0	—	—	0.8	61000	1150	70	—	—	
6AR5	Pentode Power Amp.	B.	6CC	6.3	0.4	—	—	—	Class-A_1 Amp.	250	−18	250	5.5[2]	33[2]	68000	2300	—	7600	3.4	6K6GT
										250	−16.5	250	5.5[2]	35[2]	65000	2400	—	7000	3.2	
6AS5	Beam Pentode	B.	7CV	6.3	0.8	12	6.2	0.6	Class-A_1 Amp.	150	−8.5	110	2/6.5	35/36	—	5600		4500	2.2	—
6AS6	Sharp Cut-off Pentode	B.	7CM	6.3	0.175	4.0	3.0	0.02	Class-A Amp.	120	−2	120	3.5	5.5	—	3500	—	—	—	—
6AT6	Duplex Diode Triode	B.	7BT	6.3	0.3	2.3	1.1	2.10	Class-A Amp.	250	−3	—	—	1.0	58000	1200	70	—	—	6Q7GT
6AU6	Sharp Cut-off Pentode	B.	7BK	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	−1	150	4.3	10.8	2000000	5200	—	—	—	6SH7GT
6AV6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.3	—	—	—	Class-A_1 Amp.	250	−2	—	—	1.2	62500	1600	100	—	—	6SQ7GT
6BA6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11	1500000	4400	—	—	—	6SG7GT
6BA7	Pentagrid Converter	B.	8CT	6.3	0.3	9.5	8.3	—	Converter	250	−1	100	10	3.8	1000000	3.5	—	—	—	—
6BD6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	—	—	—	Class-A Amp.	100	−1	100	5	13	120000	2350	—	—	—	6SK7GT
										250	−3	100	3.5	9	700000	2000	—	—	—	
6BE6	Pentagrid Converter	B.	7CH	6.3	0.3	Osc. Grid 50000 Ω			Converter	250	−1.5	100	7.8	3.0	1000000	475	—	—	—	6SA7GT
6BF6	Duplex-Diode Triode	B.	7BT	6.3	0.3	1.8	1.1	2.0	Class-A_1 Amp.	250	−9	—	—	9.5	8500	1900	16	10000	—	6SR7GT
6BH6	Sharp Cut-off Pentode	B.	7CM	6.3	0.15	5.4	4.4	0.0035	Class-A_1 Amp.	250	−1	150	2.9	7.4	1400000	4600	—	—	—	—
6BJ6	Remote Cut-off Pentode	B.	7CM	6.3	0.15	4.5	5.0	.0035	Class-A_1 Amp.	250	−1	100	3.3	9.2	1300000	3800	—	—	—	6SS7GT
6C4	Triode Amplifier	B.	6BG	6.3	0.15	1.8	1.3	1.60	Class-A_1 Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	6J5GT
6J4	U.h.f. Grounded-Grid R.F. Amplifier	B.	7BQ	6.3	0.4	5.5	0.24	4.0	Grounded-Grid	150	200*	—	—	15.0	4500	12000	55	—	—	—
									Class-A_1 Amp.	100	100*	—	—	10.0	5000	11000	55	—	—	
6J6	Twin Triode	B.	7BF	6.3	0.45	2.2	0.4	1.6	Class-A_1 Amp. Mixer, Oscillator	100	50*	—	—	8.5	7100	5300	38	—	—	—
6N4	U.h.f. Triode Amplifier	B.	7CA	6.3	0.2	3.0	1.6	1.10	Class-A Amp.	180	−3.5	—	—	12.0	—	6000	32	—	—	—
6T8	Triple-Diode Triode	B.	9E	6.3	0.3	1.5	1.1	2.4	Class-A_1 Amp.	250	−3	—	—	1.0	5800	1200	70	—	—	—
										100	−1	—	—	0.8	5400	1300	70	—	—	
12AL5	Twin Diode	B.	6BT	12.6	0.15	2.5	—	—	Detector		R.m.s. voltage per plate=117; d.c. output=9 ma. per plate; peak ma. per plate=54; peak inverse voltage=330.									12H6GT
12AT6	Duplex Diode Triode	B.	7BT	12.6	0.15	2.3	1.1	2.10	Class-A Amp.	250	−3.0	—	—	1.0	58000	1200	70	—	—	12Q7GT
12AT7	Double Triode	B.	9A	6.3	0.3	2.5[7]	0.45[7]	1.45[7]	Class-A_1 Amp. Each Unit	250	−2	—	—	10	10000	5500	55	—	—	—
				12.6	0.15	2.5[8]	0.35[8]	1.45[8]		180	−1	—	—	11	9400	6600	62	—	—	
12AU6	Sharp Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A_1 Amp.	250	−1.0	150	4.3	10.8	1 meg.	5200	—	—	—	12SH7GT
12AU7	Twin-Triode Amplifier	B.	9A	6.3	0.3	1.6[7]	0.5[7]	1.5[7]	Class-A_1 Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	12SN7GT
				12.6	0.15	1.6[8]	0.35[8]	1.5[8]												
12AV6	Duodiode Hi-mu Triode	B.	7BT	12.6	0.15	—	—	—	Class-A_1 Amp.	250	−2	—	—	1.2	62500	1600	100	—	—	—
12AW6	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Pentode Amp.	250	200*	150	2.0	7.0	800000	5000	—	—	—	—
									Triode Amp.[9]	250	825*	—	—	5.5	11000	3800	42	—	—	
12AW7	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Class-A_1 Amp.	250	200*	150	2.0	7.0	0.8 meg.	5000	—	—	—	—
12AX7	Double Triode	B.	9A	12.6	0.15	1.6[7]	0.46[7]	1.7[7]	Class-A_1 Amp.	250	−2	—	—	1.2[1]	62500	1600	100	—	—	—
				6.3	0.3	1.6[8]	0.34[8]	1.7[8]		100	−1	—	—	0.5[1]	8000	1250	100	—	—	

TABLE XI — MINIATURE RECEIVING TUBES — *Continued*

Type	Name	Base	Socket Connections[1]	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
12BA6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11.0	1500000	4400	—	—	—	12SG7G
12BA7	Pentagrid Converter	B.	8CT	12.6	0.15	9.5	8.3	—	Converter	250	−1	100	10	3.8	1000000	3.5	—	—	—	—
12BD6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	4.3	5.0	.004	Class-A Amp.	250	−3	100	3.5	9.0	700000	2000	—	—	—	12SK7GT
12BE6	Pentagrid Converter	B.	7CH	12.6	0.15	Osc. Grid 50000 Ω			Converter	250	−1.5	100	7.8	3.0	1000000	475	—	—	—	12SA7GT
12BF6	Duodiode Triode	B.	7BT	12.6	0.15	1.8	1.1	2.00	Class-A Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	12SR7GT
19J6	Twin Triode	B.	7BF	18.9	0.15	2.0	0.4	1.5	Class-A_1 Amp.	100	50*	—	—	8.5[1]	7100	5300	38	—	—	—
19T8	Triple-Diode Triode	B.	9E	18.9	0.15	1.5	1.1	2.4	Class-A_1 Amp.	250	−3	—	—	1.0	5800	1200	70	—	—	—
26A6	Remote Cut-off Pentode	B.	7BK	26.5	0.07	6.0	5.0	.0035	Class-A_1 Amp.	250	125*	100	4	10.5	1000000	4000	—	—	—	—
26C6	Duplex-Diode Triode	B.	7BT	26.5	0.07	1.8	1.4	2	Class-A_1 Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	—
26D6	Pentagrid Converter	B.	7CH	26.5	0.07	Osc. Grid 20000 Ω			Converter	250	−1.5	100	7.8	3.0	1000000	4 5	—	—	—	—
35B5	Beam Power Amplifier	B.	7BZ	35	0.15	11	6.5	0.4	Class-A_1 Amp.	110	−7.5	110	7[2]	41[2]	—	5800	40[6]	2500	1.5	35L6GT
35C5	Beam Power Amplifier	B.	7CV	35	0.15	12	6.2	0.57	Class-A_1 Amp.	110	−7.5	110	3/7	40/41	—	5800		2500	1.5	—
50B5	Beam Power Amplifier	B.	7BZ	50	0.15	13	6.5	0.50	Class-A Amp.	110	−7.5	110	4.0	49.0	14000	7500	—	3000	1.9	50L6GT
50C5	Beam Power Amplifier	B.	7CV	50	0.15	—	—	—	Class-A_1 Amp.	110	−7.5	110	4/8.5	49/50	10000	7500	—	2500	1.9	—
5590	Pentode	B.	7BD	6.3	0.15	3.4	2.9	0.01	Class-A_1 Amp.	90	820*	90	1.4	3.9	300000	2000	—	—	—	—
5591	R.F. Pentode	B.	7BD	6.3	0.15	3.9	2.85	0.01	Class-A_1 Amp.	180	200*	120	2.4	1.7	690000	5100	3500	—	—	—
9001	Sharp Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	−3.0	100	0.7	2.0	1 meg.+	1400	—	—	—	—
									Mixer	250	−5.0	100	Osc. peak voltage 4 volts			550	—	—	—	
9002	Triode Detector, Amplifier, Oscillator	B.	7TM	6.3	0.15	1.2	1.1	1.40	Class-A Amp.	250	−7.0	—	—	6.3	11400	2200	25	—	—	—
										90	−2.5	—	—	2.5	14700	1700	25	—	—	
9003	Remote Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	−3.0	100	2.7	6.7	700000	1800	—	—	—	—
									Mixer	250	−10.0	100	Osc. peak voltage 9 volts			600	—	—	—	
9006	U.h.f. Diode	B.	6BH	6.3	0.15	—	—	—	Detector	Max. a.c. voltage—270. Max. d.c. output current—5 ma.										—

* Cathode resistor—ohms.

[1] Per Plate.
[2] Maximum-signal current for full-power output.
[3] Values are for two tubes in push-pull.
[4] Also no-signal plate ma. when so indicated.
[5] No signal plate ma.
[6] Effective plate-to-plate.
[7] Triode No. 1.
[8] Triode No. 2.
[9] Grid No. 2 tied to plate and No. 3 to cathode.

TABLE XII — SUB-MINIATURE TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
1C8	Heptode	—	—	1.25	0.04	6.5	4.0	0.25	Converter	30	0	30	0.75	0.32	300000	100	—	—	—	1C8
1V5	Audio Pentode	[1]	[2]	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	−4.5	67.5	0.4	2.0	150000	750	—	25000	0.05	1V5
1W5	Sharp Cut-off Pentode	[1]	[2]	1.25	0.04	2.3	3.5	0.01	Class-A_1 Amp.	67.5	0	67.5	0.75	1.85	700000	735	—	—	—	1W5
2E31	R.F. Pentode	[1]	[2]	1.25	0.05				Class-A_1 Amp.	22.5	0	22.5	0.3	0.4	—	500	—	—	—	2E31
2E32	R.F. Pentode	[1]	[2]	1.25	0.05				Class-A Amp.	22.5	0	22.5	0.3	0.4	350000	500	—	—	—	2E32
2E35	Audio Pentode	[1]	[2]	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	—	385	—	—	.0012	2E35
2E36	Audio Pentode	[1]	[2]	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	220000	385	—	150000	0.0012	2E36
										45	−1.25	45	0.11	0.45	250000	500	—	100000	0.006	
2E41	Diode Pentode	[1]	[2]	1.25	0.03	—	—	—	Detector Amp.	22.5	0	22.5	0.12	0.35	—	—	—	—	—	2E41
2E42	Diode Pentode	[1]	[2]	1.25	0.03				Detector Amp.	22.5	0	22.5	0.12	0.35	250000	375	—	1 meg.	—	2E42
2G21	Triode Heptode	[1]	[2]	1.25	0.05	—	—	—	Converter	22.5	—	22.5	0.2	0.3	—	75	—	—		2G21
2G22	Converter	[1]	[2]	1.25	0.05				Converter	22.5	0	22.5	0.3	0.2	500000	60	—	—	—	2G22
6K4	Triode	[1]	[2]	6.3	0.15	2.4	0.8	2.4	Class A_1 Amp.	200	680*	—	—	11.5	4650	3450	16	—	—	6K4

TABLE XII — SUB-MINIATURE TUBES — *Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
1247	Diode	[1]	[2]	0.7	0.065	—	—	—	R.F. Probe			Max. a.c. volts—300 r.m.s.			D.C. plate current—0.4 Ma.					1247
CK501	Pentode Voltage Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.06 0.055	0.3 0.28	1000000 1500000	325 300	—	—	—	CK501
CK502	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.13	0.55	500000	400	—	60000	0.003	CK502
CK503	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.33	1.5	150000	600	—	20000	0.006	CK503
CK504	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	−1.25	30	0.09	0.4	500000	350	—	60000	0.003	CK504
CK505	Pentode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.07 0.08	0.17 0.2	1100000 2000000	140 150	—	—	—	CK505
CK506	Pentode Output Amplifier	—[1]	[2]	1.25	0.05	—	—	—	Class-A₁ Amp.	45	−4.5	45	0.4	1.25	120000	500	—	30000	0.025	CK506
CK507	Pentode Output Amplifier	—[1]	[2]	1.25	0.05	—	—	—	Class-A₁ Amp.	45	−2.5	45	0.21	0.6	360000	500	—	50000	0.010	CK507
CK509	Triode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	150000	160	16	1000000	—	CK509
CK510	Dual Space-Charge Tetrode	—[1]	[2]	0.625	0.05	—	—	—	Class-A Amp.	45	0	0.2	200 μa	60 μa	500000	65	32.5	—	—	CK510
CK512	Low Microphonic Pentode	[1]	[2]	0.625	0.02	—	—	—	Voltage Amp.	22.5	0	22.5	0.04	0.125	—	160	—	—	—	CK412
CK515BX	Triode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	—	160	24	1000000	—	CK515BX
CK520AX	Audio Pentode	[1]	[2]	0.625	0.05	—	—	—	Class-A₁ Amp.	45	−2.5	45	0.07	0.24	—	180	—	—	0.0045	CK520AX
CK521AX	Audio Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A₁ Amp.	22.5	−3	22.5	0.22	0.8	—	400	—	—	0.006	CK521AX
CK522AX	Audio Pentode	[1]	[2]	1.25	0.02	—	—	—	Class-A₁ Amp.	22.5	0	22.5	0.08	0.3	—	450	—	—	0.0012	CK522AX
CK551AXA	Diode Pentode	[1]	[2]	1.25	0.03	—	—	—	Detector-Amp.	22.5	0	22.5	0.04	0.17	—	235	—	—	—	CK551AXA
CK553AXA	R.F. Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A₁ Amp.	22.5	0	22.5	0.13	0.42	—	550	—	—	—	CK553AXA
CK556AX	U.h.f. Triode	[1]	[2]	1.25	0.125	—	—	—	R.F. Oscillator	135	−5	—	—	4.0	—	1600	—	—	—	CK556AX
CK568AX	U.h.f. Triode	[1]	[2]	1.25	0.07	—	—	—	R.F. Oscillator	135	−6	—	—	1.9	—	650	—	—	—	CK568AX
CK569AX	R.F. Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A₁ Amp.	67.5	0	67.5	0.48	1.8	—	1100	—	—	—	CK569AX
CK650AX	Sharp Cut-off Pentode	[1]	[2]	6.3	0.2	—	—	—	Class-A₁ Amp.	120	−2	120	2.5	7.5	—	5000	—	—	—	CK650AX
CK606BX	Single Diode	[1]	[2]	6.3	0.15	—	—	—	Detector	150 a.c.	—	—	—	9.0 d.c.	—	—	—	—	—	CK606BX
CK608CX	U.h.f. Triode	[1]	[2]	6.3	0.2	—	—	—	500-Mc. Osc.	120	−2	—	—	9.0	—	5000	—	—	0.75	CK608CX
CK619CX	Hi-Mu Triode	[1]	[2]	6.3	0.2	—	—	—	Class-A₁ Amp.	250	−2	—	—	4.0	—	4000	—	—	—	CK619CX
HY113 HY123	Triode Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45	−4.5	—	—	0.4	25000	250	6.3	40000	0.0065	HY113 HY123
HY115 HY145	Pentode Voltage Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	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	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45 90	−3.0 −7.5	45 90	0.2 0.5	0.9 2.6	825000 420000	310 450	255 190	50000 28000	0.0115 0.09	HY125 HY155
M54	Tetrode Power Amplifier	[1]	[2]	0.625	0.04	—	—	—	Class-A Amp.	30	0	30	0.06	0.5	130000	200	25	35000	0.005	M54
M64	Tetrode Voltage Amplifier	[1]	[2]	0.625	0.02	—	—	—	Class-A Amp.	30	0	—	—	0.03	200000	110	25	—	—	M64
M74	Tetrode Voltage Amplifier	[1]	[2]	0.625	0.02	—	—	—	Class-A Amp.	30	0	7.0	0.01	0.02	500000	125	70	—	—	M74
RK61	Gas Triode	[1]	[2]	1.4	0.05	—	—	—	Radio Control	45	—	—	—	1.5	—	—	—	—	—	RK61
SD917A 5637	Triode	[1]	[2]	6.3	0.15	2.6	0.7	1.4	Class-A₁ Amp.	100	820*	—	—	1.4	26000	2700	70	—	—	SD917A 5637
SD828A 5638	Audio Pentode	[1]	[2]	6.3	0.15	4.0	3.0	0.22	Class-A₁ Amp.	100	270*	100	1.25	4.8	150000	3300	—	—	—	SD828A 5638
SD828E 5634	Sharp Cut-off Pentode	[4]	—	6.3	0.15	4.4	2.8	0.01	Class-A₁ Amp.	100	150*	100	2.5	6.5	240000	3500	—	—	—	SD828E 5634
SN944 5633	Remote Cut-off Pentode	[4]	—	6.3	0.15	4.0	2.8	0.01	Class-A₁ Amp.	100	150*	100	2.8	7.0	200000	3400	—	—	—	SN944 5633
SN946	Diode	[1]	[2]	6.3	0.15	1.8	—	—	Rectifier	150	—	—	—	9.0	—	—	—	—	—	SN946
SN947C 5640	Audio Beam Tetrode	[1]	[2]	6.3	0.45	—	—	—	Class-A₁ Amp.	100	−9	100	2.2	31.0	15000	5000	—	3000	1.25	SN947C 5640

TABLE XII — SUB-MINIATURE TUBES — *Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	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
SN954 5641	Half-Wave Rectifier	1	2	6.3	0.45	—	—	—	Rectifier	300	—	—	—	45.0	—	—	—	—	—	SN954 5641
SN955B	Dual Triode	1	2	6.3	0.45	2.8	1.0	1.3	Class-A₁ Amp. [5]	100	100*	—	—	5.5	8000	4250	34	—	—	SN955B
SN956B 5642	H.V. Half-Wave Rectifier	4	—	1.25	0.14	—	—	—	H.V. Rectifier	Peak inverse V. = 10000 Max. Average Ip = 2 Ma. Peak Ip = 23 Ma.										SN956B 5642
SN957A 5645	Triode	1	2	6.3	0.15	2.0	1.0	1.8	Class-A₁ Amp.	100	560*	—	—	5.0	7400	2700	20	—	—	SN957A 5645
SN1006	Triode	1	2	6.3	0.15				Class-A₁ Amp.	100	820*	—	—	1.4	29000	2400	70	—	—	SN1006
SN1007A	Mixer	4	—	6.3	0.15	5.0	2.8	0.003	Mixer	100	150*	100	5.0	4.0	230000	900	—	—	—	SN1007A

* Cathode resistor ohms. [1] No base; tinned wire leads. [2] Leads identified on tube. [3] No screen connection. [4] Double-ended type. [5] Values per triode.

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
0A2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	185	150	5–30	—	—	0A2
0B2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	133	108	5–30	—	—	0B2
0A4G 1267	Gas Triode Starter-Anode Type	6-pin O.	4V 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 1267
1B47	Voltage Regulator	7-pin B.	—	—	—	—	Voltage Regulator	—	—	225	82	1–2	—	—	1B47
1C21	Gas Triode Glow-Discharge Type	6-pin O.	4V	Cold	—	—	Relay Tube	125–145	25	66 [6]	—	—	—	73	1C21
							Voltage Regulator		0.1 [6]	180 [4]				55	
2A4G	Gas Triode Grid Type	7-pin O.	5S	Fil.	2.5	2.5	Control Tube	200	100	—	—	—	—	15	2A4G
6Q5G	Gas Triode Grid Type	8-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	1.0	0.1–10 [7]	19	6Q5G
2B4		5-pin M.	5A	Htr.	2.5	1.4									2B4
2C4	Gas Triode	7-pin B.	5AS	Fil.	2.5	0.65	Control Tube	Plate volts = 350; Grid volts = −50; Avg. Ma. = 5; Peak Ma. = 20; Voltage drop = 16.							2C4
2D21	Gas Tetrode	7-pin B.	7BN	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	650	100	0.1–10 [7]	8	2D21
							Relay Tube	400	—	—	—	—	1.0 [7]	—	
3C23	Gas and Mercury Vapor Grid Type	4-pin M.	3G	Fil.	2.5	7.0	Grid-Controlled Rectifier	1000	6000	—	500	1500	−4.5 [8]	15	3C23
										—	100	1500	−2.5 [8]	15	
6D4	Gas Triode	7-pin B.	5AY	Htr.	6.3	0.25	Control Tube	Plate volts = 350; Grid volts = −50; Avg. Ma. = 25; Peak Ma. = 100; Voltage drop = 16.							6D4
17	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	7500 [5]	2000	—	—	500	200–3000	—	17
								2500		−5 [3]	1000	250	—	10–24	
874	Voltage Regulator	4-pin M.	4S	—	—	—	Voltage Regulator	—	—	125	90	10–50	—	—	874
876	Current Regulator	Mogul	—	—	—	—	Current Regulator	—	—	—	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	—	884
							Grid-Controlled Rectifier	350	300	—	—	75	25000	—	
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	—	—	—	40–60	2.05	—	—	886
967	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	2500	500	−5 [3]	—	—	—	10–24	967
991	Voltage Regulator	Bayonet	—	—	—	—	Voltage Regulator	—	—	87	55–60	2.0	—	—	991
1265	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	130	90	5–30	—	—	1265
1266	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	—	70	5–40	—	—	1266
1267	Gas Triode	6-pin O.	4V	Cold	—	—	Relay Tube	Characteristics same as OA4G							1267
2050	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	—	100	0.1–10 [7]	8	2050

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
2051	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	350	375	—	—	75	0.1–10[7]	14	2051
2523N1/ 128AS	Gas Triode Grid Type	5-pin M.	5A	Htr.	2.5	1.75	Relay Tube	400	300	—	—	1.0	300[7]	13	2523N1/ 128AS
5651	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	115	—	115	87	1.5–3.5	—	—	5651
KY21	Gas Triode Grid Type	4-pin M.	—	Fil.	2.5	10.0	Grid-Controlled Rectifier	—	—	—	3000	500	—	—	KY21
RK61	Thyratron	—[9]	—	Fil.	1.4	0.05	Radio-Controlled Relay	45	1.5	30	—	0.5–1.5	3[7]	30	RK61
RK62	Gas Triode Grid Type	4-pin S.	4D	Fil.	1.4	0.05	Relay Tube	45	1.5	—	30–45	0.1–1.5	—	15	RK62
RM208	Permatron	4-pin M.	—	Fil.	2.5	5.0	Controlled Rectifier[1]	7500[2]	1000	—	—	—	—	15	RM208
RM209	Permatron	4-pin M.	—	Fil.	5.0	10.0	Controlled Rectifier[1]	7500[2]	5000	—	—	—	—	15	RM209
OA3/VR75	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	105	75	5–40	—	—	OA3/VR75
OB3/VR90	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	125	90	5–40	—	—	OB3/VR90
OC3/VR105	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	135	105	5–40	—	—	OC3/VR105
OD3/VR150	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	185	150	5–40	—	—	OD3/VR150
KY866	Mercury Vapor Triode	4-pin M.	Fig. 8	Fil.	2.5	5.0	Grid-Controlled Rectifier	10000	1000	0–150	—	—	—	—	KY866

[1] For use as grid-controlled rectifier or with external magnetic control. RM-208 has characteristics of 866, RM-209 of 872.
[2] When under control peak inverse rating is reduced to 2500.
[3] At 1000 anode volts.
[4] Grid tied to plate.
[5] Peak inverse voltage.
[6] Grid.
[7] Megohms.
[8] Grid voltage.
[9] No base. Tinned wire leads.

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage[1]	Screen Input Power[2]	Deflection Sensitivity[6] D_1 D_2	Deflection Sensitivity[6] D_3 D_4	Anode No. 3 Voltage	Pattern Color	Type
2AP1[7]-11	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillograph Television	2″	1000	250	− 60	—	—	660	—	0.11	0.13	—	Green	2AP1-11
							500	125	− 30	—	—			0.22	0.26			
2BP1-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	2″	2000	300/560	−135	—	—	500	—	270[3]	174[3]	—	Green	2BP1-11
							1000	150/280	−67.5	—	—	500	—	135[3]	87[3]	—		
3AP1/ 906-P1-4-5-11[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	1500	430	− 50	—	—	550	10	0.22	0.23		Green Blue White	3AP1/ 906-P1-4-5-11
							1000	285	− 33	—	—			0.33	0.35			
							600	170	− 20	—	—			0.55	0.58			
3BP1-4-11	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph	3″	2000	575	− 60	—	—	550	—	0.13	0.17	—	Green	3BP1-4-11
							1500	430	− 45	—	—			0.17	0.23			
3DP1	Electrostatic Cathode-Ray	Fig. 49	6.3	0.6	Oscillograph	3″	2000	575	− 60	—	—	550	—	200[3]	148[3]	—	Green	3DP1
							1500	430	− 40	—	—			150[3]	111[3]	—		
3EP1/ 1806-P1	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph Television	3″	2000	575	− 60	—	—	550	—	0.115	0.154	—	Green	3EP1/ 1806-P1
							1500	430	− 45	—	—			0.153	0.205			
3GP1-4-5-11	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	3″	1500	350	− 50	—	—	550	—	0.21	0.24	—	White Green Blue	3GP1-4-5-11
							1000	234	− 33	—	—		—	0.32	0.36			
3JP1-2-4-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	3″	2000	575	− 60	—	—	550	—	0.13	0.17	4000	Green Blue White	3JP1-2-4-7-11
							1500	430	− 45	—	—		—	0.17	0.23	3000		
3KP1	Electrostatic Cathode-Ray	11M	6.3	0.6	Oscillograph	3″	1000	300	− 45	1000	—	500	—	68[3]	136[3]	—	Green	3KP1
							2000	600	− 90	2000	—			52[3]	104[3]			

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amp.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [1]	Screen Input Power [2]	Deflection Sensitivity [6] D1 D2	Deflection Sensitivity [6] D3 D4	Anode No. 3 Voltage	Pattern Color	Type
3MP1	Electrostatic Cathode-Ray	Fig. 2	6.3	0.6	Oscillograph	3″	1000	200/350	− 68	—	—			190 [3]	180 [3]	—	Green	3MP1
5AP1/ 1805-P1 5AP4/ 1805-P4 [7]	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5″	2000 1500	575 430	− 35 − 27	—	—	500	10	0.17 0.23	0.21 0.28	— —	Green White	5AP1/ 1805-P1 5AP4/ 1805-P4
5BP1/ 1802-P1-2-4-5-11	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph	5″	2000 1500	450 337	− 40 − 30	— —	— —	500	10	0.3 0.4	0.33 0.45	—	Green White Blue	5BP1/ 1802-P1-2-4-5-7-11
5CP1-2-4-5-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph Television	5″	2000 1500 2000	575 430 575	− 60 − 45 − 60	— — —	— — —	550	— — —	0.28 0.37 0.36	0.32 0.43 0.41	4000 3000 2000	White Green Blue	5CP1-2-4-5-11
5FP1-2-4-11 [7]	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	5″	7000 4000	250 250	− 45 − 45	— —	— —	—	— —	— —	— —	—	Green White Blue	5FP1-2-4-11
5HP1 5HP4 [7]	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	— —	Green White	5HP1 5HP4
5JP1-2-4-5-11	Electrostatic Cathode-Ray	11E	6.3	0.6	Oscillograph	5″	2000 1500	520 390	− 75 − 55	— —	— —	500	— —	0.25 0.33	0.28 0.37	4000 3000	White Green Blue	5JP1-2-4-5-11
5LP1-2-4-5-11	Electrostatic Cathode-Ray	11F	6.3	0.6	Oscillograph Television	5″	2000 1500 1000	500 375 250	− 60 − 45 − 30	— — —	— — —	500	— — —	0.25 0.33 0.49	0.28 0.37 0.56	4000 3000 2000	White Green Blue	5LP1-2-4-5-11
5MP1-4-5-11	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	5″	1500 1000	375 250	− 50 − 33	— —	— —	660	— —	0.39 0.58	0.42 0.64	—	White Green Blue	5MP1-4-5-11
5RP1-2-4-11	Electrostatic Cathode-Ray	14F	6.3	0.6	Oscillograph	5″	3000 2000	— 575	− 90 − 60	— —	— —	1200	— —	0.12 0.18	0.12 0.18	15000 10000	Green White Blue	5RP1-2-4-11
5TP4	Projection Kinescope	12C	6.3	0.6	Television	5″	27000	4900	− 70	200	—	—	—	—	—	—	White	5TP4
5UP1-7-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	5″	2500 2500 1000 1000	640 340 320 170	− 90 − 90 − 45 − 45	— — — —	— — — —	500 500 500 500	— — — —	38.5 [3] 28 [3] 31 [3] 23 [3]	77 [3] 56 [3] 62 [3] 46 [3]	— — — —	Green Yel-low Blue	5UP1-7-11
7AP4	Electromagnetic Picture Tube	5AJ	2.5	2.1	Television	7″	3500	1000	−67.5	—	—	—	2.5	—	—	—	White	7AP4
7BP1-2-4-7-11	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	7″	7000 4000	250 250	− 45 − 45	— —	— —	—	—	—	—	—	White Green Blue	7BP1-2-4-7-11
7CP1/ [5] 1811-P1	Electromagnetic Cathode-Ray	6AZ	6.3	0.6	Oscillograph	7″	7000 4000	1470 840	− 45 − 45	250 250	— —	—	—	— —	— —	—	Green	7CP1/ 1811-P1
7DP4	Kinescope	12C	6.3	0.6	Television	7″	6000	1430	− 45	250	—	—	—	—	—	—	White	7DP4
7EP4	Electrostatic Cathode-Ray	11N	6.3	0.6	Television	7″	2500	650	− 60	—	38	—		110 [3]	95 [3]	—	White	7EP4
7GP4	Electrostatic Kinescope	Fig. 47	6.3	0.6	Television	7″	3000	1200	− 84	3000	—	—	—	123 [3]	102 [3]	—	White	7GP4
7JP4	Electrostatic Kinescope	14G	6.3	0.6	Television	7″	6000	2400	−168					246 [3]	204 [3]	—	White	7JP4
9AP4/ 1804-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	9″	7000 6000	1425 1225	− 40 − 38	250	25	—	10	—	—	—	White	9AP4/ 1804-P4
9CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	9″	7000	—	−110	—	25	—	10	—	—	—	White	9CP4
9JP1/ 1809-P1	Electrostatic-Magnetic Cathode-Ray	8BR	2.5	2.1	Oscillograph	9″	5000 2500	1570 785	− 90 − 45	—	—	3000	—	0.136 0.272	— —	—	Green	9JP1/ 1809-P1

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage[1]	Screen Input Power[2]	Deflection Sensitivity[6] D_1 D_2	Deflection Sensitivity[6] D_3 D_4	Anode No. 3 Voltage	Pattern Color	Type
10BP4	Magnetic Kinescope	12D	6.3	0.6	Television	10″	—	9000	−45	250	—	—	—	—	—	—	—	10BP4
10EP4	Magnetic-Focus Cathode-Ray	12D	6.3	0.6	Television	10½″	—	8000	−45	250	.38	—	—	—	—	—	White	10EP4
10FP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	10″	—	9000	−27/−63	250	—	—	—	—	—	—	White	10FP4
12AP4/ 1803-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	12″	7000 6000	1460 1240	−75	250	25	—	10	—	—	—	White	12AP4/ 1803-P4
12CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	12″	7000	—	−110	—	25	—	10	—	—	—	White	12CP4
12DP4-7	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Television	12″	7000 4000	250 250	−45 −45	— —	— —	— —	— —	— —	— —	— —	White	12DP4
15AP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	15″	—	8000	−45	250	38	—	—	—	—	—	White	15AP4
20BP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	20″	—	15000	−45	250	36	—	—	—	—	—	White	20BP4
902[7]	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	150	−60	—	—	350	5	0.19	0.22	—	Green	902
903[5]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9″	7000	1360	−120	250	—	—	10	—	—	—	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3	2.5	2.1	Oscillograph	5″	4600	970	−75	250	—	4000	10	0.09	—	—	Green	904
905[7]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	2000	450	−35	—	—	1000	10	0.19	0.23	—	Green	905
907	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	Characteristics same as Type 905							—	—	—	Blue	907
908[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	908
908-A	Electrostatic Cathode-Ray	7CE	2.5	2.1	Oscillograph	3″	1500 1000	430 287	−50 −33	— —		500 500	— —	0.223 0.334	0.233 0.348	— —	Blue	908-A
909[5]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	Characteristics same as Type 905							—	—	—	Blue	909
910[5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	910
911[5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Green	911
912	Electrostatic Cathode-Ray	Fig. 8	2.5	2.1	Oscillograph	5″	10000	2000	−66	250	—	7000	10	0.041	0.051	—	Green	912
913	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	1″	500	100	−65	—	—	250	5	0.07	0.10	—	Green	913
914[7]	Electrostatic Cathode-Ray	Fig. 12	2.5	2.1	Oscillograph	9″	7000	1450	−50	250	—	3000	10	0.073	0.093	—	Green	914
1800[5]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	6000	1250	−75	250	25	—	10	—	—	—	Yellow	1800
1801[5]	Electromagnetic Kinescope	Fig. 13	2.5	2.1	Television	5″	3000	450	−35	—	20	—	10	—	—	—	Yellow	1801
2001	Electrostatic Cathode-Ray	4AA	6.3	0.6	Oscillograph	1″				Characteristics essentially same as 913								2001
2002	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	120	—	—	—	—	—	0.16	0.17	—	Green	2002
2005	Electrostatic Cathode-Ray	Fig. 1[4]	2.5	2.1	Television	5″	2000	1000	−35	200	—	—	10	0.5	0.55	—	—	2005
24-XH	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscilloscope	2″	600	120	−60	—	—	—	10	0.14	0.16	—	Blue	24-XH

[1] Between Anode No. 2 and any deflecting plate.
[2] In mw./sq. cm., max.
[3] D.c. Volts/in.
[4] Cathode connected to Pin 7.
[5] Discontinued.
[6] In mm./volt d.c.
[7] Superseded by same type with suffix "A."

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
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.	4H	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	HV
OY4	Half-Wave Rectifier	5-pin O.	4BU	Cold	Connect Pins 7 and 8		95	75	300	500	G
OZ4	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	350	30–75	1250	200	G
1	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	1000	400	MV
1-V	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	—	—	HV
1B3GT/8016	Half-Wave Rectifier	6-pin O.	3C	Fil.	1.25	0.2	—	2.0	4000	17	HV
1B48	Half-Wave Rectifier	7-pin B.	—	Cold	—	—	800	6	2700	50	G
1Z2	Half-Wave Rectifier	7-pin B.	7CB	Fil.	1.5	0.3	7800	2	20000	10	HV
2B25	Half-Wave Rectifier	7-pin B.	3T	Fil.	1.4	0.11	1000	1.5	—	9	HV
2V3G	Half-Wave Rectifier	6-pin O.	4Y	Fil.	2.5	5.0	—	2.0	16500	12	HV
2W3	Half-Wave Rectifier	5-pin O.	4X	Fil.	2.5	1.5	350	55	—	—	HV
2X2/879[10]	Half-Wave Rectifier	4-pin S.	4AB	Htr.	2.5	1.75	4500	7.5	—	—	HV
2X2-A	Half-Wave Rectifier	4-pin S.	4AB	Same as 2X2/879 but will withstand severe shock & vibration							HV
2Y2	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4400	5.0	—	—	HV
2Z2/G84	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	1.5	350	50	—	—	HV
3B24	Half-Wave Rectifier	4-pin M.	T-4A	Fil.	5.0 2.5[9]	3.0 3.0	— —	60 30	20000 20000	300 150	HV
3B25	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	500	4500	2000	G
3B26	Half-Wave Rectifier	8-pin O.	Fig. 31	Htr.	2.5	4.75	—	20	15000	8000	HV
DR-3B27	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	5.0	3000	250	8500	1000	HV
5AZ4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				HV
5R4GY	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	900[4] 950[7]	150[4] 175[7]	2800	650	HV
5T4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	3.0	450	250	1250	800	HV
5U4G	Full-Wave Rectifier	8-pin O.	5T	Fil.	5.0	3.0	Same as Type 5Z3				HV
5V4G	Full-Wave Rectifier	8-pin O.	5L	Htr.	5.0	2.0	Same as Type 83V				HV
5W4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	1.5	350	110	1000	—	HV
5X3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	1275	30	—	—	HV
5X4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	3.0	Same as 5Z3				HV
5Y3G	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				HV
5Y4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	2.0	Same as Type 80				HV
5Z3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	—	HV
5Z4	Full-Wave Rectifier	5-pin O.	5L	Htr.	5.0	2.0	400	125	1100	—	HV
6W4GT	Damper Service Half-Wave Rectifier	6-pin O.	4CG	Htr.	6.3	1.2	— —	125 125	2000 1250	600 600	HV
6W5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.9	350	100	1250	350	HV
6X4	Full-Wave Rectifier	7-pin B.	7CF	Htr.	6.3	0.6	325	70	1250	210	HV
6X5	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.5	350	75	—	—	HV
6Y3G	Half-Wave Rectifier	5-pin O.	4AC	Htr.	6.3	0.7	5000	7.5	—	—	HV
6Y5[10]	Full-Wave Rectifier	6-pin S.	6J	Htr.	6.3	0.8	350	50	—	—	HV
6Z3	Half-Wave Rectifier	4-pin M.	4G	Fil.	6.3	0.3	350	50	—	—	HV
6Z5[10]	Full-Wave Rectifier	6-pin S.	6K	Htr.	6.3	0.6	230	60	—	—	HV
6ZY5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.3	350	35	1000	150	HV
7Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.5	350	60	—	—	HV
7Z4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.9	450[1] 325[4]	100	1250	300	HV
12A7	Rectifier-Pentode	7-pin S.	7K	Htr.	12.6	0.3	125	30	—	—	HV
12Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
12Z5	Voltage Doubler	7-pin M.	7L	Htr.	12.6	0.3	225	60	—	—	HV
14Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	12.6	0.3	450[1] 325[4]	70	1250	210	HV
14Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
25A7G[10]	Rectifier-Pentode	8-pin O.	8F	Htr.	25	0.3	125	75	—	—	HV
25X6GT	Voltage Doubler	7-pin O.	7Q	Htr.	25	0.15	125	60	—	—	HV
25Y4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.15	125	75	—	—	HV
25Y5[10]	Voltage Doubler	6-pin S.	6E	Htr.	25	0.3	250	85	—	—	HV
25Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	25	0.3	250	50	—	—	HV
25Z4	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.3	125	125	—	—	HV
25Z5	Rectifier-Doubler	6-pin S.	6E	Htr.	25	0.3	125	100	—	500	HV
25Z6	Rectifier-Doubler	7-pin O.	7Q	Htr.	25	0.3	125	100	—	500	HV
28Z5	Full-Wave Rectifier	8-pin L.	5AB	Htr.	28	0.24	450[7] 325[4]	100	—	300	HV
32L7GT	Rectifier-Tetrode	8-pin O.	8Z	Htr.	32.5	0.3	125	60	—	—	HV
35W4	Half-Wave Rectifier	7-pin B.	5BQ	Htr.	35[2]	0.15	125	100[8]	330	600	HV
35Y4	Half-Wave Rectifier	8-pin O.	5AL	Htr.	35[2]	0.15	235	60 100[8]	700	600	HV
35Z3	Half-Wave Rectifier	8-pin L.	4Z	Htr.	35	0.15	250[6]	100	700	600	HV
35Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	35	0.15	250	100	700	600	HV
35Z5G	Half-Wave Rectifier	6-pin O.	6AD	Htr.	35[2]	0.15	125	60 100[8]	—	—	HV
35Z6G	Voltage Doubler	6-pin O.	7Q	Htr.	35	0.3	125	110	—	500	HV

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
40Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	40 [2]	0.15	125	60 100 [8]	—	—	HV
45Z3	Half-Wave Rectifier	7-pin B.	5AM	Htr.	45	0.075	117	65	350	390	HV
45Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	45 [2]	0.15	125	60 100 [8]	—	—	HV
50X6	Voltage Doubler	8-pin L.	7AJ	Htr.	50	0.15	117	75	700	450	HV
50Y6GT	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.15	125	85	—	—	HV
50Y7GT	Voltage Doubler	8-pin L.	8AN	Htr.	50 [2]	0.15	117	65	700		HV
50Z6G	Voltage Doubler	7-pin O.	7Q	Htr.	50	0.3	125	150	—	—	HV
50Z7G [10]	Voltage Doubler	8-pin O.	8AN	Htr.	50	0.15	117	65	—	—	HV
70A7GT	Rectifier-Tetrode	8-pin O.	8AB	Htr.	70	0.15	125 [5]	60	—	—	HV
70L7GT	Rectifier-Tetrode	8-pin O.	8AA	Htr.	70	0.15	117	70	—	350	HV
72	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	30	20000	150	HV
73	Half-Wave Rectifier	8-pin O.	4Y	Fil.	2.5	4.5	—	20	13000	3000	HV
80	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	350 [4] 500 [7]	125 125	1400	375	HV
81	Half-Wave Rectifier	4-pin M.	4B	Fil.	7.5	1.25	700	85	—	—	HV
82	Full-Wave Rectifier	4-pin M.	4C	Fil.	2.5	3.0	500	125	1400	400	MV
83	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	800	MV
83-V	Full-Wave Rectifier	4-pin M.	4AD	Htr.	5.0	2.0	400	200	1100	—	HV
84/6Z4	Full-Wave Rectifier	5-pin S.	5D	Htr.	6.3	0.5	350	60	1000	—	HV
117L7GT/ 117M7GT	Rectifier-Tetrode	8-pin O.	8AO	Htr.	117	0.09	117	75	—	—	HV
117N7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117P7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117Z3	Half-Wave Rectifier	7-pin B.	4BR	Htr.	117	0.04	117	90	330	—	HV
117Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	117	0.04	117	90	350	—	HV
117Z6GT	Voltage Doubler	7-pin O.	7Q	Htr.	117	0.075	235	60	700	360	HV
217-A [10]	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	3500	600	HV
217-C	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	7500	600	HV
Z225	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250	10000	1000	MV
249-B	Half-Wave Rectifier	4-pin M.	Fig. 53	Fil.	2.5	7.5	3180	375	10000	1500	MV
HK253	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10	—	350	10000	1500	HV
705A RK-705A	Half-Wave Rectifier	4-pin W.	T-3AA	Fil.	2.5 [9] 5.0	5.0 5.0	— —	50 100	35000 35000	375 750	HV
816	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	2.0	2200	125	7500	500	MV
836	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	5.0	—	—	5000	1000	HV
866A/866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
866B	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.0	—	—	8500	1000	MV
866 Jr.	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	2.5	1250	250 [3]	—	—	MV
HY866 Jr.	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.5	1750	250 [3]	5000	—	MV
RK866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
871 [10]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.0	1750	250	5000	500	MV
878	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	7100	5	20000	—	HV
879	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	1.75	2650	7.5	7500	100	HV
872A/872	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	7.5	—	1250	10000	5000	MV
975A	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10.0	—	1500	15000	6000	MV
OZ4A/ 1003	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	—	110	880	—	G
1005/ CK1005	Full-Wave Rectifier	8-pin O.	5AQ	Fil.	6.3	0.1	—	70	450	210	G
1006/ CK1006	Full-Wave Rectifier	4-pin M.	4C	Fil.	1.75	2.25	—	200	1600	—	G
CK1007	Full-Wave Rectifier	8-pin O.	T-9G	Fil.	1.0	1.2	—	110	980	—	G
CK1009/BA	Full-Wave Rectifier	4-pin M.	—	Cold	—	—	—	350	1000	—	G
1274	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.6	Same as 7Y4				HV
1275	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	1.75	Same as 5Z3				HV
1616	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	130	6000	800	HV
1641/ RK60	Full-Wave Rectifier	4-pin M.	T-4AG	Fil.	5.0	3.0	—	50 250	4500 2500	— —	HV
1654	Half-Wave Rectifier	7-pin B.	Fig. 41	Fil.	1.4	0.05	2500	1	7000	6	HV
5517	Half-Wave Rectifier	7-pin B.	5BU	Cold	—	—	1200	6	—	50	G
8008	Half-Wave Rectifier	4-pin [6]	Fig. 11	Fil.	5.0	7.5	—	1250	10000	5000	MV
8013A	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	20	40000	150	HV
8016	Half-Wave Rectifier	6-pin O.	4AC	Fil.	1.25	0.2	—	2.0	10000	7.5	HV
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	HV
RK19	Full-Wave Rectifier	4-pin M.	4AT	Htr.	7.5	2.5	1250	200 [4]	3500	600	HV
RK21	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	4.0	1250	200 [4]	3500	600	HV
RK22	Full-Wave Rectifier	4-pin M.	T-4AG	Htr.	2.5	8.0	1250	200 [4]	3500	600	HV

[1] With input choke of at least 20 henrys.
[2] Tapped for pilot lamps.
[3] Per pair with choke input.
[4] Condenser input.
[5] With 100 ohms min. resistance in series with plate; without series resistor, maximum r.m.s. plate rating is 117 volts.
[6] Same as 872A/872 except for heavy-duty push-type base. Filament connected to pins 2 and 3, plate to top cap.
[7] Choke input.
[8] Without panel lamp.
[9] Using only one-half of filament.
[10] Discontinued.

TABLE XVI—TRIODE TRANSMITTING TUBES

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
958-A	0.6	1.25	0.1	135	7	1.0	12	0.6	2.6	0.8	500	A.	5BD	Class-C Amp.-Oscillator	135	−20	7	1.0	0.035	—	0.6
3B7[2]	—	1.4 2.8	0.22 0.11	180	25	—	20	1.4	2.6	2.6	125	O.	7AP	Class-C Amp. (Telegraphy)	180	0	25	—	—	—	2.8
RK24	1.5	2.0	0.12	180	20	6.0	8.0	3.5	5.5	3.0	125	S.	4D	Class-C Amp.-Oscillator	180	−45	16.5	6.0	0.5	—	2.0
6J6[2]	1.5	6.3	0.45	300	30	16	32	2.2	1.6	0.4	250	B.	7BF	Class-C Amp. (Telegraphy)[2]	150	−10	30	16	0.35	—	3.5
9002	1.6	6.3	0.15	250	8	2.0	25	1.2	1.4	1.1	250	B.	7TM	Class-C Amp.-Oscillator	180	−35	7	1.5	—	—	0.5
955	1.6	6.3	0.15	180	8	2.0	25	1.0	1.4	0.6	250	A.	5BC	Class-C Amp.-Oscillator	180	−35	7	1.5	—	—	0.5
HY114B	1.8	1.4	0.155	180	12	3.0	13	1.0	1.3	1.0	300	O.	2T	Class-C Amp.-Oscillator	180	−30	12	2.0	0.2	—	1.4[3]
														Class-C Amp. (Telephony)	180	−35	12	2.5	0.3	—	1.4[3]
3A5[2]	2.0	1.4 2.8	0.22 0.11	150	30	5.0	15	0.9	3.2	1.0	40	B.	7BC	Class-C Amp.-Oscillator[2]	150	−35	30	5.0	0.2	—	2.2
6F4	2.0	6.3	0.225	150	20	8.0	17	2.0	1.9	0.6	500	A.	7BR	Class-C Amp.-Oscillator	150	−15 550[*] 2000[4]	20	7.5	0.2	—	1.8
HY24	2.0	2.0	0.13	180	20	4.5	9.3	2.7	5.4	2.3	60	S.	4D	Class-C Amp. (Telegraphy)	180	−45	20	4.5	0.2	—	2.7
														Class-C Amp. (Telephony)	180	−45	20	4.5	0.3	—	2.5
RK33[1, 2]	2.5	2.0	0.12	250	20	6.0	10.5	3–2	3–2	2.5	60	S.	T-7DA	Class-C Amp.-Oscillator[2]	250	−60	20	6.0	0.54	—	3.5
12AU7[2]	2.75[6]	6.3	0.3	350	12[6]	3.5[6]	18	1.5	1.5	0.5	54	B.	9A	Class-C Amp.-Oscillator[2]	350	−100	24	7	—	—	6.0
6N4	3.0	6.3	0.2	180	12	—	32	3.1	2.35	0.55	500	B.	7CA	Class-C Amp.-Oscillator	180	—	—	—	—	—	—
HY6J5GTX	3.5	6.3	0.3	330	20	4.0	20	4.2	3.8	5.0	60	O.	6Q	Class-C Amp.-Oscillator	330	−30	20	2.0	0.2	—	3.5
														Class-C Amp. (Telephony)	250	−30	20	2.5	0.3	—	2.5
2C22/7193	3.5	6.3	0.3	500	—	—	20	2.2	3.6	0.7	—	O.	4AM	Class-C Amp. (Telegraphy)	—	—	—	—	—	—	—
HY615 HY-E1148	3.5	6.3	0.175	300	20	4.0	20	1.4	1.6	1.2	300	O.	T-8AG	Class-C Amp.-Oscillator	300	−35	20	2.0	0.4	—	4.0[3]
														Class-C Amp. (Telephony)	300	−35	20	3.0	0.8	—	3.5[3]
GL-446A[1] GL-446B[1]	3.75	6.3	0.75	400	20	—	45	2.2	1.6	0.02	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
GL-2C44[1] GL-464A[1]	5.0	6.3	0.75	500	40	—	—	2.7	2.0	0.1	500	O.	Fig. 17	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
6C4	5.0	6.3	0.15	350	25	8.0	18	1.8	1.6	1.3	54	B.	6BG	Class-C Amp.-Oscillator	300	−27	25	7.0	0.35	—	5.5
1626	5.0	12.6	0.25	250	25	8.0	5.0	3.2	4.4	3.4	30	O.	6Q	Class-C Amp.-Oscillator	250	−70	25	5.0	0.5	—	4.0
2C21/ RK33[2]	5.0	6.3	0.6	250	40	12	—	1.6	1.6	2.0	—	S.	T-7DA	Class-C Amp.-Oscillator[2]	250	−60	40	12	1.0	—	7
6N7[2]	5.5[6]	6.3	0.8	350	30[6]	5.0[6]	35	—	—	—	10	O.	8B	Class-C Amp. Oscillator[2, 11]	350	−100	60	10	—	—	14.5
2C40	6.5	6.3	0.75	500	25	—	36	2.1	1.3	0.05	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	−5	20	0.3	—	—	0.075
5556	7.0	4.5	1.1	350	40	10	8.5	4.0	8.3	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	350	−80	35	2	0.25	—	6
														Class-C Amp. (Telephony)	300	−100	30	2	0.3	—	4
2C43	12	6.3	0.9	500	40	—	48	2.9	1.7	0.05	1250	O.	Fig. 19	Class-C Amp.-Oscillator	470	—	38[7]	—	—	—	9[7]
2C26A	10	6.3	1.10	—	—	—	16.3	2.6	2.8	1.1	250	O.	4BB		—	—	—	—	—	—	—
2C34/ RK34[2]	10	6.3	0.8	300	80	20	13	3.4	2.4	0.5	250	M.	T-7DC	Class-C Amp.-Oscillator[2]	300	−36	80	20	1.8	—	16
205D	14	4.5	1.6	400	50	10	7.2	5.2	4.8	3.3	6	M.	4D	Class-C Amp.-Oscillator	400	−112	45	10	1.5	—	10
														Class-C Amp. (Telephony)	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	—	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
10Y	15	7.5	1.25	450	65	15	8	4.1	7.0	3.0	8	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
843	15	2.5	2.5	450	40	7.5	7.7	4.0	4.5	4.0	6	M.	5A	Class-C Amp.-Oscillator	450	−140	30	5.0	1.0	—	7.5
														Class-C Amp. (Telephony)	350	−150	30	7.0	1.6	—	5.0
RK59[2]	15	6.3	1.0	500	90	25	25	5.0	9.0	1.0	—	M.	T-4D	Class-C Amp.-Oscillator	500	− 60	90	14	1.3	—	32
HY75A	15	6.3	2.6	450	90	25	9.6	1.8	2.6	1.0	175	O.	2T	Class-C Amp. (Telegraphy)	450	−140	90	20	5.2	—	26
														Class-C Amp. (Telephony)	400	−140	90	28	5.2	—	21
HY75	15	6.3	2.5	450	80	20	10	1.0	3.8	1.0	60	O.	2T	Class-C Amp.-Oscillator	450	− 50	80	12	—	—	21[3]
														Class-C Amp. (Telephony)	450	− 60	80	12	—	—	16[3]
1602	15	7.5	1.25	450	60	15	8.0	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	−115	55	15	3.3	—	13
														Class-C Amp. (Telephony)	350	−135	45	15	3.5	—	8.0
														Class-B Amp. Audio[7]	425	− 50	110[8]	260[9]	2.5[8]	8000	25
841	15	7.5	1.25	450	60	20	30	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	− 34	50	15	1.8	—	15
														Class-C Amp. (Telephony)	350	− 47	50	15	2.0	—	11
10 RK10[1]	15	7.5	1.25	450	65	15	8.0	3.0	8.0	4.0	— 60	M.	4D	Class-C Amp. (Telegraphy)	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
														Class-B Audio[7]	425	− 50	55[8]	130[9]	2.5[8]	8000	25
RK100[1]	15	6.3	0.9	150	250	100	40	23	19	3.0	—	M.	T-6B	Class-C Oscillator	110	—	80	8.0	—	—	3.5
														Class-C Amplifier	110	—	185	40	2.1	—	12
TUF-20	20	6.3	2.75	750	75	20	10	1.8	3.6	0.095	250	O.	2T	Class-C Amp.-Oscillator	750	−150	75	20	1.5/2.5	—	40
1608	20	2.5	2.5	425	95	25	20	8.5	9.0	3.0	45	M.	4D	Class-C Amp. (Telegraphy)	425	− 90	95	20	3.0	—	27
														Class-C Amp. (Telephony)	350	− 80	85	20	3.0	—	18
														Class-B Amp. Audio[7]	425	− 15	190[8]	130[9]	2.2[8]	4000	50
310	20	7.5	1.25	600	70	15	8.0	4.0	7.0	2.2	6	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
703-A	20	1.2	4/4.5	350	75	12	8	0.9	1.1	0.6	1400	N.	—	Class-C Amplifier	350	−120	75	12	—	—	2/2.5
801-A/801	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
														Class-B Amp. Audio[7]	600	− 75	130	320[9]	3.0[8]	10000	45
HY801-A	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−200	70	15	4.0	—	30
														Class-C Amp. (Telephony)	500	−200	60	15	4.5	—	22
T20	20	7.5	1.75	750	85	25	20	4.9	5.1	0.7	60	M.	3G	Class-C Amp (Telegraphy)	750	− 85	85	18	3.6	—	44
														Class-C Amp. (Telephony)	750	−140	70	15	3.6	—	38
TZ20	20	7.5	1.75	750	85	30	62	5.3	5.0	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	750	− 40	85	28	3.75	—	44
														Class-C Amp (Telephony)	750	−100	70	23	4.8	—	38
														Class-B Amp. Audio[7]	800	0	40/136	160[9]	1.8[8]	12000	70
15E	20	5.5	4.2	—	—	—	25	1.4	1.15	0.3	600	N.	T-4AF	—	—	—	—	—	—	—	—
3-25A3 25T	25	6.3	3.0	2000	75	25	24	2.7	1.5	0.3	60	M.	3G	Class-C Amp.-Oscillator	2000	−130	63	18	4.0	—	100
															1500	− 95	67	13	2.2	—	75
															1000	− 70	72	9	1.3	—	47
														Class-B Amp. Audio[7]	2000	− 80	16/80	270[9]	0.7[8]	55500	110
3-25D3 3C24 24G	25	6.3	3.0	2000	75	25	23	2.0 1.7	1.6 1.5	0.2 0.3	60	S.	2D	Class-C Amp.-Oscillator	2000	−170	63	17	4.5	—	100
															1500	−110	67.	15	3.1	—	75
															1000	− 80	72	15	2.6	—	47
														Class-B Audio[7]	2000	− 85	16/80	290[9]	1.1[8]	55500	110
3C28	25	6.3	3.0	2000	75	25	23	2.1	1.8	0.1	100	S.	Fig. 56	Class-C Amp. Oscillator		Characteristics same as 3C24					

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
3C34	25	6.3	3.0	2000	75	25	23	2.5	1.7	0.4	60	S.	3G	Class-C Amp. Oscillator		Characteristics same as 3C24					
RK11[1]	25	6.3	3.0	750	105	35	20	7.0	7.0	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	750	−120	105	21	3.2	—	55
														Class-C Amp. (Telephony)	600	−120	85	24	3.7	—	38
RK12	25	6.3	3.0	750	105	40	100	7.0	7.0	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	750	−100	105	35	5.2	—	55
														Class-C Amp. (Telephony)	600	−100	85	27	3.8	—	38
HK24	25	6.3	3.0	2000	75	30	25	2.5	1.7	0.4	60	S.	3G	Class-C Amp. (Telegraphy)	2000	−140	56	18	4.0	—	90
														Class-C Amp. (Telephony)	1500	−145	50	25	5.5	—	60
HY25	25	7.5	2.25	800	75	25	55	4.2	4.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	750	−45	75	15	2.0	—	42
														Class-C Amp. (Telephony)	700	−45	75	17	5.0	—	39
8025	30	6.3	1.92	1000	65	—	18	2.7	2.8	0.35	500	M.	4AQ	Class-C Amp. (Grid. Mod.)	1000	−135	50	4	3.5	—	20
	20				65	20								Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
	30				80	20								Class-C Amp. (Telegraphy)	1000	−90	50	14	1.6	—	35
HY30Z[1]	30	6.3	2.25	850	90	25	87	6.0	4.9	1.0	60	M.	4BO	Class-C Amp.-Oscillator	850	−75	90	25	2.5	—	58
														Class-C Amp. (Telephony)	700	−75	90	25	3.5	—	47
HY31Z[2]	30	6.3	3.5	500	150	30	45	5.0	5.5	1.9	60	M.	T-4D	Class-C Amp. (Telegraphy)	500	−45	150	25	2.5	—	56
HY1231Z[2]		12.6	1.7											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	500	N.	—	Class-C Amp. (Telegraphy)	450	—	80	12	—	—	7.5
														Class-C Amp. (Telephony)	400	—	80	12	—	—	6.5
809	30	6.3	2.5	1000	125	—	50	5.7	6.7	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−75	100	25	3.8	—	75
														Class-C Amp. (Telephony)	750	−60	100	32	4.3	—	55
														Class-B Amp. Audio[7]	1000	−9	40/200	155[9]	2.7[8]	11600	145
1623	30	6.3	2.5	1000	100	25	20	5.7	6.7	0.9	60	M.	3G	Class-C Amp.-Oscillator	1000	−90	100	20	3.1	—	75
														Class-C Amp. (Telephony)	750	−125	100	20	4.0	—	55
														Class-B Amp. Audio[7]	1000	−40	30/200	230[9]	4.2[8]	12000	145
53A	35	5.0	12.5	15000	—	—	35	3.6	1.9	0.4	—	N.	T-4B	Oscillator at 300 Mc.		Approximately 50 watts output					
RK30[1]	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−180	90	18	5.2	—	85
														Class-C Amp. (Telephony)	1000	−200	80	15	4.5	—	60
800	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−175	70	15	4.0	—	65
														Class-C Amp. (Telephony)	1000	−200	70	15	4.0	—	50
														Class-B Amp. Audio[7]	1250	−70	30/130	300[9]	3.4[8]	21000	106
1628[1]	40	3.5	3.25	1000	60	15	23	2.0	2.0	0.4	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	−65	50	15	1.7	—	35
														Class-C Amp. (Telephony)	800	−100	40	11	1.6	—	22
														Grid-Modulated Amp.	1000	−120	50	3.5	5.0	—	20
8012	40	6.3	2.0	1000	80	20	18	2.7	2.8	0.35	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	−90	50	14	1.6	—	35
GL-8012-A								2.7	2.5	0.4				Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
														Grid-Modulated Amp.	1000	−135	50	4.0	3.5	—	20
RK18[1]	40	7.5	3.0	1250	100	40	18	6.0	4.8	1.8	60	M.	3G	Class-C Amp. (Telegraphy)	1250	−160	100	12	2.8	—	95
														Class-C Amp. (Telephony)	1000	−160	80	13	3.1	—	64
RK31	40	7.5	3.0	1250	100	35	170	7.0	1.0	2.0	30	M.	3G	Class-C Amp. (Telegraphy)	1250	−80	100	30	3.0	—	90
														Class-C Amp. (Telephony)	1000	−80	100	28	3.5	—	70
HY40[1]	40	7.5	2.25	1000	125	25	25	6.1	5.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−90	125	20	5.0	—	94
														Class-C Amp. (Telephony)	850	−90	125	25	5.0	—	82
														Grid-Modulated Amp.	1000	—	125	—	—	—	20

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HY40Z[1]	40	7.5	2.6	1000	125	30	80	6.2	6.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−27	125	25	5.0	—	94
														Class-C Amp. (Telephony)	850	−30	100	30	7.0	—	82
														Grid-Modulated Amp.	1000	—	60	—	—	—	20
T40	40	7.5	2.5	1500	150	40	25	4.5	4.8	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	−140	150	28	9.0	—	158
														Class-C Amp. (Telephony)	1250	−115	115	20	5.25	—	104
TZ40	40	7.5	2.5	1500	150	45	62	4.8	5.0	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	−90	150	38	10	—	165
														Class-C Amp. (Telephony)	1250	−100	125	30	7.5	—	116
														Class-B Amp. Audio [7]	1500	−9	250 [8]	285 [9]	6.0 [8]	12000	250
HY57	40	6.3	2.25	850	110	25	50	4.9	5.1	1.7	60	M.	3G	Class-C Amp. (Telegraphy)	850	−48	110	15	2.5	—	70
														Class-C Amp. (Telephony)	700	−45	90	17	5.0	—	47
														Grid-Modulated Amp.	850	—	70	—	—	—	20
756[1]	40	7.5	2.0	850	110	25	8.0	3.0	7.0	2.7	—	M.	4D	Class-C Amplifier	850	—	110	25	—	—	—
830[1]	40	10	2.15	750	110	18	8.0	4.9	9.9	2.2	15	M.	4D	Class-C Amplifier	750	−180	110	18	7.0	—	55
														Grid-Modulated Amp.	1000	−200	50	2.0	3.0	—	15
3-50A4 35T	50	5.0	4.0	2000	150	50	39	4.1	1.8	0.3	100	M.	3G	Class-C Amp. (Telegraphy)	2000	−135	125	45	13	—	200
3-50D4 35TG								2.5	1.8	0.4	100	M.	2D	Class-C Amp. (Telephony)	1500	−120	100	30	5.0	—	120
														Class-B Amp. Audio [7]	2000	−40	34/167	255 [9]	4.0 [8]	27500	235
8010-R	50	6.3	2.4	1350	150	20	30	2.3	1.5	0.07	350	N.	—	Class-C Amplifier	—	—	—	—	—	—	—
RK32[1]	50	7.5	3.25	1250	100	25	11	2.5	3.4	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	100	14	4.8	—	90
														Class-C Amp. (Telephony)	1000	−310	100	21	8.7	—	70
RK35[1]	50	7.5	4.0	1500	125	20	9.0	3.5	2.7	0.4	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−250	115	15	5.0	—	120
														Class-C Amp. (Telephony)	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	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−130	115	30	7.0	—	122
														Class-C Amp. (Telephony)	1250	−150	100	23	5.6	—	90
														Grid-Modulated Amp.	1500	−50	50	—	2.4	—	26
3-50G2 UH50	50	7.5	3.25	1250	125	25	10.6	2.2	2.6	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	125	20	7.5	—	115
														Class-C Amp. (Telephony)	1250	−325	125	20	10	—	115
														Grid-Modulated Amp.	1250	−200	60	2.0	3.0	—	25
UH51[1]	50	5.0	6.5	2000	175	25	10.6	2.2	2.3	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−500	150	20	15	—	225
														Class-C Amp. (Telephony)	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	100	M.	2D	Class-C Amp. (Telegraphy)	3000	−290	100	25	10	—	250
														Class-C Amp. (Telephony)	2500	−250	100	20	8.0	—	210
														Class-B Amp. Audio [7]	2500	−85	20/150	360 [9]	5.0	40000	275
HK154[1]	50	5.0	6.5	1500	175	30	6.7	4.3	5.9	1.1	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−590	167	20	15	—	200
														Class-C Amp. (Telephony)	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	60	M.	2D	Class-C Amp.-Oscillator	2000	−150	125	25	6.0	—	200
														Class-C Amp. (Telephony)	2000	−140	105	25	5.0	—	170
WE304A[1] 304B	50	7.5	3.25	1250	100	25	11	2.0	2.5	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−200	100	—	—	—	85
														Class-C Amp. (Telephony)	1000	−180	100	—	—	—	65

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
356A	50	5.0	5.0	1500	120	35	50	2.25	2.75	1.0	60	N.	T-4BD	Class-C Amp. (Telegraphy)	1500	−60	100	—	—	—	100
														Class-C Amp. (Telephony)	1250	−100	100	35	—	—	85
808	50	7.5	4.0	1500	150	35	47	5.3	2.8	0.15	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	125	30	9.5	—	140
														Class-C Amp. (Telephony)	1250	−225	100	32	10.5	—	105
														Class-B Amp. Audio[7]	1500	−25	30/190	220[9]	4.8[8]	18300	185
834	50	7.5	3.1	1250	100	20	10.5	2.2	2.6	0.6	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	90	15	4.5	—	75
														Class-C Amp. (Telephony)	1000	−310	90	17.5	6.5	—	58
841A[1]	50	10	2.0	1250	150	30	14.6	3.5	9.0	2.5	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	85
841SW	50	10	2.0	1000	150	30	14.6	—	9.0	—	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	—
T55	55	7.5	3.0	1500	150	40	20	5.0	3.9	1.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1500	−195	125	15	5.0	—	145
811	55	6.3	4.0	1500	150	50	160	5.5	5.5	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−113	150	35	8.0	—	170
														Class-C Amp. (Telephony)	1250	−125	125	50	11	—	120
														Class-B Amp. Audio[7]	1500	−9	20/200	150[9]	3.0[8]	17600	220
812	55	6.3	4.0	1500	150	35	29	5.3	5.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−175	150	25	6.5	—	170
														Class-C Amp. (Telephony)	1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio[7]	1500	−45	50/200	232[9]	4.7[8]	18000	225
RK51	60	7.5	3.75	1500	150	40	20	6.0	6.0	2.5	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−250	150	31	10	—	170
														Class-C Amp. (Telephony)	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	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	130	40	7.0	—	135
														Class-C Amp. (Telephony)	1250	−120	115	47	8.5	—	102
														Class-B Amp. Audio[7]	1250	0	40/300	180[9]	7.5[8]	10000	250
T-60	60	10	2.5	1600	150	50	20	5.5	5.2	2.5	60	M.	2D	Class-C Amp.-Oscillator	1500	−150	150	50	9.0	—	100
826	60	7.5	4.0	1000	125	40	31	3.7	2.9	1.4	250	N.	T-9A	Class-C Amp.-Oscillator	1000	−70	125	35	5.8	—	86
														Class-C Amp. (Telephony)	800	−98	94	35	6.2	—	53
														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	15	M.	3G	Class-C Amp.-Oscillator	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
														Class-B Amp. Audio[7]	1000	−35	20/280	270[9]	6.0[8]	7600	175
812-A	65	6.3	4.0	1500	175	35	29	5.4	5.5	0.77	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	173	30	6.5	—	190
														Class-C Amp. (Telephony)	1250	−115	140	35	7.6	—	130
														Class-B Audio[7]	1500	−48	28/310	270[9]	5.0	13200	340
HY51A[1] HY51B[1]	65	7.5 10	3.5 2.25	1000	175	25	25	6.5	7.0	1.1	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−75	175	20	7.5	—	131
														Class-C Amp. (Telephony)	1000	−67.5	130	15	7.5	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
HY51Z[1]	65	7.5	3.5	1000	175	35	85	7.9	7.2	0.9	60	M.	4BO	Class-C Amp. (Telegraphy)	1000	−22.5	175	35	10	—	131
														Class-C Amp. (Telephony)	1000	−30	150	35	10	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
5514	65	7.5	3.0	1500	175	60	145	7.0	7.9	1.0	60	M.	4BO	Class-C Amp. (Telegraphy)	1500	−106	175	60	12	—	200
														Class-C Amp. (Telephony)	1250	−84	142	60	10	—	135
														Class-B Audio[7]	1500	−4.5	350[8]	88[8]	6.5[8]	10500	400
UH35[1]	70	5.0	4.0	1500	150	35	30	1.4	1.6	0.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	30	7.0	—	170
														Class-C Amp. (Telephony)	1500	−120	100	30	5.0	—	120

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
V70 V70B	70	10	2.5	1500	140	25	14	5.0	9.0	2.3	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1500	−215	130	6.0	3.0	—	140
														Class-C Amp. (Telephony)	1250	−250	130	6.0	3.0	—	120
V70A V70C	70	10	2.5	1500	140	20	25	5.0	9.5	2.0	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
50T[1]	75	5.0	6.0	3000	100	30	12	2.0	2.0	0.4	—	M.	2D	Class-C Amplifier	3000	−600	100	25	—	—	250
3-75A3 75TH	75	5.0	6.25	3000	225	40	20	2.7	2.3	0.3	40	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	150	32	10	—	225
														Class-B Amp. Audio[7]	2000	−90	50/225	350[9]	3[8]	19300	300
3-75A2 75TL						35	12	2.6	2.4	0.4			2D	Class-C Amp. (Telegraphy)	2000	−300	150	21	8	—	225
														Class-B Amp. Audio[7]	2000	−160	50/250	535[9]	5[8]	18000	350
HF-60	75	10	2.5	1600	160	—	28	5.4	5.2	1.5	30	M.	2D	Class-C Amp. (Telegraphy)	1600	−190	158	12	3.5	—	200
														Class-C Amp. (Telephony)	1250	−190	113	8	2.5	—	110
														Class-B Amp. Audio[7]	1600	−75	50/248	310[9]	3.0	13800	262
ZB-60	75	10	2.5	1600	160	40	80	6.1	5.8	1.85	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−95	158	31	6.0	—	190
														Class-B Amp. Audio[7]	1500	−9	30/305	208[9]	12.5	11200	320
111H	75	10	2.5	1500	160	30	23	5.0	4.6	2.9	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Class-B Amp. Audio[7]	1750	−62	40/270	324[9]	9.0	16000	350
HF75	75	10	3.25	2000	120	—	12.5	—	2.0	—	75	M.	2D	Class-C Oscillator-Amp.	2000	—	120	—	—	—	150
TW75	75	7.5	4.15	2000	175	60	20	3.35	1.5	0.7	60	M.	2D	Class-C Amp.-Oscillator	2000	−175	150	37	12.7	—	225
														Class-C Amp. (Telephony)	2000	−260	125	32	13.2	—	198
T-100 HF100	75	10	2.5	1500	150	30	23	4.0	4.5	2.6	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Grid-Modulated Amp.	1500	−280	72	1.5	6.0	—	42
														Class-B Amp. Audio[7]	1750	−62	40/270	324[9]	9.0[8]	16000	350
UE-100	75	10	2.5	1750	150	30	23	3.5	4.5	1.4	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	120	21	8.0	—	105
														Class-B Audio[7]	1750	−62	540[8]	—	9.0	16000	350
ZB120	75	10	2.0	1250	160	40	90	5.3	5.2	3.2	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−135	160	23	5.5	—	145
														Class-C Amp. (Telephony)	1000	−150	120	21	5.0	—	95
														Grid-Modulated Amp.	1250	—	95	8.0	1.5	—	45
														Class-B Amp. Audio[7]	1500	−9	60/296	196[9]	5.0[8]	11200	300
327B	75	10.5	10.6	—	—	—	30	3.4	2.45	0.3	—	N.	T-4AD	—	—	—	—	—	—	—	—
242A	85	10	3.25	1250	150	50	12.5	6.5	13	4.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
284D	85	10	3.25	1250	150	100	4.8	6.0	8.3	5.6	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−450	150	50	—	—	100
														Class-B Amp. Audio[7]	1250	−250	30/200	—	—	11200	140
812-H	85	6.3	4.0	1750	200	45	—	5.3	5.3	0.8	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−175	170	26	6.5	—	225
															1250	−125	125	25	5.0	—	116
														Class-C Amp. (Telephony)	1500	−125	165	21	6.0	—	180
															1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio[7]	1500	−46	42/200	—	—	18000	225

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
8005	85	10	3.25	1500	200	45	20	6.4	5.0	1.0	60	M.	3G	Class-C Amp.-Telegraphy	1500	−130	200	32	7.5	—	220
														Class-C Amp. (Telephony)	1250	−195	190	28	9.0	—	170
														Class-B Amp. Audio [7]	1500	−70	40/310	310 [9]	4.0 [8]	10000	300
V-70-D	85	7.5	3.25	1750	200	45	—	4.5	4.5	1.7	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−100	170	19	3.9	—	225
															1500	−90	165	19	3.9	—	195
														Class-C Amp. (Telephony)	1500	−90	165	19	3.7	—	185
															1250	−72	127	16	2.6	—	122
RK36 [1]	100	5.0	8.0	3000	165	35	14	4.5	5.0	1.0	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−360	150	30	15	—	200
														Class-C Amp. (Telephony)	2000	−360	150	30	15	—	200
														Grid-Modulated Amp.	2000	−270	72	1.0	3.5	—	42
RK38 [1]	100	5.0	8.0	3000	165	40	—	4.6	4.3	0.9	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	160	30	10	—	225
														Class-C Amp. (Telephony)	2000	−200	160	30	10	—	225
														Grid-Modulated Amp.	2000	−150	80	2.0	5.5	—	60
3-100A4 100TH	100	5.0	6.3	3000	225	60	40	2.9	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy) Class-C Amp. (Telephony)	3000	−200	165	51	18	—	400
														Grid-Modulated Amp.	3000	−400	70	3.0	7.0	—	100
														Class-B Amp. (Audio) [7]	3000	−65	40/215	335 [9]	5.0 [8]	31000	650
3-100A2 100TL	100	5.0	6.3	3000	225	50	14	2.3	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy) Class-C Amp. (Telephony)	3000	−400	165	30	20	—	400
														Grid-Modulated Amp.	3000	−560	60	2.0	7.0	—	90
														Class-B Amp. (Audio) [7]	3000	−185	40/215	640 [9]	6.0 [8]	30000	450
VT127A	100	5.0	10.4	3000	—	30	15.5	2.7	2.3	0.35	150	N.	T-4B	Class-C Amp.-Oscillator	Characteristics similar to 100TL						
227A	100	10.5	10.7	—	—	—	31	3.0	2.2	0.30	—	N.	T-4B	Oscillator at 200 Mc.	—	—	—	—	—	—	—
327A	100	10.5	10.7	—	—	—	31	3.4	2.3	0.35	—	N.	T-4AD	Oscillator at 200 Mc.	—	—	—	—	—	—	—
HK254	100	5.0	7.5	4000	200	40	25	3.3	3.4	1.1	50	J.	2N	Class-C Amp. (Telegraphy)	4000	−380	120	35	20	—	475
														Class-C Amp. (Telephony)	3000	−290	135	40	23	—	320
														Grid-Modulated Amp.	3000	—	51	3.0	4.0	—	58
														Class-B Amp. (Audio) [7]	3000	−100	40/240	456 [9]	7.0 [8]	30000	520
RK58	100	10	3.25	1250	175	70	—	8.5	6.5	10.5	—	J.	3N	Class-C Amp. (Telegraphy)	1250	−90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	16	—	100
HF120	100	10	3.25	1250	175	50	12	5.5	12.5	3.5	15	J.	4F	Class-C Amp.-Oscillator	1250	−300	166	8	3.5	—	148
HF125	100	10	3.25	1500	175	—	25	—	11.5	—	30	J.	—	Class-C Amp.-Oscillator	1500	—	175	—	—	—	200
HF140	100	10	3.25	1250	175	—	12	5.5	13.0	4.5	15	J.	4F	Class-C Amp.-Oscillator	1250	−300	166	8	3.5	—	148
203A 303A	100	10	3.25	1250	175	60	25	6.5	14.5	5.5	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	25	7.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	14	—	100
														Class-B Amp. (Audio) [7]	1250	−45	26/320	330 [9]	11 [8]	9000	260
203H	100	10	3.25	1500	175	60	25	6.5	11.5	1.5	15	J.	3N	Class-C Amp. (Telegraphy)	1500	−200	170	12	3.8	—	200
														Class-C Amp. (Telephony)	1250	−160	167	19	5.0	—	160
														Class-B Amp. (Audio) [7]	1500	−52	30/320	304 [9]	5.5 [8]	11000	340
211 311 835 [1]	100	10	3.25	1250	175	50	12	6.0 6.0	14.5 9.25	5.5 5.0	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−225	150	18	7.0	—	130
														Class-C Amp. (Telephony)	1000	−260	150	35	14	—	100
														Class-B Amp. (Audio) [7]	1250	−100	20/320	410 [9]	8.0 [8]	9000	260
242B 342B	100	10	3.25	1250	150	50	12.5	7.0	13.6	6.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
242C	100	10	3.25	1250	150	50	12.5	6.1	13.0	4.7	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−80	25/150	—	25[8]	7600	200
261A 361A	100	10	3.25	1250	150	50	12	6.5	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−90	20/150	—	25[8]	7200	200
276A 376A	100	10	3.0	1250	125	50	12	6.0	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	125	50	—	—	85
														Class-B Amp. (Audio)[7]	1250	−90	20/125	—	25[8]	9000	175
284B	100	10	3.25	1250	150	100	5.0	4.2	7.4	5.3	—	J.	3N	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−430	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−245	15/150	—	10[8]	7200	200
295A	100	10	3.25	1250	175	50	25	6.5	14.5	5.5	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−125	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−40	12/160	—	20[8]	9000	250
838 938	100	10	3.25	1250	175	70	—	6.5	8.0	5.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	60	16	—	100
														Class-B Amp. (Audio)[7]	1250	0	148/320	200[9]	7.5[8]	9000	260
852	100	10	3.25	3000	150	40	12	1.9	2.6	1.0	30	M.	2D	Class-C Amp. (Telegraphy)	3000	−600	85	15	12	—	165
														Class-C Amp. (Telephony)	2000	−500	67	30	23	—	75
														Class-B Amp. (Audio)[7]	3000	−250	14/160	780[9]	3.5[8]	10250	320
8003	100	10	3.25	1500	250	50	12	5.8	11.7	3.4	30	J.	3N	Class-C Amp.-Oscillator	1350	−180	245	35	11	—	250
														Class-C Amp. (Telephony)	1100	−260	200	40	15	—	167
														Class-B Amp. (Audio)[7]	1350	−100	40/490	480[9]	10.5[8]	6000	460
3X100A11 2C39	100	6.3	1.1	1000	60	40	100	6.5	1.95	0.03	500	N.	—	"Grid Isolation" Circuit	600	−35	60	40	5.0	—	20
311-CH	125	10	3.25	1750	200	50	12	5.5	8.0	4.5	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	1750	−200	200	20	4.5	—	260
														Class-C Amp. (Telephony)	1250	−200	166	8	3.5	—	148
														Class-B (Audio)[7]	1500	−110	400[8]	—	—	8200	400
3C22	125	6.3	2.0	1000	150	70	40	4.9	2.4	0.05	500	O.	Fig. 30	Class-C Amp.-Oscillator	1000	−200	150	70	—	—	65
4C36	125	5	7.5	4000	—	—	29	3.2	3.0	0.4	60	J.	Fig. 56	Class-C Amp.-Oscillator	—	—	—	—	18	—	480
F-123-A DR-123C	125	10	4.0	2000	300	75	14.5	6.5	8.5	3.3	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	1500	−250	250	30	11	—	300
														Class-C Amp. (Telephony)	1500	−290	160	25	10	—	200
														Class-B Amp. (Audio)[7]	2000	−130	30/175	217[9]	3.4[8]	13800	522
RK57/805	125	10	3.25	1500	210	70	—	6.5	8.0	5.0	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio)[7]	1500	−16	84/400	280[9]	7.0[8]	8200	370
T125	125	10	4.5	2500	250	60	25	6.3	6.0	1.3	60	J.	2N	Class-C Amp. (Telegraphy)	2500	−200	240	31	11	—	475
														Class-C Amp. (Telephony)	2000	−215	200	28	10	—	320
HF130	125	10	3.25	1250	210	—	12.5	5.5	9.0	3.5	20	J.	—	Class-C Amp.-Oscillator	1250	−250	200	10	3.5	—	170
HF150	125	10	3.25	1500	210	—	12.5	5.5	7.2	1.9	30	J.	—	Class-C Amp.-Oscillator	1500	−300	200	10	4	—	220
HF175	125	10	4.0	2000	250	—	18	4.8	6.3	2.7	25	J.	T-3AC	Class-C Amp.-Oscillator	2000	−250	200	23	9	—	320

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
GL146	125	10	3.25	1500	200	60	75	7.2	9.2	3.9	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	40	—	—	100
														Class-B Amp. (Audio) [7]	1250	0	34/320	—	—	8400	250
GL152	125	10	3.25	1500	200	60	25	7.0	8.8	4.0	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	30	—	—	100
														Class-B Amp. (Audio) [7]	1250	−40	16/320	—	—	8400	250
805	125	10	3.25	1500	210	70	40/60	8.5	6.5	10.5	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio) [7]	1500	−16	84/400	280 [9]	7.0 [8]	8200	370
3X150A3 3C37	150	6.3	2.5	1000	—	—	23	4.2	3.5	0.6	500	N.	—	—	—	—	—	—	—	—	—
150T [1]	150	5.0	10	3000	200	50	13	3.0	3.5	0.5	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−600	200	35	—	—	450
3-150A3 152TH	150	5/10	12.5[1]/6.25	3000	450	85	20	5.7	4.5	0.8	40	J.	4BC	Class-C Amp. (Telegraphy)	3000	−300	250	70	27	—	600
														Class-B Amp. (Audio) [7]	3000	−150	57/335	430 [9]	3.0 [8]	20300	700
3-150A2 152TL						75	12	4.5	4.4	0.7			4BC	Class-C Amp. (Telegraphy)	3000	−400	250	40	20	—	600
														Class-B Amp. (Audio) [7]	3000	−260	65/335	675 [9]	3.0 [8]	23400	700
TW150	150	10	4.1	3000	200	60	35	3.9	2.0	0.8	—	J.	2N	Class-C Amp.-Oscillator	3000	−170	200	45	17	—	470
														Class-C Amp. (Telephony)	3000	−250	165	40	17	—	400
HK252-L	150	5/10	13/6.5	3000	500	75	10	7.0	5.0	0.4	125	N.	4BC	Class-C Amp.-Oscillator	3000	−400	250	30	15	—	610
														Class-C Amp. (Telephony)	2500	−350	250	35	16	—	500
DR200 HF200 HV18	150	10–11	3.4	2500	200	50	18	5.2	5.8	1.2	20	J.	2N	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B Amp. (Audio) [7]	2500	−130	60/360	460 [9]	8.0 [8]	16000	600
HD203A	150	10	4.0	2000	250	60	25	—	12	—	15	J.	3N	Class-C Amplifier	—	—	—	—	—	—	375
HF250	150	10.5	4.0	2500	200	—	18	—	5.8	—	20	J.	2N	Class-C Amp.-Oscillator	2500	—	200	—	—	—	375
HK354 HK354C	150	5.0	10	4000	300	50	14	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	4000	−690	245	50	48	—	830
														Class-C Amp. (Telephony)	3000	−550	210	50	35	—	525
														Grid-Modulated Amp.	3000	−400	78	3.0	12	—	85
														Class-B Amp. (Audio) [7]	3000	−205	65/313	630 [9]	20 [8]	22000	665
HK354D	150	5.0	10	4000	300	55	22	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−490	240	50	38	—	690
														Class-C Amp. (Telephony)	3500	−425	210	55	36	—	525
HK354E	150	5.0	10	4000	300	60	35	4.5	3.8	1.1	30	J.	2N	Class C Amp. (Telegraphy)	3500	−448	240	60	45	—	690
														Class-C Amp. (Telephony)	3000	−437	210	60	45	—	525
HK354F	150	5.0	10	4000	300	75	50	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−368	250	75	50	—	720
														Class-C Amp. (Telephony)	3000	−312	210	75	45	—	525
UE-468	150	10	4.05	2500	200	60	18	8.8	7.0	1.25	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B (Audio) [7]	2500	−130	320 [8]	410 [9]	2.5	16000	500
810 1627 [1]	175	10 5.0	4.5 9.0	2500	300	75	36	8.7	4.8	12	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−180	300	60	19	—	575
														Class-C Amp. (Telephony)	2000	−350	250	70	35	—	380
														Grid-Modulated Amp.	2250	−140	100	2.0	4.0	—	75
														Class-B Amp. (Audio) [7]	2250	−60	70/450	380 [9]	13 [8]	11600	725

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
8000	175	10	4.5	2500	300	45	16.5	5.0	6.4	3.3	30	J.	2N	Class-C Amp.-Oscillator	2500	−240	300	40	18	—	575
														Class-C Amp. (Telephony)	2000	−370	250	37	20	—	380
														Grid-Modulated Amp.	2250	−265	100	0	2.5	—	75
														Class-B Amp. (Audio)[7]	2250	−130	65/450	560[9]	7.9[8]	12000	725
GL-5C24	160	10	5.2	1750	107	—	8	5.6	8.8	3.3	—	N.	Fig. 26	Class-A Amp. (Audio)	1500	−155	107	—	—	8200[6]	55
														Class-AB$_1$ Amp. (Audio)[7]	1750	−200	320[8]	390[9]	—	8000	240
RK63 RK63A	200	5.0 6.3	10 14	3000	250	60	37	2.7	3.3	1.1	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−200	233	45	17	—	525
														Class-C Amp. (Telephony)	2500	−200	205	50	19	—	405
														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	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−280	350	54	25	—	685
														Class-C Amp. (Telephony)	2000	−260	300	54	23	—	460
F-127-A	200	10	4.0	3000	325	70	38	13	4	13	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	3000	−250	250	47	18	—	600
														Class-C Amp. (Telephony)	2500	−300	200	58	25.2	—	420
														Class-B Amp. (Audio)[7]	2800	−75	20/400	175[9]	6.65[8]	16600	820
822 822S	200	10	4.0	2500	300	60	30	8.5	13.5	2.1	20 30	J.	3N 2N	Class-C Amp. (Telegraphy)	2500	−190	300	51	17	—	600
														Class-C Amp. (Telephony)	2000	−75	250	43	13.7	—	405
														Class-B Amp. (Audio)[7]	3000	−80	450[8]	362[9]	8.0[8]	16000	1000
4C32	200	10	4.5	3000	300	60	30	5.5	5.8	1.1	60	J.	2N	Class-C Amp.-Oscillator	2000	−165	275	20	10	—	400
														Class-C Amp. (Telephony)	2000	−200	250	20	15	—	375
GL-592	200	10	5.0	3500	250	50	24	3.6	3.3	0.41	110	N.	Fig. 52	Class-C Amp.-Oscillator	2600	−240	250	45	18	—	425
														Class-C Amp. (Telephony)	2000	−500	250	50	—	—	—
4C34 HF300	200	11–12	4.0	3000	275	60	23	6.0	6.5	1.4	60 20	J.	2N	Class-C Amp. (Telegraphy)	3000	−400	250	28	16	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B Amp. (Audio)[7]	3000	−115	60/360	450[9]	13[8]	20000	780
T014 HV12	200	10	4.0	2500	200	60	12	8.5	12.8	1.7	30	J.	3N	Class-C Amp. (Telegraphy)	2500	−240	300	30	10	—	575
														Class-C Amp. (Telephony)	2000	−370	300	40	20	—	485
														Class-B Amp. (Audio)[7]	2000	−160	50/275	350[9]	7.0[8]	14400	400
T022 HV27	200	10	4.0	2500	300	60	27	8.5	13.5	2.1	30	J.	3N	Class-C Amp. (Telegraphy)	2500	−175	300	50	15	—	585
														Class-C Amp. (Telephony)	2000	−195	250	45	15	—	400
T-300	200	11	6.0	3000	300	—	23	6.0	7.0	1.4	—	—	—	Class-C Amp. (Telegraphy)	3000	−400	250	28	20	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B (Audio)[7]	2500	−100	60/450	—	7.5[8]	—	750
806	225	5.0	10	3300	300	50	12.6	6.1	4.2	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3300	−600	300	40	34	—	780
														Class-C Amp. (Telephony)	3000	−670	195	27	24	—	460
														Class-B Amp. (Audio)[7]	3300	−240	80/475	930[9]	35[8]	16000	1120
3-250A4 250TH	250	5.0	10.5	4000	350	100	37	5.0	2.9	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	2000	−120	350	100	34	—	500
														Class-C Amp. (Telephony)	3000	−210	330	75	42	—	750
														Grid-Modulated Amp.	3000	−160	125	4.5	20	—	125
														Class-B Amp. (Audio)[7]	3000	−65	100/560	460[9]	24[8]	12250	1150
3-250A2 250TL	250	5.0	10.5	4000	350	50	14	3.7	3.1	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	3000	−350	335	45	29	—	750
														Class-C Amp. (Telephony)	3000	−350	335	45	29	—	750
														Grid-Modulated Amp.	3000	−450	125	2.0	15	—	125
														Class-B Amp. (Audio)[7]	3000	−175	100/500	840[9]	17[8]	13000	1000

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	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.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
GL159	250	10	9.6	2000	400	100	20	11	17.6	5.0	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−200	400	17	6.0	—	620
														Class-C Amp. (Telephony)	1500	−240	400	23	9.0	—	450
														Class-B Amp. (Audio)[7]	2000	−100	30/660	400[9]	4.0[8]	6880	900
GL169	250	10	9.6	2000	400	100	85	11.5	19	4.7	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−100	400	42	10	—	620
														Class-C Amp. (Telephony)	1500	−100	400	45	10	—	450
														Class-B Amp. (Audio)[7]	2000	−18	30/660	220[9]	6.0[8]	7000	900
204A 304A	250	11	3.05	2500	275	80	23	12.5	15	2.3	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−200	250	30	15	—	450
														Class-C Amp. (Telephony)	2000	−250	250	35	20	—	350
														Class-B Amp. (Audio)[7]	3000	−100	80/372	500[9]	18[8]	20000	700
308B	250	14	4.0	2250	325	75	8.0	13.6	17.4	9.3	1.5	N.	T-2A	Class-C Amp. (Telegraphy)	1750	−345	300	—	—	—	350
															1500	−300	300	—	—	—	300
														Class-B Amp. (Audio)[7]	1750	−215	30/300	—	35[8]	5200	575
HK454H	250	5.0	11	5000	375	85	30	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
HK454-L	250	5.0	11	5000	375	60	12	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
212E 241B 312E	275	14	4.0	3000	350	75	16	14.9	18.8	8.6	1.5	N.	T-2A T-2AA	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
														Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
														Class-B Amp. (Audio)[7]	2000	−105	40/300	—	50[8]	8000	650
300T[1]	300	8.0	11.5	3500	350	75	16	4.0	4.0	0.6	—	J.	2N	Class-C Amp. (Telegraphy)	2000	−225	300	—	—	—	400
HK304-L	300	5/10	26/13	3000	1000	150	10	12	9.0	0.8	—	N.	4BC	Class-C Amp. (Telephony)	1500	−200	300	75	—	—	300
527	300	5.5	135.0	—	—	—	38	19.0	12.0	1.4	200	N.	T-4B	Oscillator at 200 Mc.		Approximately 250 watts output					
HK654	300	7.5	15	4000	600	100	22	6.2	5.5	1.5	20	J.	2N	Class-C Amp. (Telegraphy)	2000	−380	500	75	57	—	720
														Class-C Amp. (Telephony)	2000	−365	450	110	70	—	655
														Grid-Modulated Amp.	3500	−210	150	15	15	—	210
3-300A3 304TH	300	5/10	25/12.5	3000	900	170	20	13.5	10.2	0.7	40	N.	4BC	Class-C Amplifier	1500	−125	667	115	25	—	700
														Class-B Amp. (Audio)[7]	3000	−150	134/667	420[9]	6.0[8]	10200	1400
3-300A2 304TL						150	12	8.5	9.1	0.6	40	N.	4BC	Class-C Amplifier	1500	−250	665	90	33	—	700
														Class-B Amp. (Audio)[7]	3000	−260	130/667	650[9]	6.0[8]	10200	1400
833A	300	10	10	3000	500	100	35	12.3	6.3	8.5	30	N.	T-1AB	Class-C Amp. (Telegraphy)	2000	−200	475	65	25	—	740
														Class-C Amp. (Telephony)	2500	−300	335	75	30	—	635
270A	350	10	4.0	3000	375	75	16	18	21	2.0	7.5	N.	T-1A	Class-C Amp. (Telegraphy)	3000	−375	350	—	—	—	700
														Class-C Amp. (Telephony)	2250	−300	300	80	—	—	450
849[1]	400	11	5.0	2500	350	125	19	17	33.5	3.0	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−250	300	20	8.0	—	560
														Class-C Amp. (Telephony)	2000	−300	300	30	14	—	425
831[1]	400	11	10	3500	350	75	14.5	3.8	4.0	1.4	—	N.	T-1AA	Class-C Amp. (Telegraphy)	3500	−400	275	40	30	—	590
														Class-C Amp. (Telephony)	3000	−500	200	60	50	—	360

* Cathode resistor in ohms.

1 Discontinued.
2 Twin triode. Values, except interelement capacities, are for both sections in push-pull.
3 Output at 112 Mc.
4 Grid-leak resistor in ohms.
5 Max. peak volts, plate pulsed.
6 Per section.
7 Values are for two tubes in push-pull.
8 Max. signal value.
9 Peak a.f. grid-to-grid volts.
10 For single tube.
11 Class-B data in Table I.

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
3A4	2.0	1.4 2.8	0.2 0.1	150	135	0.9	4.8	0.2	4.2	10	B.	7BB	Class-C Amp. (Telegraphy)	150	135	0	−26	18.3	6.5	0.13	2300	—	—	1.2
3D6	—	2.8 1.4	0.11 0.22	150	135	—	7.5	0.3	6.5	50	L.	6BB	Class-C Amp. (Telegraphy)	150	135	—	−20	23	6.0	—	—	—	—	1.4
HY63[1]	3.0	2.5 1.25	0.1125 0.225	200	100	0.6	8.0	0.1	8.0	60	O.	T-8DB	Class-C Amp. (Telegraphy)	200	100	—	−22.5	20	4.0	2.0	—	0.1	—	3.0
													Class-C Amp. (Telephony)	180	100	—	−35	15	3.0	2.0	—	0.2	—	2.0
6AK6	3.5	6.3	0.15	375	250	1.0	3.6	0.12	4.2	54	B.	7BK	Class-C Amp. (Telegraphy)	375	250	—	−100	15	4.0	3.0	—	—	—	4.0
5618	5.0	6.0 3.0	0.23 0.46	300	125	2.0	7.0	0.24	5.0	80	B.	7CU	Class-C Amp. (Telegraphy)	300	75	0	−45	25	7.0	1.5	32000	0.3	—	5.4
6AQ5	8.0	6.3	0.45	350	250	2.0	7.6	0.35	6.0	54	B.	7BZ	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6V6GT	8.0	6.3	0.45	350	250	2.0	9.5	0.7	7.5	10	O.	7AC	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6AG7	9.0	6.3	0.65	375	250	1.5	13	0.06	7.5	10	O.	8Y	Class-C Amp. (Telegraphy)	375	250	—	−75	30	9.0	5.0	—	—	—	7.5
RK64[1]	6.0	6.3	0.5	400	100	3.0	10	0.4	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	100	30	−30	35	10	3.0	—	0.18	—	10
													Class-C Amp. (Telephony)	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	20	M.	T-5CA	Class-C Amp. (Telegraphy)	400	150	—	−50	22.5	7.0	1.5	—	0.1	—	5.0
RK56	8.0	6.3	0.55	300	300	4.5	10	0.2	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	300	—	−40	62	12	1.6	—	0.1	—	12.5
													Class-C Amp. (Telephony)	250	200	—	−40	50	10	1.6	2800	0.28	—	8.5
RK23[1] RK25 RK25B[1]	10	2.5 6.3	2.0 0.9	500	250	8	10	0.2	10	—	M.	6BM	Class-C Amp. (Telegraphy)	500	200	45	−90	55	38	4.0	—	0.5	—	22
													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	45	O	7S	Class-C Amp. (Telegraphy)	350	200	—	−35	50	10	3.5	20000	0.22	—	9
													Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	10000	0.16	—	6.0
2E30	10	6.0	0.7	250	250	2.5	10	0.5	4.5	160	B.	7CQ	Class-C Amp. (Telegraphy)	250	200	—	−50	50	10	2.5	—	0.2	—	7.5
													Class-AB$_2$ Amp. (Audio)[6]	250	250	—	−30	40/120	4/20	2.3[7]	87[8]	0.2	3800	17
6F6 6F6G	12.5	6.3	0.7	400	275	3.0	6.5 8.0	0.2 0.5	13 6.5	10	O.	7AC	Class-C Amp. (Telegraphy)	400	275	—	−100	50	11	5.0	—	—	—	14
													Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	—	0.16	—	6.0
837 RK44[1]	12	12.6	0.7	500	300	8	16	0.2	10	20	M.	6BM	Class-C Amp. (Telegraphy)	500	200	40	−70	80	15	4.0	20000	0.4	—	28
													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
2E24	9.0	6.3[5]	0.65	500	200	2.3	8.5	0.11	6.5	125	O.	7CL	Class-C Amp. (Telephony)	400	180	—	−45	50	8.0	2.5	27500	0.15	—	13.5
														500	180	—	−45	54	8.0	2.5	40000	0.16	—	18.0
	13.5			600	200	2.5							Class-C Amp. (Telegraphy)	400	200	—	−45	75	10.0	3.0	20000	0.19	—	20
														600	195	—	−50	66	10	3.0	40500	0.21	—	27
2E26	13.5	6.3	0.8	600	200	2.5	13	0.2	7.0	125	O.	7CK	Class-C Amp. (Telegraphy)	600	185	—	−45	66	10	3.0	41500	0.17	—	27
													Class-C Amp. (Telephony)	500	180	—	−50	54	9.0	2.5	35500	0.15	—	18
	9.0			500	200	2.3							Class-AB$_2$ Amp. (Audio)[6]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54
802	13	6.3	0.9	600	250	6.0	12	0.15	8.5	30	M.	6BM	Class-C Amp. (Telegraphy)	600	250	40	−120	55	16	2.4	22000	0.30	—	23
													Class-C Amp. (Telephony)	500	245	40	−40	40	15	1.5	16300	0.10	—	12
													Suppressor-Modulated 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	60	O.	7AC	Class-C Amp. (Telegraphy)	300	200	—	−45	60	7.5	2.5	—	0.3	—	12
													Class C Amp. (Telephony)	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	60	M.	5AW	Class-C Amp. (Telegraphy)	425	200	—	−62.5	60	8.5	3.0	—	0.3	—	18
													Class-C Amp. (Telephony)	325	200	—	−45	60	7.0	2.5	—	0.2	—	14

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HY65[1]	15	6.3	0.85	450	250	4.0	9.1	0.18	7.2	60	O.	T-8DB	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.5	—	24
													Class-C Amp. (Telephony)	350	200	—	−45	63	12	3.0	—	0.5	—	16
2E25	15	6.0	0.8	450	250	4.0	8.5	0.15	6.7	125	O.	5BJ	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.4	—	24
													Class-C Amp. (Telephony)	400	200	—	−45	60	12	3.0	—	0.4	—	16
306A	15	2.75	2.0	300	300	6.0	13	0.35	13	—	M.	T-5CB	Class-AB₂ Amp. (Audio)[6]	450	250	—	−30	44/150	10/40	3.0	142[8]	0.9[7]	6000	40
													Class-C Amp. (Telephony)	300	180	—	−50	36	15	3.0	8000	—	—	7.0
307A / RK-75	15	5.5	1.0	500	250	6.0	15	0.55	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	500	250	0	−35	60	13	1.4	20000	—	—	20
													Suppressor-Modulated Amp.	500	200	−50	−35	40	20	1.5	14000	—	—	6.0
832[3]	15	6.3 / 12.6	1.6 / 0.8	500	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−65	72	14	2.6	21000	0.18	—	26
													Class-C Amp. (Telephony)	425	200	—	−60	52	16	2.4	14000	0.15	—	16
832A[3]	15	6.3 / 12.6	1.6 / 0.8	750	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	750	200	—	−65	48	15	2.8	36500	0.19	—	26
													Class-C Amp. (Telephony)	600	200	—	−65	36	16	2.6	25000	0.16	—	17
844[1]	15	2.5	2.5	500	180	3.0	9.5	0.15	7.5	—	M.	5AW	Class-C Amp. (Telegraphy)	500	175	—	−125	25	—	5.0	—	—	—	9.0
													Class-C Amp. (Telephony)	500	150	—	−100	20	—	—	—	—	—	4.0
865	15	7.5	2.0	750	175	3.0	8.5	0.1	8.0	15	M.	T 4C	Class-C Amp. (Telegraphy)	750	125	—	−80	40	—	5.5	—	1.0	—	16
													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	45	O.	T9H	Class-C Amp. (Telegraphy)	400	300	—	−55	75	10.5	5.0	9500	0.36	—	19.5
													Class-C Amp. (Telephony)	325	285	—	−50	62	7.5	2.8	5000	0.18	—	13
													Class-AB₂ Amp. (Audio)[6]	400	300	0	−16.5	75/150	6.5/11.5	—	77[8]	0.4[7]	6000	36
5516	15	6.0	0.7	600	250	5.0	8.5	0.12	6.5	80	O.	7CL	Class-C Amp. (Telegraphy)	600	250	—	−60	75	15	5.0	—	0.5	—	32
													Class-C Amp. (Telephony)	475	250	—	−90	63	10	4.0	22500	0.5	—	22
													Class-AB₂ (Audio)[6]	600	250	—	−25	36/140	1/24	4[7]	80[8]	0.16	10500	67
254A	20	5.0	3.25	750	175	5.0	4.6	0.1	9.4	—	M.	T-4C	Class-C Amplifier	750	175	—	−90	60	—	—	—	—	—	25
6L6 / 6L6G	21	6.3	0.9	400	300	3.5	10 / 11.5	0.4 / 0.9	12 / 9.5	10	O.	7AC	Class-C Amp.-Oscillator	400	300	—	−125	100	12	5.0	—	—	—	28
													Class-C Amp. (Telephony)	325	250	—	−70	65	—	9.0	—	0.8	—	11
6L6GX	21	6.3	0.9	500	300	3.5	11	1.5	7.0	—	O.	7AC	Class-C Amp. (Telegraphy)	500	250	—	−50	90	9.0	2.0	—	0.25	—	33
													Class-C Amp. (Telephony)	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	60	O.	7AC	Class-C Amp.-Oscillator	500	250	—	−50	90	9.0	2.0	—	0.5	—	30
													Class-C Amp. (Telephony)	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	30	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	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	—	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	300	200	—	−45	60	15	5.0	6700	0.34	—	12
1614	25	6.3	0.9	450	300	3.5	10	0.4	12.5	80	O.	7AC	Class-C Amp. (Telegraphy)	450	250	—	−45	100	8	2.0	12500	0.15	—	31
													Class-C Amp. (Telephony)	375	250	—	−50	93	7.0	2.0	10000	0.15	—	24.5
													Class-AB₁ Amp. (Audio)[6]	530	340	—	−36	60/160	20[7]	—	72[8]	—	7200	50
RK41[1] / RK39	25	2.5 / 6.3	2.4 / 0.9	600	300	3.5	13	0.2	10	30	M.	5AW	Class-C Amp. (Telegraphy)	600	300	—	−90	93	10	3.0	—	0.38	—	36
													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	60	M.	5AW	Class-C Amp. (Telegraphy)	600	250	—	−50	85	9.0	4.0	39000	0.4	—	40
													Class-C Amp. (Telephony)	475	250	—	−50	100	9.0	3.5	25000	0.2	—	27
													Class-AB₂ Amp. (Audio)[6]	600	300	—	−30	200[7]	10[7]	—	—	0.1[7]	—	80
815[3]	25	12.6 / 6.3	0.8 / 1.6	500	200	4.0	13.3	0.2	8.5	125	O.	8BY	Class-C Amp.-Oscillator	500	200	—	−45	150	17	2.5	—	0.13	—	56
													Class-C Amp. (Telephony)	400	175	—	−45	150	15	3.0	—	0.16	—	45
													Class-AB₂ Amp. (Audio)[3]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
254B	25	7.5	3.25	750	150	5.0	11.2	0.085	5.4	—	M.	T-4C	Class-C Amplifier	750	150	—	−135	75	—	—	—	—	—	30
1624	25	2.5	2.0	600	300	3.5	11	0.25	7.5	60	M.	T-5DC	Class-C Amp. (Telegraphy)	600	300	—	−60	90	10	5.0	30000	0.43	—	35
													Class-C Amp. (Telephony)	500	275	—	−50	75	9.0	3.3	25000	0.25	—	24
													Class-AB2 Amp. (Audio) [6]	600	300	—	−25	42/180	5/15	106 [8]	—	1.2 [7]	7500	72
3DX3	25	6.3	3.0	1500	200	—	—	—	—	250	S.	Fig. 40	Class-C Amp. (Telegraphy)	1000	200	—	−155	75	—	2.8	—	0.57	—	50
3E22 [3]	30	12.6 6.3	0.8 1.6	560	225	6.0	14	0.22	8.5	—	O.	8BY	Class-C Amp. (Telegraphy) [3]	600	200	—	−55	160	20	7.0	20000	0.45	—	72
													Class-C Amp. (Telephony) [3]	560	200	—	−50	160	20	6.5	18000	0.4	—	67
RK66	30	6.3	1.5	600	300	3.5	12	0.25	10.5	60	M.	T-5C	Class-C Amp.-Oscillator	600	300	—	−60	90	11	5.0	—	0.5	—	40
													Class-C Amp. (Telephony)	500	—	—	−50	75	8.0	3.2	25000	0.23	—	25
807 1625	30	6.3 12.6	0.9 0.45	750	300	3.5	11	0.2	7.0	60	M.	5AW 5AZ	Class-C Amp. (Telegraphy)	750	250	—	−45	100	6	3.5	85000	0.22	—	50
													Class-C Amp. (Telephony)	600	275	—	−90	100	6.5	4.0	50000	0.4	—	42.5
													Class-AB2 Amp. (Audio) [6]	750	300	—	−32	60/240	5/10	92 [8]	—	0.2 [7]	6950	120
2E22	30	6.3	1.5	750	250	10	13	0.2	8.0	—	M.	5J	Class-C Amp.-Oscillator	500	250	22.5	−60	100	16	6.0	15000	0.55	—	34
													Class-C Amp.-Oscillator	750	250	22.5	−60	100	16	6.0	30000	0.55	—	53
													Suppressor-Modulated Amp.	750	250	−90	−65	55	29	6.5	17000	0.6	—	16.5
3D23 TB-35	35	6.3	3.0	—	—	—	6.5	0.2	1.8	250	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	110	22	15	—	4.5	—	130
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
RK20 [1] RK20A RK46 [1]	40	7.5 7.5 12.6	3.0 3.25 2.5	1250	300	15	14	0.01	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	1250	300	45	−100	92	36	11.5	—	1.6	—	84
													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	40	6.3	1.5	600	300	5.0	15.4	0.23	6.5	60	M.	T-5D	Class-C Amp.-Oscillator	600	250	—	−60	100	12.5	4.0	30000	0.25	—	42
													Class-C Amp. (Telephony)	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
													Class-AB2 Amp. (Audio) [6]	600	300	—	−35	200 [7]	18 [7]	5.0 [7]	—	0.3 [7]	—	80
829 [1,3]	40	6.3 12.6	2.25 1.12	500	225	40	14.5	0.1	7.0	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	—	83
													Class-C Amp. (Telephony)	425	200	—	−60	212	35	11	6400	0.8	—	63
													Grid-Modulated Amp.	500	200	—	−38	120	10	2.0	—	0.5	—	23
829A [1,3]	40	6.3 12.6	2.25 1.12	750	240	7.0	14.4	0.1	7.0	200	N.	7BP	Class-C Amp.-Oscillator	750	200	—	−55	160	30	12	18300	0.8	—	87
													Class-C Amp. (Telephony)	600	200	—	−70	150	30	12	13300	0.9	—	70
													Grid-Modulated Amp.	750	200	—	−55	80	5.0	0	—	0.7	—	24
829B [3] 3E29 [3]	30	12.6 6.3	1.125 2.25	750	225	6	14.5	0.12	7.0	200	N.	7BP	Class-C Amp. (Grid Mod.)	500	200	—	−38	120	10	2	—	0.5	—	23
	28			600	225	7							Class-C Amp. (Telephony)	425	200	—	−60	212	35	11.0	6400	0.8	—	63
	40			750	240	7							Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12.0	9300	0.7	—	83
HY1269	40	6.3 12.6	3.5 1.75	750	300	5.0	16.0	0.25	7.5	6	M.	T-5DB	Class-C Amp.-Oscillator	750	300	—	−70	120	15	4	—	0.25	—	63
													Class-C Amp. (Telephony)	600	250	—	−70	100	12.5	5	35000	0.5	—	42
													Grid-Modulated Amp.	750	300	—	—	80	—	—	—	—	—	20
													Class-AB2 Amp. (Audio) [6]	600	300	—	−35	200 [7]	—	—	—	0.3	—	80
3D24	45	6.3	3.0	2000	400	10	6.5	0.2	2.4	125	L.	T-9J	Class-C Amp.-Oscillator	2000	375	—	−300	90	20	10	—	4.0	—	140
														1500	375	—	−300	90	22	10	—	4.0	—	105
715-B	50	26/28	—	—	—	—	—	—	—	—	—	—	Class-C Amp. (Telegraphy)	1500	300	—	—	125	—	—	—	—	—	—

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
5562	45	6.3	3.0	2000	400	8	6.5	0.2	1.8	120	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	116	21	12	—	3.6	—	135
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
RK47	50	10	3.25	1250	300	10	13	0.12	10	—	M.	T-5D	Class-C Amp. (Telegraphy)	1250	300	—	−70	138	14	7.0	—	1.0	—	120
													Class-C Amp. (Telephony)	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	—	M.	T-6C	Class-C Amp. (Telegraphy)	1250	300	20	−55	100	36	5.5	—	0.7	—	90
													Class-C Amp. (Telephony)	1000	—	40	−40	95	35	7.0	22000	1.0	—	65
													Suppressor-Modulated Amp.	1250	—	−85	−50	50	42	5.0	22000	0.55	—	23
804	50	7.5	3.0	1500	300	15	16	0.01	14.5	15	M.	T-5C	Class-C Amp. (Telegraphy)	1500	300	45	−100	100	35	7.0	34000	1.95	—	110
													Class-C Amp. (Telephony)	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-Modulated Amp.	1500	300	−50	−115	50	32	7.0	—	0.95	—	28
4D22	50	25.2 12.6	0.8 1.6	750	350	14	28	0.27	13	60	N.	Fig. 50	Class-C Amp. (Telegraphy)	750	300	—	−100	240	26	12	—	1.5	—	135
														600	300	—	−100	215	30	10	—	1.25	—	100
4D32		6.3	3.75									Fig. 51	Class-C Amp. (Telephony)	600	—	—	−100	220	28	10	10000	1.25	—	100
														550	—	—	−100	175	17	6	15000	0.6	—	70
													Class-AB₂ Amp. (Audio) [6]	600	250	—	−25	100/365	26 [7]	70 [8]	—	0.45 [7]	3000	125
305A	60	10	3.1	1000	200	6	10.5	0.14	5.4	—	M.	T-4CE	Class-C Amp. (Telegraphy)	1000	200	—	−200	125	—	—	—	—	—	85
													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	—	M.	T-5DB	Class-C Amp. (Telegraphy)	1250	300	—	−80	175	22.5	10	—	1.5	—	152
													Class-C Amp. (Telephony)	1000	300	—	−150	145	17.5	14	—	2.0	—	101
													Grid-Modulated Amp.	1250	300	—	—	78	—	—	—	—	—	32.5
814	65	10	3.25	1500	300	10	13.5	0.1	13.5	30	M.	T-5D	Class-C Amp. (Telegraphy)	1500	300	—	−90	150	24	10	50000	1.5	—	160
													Class-C Amp. (Telephony)	1250	300	—	−150	145	20	10	48000	3.2	—	130
													Grid-Modulated Amp.	1500	250	—	−120	60	3.0	2.5	—	4.2	—	35
4-65A	65	6.0	3.5	3000	400	10	8.0	0.08	2.1	160 [9]	N.	5BK	Class-C Amp. (Telegraphy)	3000	250	—	−90	115	20	10	—	1.7	—	280
				2500	400								Class-C Amp. (Telephony)	2500	250	—	−150	108	16	8	—	1.9	—	225
				3000	600								Class-B Linear Amp.	2500	500	—	−100	20/230	0/35	6 [10]	—	1.8 [10]	—	325 [7]
				3000	600								Class-AB₂ Amp. (Audio) [6]	1800	250	—	−35	50/220	0/25	180 [8]	—	2.2 [7]	20000	270
282A	70	10	3.0	1000	250	5	12.2	0.2	6.8	—	M.	T-4C	Class-C Amp. (Telegraphy)	1000	150	—	−160	100	—	—	—	—	—	33
													Class-C Amp. (Telephony)	750	150	—	−180	100	—	50	—	—	—	50
4E27/ 8001	75	5.0	7.5	4000	750	30	12	0.06	6.5	75	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6	136000	1.4	—	230
													Class-C Amp. (Telephony)	1800	400	60	−130	135	11	8	125000	1.7	—	178
													Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	—	35
HK257 HK257B	75	5.0	7.5	4000	750	25	13.8	0.04	6.7	75 120	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6.0	—	1.4	—	230
													Class-C Amp. (Telephony)	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	80	10	3.25	2000	750	23	13.5	0.05	14.5	30	M.	5J	Class-C Amp. (Telegraphy)	1500	400	75	−100	180	28	12	40000	2.2	—	200
													Class-C Amp. (Telephony)	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
													Class-AB₁ Amp. (Audio) [6]	2000	750	60	−120	50/270	2/60	240 [8]	—	0	18500	385

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
RK28	100	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	150	55	13	21000	2.0	—	210
													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 RK48A	100	10	5.0	2000	400	22	17	0.13	13	—	J.	T-5D	Class-C Amp. (Telegraphy)	2000	400	—	−100	180	40	6.5	—	1.0	—	250
													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	2250	400	22	16.3	0.2	14	30	J.	5BA	Class-C Amp. (Telegraphy)	2250	400	0	−155	220	40	15	46000	4.0	—	375
													Class-C Amp. (Telephony)	2000	350	0	−175	200	40	16	41000	4.3	—	300
													Grid-Modulated Amplifier	2000	400	—	−120	75	3.0	—	—	—	—	50
													Class-B Amp. (Audio) [6]	2500	750	0	−95	35/360	1.2/55	—	—	0.35	17000	650
850	100	10	3.25	1250	175	10	17	0.25	25	15	J.	T-3B	Class-C Amp. (Telegraphy)	1250	175	—	−150	160	—	35	—	10	—	130
													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	7.5	30	M.	T-4CB	Class-C Amp.-Oscillator	3000	300	—	−150	85	25	15	—	7.0	—	165
													Class-C Amp. (Telephony)	2000	220	—	−200	85	25	38	100000	17	—	105
4-125A	125	5.0	6.2	3000	400	20	10.3	0.03	3.0	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	350	—	−150	167	30	9	—	2.5	—	375
													Class-C Amp. (Telegraphy)	2500	350	—	−330	150	30	13	—	6	—	300
RK28A	125	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	170	60	10	—	1.6	—	250
													Class-C Amp. (Telegraphy)	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-Modulated Amp.	2000	—	−45	−115	90	52	11.5	30000	1.5	—	60
803	125	10	5.0	2000	600	30	17.5	0.15	29	20	J.	5J	Class-C Amp. (Telegraphy)	2000	500	40	−90	160	45	12	—	2.0	—	210
													Class-C Amp. (Telephony)	1600	400	100	−80	150	45	25	27000	5.0	—	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
4X-150A [9]	150	6.0	2.8	1000	300	15	14.1	0.02	4.7	165	N.	T-9J	Class-C Amp. (Telegraphy)	1000	250	—	−80	200	39	7	—	0.69	—	148
														750	250	—	−80	200	37	6.5	—	0.63	—	110
														600	250	—	−75	200	35	6	—	0.52	—	85
PE340/ 4D23 [9]	150	5.0	7.5	4000	400	—	11.6	0.06	4.35	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	400	—	−290	200	27	7	—	2.6	—	450
													Class-C Amp. (Telephony)	2500	400	—	−425	180	27	9	—	4	—	350
													Class AB_2 Audio [8]	2500	400	—	−95	284 [7]	7 [7]	—	—	1.8 [7]	19100	460
AT-340	150	5	7.0	4000	400	—	9.04	0.19	4.16	120	J.	5BK	Class-C Amp.-Oscillator	3000	400	—	−500	165	75	—	—	2.4	—	—
RK65	215	5.0	14	3000	500	35	10.5	0.24	4.75	60	J.	T-3BC	Class-C Amp. (Telegraphy)	3000	400	—	−100	240	70	24	—	6.0	—	510
													Class-C Amp. (Telephony)	2500	—	—	−150	200	70	22	30000	6.3	—	380
4-250A	250	5.0	14.5	4000	600	50	12.7	0.06	4.5	85	N.	5BK	Class-C Amp. (Telegraphy)	4000	500	—	−250	250	22	13	—	4.1	—	750
														2500	500	—	−100	325	70	22	—	3.7	—	562
GL-5D24	250	5.0	14.1	4000	350	50	12.7	0.06	4.5	85	N.	5BK	Class-C Amp. (Telegraphy)	Same as 4-250A										GL-5D24
4-400A [9]	400	5.0	14.5	4000	600	35	12.5	0.12	4.7	110	N.	5BK	Class-C Teleg. or Telephony	4000	300	—	−170	270	22.5	10	—	10	—	720
861	400	11	10	3500	750	35	14.5	0.1	10.5	20	N.	T-1B	Class-C Amp. (Telegraphy)	3500	500	—	−250	300	40	40	—	30	—	700
													Class-C Amp. (Telephony)	3000	375	—	−200	200	—	55	70000	35	—	400

[1] Discontinued.
[2] Triode connection—screen grid tied to plate.
[3] Dual tube. Values for both sections, in push-pull. Interelectrode capacitances, however, are for each section.
[4] Terminals 3 and 6 must be connected together.
[5] Filament limited to intermittent operation.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Peak grid-to-grid a.f. volts.
[9] Forced-air cooling required.
[10] Average value.

TABLE XVIII—KLYSTRONS

Type	Freq. Range-Mc.	Cathode Volts	Cathode Amp.	Base Connections	Typical Operation	Beam Volts	Beam Ma. (Max.)	Beam Watts (Max.)	Control-Electrode Volts	Reflector Volts	Cathode Ma.	R.F. Driving Power Watts [4]	Output Watts
2K25/ 723A-B	8702–9548	6.3	0.44	Fig. 60	Reflex Oscillator	300	32	—	—	−130/−185	25	—	0.033
2K26	6250-7060	6.3	0.50	Fig. 60	Reflex Oscillator	300	25	—	—	−65/−120	—	—	0.120
2K-28 [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
2K33	23500–24500	6.3	0.65	Fig. 62	Reflex Oscillator	1800 [7]	—	—	−20/−100	−80/−220	[6]	—	0.04
2K34	2730–3330	6.3	1.6	Fig. 58	Oscillator-Buffer *	1900	150	450	−45	—	75	—	10–14
2K35	2730–3330	6.3	1.6	Fig. 58	Cascade Amplifier *	1500	150	450	0	—	75	0.005	5
2K41	2660–3310	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	+24	−510	60	—	0.75
2K42 [3]	3300–4200	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−650	45	—	0.75
2K43 [3]	4200–5700	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−320	40	—	0.8
2K44 [3]	5700–7500	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−700	43	—	0.9
2K39 [3]	7500–10300	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−660	30	—	0.46
2K46	2730–3330 [1] 8190–10000 [2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1500	60	60	−90	—	30	0.01/0.07	0.01–0.07
2K47	250–280 [1] 2250–3360 [2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1000	60	60	−35	—	50	3.5	0.15
2K56	3840–4460	6.3	0.5	Fig. 60	Reflex Oscillator	300	25	—	—	−85/−150	—	—	0.090
3K21 [3]	2300–2725	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K22 [3]	3320–4000	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K23 [3]	950–1150	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K27 [3]	750–960	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K30 (410R)[3]	2700–3300	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
707B [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
QK159	2950–3275	6.3	0.65	Fig. 63	Reflex Oscillator	300	45	—	300	−100/−175	20	—	0.150
Z-668	21900–26100	—	—	—	Reflex Oscillator *	1700	—	15	—	−1700/−2300	—	—	0.02

[1] Input frequency.
[2] Output frequency.
[3] Tuner required.
[4] At max. ratings.
[5] Has demountable tuning cavity.
[6] Cathode current specified on each tube.
[7] G2 and G3 voltage.
* Forced-air cooling required.

TABLE XIX—CAVITY MAGNETRONS

Type	Class	Band or Range Mc.	Heater		Maximum Ratings				Typical Operation					
			Volts	Amps.	Anode KV.	Anode Amps.	Duty Cycle	Input Watts	Anode KV.	Anode Amps.	Field Gauss	Pulse µ Sec.	P.P.S.	Peak Pwr. Output KW.
RK2J22	1	3267–3333	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2250	1.0	1000	265
RK2J23	1	3071–3100	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J24	1	3047–3071	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J25	1	3019–3047	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J26	1	2992–3019	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J27	1	2965–2992	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J28	1	2939–2965	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J29	1	2914–2939	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J30	1	2860–2900	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J31	1	2820–2860	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J32	1	2780–2820	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J33	1	2740–2780	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J34	1	2700–2740	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J36	1	9003–9168	6.3	1.3	13.5	12.0	.002	200	11.5	10.0	2500	1.0	1000	15.0
RK2J38	1	3249–3263	6.3	1.25	6.0	8.0	.012	200	4.9	3.0	Pkg.	1.0	2000	5.0
RK2J39	1	3267–3333	6.3	1.25	6.0	8.0	.002	200	5.4	5.0	Pkg.	1.0	2000	8.7
RK2J48	1	9310–9320	6.3	1.0	16.0	16.0	.002	230	12.0	12.0	4850	1.0	1000	50.0
RK2J49	1	9000–9160	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J50	1	8740–8890	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J54	2	3123–3259	6.3	1.5	14.0	15.0	.002	250	11.6	12.5	1400	1.0	2000	45.0
RK2J55	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J56	1	9215–9275	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J58	2	2992–3100	6.3	1.5	22.0	15.0	.002	600	10.5	12.5	1450	1.0	2000	50.0
RK2J61A	2	3000–3100	6.3	1.5	15.0	15.0	.002	250	10.7	12.5	1300	1.0	2000	35.0
RK2J62A	2	2914–3010	6.3	1.5	15.0	15.0	.002	250	10.2	12.5	1300	1.0	2000	35.0
RK2J66	2	2845–2905	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J67	2	2795–2855	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J68	2	2745–2805	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J69	2	2595–2755	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK4J31	1	2860–2900	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J32	1	2820–2860	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J33	1	2780–2820	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J34	1	2740–2780	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J35	1	2700–2740	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J36	1	3650–3700	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J37	1	3600–3650	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J38	1	3550–3600	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J39	1	3500–3550	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J40	1	3450–3500	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J41	1	3400–3450	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J43	1	2992–3019	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J44	1	2965–2992	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J53	1	2793–2813	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J54	1	6875–6775	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J55	1	6775–6675	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J56	1	6675–6575	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J57	1	6575–6475	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J58	1	6475–6375	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J59	1	6375–6275	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK725A	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.0	12.0	5400	1.0	1000	50.0

[1] Fixed-frequency—Pulsed. [2] Tunable—Pulsed.

Index

Charts and Tables

Formulas

Text

The Catalog Section

☆ ☆ ☆

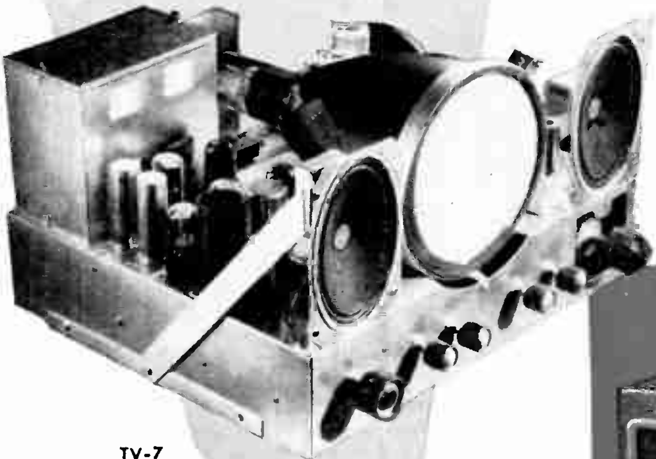
In the following pages is a catalog-file of products of the principal manufacturers who serve the short-wave field. Appearance in these pages is by invitation—space has been sold only to those dependable firms whose established integrity and whose products have met with the approval of the American Radio Relay League.

INDEX OF ADVERTISERS

CATALOG SECTION

The Radio Amateur's Handbook

see inside . . . then decide on NATIONAL TELEVISION



TV-7

Clearer, Steadier Pictures . . .
Better Designed and Built

National Television is a product of the same engineering skill, painstaking craftsmanship and years of experience that have made National receivers so widely respected among radio amateurs. Compare the chassis, components and wiring with any other set at any price. Compare the clarity, stability and realism of the picture. In both construction and performance, you'll find National gives you more honest value for every dollar.

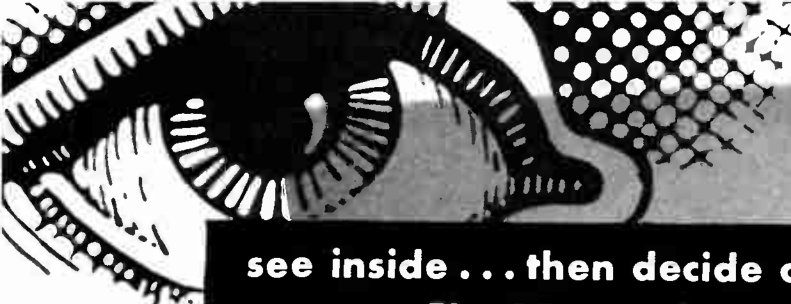
$189.50*

Also available in handsome mahogany cabinet at **$199.50***

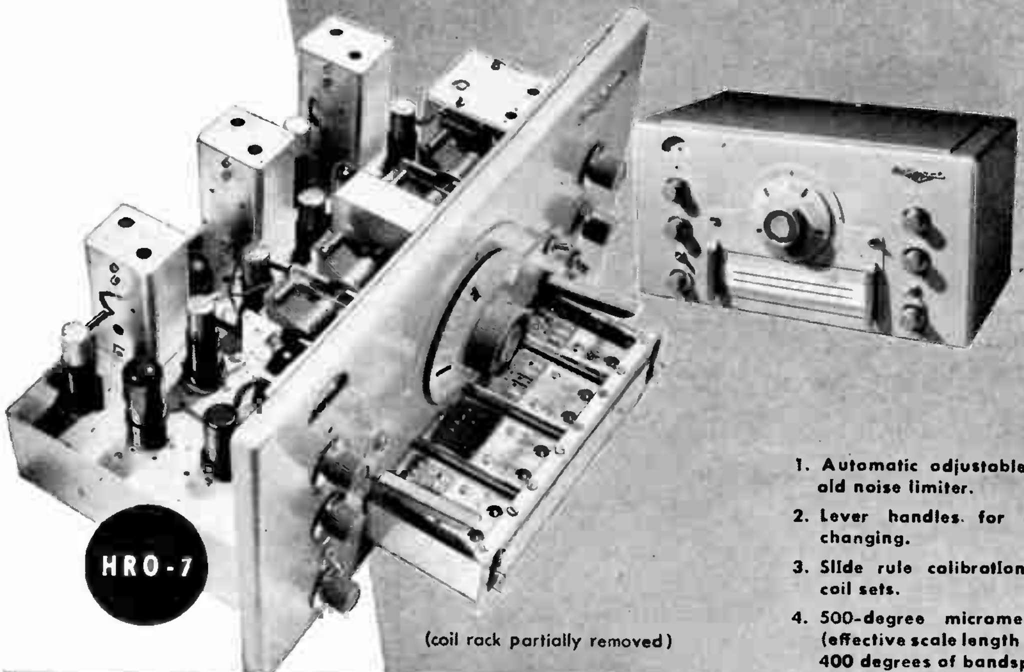
- Coil switching assures equivalent of separate, high-Q tuned circuits for each channel to improve sensitivity, stability.
- Automatic Gain Control corrects for variations in signal strength.
- Greater resolution gives finer, more realistic shading.
- Extra-stable synchronizing circuit locks picture in place, eliminates need for constant re-tuning.
- Uses 3-stage 37 mc IF instead of conventional 21 mc — minimizing picture interference caused by other radio services.
- Specially designed, double-tuned RF bandpass circuits improve selectivity and image ratio.
- Hum-free power supply and flanking dual speakers result in amazingly realistic binaural sound.
- Automatic Station Selector selects picture and sound simultaneously. Fine tuning control makes possible pin-point accuracy.

*Prices slightly higher west of the Rockies.

NATIONAL COMPANY inc.
61 Sherman St., Malden, Mass.



see inside . . . then decide on world-famous National RECEIVERS



(coil rack partially removed)

1. Automatic adjustable-threshold noise limiter.
2. Lever handles for coil set changing.
3. Slide rule calibration on all coil sets.
4. 500-degree micrometer dial (effective scale length 12 feet). 400 degrees of bandspread on 80, 40, 20, 11-10 meters!
5. Accessory socket and switch for NFM adaptor or phonograph.
6. Two tuned RF stages.
7. Two IF stages.
8. Precision gear drive eliminates backlash.
9. Voltage-regulated high frequency oscillator for exceptionla stability.

the finest amateur receiver National has ever made!

Subjected to the severest tests of government, commercial and amateur use for 14 years, the basic HRO design has set a new high in receiver performance. Now, here it is in its newest, finest form. As always, the major components are National designed and made.

RANGE: 1.7 to 30 mcs (Additional coils available for 50 to 430 kcs. 480 to 2050 kcs, 30 to 35 mcs.)

SENSITIVITY: 1 microvolt or better.

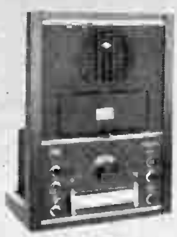
IMAGE REJECTION: Better than 30 db at 30 mcs.

SIGNAL-TO-NOISE RATIO: Exceeds 16 db with 5 microvolts input.

AVC CHARACTERISTIC: to ± 10 db between 1.0 and 100,000 microvolts input.

$312.86*

(Complete with coils and power supply, less speaker)

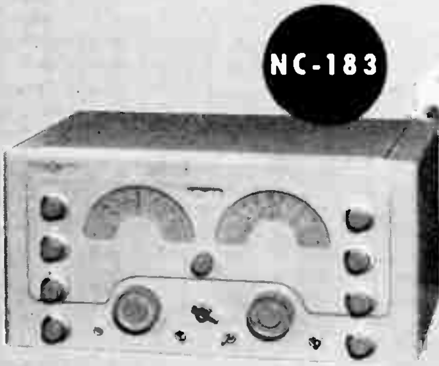


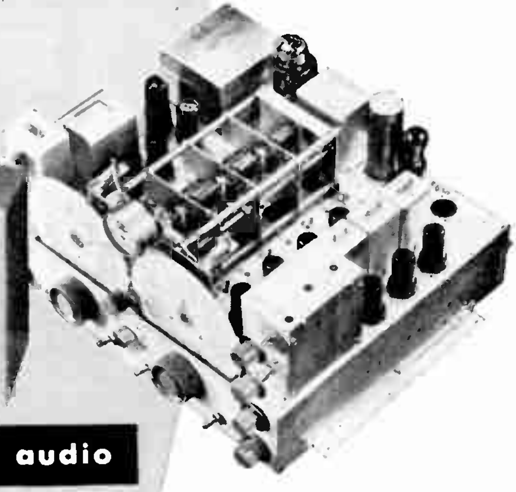
Deluxe HRO-7C

The incomparable HRO-7 power supply 10" speaker, coils and coil compartment all in one convenient table unit.

$358.50*

NAT

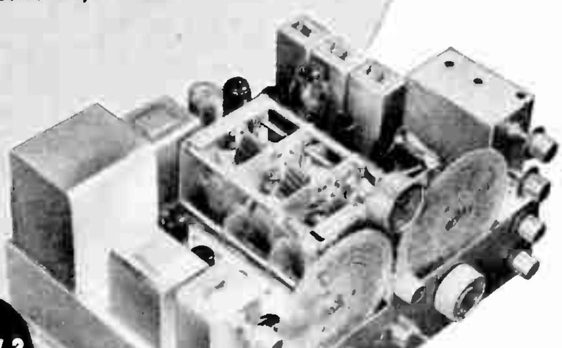


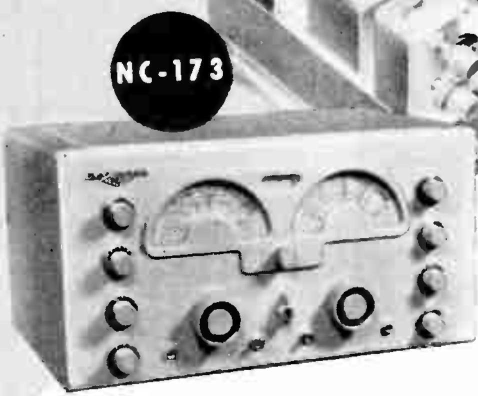


2 RF stages plus high-fidelity audio

- Covers 540 kcs to 31 mcs plus 48-56 mcs. Calibrated amateur bandspread for 6, 10-11, 20, 40 and 80 meter bands. Gear drive tuning dials.
- Two RF stages on all bands! Image rejection 40 db at 28 mc!
- New automatic "double-diode" noise limiter, effective on both phone and CW!
- New Crystal filter provides 6 steps of selectivity!
- S-meter with adjustable sensitivity for both phone and CW!
- Temperature compensation and voltage regulator provide outstanding stability!
- High-fidelity push-pull audio output! Ideal for phonograph attachment. Continuously variable tone control.
- Accessory socket for NFM adaptor!

$268*
(less speaker)





exceptional sensitivity, stabili

- Covers 540 kcs to 31 mcs plus 48-56 mcs. Calibrated electrical bandspread for 6, 10-11, 20, 40 and 80 meter amateur bands!
- Automatic noise limiter effective on both phone and CW, with adjustable threshold!
- Highly flexible crystal filter provides 6 steps of selectivity!
- S-meter for both phone and CW!
- New temperature compensation and voltage regulation assure exceptional stability!
- Accessory socket for NFM-73 adaptor!
- Trimmer control permits panel adjustment of RF stage!
- Tone control. Phono input jack also provided.

$189.50*
(less speaker)

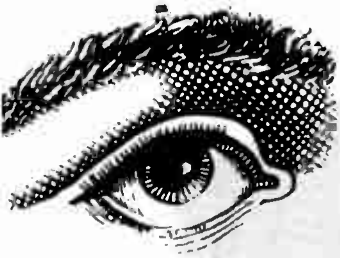
**Prices slightly higher west of the Rockies.*

NAL COMPANY inc.

MALDEN, MASS.

see inside . . . then decide on world-famous

National RECEIVERS



HFS

EXPLORE VHF

Check MUF! Be ready for those DX contacts whether it's on 1, 2, 6 or 10 meters! Here is the latest in VHF design — compact, dependable, modestly-priced — ideal for both your car and your shack.

(less power supply) **$142.00***

COMPLETE COVERAGE 27 MCS — 250 MCS!

. . . in 6 bands, including 1¼, 2, 6, 10 and 11 meter amateur bands.

AM — FM — CW!

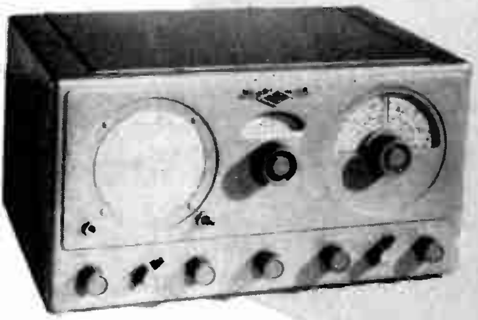
Operation assures optimum signal-to-noise ratio.

MOBILE, PORTABLE OR FIXED!

Operates from standard National 5886 power supply, National 686S vibrator power supply or "A" and "B" batteries! Built-in speaker. Light.

RECEIVER OR CONVERTER!

Makes any receiver capable of tuning to 10.6 mcs a top VHF receiver. All features of connected receiver are usable on VHF.

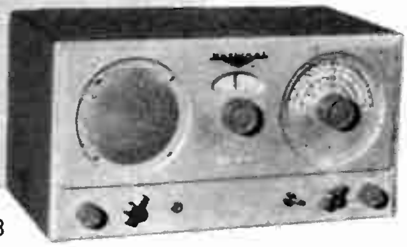


NC - 57

REATEST RANGE IN ITS CLASS!

Complete coverage 540 kc to 55 mc. Separate 6SG7 tuned RF amplifier. Bandspread tuning over entire range. Separate RF gain control for adjusting sensitivity. Pitch control to adjust beat note on CW. Voltage regulated oscillator circuit. Automatic threshold noise limiter to minimize ignition noise, static, etc. Simple 5-position switch for band switching. RF trimmer control to match various types of antenna for maximum efficiency. Provision for battery operation. Accessory socket for SM-57 signal strength meter.

$89.50*



NC - 33

Operates from 110-120 volts AC or DC. Ideal for shipboard and other uses where DC only is available. Covers from 500 kcs distress frequency to 35 mcs. Electrical bandspread on all bands! Broadcast, amateur, police and foreign bands plainly marked. Automatic noise limiter assures optimum reception under all operating conditions. CW oscillator with pitch control provides superb CW reception.

$57.50*

(with built-in speaker)

* Prices slightly higher west of the Rockies.

NATIONAL COMPANY inc.

61 Sherman St., Malden, Mass.

FWG **Net $.60**
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 **Net $.66**
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 **Net $.54**
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 **Net $.70**
This molded R-39 plug has two banana plugs on 3/4" centers and fits FWG, FWH or FWJ above. Leads may be brought out through the top or side.

FWA, Post **Net, each $.20**
Brass Nickel plated

FWE, Jack **Net, each $.15**
Brass Nickel Plated

BWA (not illustrated)
Net $.10
Standard banana plug, silver plated to reduce contact resistance in r.f. circuits.

BWE (not illustrated)
Net $.15
Matching jack for BWA, silver plated.

FWC, Insulator
Net, per pair $.24
R-39 Insulation.

FWB, Insulator
Net, each $.15
Polystyrene insulation.

XS-6 **Net, each $.12**
A low-loss steatite bushing for 1/2" holes. Passes 6-32 screw.

XP-6 **Net, box of ten $.51**
Same as above but polystyrene.

TPB **Net, per dozen $.75**
A threaded polystyrene bushing with removable .093 conductor moulded in, 1/4" diam., 32 thread.

XS-7, (3/8" Hole) **Net $.36**
XS-8, (1/2" Hole) **Net $.48**
Steatite bushings. Prices include male and female bushings with metal fittings.

XS-1, (1" Hole) **Net $.72**
XS-2, (1 1/2" Hole) **Net $.81**
Prices listed are per pair, including metal fittings. Insulation steatite.

AA-3 **Net $.36**
A low-loss steatite spreader for 6 inch line spacing. (600 ohms impedance with No. 12 wire.)

AA-5 **Net $.30**
A low-loss steatite aircraft-type strain insulator.

AA-6 **Net $.54**
A general purpose strain insulator of low-loss steatite.

GS-1, 1/2" x 1 3/8" **Net $.24**
GS-2, 1/2" x 2 7/8" **Net $.30**
GS-3, 3/4" x 2 7/8" **Net $.60**
GS-4, 3/4" x 4 7/8" **Net $.75**

GS-4A, 3/4" x 6 7/8"
Net $1.05
Cylindrical low-loss steatite standoff insulators with nickel plated caps and bases.

GSJ, (not illustrated)
Net $.10
A special nickel plated jack top threaded to fit the 3/4" diameter insulators GS-3, GS-4 & GS-4A.

GS-10, 3/4" high
Net, box of ten $.90

GS-10S (not illustrated) but same as GS-10 except includes threaded stud in top end. **Net, box of ten $1.00**

GS-5, 1 1/4" high **Net $.30**
GS-6, 2" high **Net $.42**
GS-7, 3" high **Net $.75**
These cone type standoff insulators are of low loss steatite. They are molded with a tapped hole in each end for mounting as follows:

GS-5, 8-32 tap 7/16" deep; GS-6 & GS-7, 10-24 tap 11/16" deep; GS-10, 6-32 tap 1/4" deep and GS-10S as noted above.

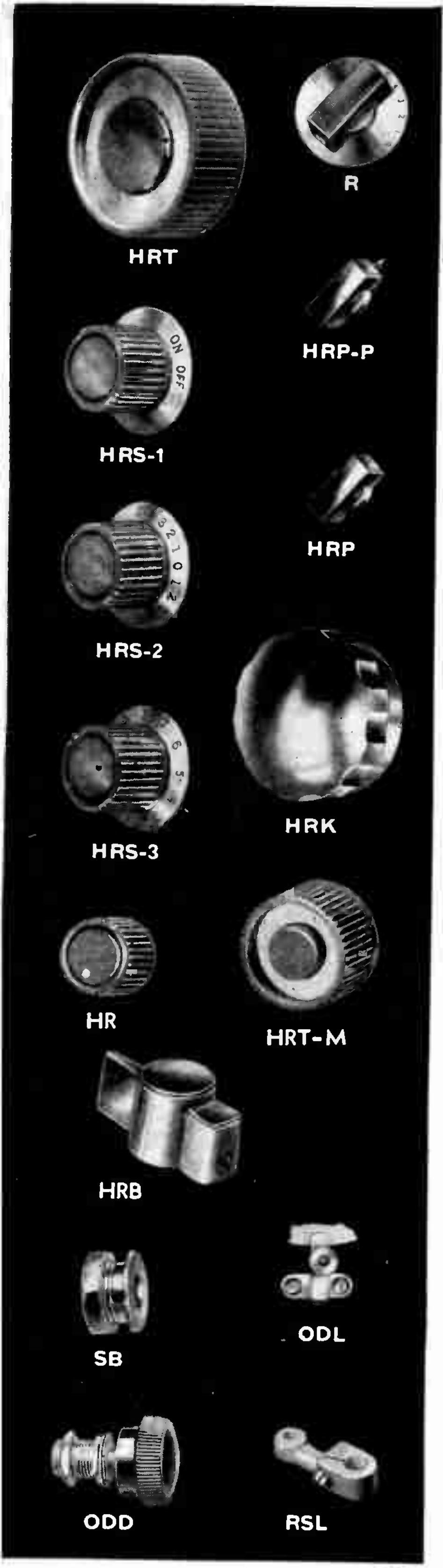
GS-8, with terminal **Net $.54**
GS-9, with jack **Net $.75**
These low-loss steatite standoff Insulators are also useful as lead-through bushings.

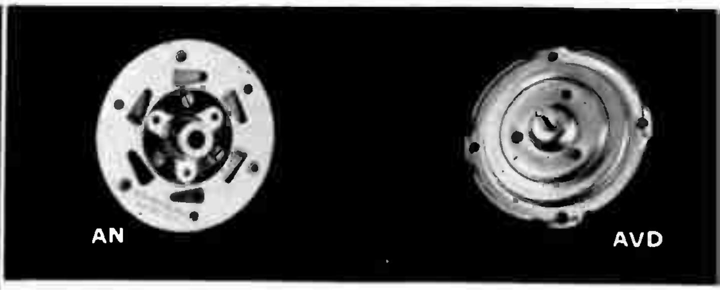
XS-3, (2 3/4" hole) **Net $3.60**
XS-4, (3 3/4" hole) **Net $4.35**
Prices are per pair and include nickel plated spindles, lugs and hardware. These low-loss steatite bowls are ideal for lead-in purposes at high voltages.

XS-5, Without Fittings
Net, each $ 4.95

XS-5F, With Fittings
Net, per pair $10.20
These big low-loss bowls have an extremely long leakage path and a 5 1/4" flange for bolting in place. Insulation steatite. Fittings include nickel plated brass spindles, lugs, nuts and washers.

National





HRT (gray or black) **Net $.75**

The HRT knob is $2\frac{1}{8}$" in dia. and fits $\frac{1}{4}$" shafts. This knob has a chrome appearance circle and combined with the HRS series shown below gives the new look to panel layouts.

HRS (gray or black) **Net $.50**

The HRS series knobs are a popular easy to grip knob. They are molded of high quality plastic and have $1\frac{3}{8}$" dia. chrome plated bevel skirts fit $\frac{1}{4}$" shafts available in the following scales:

HRS-1	ON-OFF	through 30°
HRS-2	5-0-5	through 180°
HRS-3	0-10	through 300°
HRS-4		Single etched line

HR (gray or black) **Net $.30**

An HRS type knob without the chrome plated skirt but with a white dot for spotting relative control settings.

HRB **Net $.45**

Ideal for bandswitching or other applications where a switch is turned to several index positions, the new HRB lever knob has just the right feel — a bright zinc alloy die casting.

SB **Net $.18**

A nickel plated brass bushing $\frac{1}{2}$" dia. (Fits $\frac{1}{4}$" shaft).

ODL **Net $.33**

A locking device which clamps the rim of O, K, L and M Dials. Brass, nickel plated.

ODD **Net $.42**

Vernier pinch drive for O, L, or other plain dials.

AN Vernier Mechanism **Net $**

A vernier mechanism ratio 5-1 an insulated output shaft coup for $\frac{1}{4}$" shafts. Drive Shaft 3/16" knob.

AVD Vernier Mechanism **Net $**

Similar to AN-Output shaft co ling is non insulated.

For commercial uses many variati available. Write for further p ticulars.

R **Net $**

This small dial has a $1\frac{5}{8}$" c scale calibrated 0-10 in 180° increased reading with clockw rotation. Black bakelite knob. I $\frac{1}{4}$" shaft.

HRP-P **Net $**

Black bakelite knob $1\frac{1}{4}$" long a $\frac{1}{2}$" wide. Equipped with poin Especially suitable for use on wa and other rotary switches on la oratory equipment and the li (Fits $\frac{1}{4}$" shaft).

HRP **Net $.**

The type HRP knob has no poin but is otherwise the same as t knob above. Recommended for u calibrated or hard-tuning contro (Fits $\frac{1}{4}$" shaft).

HRK **Net $.**

Black bakelite knob $2\frac{3}{8}$" dial - extremely rugged. This is the kno used on National type O and typ L dials.

HRT-M **Net $.**

This is a smaller version of the HI and was designed originally f use on the NC-57 Receiver — no available in choice of gray or bla — is 1-7/16" in diameter.

ial Net $4.50

Dial Net $3.00

four-inch N and AD Dials have
ne divided and die stamped
es respectively. The N Dial has
ecimal vernier; the AD Dial em-
's a pointer. The planetary drive
a ratio of 5 to 1, and is con-
ed within the body of the dial.
3, 4 or 5 scale. Fits 1/4" shaft.
cify scale.

ial Net $2.70

lvet Vernier" Dial, Type B, has a
ipact veriable ratio 6 to 1 min.,
to 1 max. drive that is smooth
trouble free. The case is black
elite. 1 or 5 scale. 4" dia. Fits
shaft. **Specify scale.**

Dial Net $2.10

BM Dial is a smaller version
the B for use where space is limi-
. The drive ratio is fixed. Al-
ugh small in size, the BM Dial
the same smooth action as the
ger units. 1 or 5 scale. 3" dia.
; 1/4" shaft. **Specify scale.**

1 Dial Net $2.25

original "Velvet Vernier" mech-
sm in a metal skirted dial 3" in
. ratio 5 to 1. It is available
h 2, 3, 4, 5 or 6 scale and fits
' shaft.

Dial Net $1.00

new P dial is the same as the
1 except direct drive.

pe O, 3 1/2" dia., scale 2, with
RK knob, fits 1/4" shafts. Net $1.00

pe L, same as O except 5" dia.,
ale 2 only. Net $1.95

pe K, same as O except less knob,
mplete with ODD vernier drive,
ale 2 only. Net $1.50

pe M, same as K except 5" dia.,
ale 2 only. Net $2.25

The dials at the right are for individual calibration: all four employ the noted 5:1 drive ratio Velvet Vernier mechanism and are of excellent quality.

MCN Dial **Net $2.70**

The MCN dial has been scaled down to lend itself ideally to mobile installations and small converters and tuners. It may also be mounted on the standard 3 1/2" rack panel where such mounting may be desirable. The dial provides three calibrating scales and a 0-100 logging scale. On the rear side of the dial, the mechanism extends 1/4" below the dial frame. 2 3/4" H. x 3 7/8" W.

SCN Dial **Net $3.00**

The SCN dial provides the same dial scales as the ACN dial but in a reduced size. It is used where economy of panel-mounting space is desirable and where a smaller dial would be out of proportion with the size of the panel. 4-7/16" H. x 6 1/4" W.

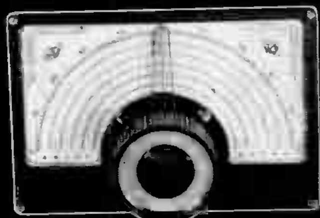
ICN Dial **Net $6.00**

The ICN dial meets those hundreds of requests from amateurs the world over for an illuminated ACN dial. Two dial lights mounted on the top corners of the dial provide efficient and even illumination on all bands. The dial window has been blanked out in semi-circular shape to prevent shadow casting. Dial scales are the same as those used on the ACN dial. 5 1/8" H. x 7 1/4" W.

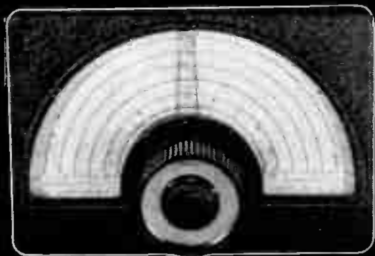
ACN Dial **Net $3.30**

The ACN is the original of this type dial, a National design for the benefit of experimenters who "build their own" and desire direct calibration 5" H. x 7 1/4" W.

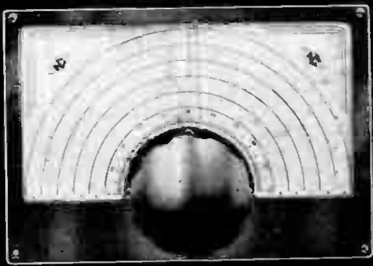
MCN



SCN



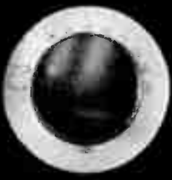
ICN



ACN

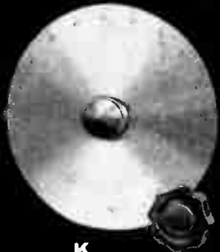
DIAL SCALES

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



AM

O
L

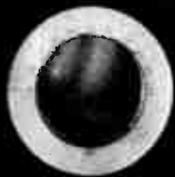


K
M

N
AD

B

BM



P

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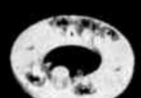
XLA

XLA-S

XLA-C



XCA

XMA

XOA-7 (Axial)
XOA-C-7

XOR-7 (Radial)
XOR-C-7

XOA-C-9

XLA **Net $.99**
A low-loss socket for the 6F4 and 950 series acorn tubes for frequencies as high as 600 Mc. Conventional by-pass condensers may be compactly mounted between the contact terminals and the chassis. Low contact resistance, short and direct leads and low and constant inductance are features.

XLA-S **Net $.36**
An internal shield fitting the XLA socket and suitable for tubes such as the 956.

XLA-C **Net $.36**
This miniature by-pass condenser may be mounted inside the socket, directly below the contact. Capacities of 50 or 100 mmf. available.

XCA **Net $.99**
A low-loss steatite socket for acorn triodes. Pin grips are designed to accept tube prongs with minimum strain but exert maximum pressure when seated.

XMA **Net $1.32**
For pentode acorn tubes, this socket has built-in by-pass condensers. The base is a copper plate.

XOA-7 (mica-filled bakelite) **Net $.50**

XOA-C-7 (ceramic) **Net $.50**

XOR-7 (mica-filled bakelite) **Net $.50**

XOR-C-7 (ceramic) **Net $.50**
These high quality sockets for the 7 pin miniature tubes have silver plated beryllium copper contacts that correctly grip the tube pins close to the base of the tube to provide the short leads and low inductance so necessary in ultra-high frequency design.
A novel feature of these new sockets is the interchangeability of the contacts, which are easily removed for replacement. This permits the use of a mixture of axial (XOA) and radial (XOR) type contacts in the same socket to obtain the shortest possible leads, or minimum size in tight places. The above sockets all mount with two 4-40 screws on .875" centers. Chassis cutout should be ¾" dia. Shields for use with these sockets are on page 21.

XOA-C-9 (ceramic) **Net $.57**

XOR-C-9 (ceramic) **Net $.57**
These sockets are for the new 9-pin miniature tubes. The XOR-C-9 (not illustrated) has radial contacts. Both have all of the features described above for the 7-pin types and they also mount with 4-40 screws. Mounting center dimension is 1⅛", the chassis cutout should be 13/16" dia.

CIR SERIES SOCKETS

Any Type **Net $.30**
Always a popular National component, type CIR Sockets feature low-loss steatite insulation, a contact that grips the tube prong for its entire length, and a metal ring for six position mounting.
XC-4, 5, 6, 7S, 7L and CIR-4, 5, 6, 7S and 7L all have 1-27/32" mounting centers. CIR-8E has slotted holes in plate but will mount on 1-27/32" center. CIR-8 and XC-8 have 1½" mounting centers.

XC SERIES SOCKETS

XC-4	Net $.36
XC-5	Net $.39
XC-6	Net $.42
XC-7S	Net $.45
XC-7L	Net $.45
XC-8	Net $.39

National wafer sockets have exceptionally good contacts with high current capacity together with low loss steatite insulation. All types have a locating groove to make tube insertion easy. The XC-6 is ideal for use with AR-17 coils shown on page 24.

HX-29 **Net $.81**
A low-loss wafer socket with steatite insulation for the popular 829 and 832 tubes.

JX-51 **Net $.81**
A low loss steatite wafer socket for the 813 and other tubes having the Giant 7-pin base. (not illustrated)

XM-10 **Net $.90**
A heavy duty metal shell socket for tubes having the XU 4-pin base.

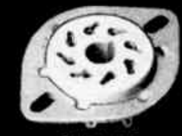
XM-50 **Net $1.20**
(see XM-10 for style)
A heavy duty metal shell socket for tubes having the Jumbo 4-pin base ("fifty watters").

HX-100S **Net $1.65**
With Standoff Insulators
A low loss wafer socket suitable for the type 4-125-A, 4-250-A and other tubes using the Giant 5-pin base. Shield grounding clips are supplied which mount on the chassis with the socket mounting screws to ground the tube shield at three points. Air holes are provided in the socket to permit forced air cooling.

HX-100 **Net $.99**
Same as above less standoff insulators.

CIR-5

CIR-8



CIR-8E

XC-5

XC-8

HX-29

XM-10

HX-100S

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SHAFT COUPLINGS

TX-19 Net $1.25

A steatite insulated flexible coupling for 1/4" shafts. Conservatively rated at 5000 volts peak. Diameter 1 3/8", length 1". Length and flashover voltage can be increased by turning collars outboard.

TX-11 Net $.42

The flexible shaft of this coupling connects shafts at angles up to 90 degrees, and eliminates misalignment problems. Fits 1/4" shafts. Length 4 1/4".

TX-12, Length 4 5/8" Net $ .90
TX-13, Length 7 1/8" Net $1.05

These couplings use flexible shafting like the TX-11 above, but are also provided with steatite insulators at each end.

TX-1, Leakage path 1" Net $.65

TX-2, Leakage path 2 1/2" Net $.75

Flexible couplings with glazed steatite insulation which fit 1/4" shafts.

TX-20 Net $1.25

A small bakelite insulated flexible coupling of the "Hooke's joint" type. Accommodates up to five degrees angular misalignment as well as 1/64" offset of centers. For 1/4" shafts.

TX-8 Net $.60

A non-flexible rigid coupling with steatite insulation. 1" diam. Fits 1/4" shaft.

TX-10 Net $.40

A very compact insulated coupling free from backlash. Insulation is canvas bakelite. 1-1/16" diam. Fits 1/4" shaft.

TX-10F (Not illustrated) Net $.45

A new version of the TX-10 which employs thin canvas bakelite strips for flexibility.

TX-22 (not illustrated) Net $.40

A non-insulated coupling identical to TX-10 except of all metal construction. Makes good electrical connection between coupled shafts.

TX-9 Net $.75

This small insulated flexible coupling provides high electrical efficiency when used to isolate circuits. Insulation is steatite. 1 5/8" diam. Fits 1/4" shaft.

TX-21 (not illustrated) Net $.40

Similar to TX-10 except 13/16" long and couples 1/4" shaft to 5/32" shaft.

SAFETY GRID AND PLATE CAPS

SPP-9 Net $.21

Ceramic insulation. Fits 9/16" diameter.

SPP-3 Net $.21

Ceramic insulation. Fits 3/8" diameter.

National Safety Grid and Plate Caps have a ceramic body which offers protection against accidental contact with high voltage caps on tubes.

GRID AND PLATE GRIPS

Type 12, for 9/16" Caps Net $.06

Type 24, for 3/8" Caps Net $.03

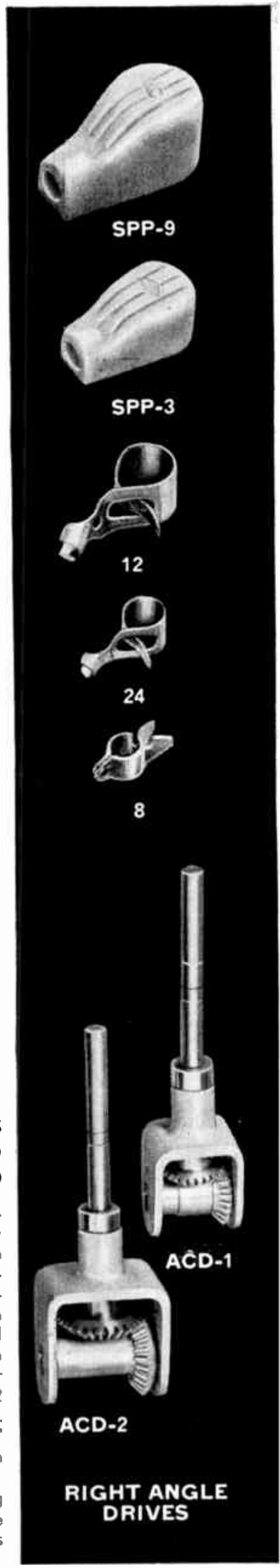
Type 8, for 1/4" Caps Net $.03

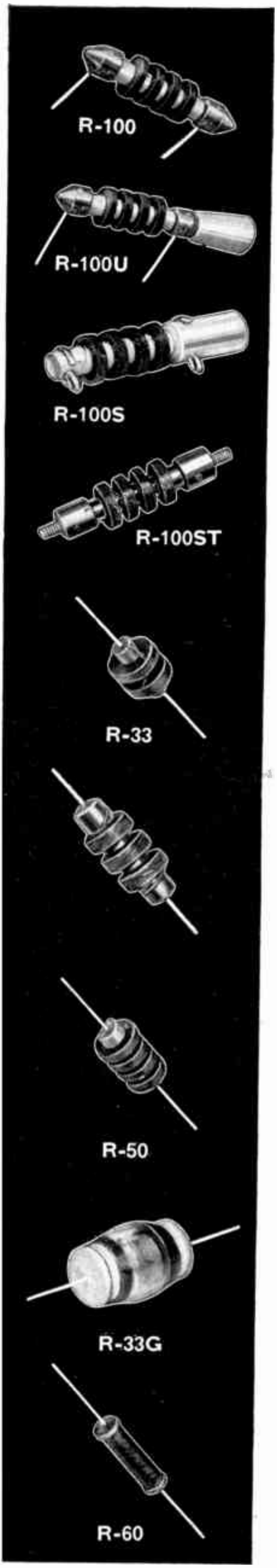
National Grid and Plate Grips provide a secure and positive contact with the tube cap and yet are released easily by a slight pressure on the ear.

RIGHT ANGLE DRIVES

ACD-1 ....Net $3.75
ACD-2 ....Net $3.90
ACD-3 ....Net $3.90

These sturdy drives were developed for use with the new National AMT condensers (see page 26). They are as compact as the torque requirements will allow and have nickel plated cast frames and bronze gears which operate smoothly without chatter or binding. The ACD-1 has 32 pitch gears and a 1/4" dia. dial shaft and drives 1/4" shafts. ACD-2 has 24 pitch gears (for heavier service) and 1/4" dia. shaft driving 1/4" shafts. ACD-3 is the same as ACD-2 except that it drives 3/8" diameter shafts.





R-100 ....Net $ .35
R-100U ....Net $ .42
R-100S ....Net $ .42
R-100ST ....Net $ .40

These RF chokes are identical electrically, but differ in mounting provisions. The R-100 employs pigtail leads; the R-100U has pigtail leads and a removable stand-off insulator; the R-100S has cotter-pin lug terminals and a non-removable stand-off insulator; the R-100ST has a 6-32 threaded stud at each end. These chokes are available in 2.5, 5 and 10 millihenry sizes and are rated at 125 milliamperes.

R-33 Net $ .35

The R-33 series chokes are 2-section RF chokes available in 10, 50, 100 and 750 microhenry sizes. Also available in this series is a single layer solenoid choke of 1 microhenry inductance. All are rated at 33 milliamperes. The chokes are wound on a 5/8" long form and range in diameter up to 5/16" maximum.

R-50 Net $ .35
R-50-1 Net $ .53

The R-50 series chokes are 3 and 4-section RF chokes and available in 0.5, 1, 2.5, and 10 millihenry sizes. They are rated at 50 milliamperes. The chokes are wound on a 1" long form and have a maximum diameter of 15/32". The 10 millihenry R—50-1 choke is wound on an iron core.

R-33G Net $3.60

The R-33G choke is a 2-section 750 microhenry RF choke hermetically sealed in glass with a current rating of 33 milliamperes. The choke body is 1" long by 5/8" diameter.

R-60 Net $ .35

The R-60 choke is a high current RF choke (500 milliamperes) available in 2 and 4 microhenry sizes. The choke is 1 1/8" long by 5/16" diameter.

R-300 ....Net $ .38
R-300U ....Net $ .42
R-300S ....Net $ .42
R-300ST ....Net $ .40

These RF chokes are similar in size to R-100 series but have higher current capacity. The R-300U is provided with a removable stand-off insulator at one end. The R-300S has a non-removable stand-off insulator and cotter-pin lug terminals. The R-300ST has a 6-32 threaded stud at each end. Inductance values of 0.5, 1.0, 2.5 and 5.0 millihenries are available with a current rating of 300 milliamperes. R-300, R-300U, R-300S and R-300ST are identical electrically.

R-152 Net $1.75

For use in the range between 2 and 4 Mc. Ideal for high power transmitter stages operated in the 80 meter amateur band. Inductance 4 m.h., DC resistance 10 ohms, DC current 600 ma. Coils honeycomb wound on steatite core.

R-154 Net $1.75
R-154U Net $1.40

For the 20, 40 and 80 meter bands. Inductance 1 m.h., DC resistance 6 ohms, DC current 600 ma. Coils honeycomb wound on steatite 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 Net $2.25

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 and 80 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.

Manufacturers: We have facilities for quantity production of RF chokes of practically any type. Send us your specifications.

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I. F. TRANSFORMERS

FC
FCO

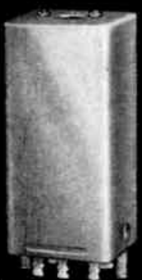
IFC, Transformer, **Net $4.25**
IFCO, Oscillator, **Net $4.25**
Litz coils wound on a polystyrene form and ceramic insulated air-dielectric trimming condensers make these transformers inherently stable and exceptionally retentive of tuning. The 4½" x 2⅜" x 2" shield can has two 6-32 spade bolts for mounting. Available for either 175 KC or 450-550 KC. Specify frequency.

IFL FM Discriminator **Net $6.90**
IFM IF Transformer **Net $6.45**
IFN IF Transformer **Net $6.45**
IFO FM Ratio Discriminator **Net $6.98**

IFL, IFM, IFN and IFO transformers operate at 10.7 Mc. and are designed for use in FM Superheterodyne receivers. Coils are precision wound on grooved polystyrene forms and tuning is accomplished by movable iron cores. Bandwidth is not affected by tuning slug position. The transformer cans are 1⅜" square and stand 3⅛" above the chassis. Two 6-32 spade bolts are provided for mounting.

The IFL transformer is a 10.7 Mc. FM discriminator transformer suitable for use in conventional FM receiver discriminator circuit and is linear over a band of ± 100 Kc.

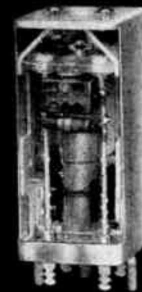
The IFM transformer is a 10.7 Mc. IF transformer with a 150 Kc. bandwidth at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFM Transformer and 6SG7 tube.



IFL
IFM
IFN
IFO

The IFN transformer is a 10.7 Mc. IF transformer with a 100 Kc. pass band at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFN Transformer and 6SG7 tube.

The IFO transformer is a 10.7 Mc. FM discriminator transformer of the ratio type and is linear over a band of ± 100 Kc.



IFJ
IFK

IFJ, with variable coupling **Net $8.25**

IFK, with fixed coupling **Net $7.25**

15 Mc. IF transformers suitable for ultra high frequency superheterodynes. They are made in two models with and without variable coupling. Approximate stage gain of 10 is obtained with IFJ or IFK Transformer and 6AB7 tube.

SA-4842

SA:4842 **Net $4.50**
A 456 kc. discriminator transformer for narrow band frequency modulation. This unit is the nucleus of the NFM adapter described by Harrington and Bartell in November 1947 QST. Two slug-tuned secondaries are employed and discrimination is accomplished by resonating one at approximately 10 kc. above, the other at approximately 10 kc. below the center frequency of the i.f. channel.

CD-1

CD-1, ¼ pint can **Net $.95**
Liquid Polystyrene Cement — is ideal for windings as it will not spoil the properties of the best coil form.

COILS AND COIL FORMS

OSR

XR-50
AR-5
AR-2

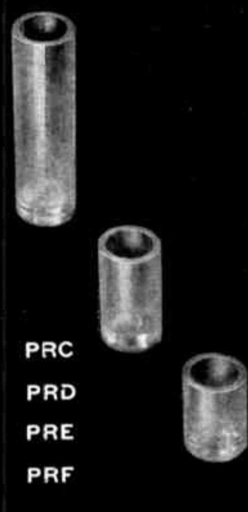
AR-2 High Frequency Coil **Net $1.13**
AR-5 High Frequency Coil **Net $.97**
The AR-2 and AR-5 coils are high Q permeability tuned RF coils on low loss mica-filled bakelite forms. The AR-2 coil tunes from 75 Mc. to 220 Mc. with capacities from 100 to 10 mmfd. The AR-5 coil tunes from 37 Mc. to 110 Mc. with capacities from 100 to 10 mmfd. The inductive windings supplied may be replaced by other windings as desired to modify the tuning range.

XR-50 **Net $.60**
These mica-filled bakelite coil forms may be wound as desired to provide a permeability tuned coil. The form winding length is 11/16" and the form winding diameter is ½ inch. The iron slug is ⅜" dia. by ½" long.

OSR **Net $1.80**
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.

Symbol	Outside Diameter	Length	Net
PRC-1	⅜"	⅜"	.15
PRC-2	⅜"	½"	.15
PRC-3	⅜"	¾"	.15
PRD-1	½"	½"	.15
PRD-2	½"	1"	.15
PRE-1	9/16"	¾"	.18
PRE-2	9/16"	1"	.18
PRE-3	9/16"	2"	.24
PRF-1	¾"	¾"	.24
PRF-2	¾"	1¼"	.30

These small coil forms are of molded polystyrene, open at one end and closed at the other except for a hole which permits mounting by a single 6-32 screw. A size for every application.



PRC
PRD
PRE
PRF

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Coil Forms molded of R-39 mica-filled bakelite permitting them to be grooved and drilled. Coil Form diameter 1", length 1½".
XR-1 Four Prong, **Net $.35**
XR-2, without Prongs
Net $.25

XR-3, molded of R-39 Diameter 9/16", length ¾" without prongs. **Net $.20**

XR-4, Four Prong, **Net $.51**
XR-5, Five Prong, **Net $.51**
XR-6, Six Prong, **Net $.60**
Molded of R-39 permitting them to be grooved and drilled. Coil Form Diameter 1½", length 2¼". A special socket is required for the XR-6.
National type **XC-6C**
Net $.51

SC, Crystal Sockets
Net $.32
The SC-1, SC-2, and SC-3 are crystal mounting sockets for crystal holders with mounting pins spaced 0.5000", 0.486", and .750" respectively and pin diameters of ⅛" and 3/32" and ⅛" respectively, steatite insulation. Single 4-36 or 4-40 screw mounting for SC-1 and SC-2; single 6-32 screw mounting for SC-3.

CFA **Net $.35**
The National chart frame is supplied with a celluloid sheet to cover the chart size 2¼" x 3¼" with sides ¼" wide. Durable finish.

PH-1 An attractive and rugged pull handle of cast zinc alloy chrome plated, with 10-32 Tapped Holes on 3¾" mounting centers.
Net $.45

PH-2 same as PH-1 but with black or gray finish.
Net $.25
The plug in base and shield includes the low loss R-39 base which is ideal for mounting condensers and coils when it is desirable to have them shielded and easily removable. Shield is 2" x 2⅜" x 4½".
5 Prong base and shield
PB-10-5 **Net $.75**
6 Prong base and shield
PB-10-6 **Net $.75**
5 Prong base only
PB-10-A-5 **Net $.51**
6 Prong base only
PB-10-A-6 **Net $.51**

RZ Coil Shield **Net $.35**
1⅜" square x 4" high.
RS Coil Shield **Net $.35**
1-7/16" x 1⅞" x 3½" high.

RO Coil Shield **Net $.35**
2" x 2⅜" x 4⅛" high. National Coil Shields are formed from a single piece of pure aluminum. They are mechanically strong and have ample thickness to mount small parts on the walls, and include spade belts, for chassis mounting.

T-78 Tube Shield - **Net $.27**
National Tube Shield type T-78 is a three-piece pure aluminum shield suitable for shielding glass tubes with ST-12 bulb, such as the 6C6 and 6D6 tubes.

JS-1 Jack Shield **Net $.30**
For shielding small standard jacks mounted behind a panel, or on the ends of extension coils. Indispensable for reducing hum pickup.

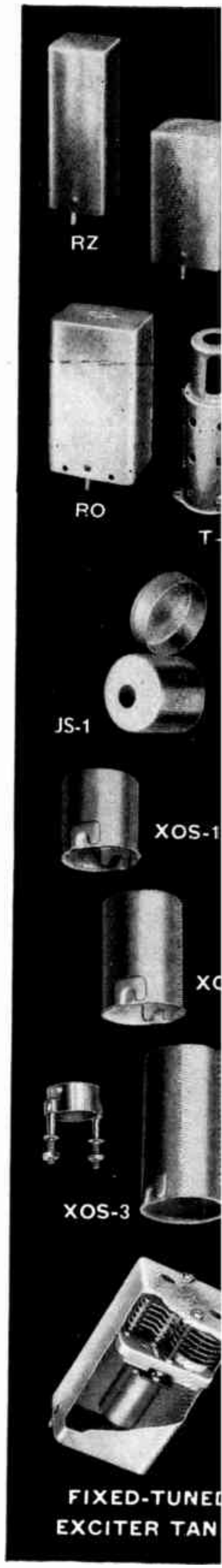
XOS Tube Shields **Net $.48**
The XOS tube shield is a two-piece shield for the miniature Button 7 pin base tubes. The shield is available in three sizes corresponding to the tube body heights XOS-1 for 1-5/16", XOS-2 for 1½", XOS-3 for 2".

The shield contains a spring which centers tube in shield and holds tube and shield firmly in place. The two 4-40 spade bolts serve to mount the XOA or XOR Socket and the XOS Tube Shield.

FXT 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)
Net $3.45
FXTB-5 (with 5 prong base)
Net $3.90
FXTB-6 (with 6 prong base)
Net $3.90

Paint (not illustrated)

CP-1, dark gray **Net $.40**
CP-2, black **Net $.40**
A high quality air-drying paint that may be applied with a brush.
CP-3, light gray, matches newest National receivers—for spraying and baking.
Net $.50



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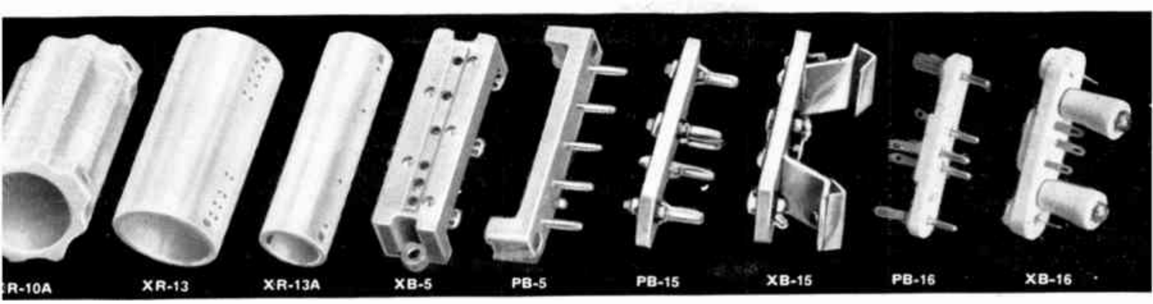


NSMITTER COIL FORMS

ransmitter Coil Forms and Mounting are designed as a group, and · conveniently on the bars of a TMA condenser. The larger coil form, XR-14A, (not illustrated) has a winding diameter of 5", a winding of 3¾" (30 turns total) and is intended for the 80 meter band. maller form, Type XR-10A, has a winding length of 3¾" and a ıg diameter of 2½" (26 turns total). It is intended for the 20 and ɜter bands.

er coil form fits the PB-15 plug. For higher frequencies, the plug may ed with a self-supporting coil of copper tubing. The XB-15 Socket be mounted on breadboards or chassis, as well as on the TMA ɜnser.

SINGLE UNITS

XR-10A, Coil Form only	Net $ .99
XR-14A, Coil Form only	Net $2.4C
PB-15, Plug only	Net $1.05
XB-15, Socket only	Net $1.2C

ASSEMBLIES

UR-10A, Assembly (including small Coil Form, Plug and Socket)	Net $3.24
UR-14A, Assembly (including large Coil Form, Plug and Socket)	Net $3.6C

FER COIL FORMS

nal Buffer Coil Forms are designed to mount directly on the tie bars TMC condenser using the PB-5 Plug and XB-5 Socket. Plug and t are of molded R-39.

two coil forms are of steatite, left unglazed to provide a tooth for lope. The larger form, Type XR-13, is 1¾" in diameter and has a ng length of 2¾". The smaller form, Type XR-13A, is 1" in diameter ›rovides a winding length of 2¾". Both forms have holes for mount- nd for leads.

SINGLE UNITS

XR-13, Coil Form only	Net $.7!
XR-13A, Coil Form only	Net $.6(
PB-5, Plug only	Net $.5
XB-5, Socket only	Net $.5

ASSEMBLIES

UR-13A, Assembly (including small Coil Form, Plug and Socket)	Net $1.6!
UR-13, Assembly (including large Coil Form, Plug and Socket)	Net $1.6!

EXCITER COILS

is a National exciter coil for every application. AR-15 coils are mounted on 5 pin bases which fit any standard ct tube socket. AR-16 coils are mounted on the well known National PB-16 plug which fits the National XB-16 socket. Th 7 coils have 6 pin bases which fit standard 6 contact tube sockets and the link windings of this series have center tap may be grounded for harmonic reduction. All center link models are center tapped for use in balanced circuits. Insula polystyrene and steatite. For use where plate power input does not exceed 50 watts. Available with fixed or swingin ›r center links for all amateur bands, 6 through 80 meters.

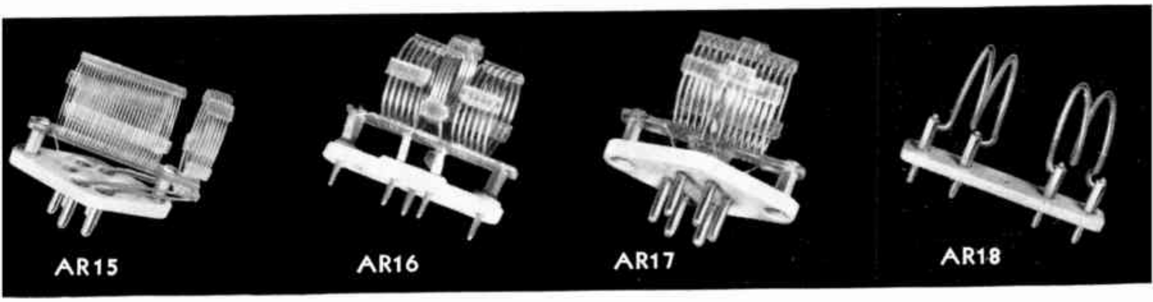
‹R-16 Coil Form (not illustrated) fits the PB-16 Plug-in Base; it has a winding length of 1¾", diameter 1¼".

AR-15, AR-16, AR-17 Coil, any type	Net $1.25
XR-16 Coil Form	Net $ .42
PB-16 Plug-in Base	Net $ .45
XB-16 Socket for PB-16	Net $ .45

500 WATT COILS

›ound coils designed to mount on the split stator models of National AMT condensers. The AR18-C coils have fixed ce nks and require the XB18-C socket. The AR18-S coils are designed to accommodate the swinging link furnished with th -S socket. Link winding of the XB18-S has a center tap which may be grounded for harmonic reduction. Plugs and jac silver plated to insure low contact resistance. Insulation, steatite. The sockets (not illustrated) are 7¼" in length. AR- are available for all amateur bands, 6 through 80 meters.

your National distributor for prices)



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TYPE TMS TRANSMITTING CONDENSERS

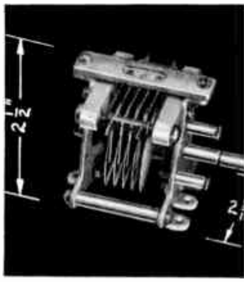
This is a condenser designed for transmitter use in low power stages. It is compact, rigid, and dependable. Provision been made for mounting either on the panel, on the chassis, or on two stand-off insulators. Insulation is steatite. Voltage ings listed are conservative.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
100 Mmf.	9.5	3"	.026"	1000v.	9	TMS-100	$2.6
150	11	3"	.026"	1000v.	14	TMS-150	2.8
250	13.5	3"	.026"	1000v.	22	TMS-250	3.3
300	15	3"	.026"	1000v.	27	TMS-300	3.8
35	8	3"	.065"	2000v.	7	TMSA-35	3.9
50	11	3"	.065"	2000v.	11	TMSA-50	4.4
DOUBLE STATOR MODELS							
50–50 Mmf.	6–6	3"	.026"	1000v.	5–5	TMS-50D	$3.0
100–100	7–7	3"	.026"	1000v.	9 9	TMS-100D	3.2
50–50	10.5–10.5	3"	.065"	2000v.	11–11	TMSA-50D	4.4

TYPE TMK TRANSMITTING CONDENSERS

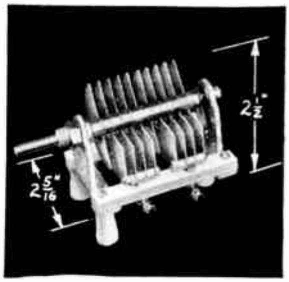
This is a new condenser for exciters and low power transmitters. Special provision has been made for mounting AR-16 c in a swivel plug-in mount on either the top or rear of the condenser. For stand-off or panel mounting-steatite insulation.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
35 Mmf.	7.5	2 7/32"	.047"	1500v.	7	TMK-35	$3.45
50	8	2 3/8"	.047"	1500v.	9	TMK-50	3.55
75	9	2 11/16"	.047"	1500v.	13	TMK-75	3.80
100	10	3"	.047"	1500v.	17	TMK-100	3.95
150	10.5	3 5/8"	.047"	1500v.	25	TMK-150	4.65
200	11	4 1/4"	.047"	1500v.	33	TMK-200	5.25
250	11.5	4 7/8"	.047"	1500v.	41	TMK-250	5.75
DOUBLE STATOR MODELS							
35–35 Mmf.	7.5–7.5	3"	.047"	1500v.	7–7	TMK-35D	$3.80
50–50	8–8	3 5/8"	.047"	1500v.	9–9	TMK-50D	3.95
100–100	10–10	4 1/4"	.047"	1500v.	17–17	TMK-100D	5.25
Swivel Mounting Hardware for AR 16 Coils						SMH	$ .10



TYPE TMH TRANSMITTING CONDENSERS

A condenser that features very compact construction. Excellent power factor, and aluminum plates .0400" thick polished edges. It mounts on the panel or on removable stand-off insulators. Steatite insulators have long leakage p Stand-offs included in listed price.

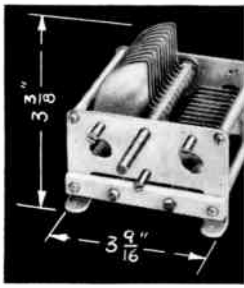


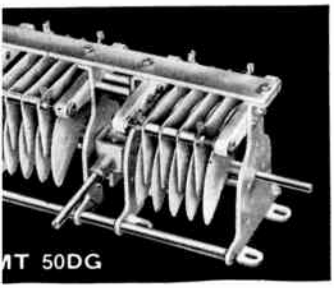
Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	9	3 3/4"	.085"	3500v.	15	TMH-50	$3.9
75	11	3 3/4"	.085"	3500v.	19	TMH-75	4.1
100	12.5	5 1/8"	.085"	3500v.	25	TMH-100	4.3
150	18	6 1/2"	.085"	3500v.	37	TMH-150	4.9
35	11	5 1/8"	.180"	6500v.	17	TMH-35A	4.2
DOUBLE STATOR MODELS							
35–35 Mmf.	6–6	3 3/4"	.085"	3500v.	9 9	TMH-35D	$4.1
50–50	8–8	5 1/8"	.085"	3500v.	13–13	TMH-50D	4.3
75–75	11–11	6 1/2"	.085"	3500v.	19–19	TMH-75D	4.9

TYPE TMC TRANSMITTING CONDENSERS

A condenser designed for use in the power stages of transmitters where peak voltages do not exceed 3000 volts. The fram extremely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are aluminum with bu edges. Insulation is steatite. 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	Net
SINGLE STATOR MODELS							
50 Mmf.	10	3"	.077"	3000v.	7	TMC-50	$3.60
100	13	3 1/2"	.077"	3000v.	13	TMC-100	4.25
150	17	4 5/8"	.077"	3000v.	21	TMC-150	5.25
250	23	6"	.077"	3000v.	32	TMC-250	5.70
300	25	6 3/4"	.077"	3000v.	39	TMC-300	6.10
DOUBLE STATOR MODELS							
50–50 Mmf.	9–9	4 5/8"	.077"	3000v.	7–7	TMC-50D	$4.35
100–100	11–11	6 3/4"	.077"	3000v.	13–13	TMC-100D	5.95
200–200	18.5–18.5	9 1/4"	.077"	3000v.	25–25	TMC-200D	7.25



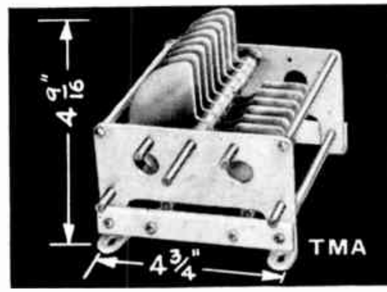


TYPE AMT

A larger and sturdier model of the TMK condenser. The frame is extremely rigid, with mounting feet a part of the end plates. Heavy steatite insulation.

The solid aluminum tie bar across the top of the condenser acts as a mounting for AR-18 series coils in the double stator models.

The double stator models are available in either standard end drive (D series) or center-drive (DG series) with 1/4" dia. shaft extension.



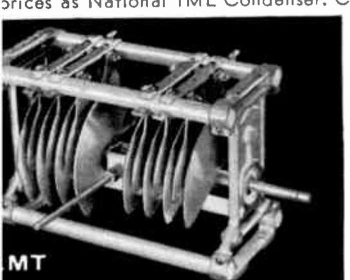
TYPE TMA

. a larger model of the popular TMC. The frame is extremely rigid and arranged for mounting on panel, chassis or stand-
;ulators. The plates are of heavy aluminum with rounded and buffed edges. Insulation is steatite located outside of the
ntrated field.

Maximum Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	13	4 3/4"	.177"	6000 v.	9	AMT-50	$ 5.20
100	20	6 3/4"	.177"	6000 v.	17	AMT-100	6.10
300	19.5	4 9/16"	.077"	3000 v.	23	TMA-300	7.60
50	15	4 9/16"	.171"	6000 v.	7	TMA-50A	4.95
100	19.5	6 7/8"	.171"	6000 v.	15	TMA-100A	5.85
150	22.5	6 7/8"	.171"	6000 v.	21	TMA-150A	6.45
230	33	9 5/16"	.171"	6000 v.	33	TMA-230A	7.95
100	30	9 1/4"	.265"	9000 v.	23	TMA-100B	8.50
150	40.5	12 1/2"	.265"	9000 v.	33	TMA-150B	9.95
50	21	7 1/8"	.359"	12,000 v.	13	TMA-50C	5.55
100	37.5	12 7/8"	.359"	12,000 v.	25	TMA-100C	8.95
75	25	18 1/16"	.719"	20,000 v.	17	TML-75E	18.35
150	60	18 1/16"	.469"	15,000 v.	27	TML-150D	18.50
100	45	13 5/8"	.469"	15,000 v.	19	TML-100D	16.60
50	22	8 5/16"	.469"	15,000 v.	9	TML-50D	11.50
245	54	18 1/16"	.344"	10,000 v.	35	TML-245B	20.15
150	45	13 5/8"	.344"	10,000 v.	21	TML-150B	18.35
100	32	10 15/16"	.344"	10,000 v.	15	TML-100B	17.55
75	23.5	8 5/16"	.344"	10,000 v.	11	TML-75B	12.80
500	55	18 1/16"	.219"	7,500 v.	49	TML-500A	24.60
350	45	13 5/8"	.219"	7,500 v.	33	TML-350A	19.65
250	35	10 15/16"	.219"	7,500 v.	25	TML-250A	18.35
DOUBLE STATOR MODELS			D—End drive		DG—Center drive		
50–50	13–13	9 3/8"	.177"	6000 v.	18	AMT-50D	7.00
100–100	20–20	13 3/8"	.177"	6000 v.	34	AMT-100D	9.00
50–50	13–13	9 3/8"	.177"	6000 v.	18	AMT-50DG	10.75
100–100	20–20	13 3/8"	.177"	6000 v.	34	AMT-100DG	12.75
200–200	15–15	6 7/8"	.077"	3000 v.	16–16	TMA-200D	9.40
180–180	10–10	12 3/4"	.140"	4000 v.	24–24	TMA-180D	12.90
50–50	12.5–12.5	6 7/8"	.155"	6000 v.	8–8	TMA-50DA	6.75
100–100	17–17	9 5/16"	.155"	6000 v.	14–14	TMA-100DA	8.75
60–60	19.5–19.5	12 1/2"	.249"	9000 v.	15–15	TMA-60DB	8.95
40–40	18–18	12 7/8"	.343"	12,000 v.	11–11	TMA-40DC	8.50
30–30	12–12	18 1/16"	.719"	20,000 v.	7–7	TML-30DE	18.55
60–60	26–26	18 1/16"	.469"	15,000 v.	11–11	TML-60DD	20.15
100–100	27–27	18 1/16"	.344"	10,000 v.	15–15	TML-100DB	12.35
60–60	20–20	13 5/8"	.344"	10,000 v.	9–9	TML-60DB	19.15
200–200	30–30	18 1/16"	.219"	7,500 v.	21–21	TML-200DA	24.60
100–100	17–17	10 15/16"	.219"	7,500 v.	11–11	TML-100DA	20.15

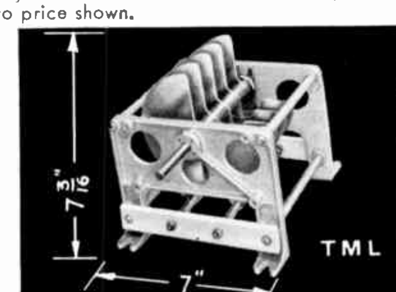
TYPE LMT

ɪavy duty transmitting condenser that completely eliminates troublesome closed loops, vastly simplifying the problem
ıwanted harmonics. The rotor shaft is completely insulated from the end plates. Long leakage path (higher safety factor).
s and parts are extra heavy with highly polished rounded edges to prevent flash-over. Adjustable stator plate mounting
end bearings. Available in single-stator, double-stator, or double-stator right angle center drive models. Same capacities
prices as National TML Condenser. Condensers with right angle drive add $3.90 to price shown.



TYPE TML

is a heavy duty job throughout. The frame structure (rugged aluminum castings with dural tie bars) and precision bearings assure permanent rotor alignment. All plates are extra thick with rounded and polished edges. This, plus specially treated steatite insulators and a husky self-cleaning rotor contact, provides high flashover, current and voltage ratings.



National
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PSR

M30

W100

NC-600U

STN

MINIATURE CONDENSERS:

Type PS variable condensers are compact silver plated units of soldered construction for use as semi-fixed bandsets or padders. Base is steatite — bearing is "snug" but smooth. PSR models are screwdriver adjust type; PSE have ¼" diameter shafts both ends; PSL are similar to PSR but include rotor shaft lock.

Type M-30 **Net $.22**
The M-30 is a tiny (13/16" x 9/16" x ½") mica trimmer — 30 mmf. max. — steatite base.

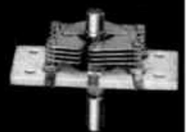
Type W-75, 75 mmf. **Net $1.60**
Type W-100, 100 mmf. **Net $1.76**
Small air-dielectric padding condensers having a very low temperature coefficient. They are mounted in 1¼" diameter aluminum shields and have ¼" hex heads for socket-wrench adjustment.

The UM condensers are low-loss, aluminum plate staked construction miniature variables designed for UHF converters, VFOs and the like — minimum capacity is exceptionally low. The UMs can be mounted in PB-10 or RO shield cans and have ¼" dia. shafts front and rear for ganging (see pages 21, 23 and 24 for shield cans and couplings). Plates: straight-line-cap., 180° rotation. Dimensions: Base 1" x 2¼", mtg. holes on ⅝" x 1-23/32" centers, 2-5/16" max. length.

The UMB-25 and UMB-50 are differential (balanced stator) models. UM-10D and UMA-25 are double-spaced and the latter is bolted construction for experimental capacity reduction. Hardware for panel or chassis mounting is supplied with all UM condensers.

UM

UMA-25



UMB-25

Capacity	Catalog Symbol			Net
25 mmf.	PSR-25	PSE-25	PSL-25	$1.70
50	PSR-50	PSE-50	PSL-50	1.85
75	PSR-75	PSE-75	PSL-75	2.00
100	PSR-100	PSE-100	PSL-100	2.15

Capacity	Minimum Capacity	No. of Plates	Air Gap	Catalog Symbol	Net
15 mmf.	1.5	6	.017"	UM-15	$1.02
35	2.5	12	.017"	UM-35	1.15
50	3	16	.017"	UM-50	1.25
75	3.5	22	.017"	UM-75	1.45
100	4.5	28	.017"	UM-100	1.60
10	1	8	.042"	UM-10D	1.40
25	3.4	14	.042"	UMA-25	1.75
BALANCED STATOR MODEL					
25	2	4-4-4	.017"	UMB-25	$2.40
50	5	8-8-8	.017"	UMB-50	2.70

NEUTRALIZING CONDENSERS:

NC-600U **Net $.38**
With standoff insulator
NC-600 **Net $.32**
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 **Net $2.07**
The Type STN has a maximum capacity of 18 mmf. (3000 V), making it suitable for such tubes as the 809. It is supplied with two standoff insulators.

NC-800A **Net $3.00**
The NC-800A disk-type neutralizing condenser is suitable for the T40, 35TG, 808 and similar tubes. It is equipped with a clamp for locking. The chart below gives capacity and air gap for different settings.

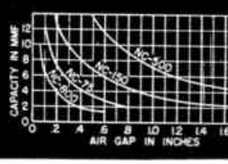
NC-75 **Net $3.60**
For 812, 75TH and similar tubes.

NC-150 **Net $5.25**
For RK36, 100TH, HK354, 250TH, etc.

NC-500 **Net $8.75**
For WE-251, 304TH, 833A and the like. These large disk-type neutralizing condensers are for the higher powered tubes. Disks are aluminum, insulation steatite.

NC-800A

NC-75
NC-150
NC-500



National
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'RECISION CONDENSERS

ginally developed for the famous
) and NC-100 receivers, National
and NPW condensers and drive
s are well known to professional
amateur radio men throughout the
ld. Sturdily constructed of the finest
erials and carefully adjusted by
ed hands, they have become "stand-
specifications" for applications re-
·ing smooth, precise control and high
et accuracy.

Micrometer Dial reads direct to
part in 500. Division lines are
roximately 1/4" apart. The drive, at
mid-point of the rotor, is through
enclosed preloaded worm gear with
to 1 ratio. Each rotor is individually
ılated from the frame, and each
its own individual rotor contact.
tor insulation is steatite. Plate shape
straight-line frequency when the fre-
ıncy range is 2:1.

' Condensers are available in 1, 2,
or 4 sections, in either 160 or 225
ıf per section. Larger capacities can-
be supplied.

'-1R	Single section right	Net **$13.50**
'-1L	Single section left	Net **$13.50**
'-2R	Double section right	Net **$18.00**
'-2L	Double section left	Net **$18.00**
'-2S	Single section each side	Net **$18.00**
'-3R	Double section right; single left	Net **$24.00**
'-3L	Double section left; single right	Net **$24.00**
'-4	Double section each side	Net **$27.00**
W-3	Three sections, each 225 mmf.	Net **$24.00**

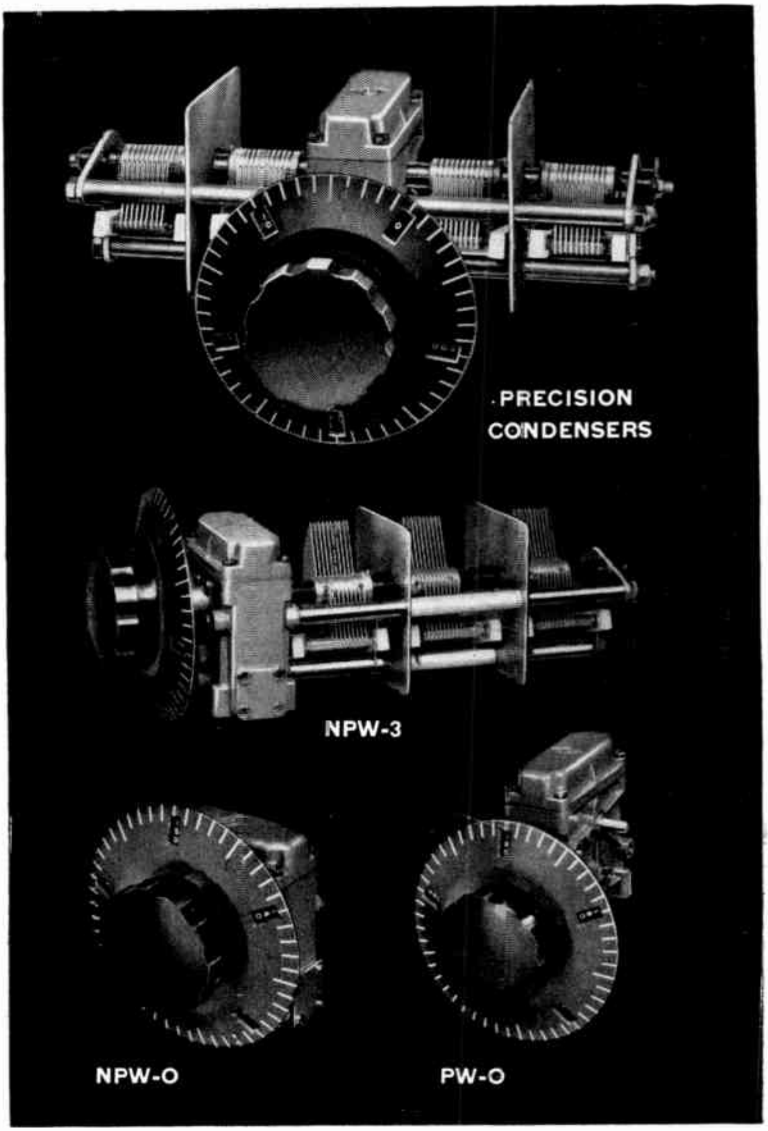
nilar to PW models, except that rotor
ıft is perpendicular to panel.

'W-O Net **$9.00**

es parts similar to the NPW condens-
Drive shaft perpendicular to panel.
ıe TX-9 coupling supplied.

'-O Net **$9.90**

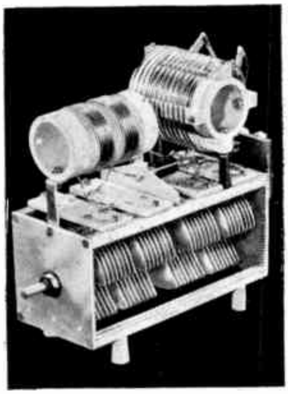
es parts similar to the PW condenser.
ive shaft parallel to panel. Two TX-9
uplings supplied.



'-D Net **$5.25**

e Micrometer Dial used on the con-
nsers and drives above is available
separately. It revolves ten times in
covering the complete range and as
there is no gear reduction unit fur-
nished, the driven shaft will revolve
ten times, also. The PW-D dial fits a
shaft 5/16" in diameter.

MULTI-BAND TANK ASSEMBLY



e unique MB-150 Multi-Band Tank tunes all amateur bands from 80 through 10 meters
th 180° rotation of the shaft; the coils are never changed. The unit is built around a
cuit which tunes to two harmonically unrelated frequencies at the same time. Thus, it
comes possible to cover a wide frequency range and yet maintain a reasonably constant
C ratio. 3" wide x 8 1/4" high (including the GS-10 standoffs) x 9" long overall including
e 1/4" dia. shaft and output terminals.

atures of the MB-150:

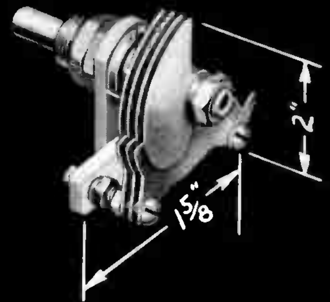
) For use as the all-band plate tank in push-pull or single-ended stages running up to 150-watts input (1500 volts peak). It is ideal for a pair of 807s or 809s or a single 829B.

) Separate link coupling coil has special clips which adjust to match impedances up to 600 ohms directly. Output couples into a higher powered amplifier, an antenna or an antenna tuning network.

) Fast band changing is accomplished without handling coils, thus removing one of the danger points in the amateur station.

MB-150 Multi-Band Tank Assembly Net **$18.75**

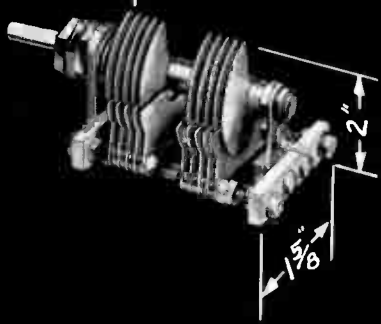
TYPE STHS
STRAIGHT-LINE WAVELENGTH
180° Rotation



The ST Type condenser has Straight-Line Wavelength plate
All double-bearing models have the front bearing insulated
prevent noise. On special order a shaft extension at each end
available, for ganging. On double-bearing single shaft mode
the rotor contact is through a constant impedance pigta
Steatite insulation.

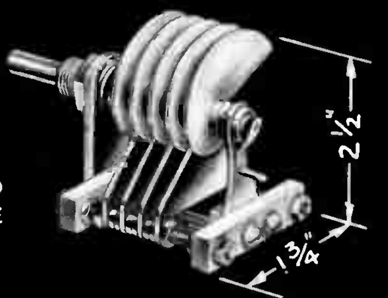
Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	Net
SINGLE BEARING MODELS						
15 Mmf.	3 Mmf.	3	.018"	1 3/16"	STHS- 15	$1.6
25	3.25	4	.018"	1 3/16"	STHS- 25	1.9
50	3.5	7	.018"	1 3/16"	STHS- 50	2.1
SPLIT STATOR DOUBLE BEARING MODELS						
50-50	5-5	11-11	.026"	2 3/4"	STD- 50	$3.60
100-100	5.5-5.5	14-14	.018"	2 3/4"	STHD-100	3.90
DOUBLE BEARING MODELS						
35 Mmf.	6 Mmf.	8	.026"	2 1/4"	ST- 35	$1.85
50	7	11	.026"	2 1/4"	ST- 50	1.90
75	8	15	.026"	2 1/4"	ST- 75	2.00
100	9	20	.026"	2 1/4"	ST-100	2.10
140	10	27	.026"	2 3/4"	ST-140	2.30
150	10.5	29	.026"	2 3/4"	ST-150	2.30
200	12.0	27	.018"	2 1/4"	STH-200	2.50
250	13.5	32	.018"	2 3/4"	STH-250	2.70
300	15.0	39	.018"	2 3/4"	STH-300	2.90
335	17.0	43	.018"	2 3/4"	STH-335	3.10

TYPE ST
(Type STD Illustrated)
STRAIGHT-LINE WAVELENGTH
180° Rotation



NOTE — Type SS Condensers, having straight-line c
pacity plates but otherwise similar to the Type ST, are ava
able. Capacities and Prices same as Type ST.

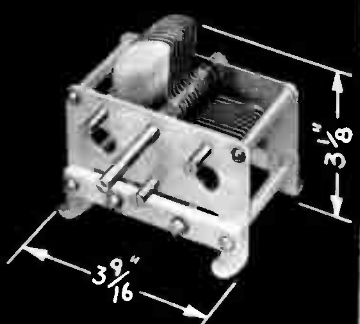
TYPE SE
(Type SEU Illustrated)
STRAIGHT-LINE FREQUENCY
270° Rotation



TYPE SE — All models have two rotor bearings, the front bear
ing being insulated to prevent noise. A shaft extension at eac
end, for ganging, is available on special order. On models wit
single shaft extension, the rotor contact is through a constar
impedance pigtail. The SEU models (illustrated) are suitable f
high voltages as their plates are thick polished aluminum wit
rounded edges. Other SE condensers do not have polished edge
on the plates. Steatite insulation.

15 Mmf.	7 Mmf.	6	.055"	2 1/4"	SEU- 15	$2.80
20	7.5	7	.055"	2 1/4"	SEU- 20	2.95
25	8	9	.055"	2 1/4"	SEU- 25	3.10
50	9	11	.026"	2 1/4"	SE- 50	2.30
75	10	15	.026"	2 1/4"	SE- 75	2.40
100	11.5	20	.026"	2 1/4"	SE-100	2.60
150	13	29	.026"	2 3/4"	SE-150	2.75
200	12	27	.018"	2 1/4"	SEH-200	2.80
250	14	32	.018"	2 3/4"	SEH-250	3.00
300	16	39	.018"	2 3/4"	SEH-300	3.25
335	17	43	.018"	2 3/4"	SEH-335	3.50

TYPE EMC
STRAIGHT-LINE WAVELENGTH
180° Rotation



TYPE EMC — A general purpose condenser available in larg
sizes and having Straight-Line wavelength plates. They ar
similar in construction to the TMC Transmitting condenser
and have high efficiency and rugged frames. Insulation i
Steatite, and Peak Voltage Rating is 1000 volts. Same sizes avai
able with straight line capacity plates, type DXC condenser.

Capacity	Minimum Capacity	No. of Plates	Length	Catalog Symbol	Net
150 Mmf.	9 Mmf.	9	2 15/16"	EMC-150	$4.50
250	11	15	2 15/16"	EMC-250	4.75
350	12	20	2 15/16"	EMC-350	6.00
500	16	29	4 3/8"	EMC-500	6.75
1000	22	56	6 3/4"	EMC-1000	10.35

National
© EST. 1914

McElroy Manufacturing Corporation

Littleton, Massachusetts, U. S. A.

I'd like to use lots of pages in this 1949 Handbook.
Say lots of things about our wireless telegraph equipment.
Use lots of photos to show the excellence of design and construction.
All that would take time and painstaking work we can't spare.

All I can do, then, is hope that communication men everywhere will take a minute or two to write for our complete catalog. Our new Wheatstone code tape perforator is the best I've ever used and I'm in this business thirty years. Our new transmitter, the ADK, performs faultlessly at speeds ranging from 5 words per minute to 500 words per minute.

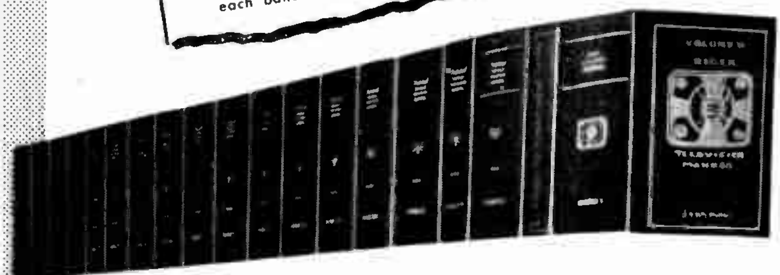
One of the most gratifying pieces of writing I've ever read is the Communications Report on the Ronne Antarctic Research Expedition written by Larry Kelsey, the Expedition Radio Operator. He tells about the dependability of our transmitter under especially severe conditions during the years 1946–48. We'd be glad to make extra copies and send them to commercial or Government communications men anywhere in the world. I'd like to tell you about our new high speed ink recorder which produces beautiful inked signals at speeds way over one thousand words per minute.

But, as I said in the beginning, I guess we'll have to admit we're a group of wireless operators and designers and builders of communications equipment. We'll let the job of telling be done by the advertising men who have prepared our new catalog — and use this page to express the hope that you readers will send for the catalog. Really, it is quite good!

Ted McElroy

RIDER MANUALS.. the world's most comprehensive compilation of servicing data —AM-FM-Television . . . and Allied Fields!

- **AUTHORIZED FACTORY-FACTS**
  Rider Servicing Techniques are compiled from data supplied by the manufacturers themselves.
- **GREATEST COVERAGE**
  Rider Servicing Data covers 27,000 models and 14,000 chassis in 25,830 pages.
- **CLARIFIED SCHEMATIC**
  Beginning with Volume XV . . . every multi-band receiver is broken down to show the circuit wiring of each band.
- **CIRCUIT THEORY EXPLAINED**
  Separate "How It Works" Books (with each Volume from XIII on) explain electronic and mechanical innovations in sets covered by the Manuals.
- **LOWER COST**
  Rider Manuals deliver data at less than a penny per receiver model, the greatest value anywhere!



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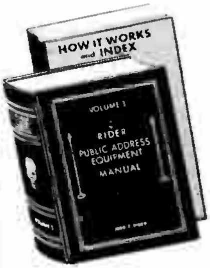

RIDER PA MANUAL

The first industry-wide public address equipment Manual, incorporating the amplifier production of 147 manufacturers from 1938 to 1948. 2024 Pages plus "How It Works" Book which explains theory of various PA circuits and method of rapidly locating faults. Index also included. **$18.00 complete.**

The RADIO AMATEUR'S BEAM POINTER GUIDE

. . . **by John F. Rider (W2RID)**
By merely consulting the charts of this book you can accurately point your beam to any country or island in the world containing amateur prefix listing. The book contains a country-prefix list, a cross index of prefixes and countries . . . also gives complete explanation of great circle maps, magnetic compass variations, etc.
32 pages. 8½" x 11"........ **$1.00**

Learn RIGHT with RIDER

- FM Transmission and Reception, by John F. Rider and Seymour D. Uslan, 416 pages, illustrated.............. $3.60
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Your Jobber Carries a Full Line of RIDER MANUALS

RIDER MANUALS mean SUCCESSFUL SERVICING

NOTE: The Mallory Radio Service Encyclopedia, 6th edition, makes reference to only one source of all radio receiver schematics — Rider Manuals. ANOTHER NOTE: The C-D Capacitor Manual for Radio Servicing, 1948 edition No. 4, makes reference to only one source of receiver schematics — Rider Manuals.

Any book which is not available at your Jobber may be ordered direct from

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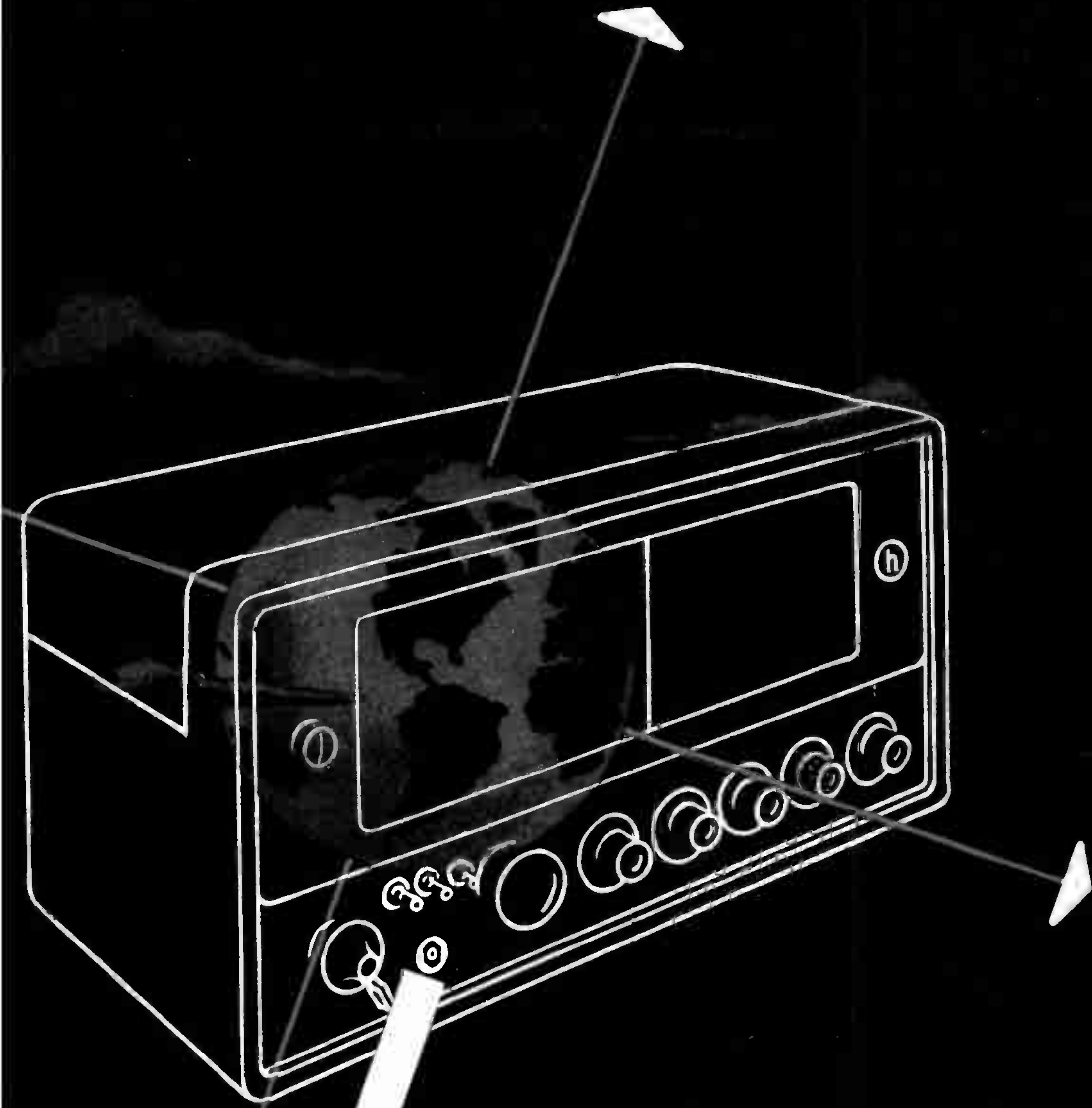


hallicrafters

the radio man's radio

Communications Equipment • Catalog No. 39 • 1949

SX-62 .... $269.50
Amateur Net

SX-42 .... $275.00
Amateur Net

"SWL" Version of Famous SX-42

newest addition to Hallicrafters line and just what the All-Wave listener has been
ting for. Will outperform any ordinary broadcast receiver on any frequency—
ndard Broadcast, Short-Wave or FM. Continuous coverage from 540 kc to 109 Mc.
Having basically the same chassis as our best communications receiver, the SX-62
vides communications-receiver performance in simplified form. A single tuning control
ers the wide-vision dial. Only one band lights up at a time—you always know just
ere you are tuning. In addition a 500 kc crystal calibration oscillator is built in, en-
ng you to adjust the dial pointer to show the exact frequency being tuned at any time.

FORMANCE: Continuous AM reception 540 kc to 109 Mc; FM reception 27—109
Temperature compensated, voltage regulated. Two RF, three IF stages; dual IF
nnels (455 kc and 10.7 Mc.). Audio flat 50—15,000 cycles; 10-watt push-pull output.

NTROLS: Band Selector 540—1620 kc, 1.62—4.9 Mc, 4.9—15 Mc, 15—32 Mc, 27—56 Mc,
–109 Mc: Receive/Standby, Calibration Osc. On/Off, Noise Limiter, Tuning, AF Gain,
no/FM/AM/CW, six-position Selectivity, four-position Tone, RF Gain, Calibration Reset.

YSICAL DATA: Gray steel cabinet with satin chrome trim. Top opens on piano hinge.
inet 20 in. wide by 10½ in. high by 16 in. deep. Ship. Wt. 70 lbs.

ERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs. Phone
Phonograph input jack. Socket for external power. Remote control connections. 105—
V. 50/60 cycle AC line.

TUBES PLUS VOLTAGE REGULATOR AND RECTIFIER: Two 6AG5 RF Amps., 7F8
v., 6SK7 IF Amp., 6SG7 IF Amp., 7H7 IF Amp., 7H7 FM Limiter and AM Det.,
FM Det., 7A4 BFO, 6H6 ANL, 6SL7 AF Amp., two 6V6 Push-pull Output, 6C4
libration Osc., VR-150 Regulator, 5U4G Rectifier.

42 Bass Reflex Speaker $34.50

tches either SX-62 or SX-42. Two-position
e switch. 500-ohm input. 8-in. heavy-duty
[type. Satin-finish gray metal cabinet 17
wide by 11¾ in. high by 12½ in. deep.
p. wt. 30 lbs.

s in Performance and Versatility

ferred by discriminating Amateurs and SWL's everywhere . . . our best communica-
ns receiver! Unsurpassed in versatility and coverage, outstanding in performance.
ntinuous coverage 540 kc to 110 Mc—Standard Broadcast, Short-Wave and FM.

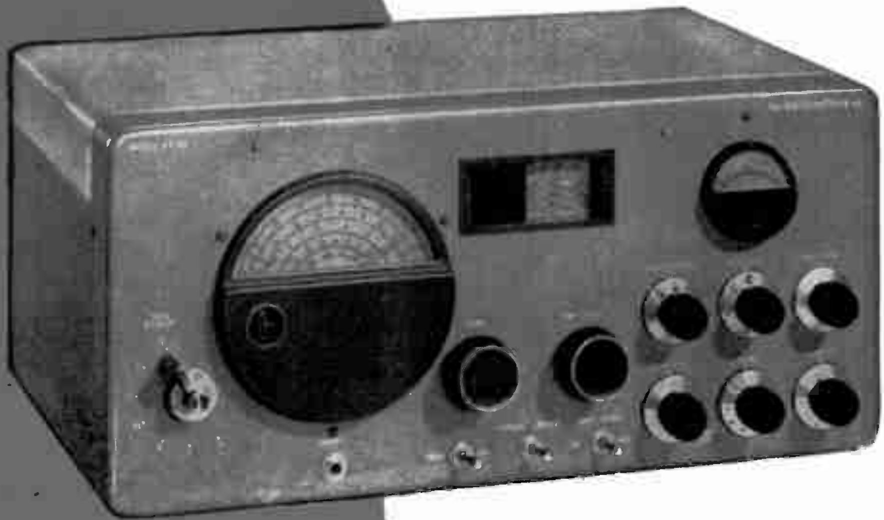
RFORMANCE: Continuous AM reception 540 kc to 110 Mc; FM reception 27—110 Mc.
mperature compensated oscillator with voltage regulator. Two RF, three IF stages, dual
channels (455 kc and 10.7 Mc). Audio flat 50—15,000 cycles, 10-watt push-pull output.

NTROLS: Band Switch 540—1620 kc, 1620—5000 kc, 5.0—15 Mc, 15—50 Mc, 27—55 Mc,
—110 Mc. Main tuning dial with logging scale on knob. Band-spread dial calibrated for 3.5,
14, and 28 Mc Bands plus logging scale. Two-position dial lock secures either main or
nd-spread knobs. AF Volume Control with power switch; AVC, Noise Limiter, Receive/
nd-by switches. Crystal Phasing, AM/FM/CW Phono, CW Pitch, six-position Selec-
ity, four-position Tone, RF Gain. "S" Meter adjustment on rear. Control settings for
oadcast and FM Bands marked in color for simplified use by others in family.

YSICAL DATA: Gray steel cabinet with satin chrome trim. Top opens on piano hinge.
binet 20 in. wide by 10½ in. high by 16 in. deep. Ship. wt. 71 lbs.

TERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs.
one jack. Phonograph input jack. Socket for external power. Remote control connec-
ns. 105—125 V. 50/60 cycle AC line.

TUBES PLUS VOLTAGE REGULATOR AND RECTIFIER: Two 6AG5 RF Amps., 7F8
nv., 6SK7 IF Amp., 6SG7 IF Amp., 7H7 IF Amp., 7H7 FM Limiter and AM Det., 6H6
Det., 7A4 BFO, 6H6 ANL, 6SL7 AF Amp., two 6V6 Push-pull Output, VR-150
gulator, 5U4G Rectifier.



W

SX-43
$189.50
Amateur Net

A

S-40A
$99.50
Amateur Net

est Coverage in Its Class

ere is all you would expect from a truly fine communications receiver plus extra coverage include the 6-Meter Band and the FM Broadcast Band. Offers coverage, versatility, and :rformance second only to our SX-42.

:RFORMANCE: AM reception 540 kc to 55 Mc; FM 44—55 Mc and 86—109 Mc. Temperare compensated oscillator. One RF and two IF stages (3rd IF stage above 44 Mc). Dual IF annels (455 kc and 10.7 Mc). Audio response to 10,000 cycles; 3-watt output.

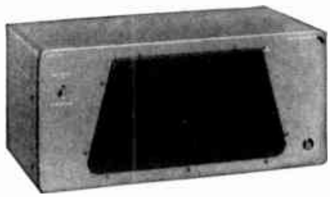
)NTROLS: Band Switch 540—1700 kc, 1700—5000 kc, 5—16 Mc, 14—14.4 Mc, 15.5—44 Mc, —55 Mc, 86—109 Mc. Main tuning in Mc; band-spread dial calibrated for 3.5, 7, 14, and Mc bands. Two-position Tone, Receive/Standby and Noise Limiter switches. Crystal ıasing, RF Gain, Phono/FM/AM-AVC/AM-MVC/CW, four-position Selectivity, AF Gain, ıd CW Pitch controls. Adjustment on rear for "S" meter.

IYSICAL DATA: Gray steel cabinet with satin chrome trim. Piano hinge top. Size 18½ in. ide by 8⅞ in. high by 12 in. deep. Ship. wt. 44 lbs.

(TERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs. ıono jack. Phonograph input jack. Socket for external power supply. Remote control ınnections. 105—125 V. 50/60 cycle AC line.

) TUBES PLUS RECTIFIER: 6BA6 RF Amp., 7F8 Conv., 6SG7 IF Amp., 6SH7 IF mp., 6SH7 IF Amp., 6H6 AM Det. and ANL, 6AL5 FM Det., 6J5 BFO, ;Q7 AF Amp., 6V6 Output, 5Y3GT Rectifier.

Matching speaker for SX-43. Two-position tone switch. 500-ohm input. Heavy-duty PM type, 6 by 9-inch oval size. Cabinet size 18½ in. wide by 8½ in. high by 9⅝ in. deep. Ship. Wt. 19 lbs.



R-44

$24.50

azing Sensitivity and Value

ffers superior performance in the medium price range, born of Hallicrafters long experience high-quality communications equipment. Complete in itself, with built-in PM speaker.

ERFORMANCE: AM reception 540 kc to 43 Mc. Temperature compensated oscillator. One F and two IF stages. Audio response to 10,000 cycles.

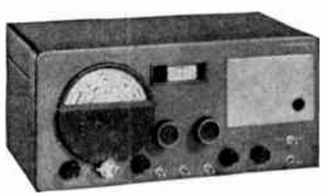
ONTROLS: Band Switch 540—1700 kc, 1700—5300 kc, 5.3—15.7 Mc, 15.7—43.0 Mc. Main tunıg in Mc; band-spread dial has arbitrary scale. AF and RF Gain controls; AVC, BFO, and Noise imiter switches; three-position Tone, BFO Pitch, and Receive/Standby controls. Settings ır Broadcast Band marked in color for simplified use by others in your family.

HYSICAL DATA: Satin black steel cabinet with brushed chrome trim. Top opens on iano hinge. Size 18½ in. wide by 9 in. high by 9½ in deep. Ship. wt. 32 lbs.

KTERNAL CONNECTIONS: Doublet or single wire antenna. Phone jack. Socket for cternal power supply. Remote control connections. 105—125 V. 50/60 cycle AC line.

TUBES PLUS RECTIFIER: 6SG7 RF Amp., 6SA7 Conv., two 6SK7 IF Amps., H6 ANL and AVC, 6J5GT BFO, 6SQ7, Det. and AF Amp., 6F6G utput, 5Y3GT Rectifier.

Same as the S-40A above only designed for AC or DC operation. 7 tubes plus rectifier and ballast tube: RF and IF tubes like S-40A; then 6H6 Det., 6SC7 BFO and AF Amp., 25L6GT Output, 25Z6GT Rectifier, and Ballast tube. Ship. Wt. 30 lbs.



S-52

$99.50

Offers Maximum

S-53 .... **$89.50**
Amateur Net

The Radio That

S-38 .... **$49.95**
Amateur Net

rformance in Compact Size

New 2 Mc IF Improves Image Rejection

A recent addition to the Hallicrafters line and a model that is rapidly gaining popularity because of its excellent performance and moderate price. Complete in itself, including built-in PM speaker.

PERFORMANCE: AM reception 540 kc to 31 Mc plus 48—54.5 Mc. Two stages IF with new 2 Mc IF,—high enough to avoid all possible images from amateur stations when operating within the amateur bands.

CONTROLS: Main tuning in Mc; separate band-spread dial with logging scale plus Mc calibration for 48—54.5 Mc band; Receive/Standby switch; Band switch 540—1630 kc, 2.5—6.3 Mc, 6.3—16 Mc, 14—31 Mc, and 48—54.5 Mc; AM/CW; RF Gain, Noise Limiter, AF Gain, two-position Tone; Speaker/Phones switch on rear.

PHYSICAL DATA: Satin black steel cabinet with brushed chrome trim. Top opens on piano hinge. Size 12⅞ in. wide by 7 in. high by 7½ in. deep. Ship. wt. 19 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. Phone tip jacks. Phonograph input jack. 105—125 V. 50/60 cycle AC line.

7 TUBES PLUS RECTIFIER: 6C4 Osc., 6BA6 Mixer, two 6BA6 IF Amps., 6H6 Det., AVC and ANL, 6SC7 BFO and AF Amp., 6K6GT Output, 5Y3GT Rectifier.

mazes Even the Experts

Exceptional Performance at a Moderate Price

The lowest priced communications receiver on the market . . . with many features found in much higher priced sets. Standard Broadcast plus three Short-Wave bands. Built-in PM speaker.

PERFORMANCE: Continuous AM reception 540 kc to 32 Mc. Maximum sensitivity and selectivity from expertly engineered chassis.

CONTROLS: Main Tuning in Mc; separate band-spread dial with arbitrary scale; Speaker/Phones, AM/CW, and Noise Limiter switches; Band Switch 540—1650 kc, 1.65—5 Mc, 5—14.5 Mc, 13.5—32 Mc; AF Gain, Receive/Standby.

PHYSICAL DATA: Steel cabinet in black wrinkle finish with brushed chrome trim. Size 12⅞ in. wide by 7 in. high by 7½ in. deep. Ship. wt. 14 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. Phone tip jacks. 105—125 V. DC or 50/60 cycle AC.

5 TUBES PLUS RECTIFIER: 12SA7 Conv., 12SK7 IF Amp., 12SQ7 BFO and Noise Limiter, 12SQ7 Det. and AVC, 35L6GT Output, 35Z5GT Rectifier.

The Newest and Most Versatile Transmitter Available

HT-19
$359.50
Amateur Net

Offers Narrow Band FM and CW, plus provisions for AM, to give maximum flexibilit on 5 bands. A completely self-contained, medium-power unit for the modern minde amateur. In addition, its compact size and smartly styled cabinet make it especiall desirable wherever appearance and space are to be considered.

Consists of an oscillator (crystal controlled or VFO), a frequency modulator wit speech amplifier, a buffer and a final amplifier. Extremely high stability and low FM distortion are obtained. The 4-65A in the final, cooled by a 3-inch 800-rpm fan, ha a plate input of 185 watts for approximately 125 watts output.

CONTROLS: Operation Switch has three crystal positions plus VFO and NBFM; tw pilot lamps show plate and filament power on/off; Band Selector switch changes multi plier coils 3.5—4, 7—7.3, 14—14.4, 21—21.45, and 27.16—29.7 Mc; final coils ar changed inside the unit, with dummy positions provided for four coils not in use. Chec switch turns on oscillator for spotting signals on receiver. Plate switch controls all "B" power and makes connections for remote control. Power switch is in 115-volt line Deviation Control adjusts for 0.4 ratio on all bands. Osc. Plate Tuning operates osc gang and calibrated dial. Power Amp. Tuning tunes final plate. Push-button mete switch throws milliammeter from final cathode to final grid.

PHYSICAL DATA: Gray steel cabinet with satin chrome trim. Piano-hinge top wit interlock. Size 20 by 10¼ by 16 in. deep. Ship. wt. 98 lbs.

EXTERNAL CONNECTIONS: Microphone connector; keying terminals (osc. keying) 50—600 ohm output (pi-section coupling); six terminals for remote control of eithe trans. or rec.; four terminals in final screen and plate circuits for applying audio fron external modulator for AM. Cord for 105—125 V. 50/60 cycle AC.

5 TUBES PLUS 2 VOLTAGE REGULATORS AND 3 RECTIFIERS: Three 6BA6—Osc., Freq Modulator, and Speech Amp., 6L6 Buffer, 4-65A Final, VR-150 and VR-105 Regulators 5Y3GT and two 866 Rectifiers.

alibator VFO CW/NBFM

Modernize your present transmitter with this famous Hallicrafters exciter. Crystal or VFO, NBFM or CW on 5 Bands with all coils, speech amplifier, and power supply built in. Features never before available in one low-priced unit. Low frequency drift, low FM distortion, low hum level, excellent keying. Output 2.5 to 4.5 watts. Chassis similar to HT-19 above, less final amplifier.

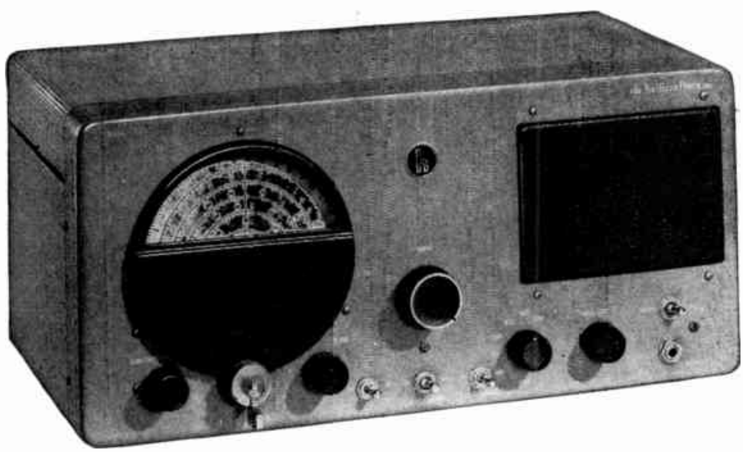
CONROLS: Operation Switch, Band Selector (ranges like HT-19), Check, Plate, Power, and Deviation switches. Single Tuning control.

PHYSICAL DATA AND CONNECTIONS: Satin gray cabinet, 12⅞ by 7 by 7¾ in. deep. Shipping weight 24 lbs. Microphone, keying, remote control connections. 72-ohm output terminals.

TUBES: Three 6BA6, 6L6, VR-150, VR-105, 5Y3GT.

HT-18
$110.00
Amateur Net

ependable, All-Purpose Marine Receiver



S-51
$149.50
Amateur Net

PERFORMANCE: Ruggedly constructed for sea or air use with special components to resist salt air, etc. Range 132 kc to 13 Mc. Three pre-tuned channels for fixed-frequency operation. 1020-cycle range filter for better voice reception on ranges. One RF, two IF stages. Temperature compensated.

CONTROLS: RF Gain; Band Selector 132—405 kc, 485—1530 kc, 1450—4550 kc, 4.2—13 Mc plus three fixed frequencies in 200—300 kc and 2—3 Mc ranges; AF Gain, CW/AM, Range Filter, ANL, Tuning, three-position Tone, CW Pitch, Rec./Send.

PHYSICAL DATA AND CONNECTIONS: Gray steel cabinet, 18½ by 9 by 9½ in deep. Ship. wt. 30 lbs. Piano hinge top. Doublet or single wire antenna. Phone jack. Socket for 6, 12, or 32-V. vibrator pack (available separately). 105—125 V. DC or 50/60 cycle AC.

9 TUBES PLUS RECTIFIER: 6SS7 RF Amp., 7A8 Conv., two 6SS7 IF Amp., 7C6 Det., 7A6 ANL, 6SS7 BFO, 35L6GT or 6V6GT Output, 35Z5GT Rect.

Top-quality FM/AM broadcast radio

S-47
$229.50
Amateur Net

Hallicrafters best radio for conventional home reception . . . comparable to chassis found in consoles in the $600—$800 price class. Special features include Automatic Frequency Control, push-button tuning on both Broadcast and FM, and high-fidelity audio.

Automatic Frequency Control gives unprecedented ease of tuning on FM—with unequalled accuracy. As a station is approached, this circuit "takes over" electronically and holds the station in perfect tune with knife-like precision.

PERFORMANCE: Covers Standard Broadcast, FM and three Short-Wave Bands. Two "band-spread" Short-Wave bands spread out stations across the dial for easier tuning of popular foreign broadcasts. Temperature compensated oscillator. One RF, three IF stages. 10-watt push-pull output; audio reponse 30—15,000 cycles for rich, resonant tone.

CONTROLS: Five push buttons for AM and five for FM; Band selector switch FM 88—108 Mc, AM 540—1720 kc, 5.9—18.2 Mc, 9—12 Mc, and 15—18 Mc. Three-position Bass Tone, four-position Treble Tone, Volume, FM tuning, AM tuning.

PHYSICAL DATA: Gray steel, satin chrome trim. Piano hinge top. Size 20 x 10½ x 16 in. deep. Ship. wt. 66 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 500-ohm output (speaker not included—use R-42 on Page 3). Phonograph jack; 115 V. outlet for phono motor. Cord for 105—125 V. 50/60 cycle AC.

14 TUBES PLUS RECTIFIER: 6BA6 RF Amp., 6BE6 Mixer, 6J6 HFO and Auto. Freq. Control, two 6SG7 IF Amps., 6SG7 FM 3rd IF Amp. and AM Det., 6SH7 FM 4th IF, 6AL5 FM Det., two 6J5 and two 6SQ7 AF Amps., two 6V6GT Output, 5U4G Rect.

S-47C **$209.50**
Amateur Net

Chassis Only for Custom Installations

Same FM/AM chassis as in S-47 radio above. Offers superb performance with high-fidelity audio. For custom installations for those who prefer the finest. Size 18½ in. wide by 8¾ in. high by 16 in. deep. Ship. wt. 47 lbs. Fits relay rack.

the hallicrafters company

There is no other radio like a Hallicrafters. Its precision construction and skillful engineering will bring you superb performance on the short wave bands plus fine quality reception of your favorite broadcast programs. Thrilling land, sea, and air communications from all parts of the world plus hours of enjoyment on the amateur bands are yours with a Hallicrafters.

These world famous precision instruments have been sold in 89 different countries, used by 33 governments. They are remembered by veterans, prized by experts, and preferred by radio amateurs who want a radio that is all radio.

hallicrafters TELEVISION

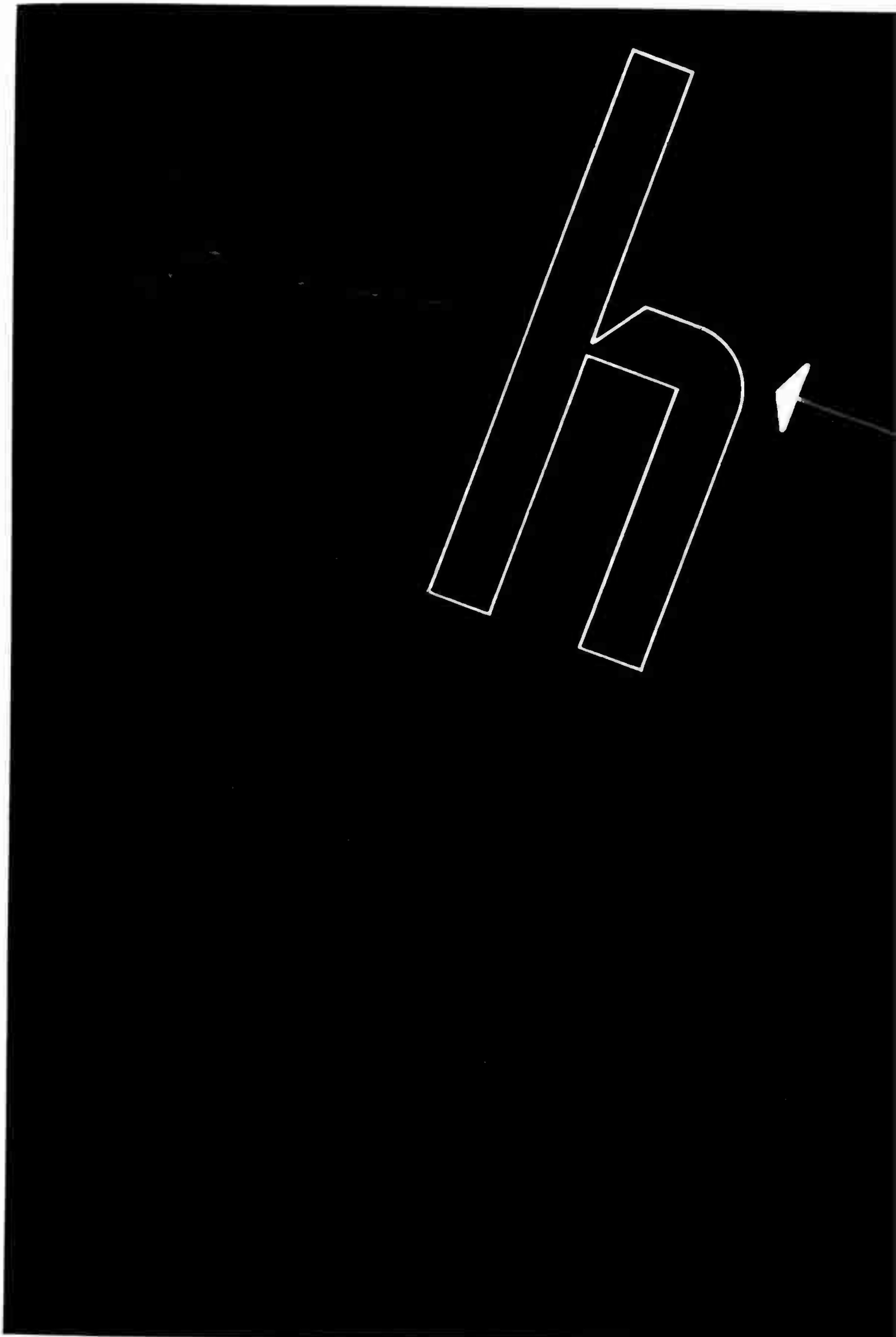
In addition to the high quality communications instruments in this catalog, Hallicrafters also make a complete line of precision-built television receivers—from 7-inch table models to large 16 by 12-inch console models. The same advanced engineering that has characterized "the radio man's radio" now brings you improved television with pictures that are exceptionally sharp, bright, and stable.

Complete information on Hallicrafters television models is available in separate folders. Ask wherever Hallicrafters equipment is sold or write direct to—

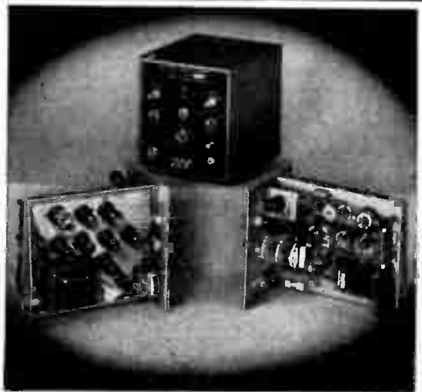
the hallicrafters co.

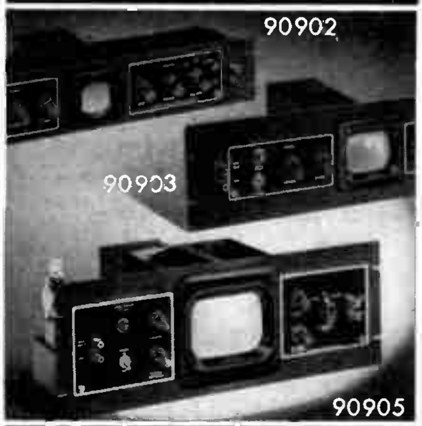
4401 West Fifth Ave., Chicago 24, Illinois

All prices in this catalog include Federal Excise Tax, if any. Prices are subject to change without notice.



JAMES MILLEN
MALDEN · MASSACHUSETTS





SECONDARY FREQUENCY STANDARD

A precision frequency standard for both laboratory and production uses, adjustable output, provided at intervals of 10, 25, 100 and 1000 kc, with magnitude useful to 50 mc. Harmonic amplifier with tuned plate circuit and panel range switch. 800 cycle modulator with panel control switch. In addition to oscillators, multivibrators, modulators and amplifiers, a built-in detector with phone jack and gain control is incorporated. Self-contained power supply.

Model 90505, with tubes............ $155.00

ABSORPTION WAVEMETERS

The 90600 series of absorption wavemeters are available in several styles and many different ranges. Most popular is kit of four units, covering range of 3.0 to 140 mc.

Model 90600..................... $18.00

FREQUENCY CALIBRATORS

The cavity type frequency calibrator covers a range of 200 to 700 mc., with a maximum error of not over 0.25%. This range is covered by two plug-in cavity type tuning units, which may be easily interchanged. The calibrator consists of an accurately calibrated cavity-type tuning unit, a crystal detector, a two-stage video amplifier and a peak reading VT voltmeter.

Model 90630, with tubes............ $375.00

LABORATORY SYNCHROSCOPES

The 5" laboratory synchroscopes are available with and without detector-video strips.

Model P-4-2, with tubes............. $350.00
Model P-4E-2, with tubes............ 445.00

MINIATURE SYNCHROSCOPE

The compact design of the No. 90952, measuring only 7½" x 5⅝" x 13", and weighing only 17 lbs., makes available for the first time a truly DESIGNED FOR APPLICATION "field service" Synchroscope.

No. 90952, with tubes.............. $375.00

CATHODE RAY OSCILLOSCOPES

The No. 90902, No. 90903 and No. 90905 Rack Panel Oscilloscopes, for two, three and five inch tubes, respectively, are inexpensive *basic units* comprising power supply, brilliancy and centering controls, safety features, magnetic shielding, switches, etc. As a transmitter monitor, no additional equipment or accessories are required. The well-known trapezoidal monitoring patterns are secured by feeding modulated carrier voltage from a pickup loop directly to vertical plates of the cathode ray tube and audio modulating voltage to horizontal plates. By the addition of such units as sweeps, pulse generators, amplifiers, servo sweeps, etc., all of which can be conveniently and neatly constructed on companion rack panels, the original basic 'scope unit may be expanded to serve any conceivable industrial or laboratory application.

No. 90902, less tubes.............. $ 42.50
No. 90903, less tubes.............. 49.50
No. 90905, less tubes.............. 100.00

'SCOPE AMPLIFIER—SWEEP UNIT

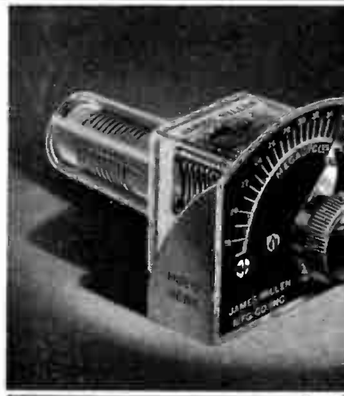
Vertical and horizontal amplifiers along with hard-tube, saw tooth sweep generator. Complete with power supply mounted on a standard 5¼" rack panel.

No. 90921, with tubes............... $75.00

REGULATED POWER SUPPLIES

A compact; uncased, regulated power supply, either for table use in the laboratory or for incorporation as an integral part of larger equipments. 50 watts, with regulated voltage from 0 to 200 volts.

Model 90201, less tubes............ $100.00



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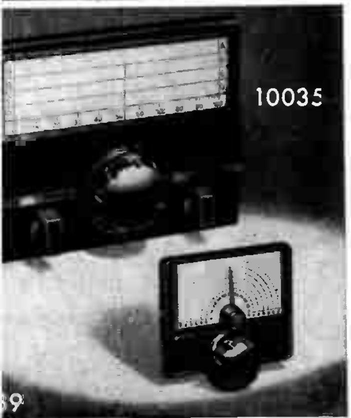
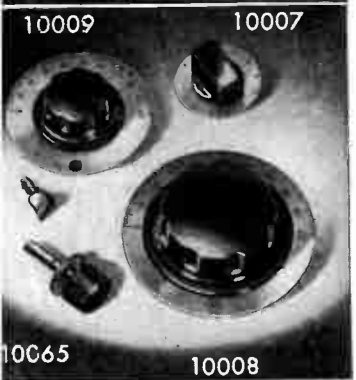


INSTRUMENT DIALS

The No. 10030 is an extremely sturdy instrument type indicator. Control shaft has 1 to 1 ratio. Veeder type counter is direct reading in 99 revolutions and vernier scale permits readings to 1 part in 100 of a single revolution. Has built-in dial lock and ¼" drive shaft coupling. May be used with multi-revolution transmitter controls, etc., or through gear reduction mechanism for control of fractional revolution capacitors, etc., in receivers or laboratory instruments.

The No. 10035 illuminated panel dial has 12 to 1 ratio; size, 8½" x 6½". Small No. 10039 has 8 to 1 ratio; size, 4" x 3¼". Both are of compact mechanical design, easy to mount and have totally self-contained mechanism, thus eliminating back of panel interference. Provision for mounting and marking auxiliary controls, such as switches, potentiometers, etc., provided on the No. 10035. Standard finish, either size, flat black art metal.

No. 10039	$ 2.70
No. 10035	6.00
No. 10030	25.00



DIALS AND KNOBS

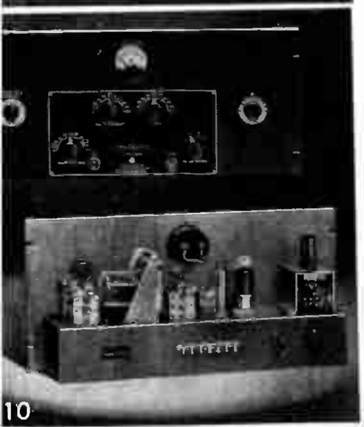
Just a few of the many stock types of small dials and knobs are illustrated herewith. 10007 is 1⅝" diameter, 10009 is 2½" and 10008 is 3½".

No. 10007	$ .60
No. 10008	1.00
No. 10009	.85
No. 10021	.15
No. 10065	.45

PANEL MARKING TRANSFERS

The panel marking transfers have ⅛" black letters. Special solution furnished. Must not be used with water. Equally satisfactory on smooth or wrinkle finished panels or chassis. Ample supply of every popular word or marking required for amateur or commercial equipment.

No. 59001, white letters	$1.25
No. 59002, black letters	1.25



HIGH FREQUENCY TRANSMITTER

The No. 90810 crystal control transmitter provides 75 watt output (higher output may be obtained by the use of forced cooling) on the 20, 10–11, 6 and 2 meter amateur bands. Provisions are made for quick band shift by means of the new 48000 series high frequency plug-in coils.

No. 90810, less tubes and crystals $69.75

HIGH FREQUENCY RF AMPLIFIER

A physically small unit capable of a power output of 70 to 85 watts on 'phone or 87 to 110 watts on C-W on 20, 15, 11, 10, 6 or 2 meter amateur bands. Provision is made for quick band shift by means of the new No. 48000 series VHF plug-in coils. The No. 90811 unit uses either an 829-B or 3E29.

No. 90811 with 10 meter band coils, less tube $33.00

HIGH VOLTAGE POWER SUPPLY

The No. 90281 high voltage power supply has a d.c. output of 700 volts, with maximum current of 250 ma. In addition, a.c. filament power of 6.3 volts at 4 amperes is also available so that this power supply is an ideal unit for use with transmitters, such as the Millen No. 90800, as well as general laboratory purposes. The power supply uses two No. 816 rectifiers and has a two section pi filter with 10 henry General Electric chokes and a 2-2-10 mfd. bank of 1000 volt General Electric Pyranol capacitors. The panel is standard 8¾" x 19" rack mounting.

No. 90281, less tubes $84.50

RF POWER AMPLIFIER

This 500 watt amplifier may be used as the basis of a high power amateur transmitter or as a means for increasing the power output of an existing transmitter. As shipped from the factory, the No. 90881 RF power amplifier is wired for use with the popular RCA or G.E. "812" type tubes, but adequate instructions are furnished for readjusting for operation with such other popular amateur style transmitting tubes as Taylor TZ40, Eimac 35T, etc. The amplifier is of unusually sturdy mechanical construction, on a 10½" relay rack panel. Plug-in inductors are furnished for operation on 10, 20, 40 or 80 meter amateur bands. The standard Millen No. 90800 exciter unit is an ideal driver for the new No. 90881 RF power amplifier.

No. 90881, with one set of coils, but less tubes $89.50

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92101

R9'er MATCHING PREAMPLIFIER

The Millen 92101 is an electronic impedance matching device and a broad-band preamplifier combined into a single unit, designed primarily for operation on 6 and 10 meters. Coils for 20 meter band also available.

No. 92101, less tubes $24.75


92105

SINGLE SIDEBAND SELECTOR

The No. 92105 is designed to permit Single Side-band Selection with existing receivers. Full technical details in April 1948 QST. Produced in co-operation and under exclusive U. S. patent license (2,364,863 and others) with the J. L. A. McLaughlin Research Laboratories.

No. 92105, with tubes and crystals $75.00


90700

FREQUENCY SHIFTER

A favorite frequency shifter, plugs in, in place of crystal, for instant finger-tip control of carrier frequency. Low drift, chirpless keying, vibration immune, big band spread, accurate calibration.

Model 90700, with tubes $42.50


90711

VARIABLE FREQUENCY OSCILLATOR

The No. 90711 is a complete transmitter control unit with 6SK7 temperature-compensated, electron coupled oscillator of exceptional stability and low drift, a 6SK7 broad-band buffer or frequency doubler, a 6A67 tuned amplifier which tracks with the oscillator tuning, and a regulated power supply. Output sufficient to drive an 807 is available on 160, 80 and 40 meters and reduced output is available on 20 meters. Close frequency setting is obtained by means of the vernier control arm at the right of the dial. Since the output is isolated from the oscillator by two stages, zero frequency shift occurs when the output load is varied from open circuit to short circuit. The entire unit is unusually solidly built so that no frequency shift occurs due to vibration. The keying is clean and free from all annoying chirp, quick drift, jump, and similar difficulties often encountered in keying variable frequency oscillators.

No. 90711, with tubes $89.75


90800

50 WATT TRANSMITTER

Based on an original Handbook design, this flexible unit is ideal for either low power amateur band transmitter use or as an exciter for high power PA stages.

Model 90800, less tubes $42.50


74400

OCTAL BASE AND SHIELD

Low loss phenolic base with octal socket plug and aluminum shield can 1⅞ x 1⅞ x 3$^{15}/_{16}$.

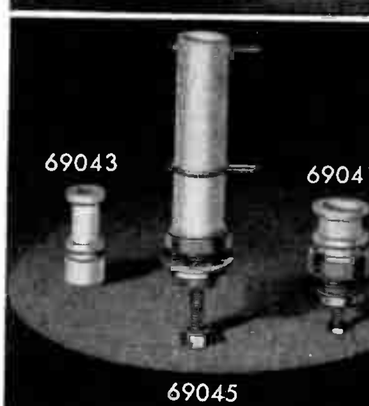
No. 74400 . . . . $.75


37412 · 33102 · FULL SIZE

TRANSMISSION LINE PLUG

An inexpensive, compact, and efficient polyethylene unit for use with the 300 ohm ribbon type polyethylene transmission lines. Fits into standard Millen No. 33102 (crystal) socket. Pin spacing ½", diameter .095".

No. 37412 . . . . $.21


69043 · 69041 · 69045 69046

PERMEABILITY TUNED CERAMIC FORMS

In addition to the popular shielded plug-in permeability tuned forms, 74000 series, the 69040 series of ceramic permeability tuned unshielded forms are available as standard stock items. Winding diameters and lengths of winding space are $^{13}/_{32}$ x $^{7}/_{32}$ for 69041–2; ¼ x ⅜ for 69043–7–8; ½ x 1$^{1}/_{16}$ for 69045–6; $^{3}/_{16}$ x $^{3}/_{16}$ for 69044.

No. 69041—(Copper Slug)	$.75
No. 69042—(Iron Core)	.75
No. 69043—(Iron Core)	.75
No. 69044—(Copper Slug)	.75
No. 69045—(Copper Slug)	.90
No. 69046—(Iron Core)	.90
No. 69047—(Copper Slug)	.90
No. 69048—(Iron Core)	.90

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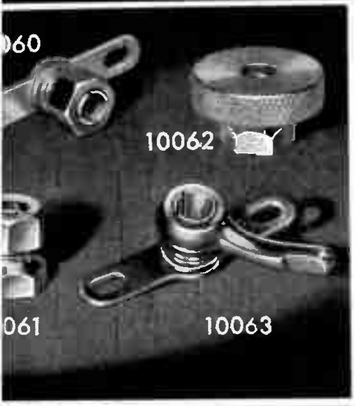
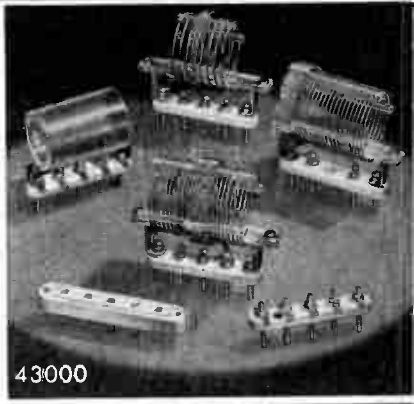


SHAFT LOCKS

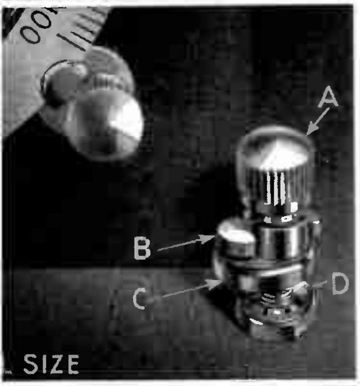
In addition to the original No. 10060 and No. 10061 "DESIGNED FOR APPLICATION" shaft locks, we can also furnish such variations as the No. 10062 and No. 10063 for easy thumb operation as illustrated above. The No. 10061 instantly converts any plain "¼ shaft" volume control, condenser, etc. from "plain" to "shaft locked" type. Each to mount in place of regular mounting nut.

No. 10060 .... $.36
No. 10061 .... .36
No. 10062 .... .45
No. 10063 .... .45



TRANSMITTING TANK COILS

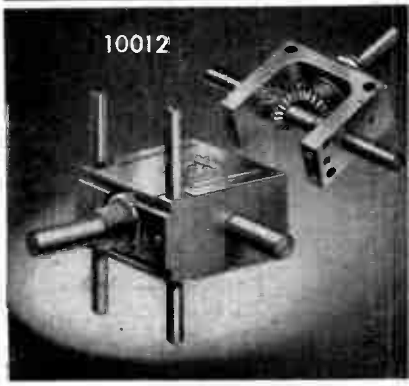
A full line—all popular wattages for all bands. Send for special catalog.



DIAL LOCK

Compact, easy to mount, positive in action, does not alter dial setting in operation! Rotation of knob "A" depresses finger "B" and "C" without imparting any rotary motion to Dial. Single hole mounted.

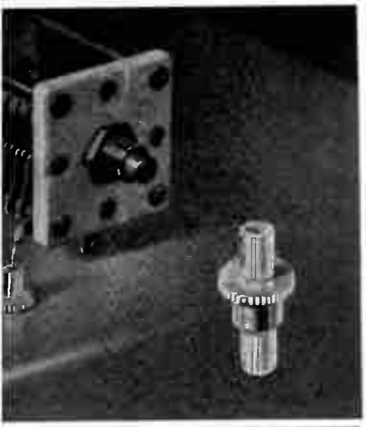
No. 10050 .... $.45



RIGHT ANGLE DRIVE

Extremely compact, with provisions for many methods of mounting. Ideal for operating potentiometers, switches, etc., that must be located, for short leads, in remote parts of chassis.

No. 10012 .... $3.75



THRU-BUSHING

Efficient, compact, easy to use and neat appearing. Fits ¼" hole in chassis. Held in place with a drop of solder or a "nick" from a crimping tool.

No. 32150 .... $.05

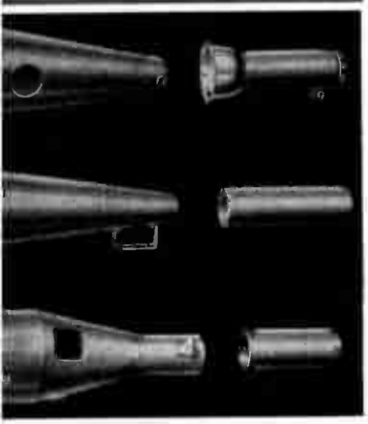
FLEXIBLE COUPLINGS

The No. 39000 series of Millen "Designed for Application" flexible coupling units include, in addition to improved versions of the conventional types, also such exclusive original designs as the No. 39001 insulated universal joint and the No. 39006 "slide-action" coupling (in both steatite and bakelite insulation).

The No. 39006 "slide-action" coupling permits longitudinal shaft motion, eccentric shaft motion and out-of-line operation, as well as angular drive without backlash.

The No. 39005 is similar to the No. 39001, but is not insulated and is designed for applications where relatively high torque is required. The steatite insulated No. 39001 has a special anti-backlash pivot and socket grip feature. All of the above illustrated units are for ¼" shaft and are standard production type units.

No. 39001 .... $.42
No. 39002 .... .42
No. 39003 .... .21
No. 39005 .... .42
No. 39006 .... .42



CATHODE RAY TUBE SHIELDS

For many years we have specialized in the design and manufacture of magnetic metal shields of nicaloi and mumetal for cathode ray tubes in our own complete equipment, as well as for applications of all other principal complete equipment manufacturers. Stock types as well as special designs to customers' specifications promptly available.

No. 80045—Nicaloi for 5" tube $10.50
No. 80043—Nicaloi for 3" tube 6.00
No. 80042—Nicaloi for 2" tube 5.25

BEZELS FOR CATHODE RAY TUBES

Five inch bezel is of cast aluminum with black wrinkle finish. Complete with neoprene cushion, green lucite filter scale and four screws for quick detachment from panel when inserting tube.

No. 80075—5" $7.50
No. 80073—3" 3.90
No. 80072—2" 1.25

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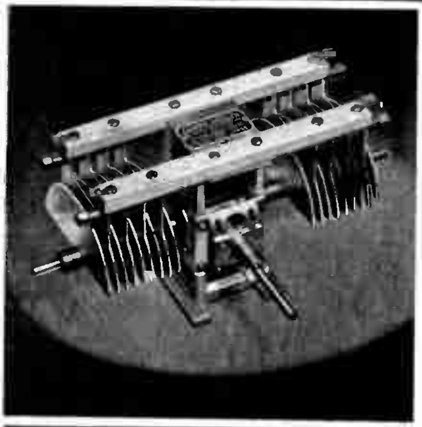
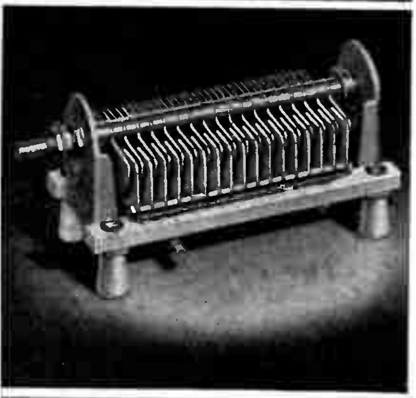
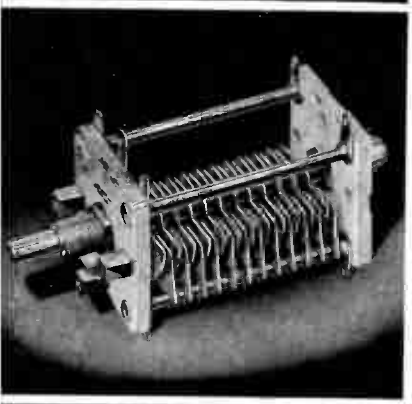
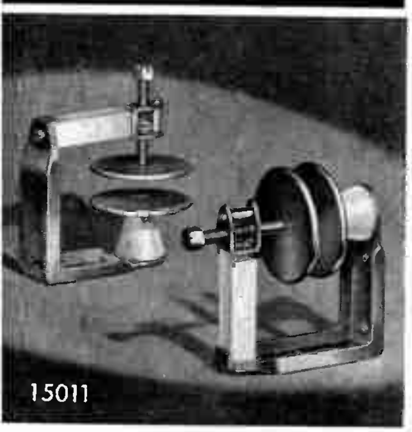








15011

04000 and 11000 SERIES TRANSMITTING CONDENSERS

A new member of the "Designed for Application" series of transmitting variable air capacitors is the 04000 series with peak voltage ratings of 3000, 6000, and 9000 volts. Right angle drive, 1–1 ratio. Adjustable drive shaft angle for either vertical or sloping panels. Sturdy construction, thick, round-edged, polished aluminum plates with 1¾" radius. Constant impedance, heavy current, multiple finger rotor contactor of new design. Available in all normal capacities.

The 11000 series has 16/1 ratio center drive and fixed angle drive shaft.

Code	*Volts*	*Capacity*	*Price*
11035	3000	35	$ 6.90
11050	3000	50	7.14
11070	3000	70	7.80
04050	6000	50	16.00
04060	9000	60	18.00
04100	6000	90	18.00
04200	3000	205	20.00

12000 and 16000 SERIES TRANSMITTING CONDENSERS

Rigid heavy channeled aluminum end plates Isolantite insulation, polished or plain edges. One piece rotor contact spring and connection lug. Compact, easy to mount with connector lugs in convenieni locations. Same plate sizes as 11000 series above.

The 16000 series has same plate sizes as 04000 series. Also has constant impedance, heavy current, multiple finger rotor contactor of new design. Both 12000 and 16000 series available in single and double sections and many capacities and plate spacing.

THE 28000–29000 SERIES VARIABLE AIR CAPACITORS

"Designed for Application," double bearings, steatite end plates, cadmium or silver plated brass plates. Single or double section .022" or .066" air gap. End plate size: 19/16" x 1 1 16". Rotor plate radius: ¾". Shaft lock, rear shaft extension, special mounting brackets, etc., to meet your requirements. The 28000 series has semi-circular rotor plate shape. The 29000 series has approximately straight frequency line rotor plate shape. Prices quoted on request. Many stock sizes.

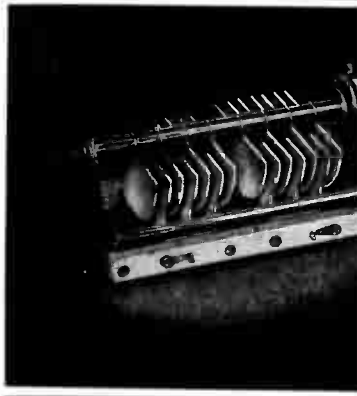
NEUTRALIZING CAPACITOR

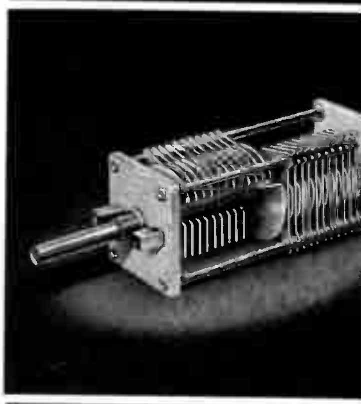
Designed originally for use in our own No. 90881 Power Amplifier, the No. 15011 disc neutralizing capacitor has such unique features as rigid channel frame, horizontal or vertical mounting, fine thread over-size lead screw with stop to prevent shorting and rotor lock. Heavy rounded-edged polished aluminum plates are 2" diameter. Glazed Steatite insulation.

No. 15011 .... $3.15

I.F. TRANSFORMERS

The Millen "Designed for Application" line of I.F. transformers includes air condenser tuned, and permeability tuned types for all applications. Standard stock units are for 456, 1600 and 5000 kc. B.F.O. also available.





68161

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TUBE SOCKETS DESIGNED FOR APPLICATION

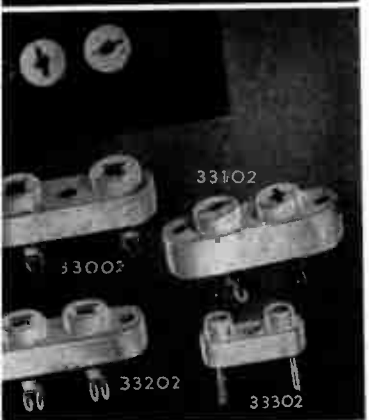
MODERN SOCKETS for MODERN TUBES! Long Flashover path to chassis permits use with transmitting tubes, 866 rectifiers, etc. Long leakage path between contacts. Contacts are type proven by hundreds of millions already in government, commercial and broadcast service, to be extremely dependable. Sockets may be mounted either with or without metal flange. Mounts in standard size chassis hole. All types have barrier between contacts and chassis. All but octal and crystal sockets also have barriers between individual contacts in addition.

The No. 33888 shield is for use with the 33008 octal socket. By its use, the electrostatic isolation of the grid and plate circuits of single-ended metal tubes can be increased to secure greater stability and gain.

The 33087 tube clamp is easy to use, easy to install, effective in function. Available in speciol sizes for all types of tubes. Single hole mounting. Spring steel, cadmium plated.

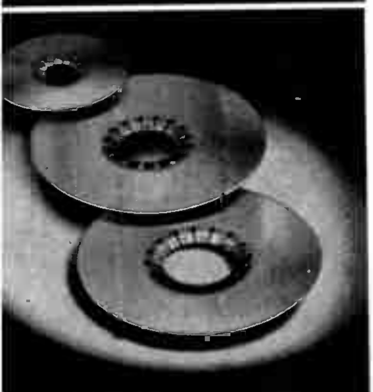
Covity Socket Contact Discs, 33446 are for use with the "Lighthouse" ultra high frequency tube. This set consists of three different size unhardened beryllium copper multi-finger contact discs. Heat treating instructions forwarded with each kit for hardening after spinning or forming to frequency requirements.

Voltage regulator dual contact bayonet socket, 33991 black Bakelite insulation and 33992 with low loss high leakage mica filled Bakelite insulation.



No. 33004	$ .30
No. 33005	.30
No. 33006	.30
No. 33007	.34
No. 33008	.30
No. 33888	.18
No. 33087	.30
No. 33002	.30
No. 33102	.30
No. 33202	.30
No. 33302	.21
No. 33446*	5.00
No. 33991	.45
No. 33992	.55

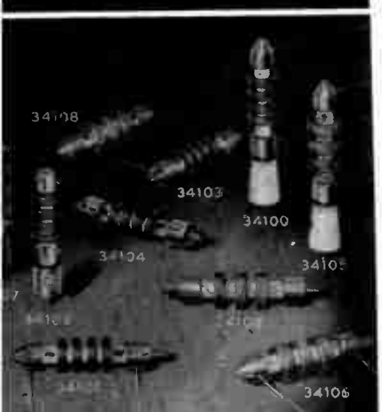
* For set of 3. Single discs $2.00 each.



RF CHOKES

Many have copied, few have equalled, and none have surpassed the genuine original design Millen Designed for Application series of midget RF Chokes. The more popular styles now in constant production are illustrated herewith. Special styles and variations to meet unusual requirements quickly furnished.

General Specifications: 2.5 mH, 250 mA for types 34100, 34101, 34102, 34103, 34104, and 1 mH, 300 mA for types 34105, 34106, 34107, 34108, 34109.



No. 34100	$.42
No. 34101	.36
No. 34102	.42
No. 34103	.36
No. 34104	.42

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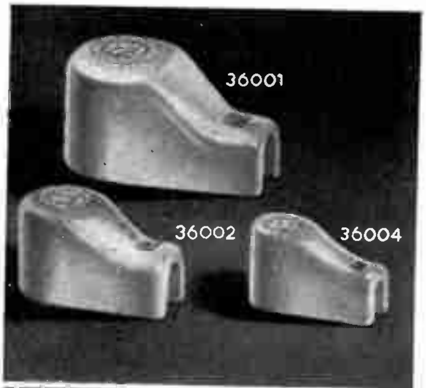
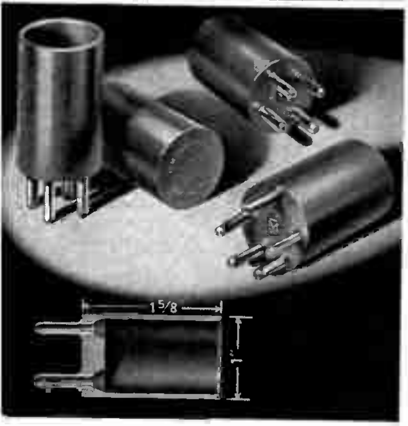




CERAMIC PLATE OR GRID CAPS

Soldering lug and contact one-piece. Lug ears annealed and solder dipped to facilitate easy combination "mechanical plus soldered" connection of cable.

No. 36001—9/16" .... $.21
No. 36002—3/8" .... .21
No. 36004—1/4" .... .21

SNAP LOCK PLATE CAP

For Mobile, Industrial and other applications where tighter than normal grip with multiple finger 360° low resistance contact is required. Contact self-locking when cap is pressed into position. Insulated snap button at top releases contact grip for easy removal without damage to tube.

No. 36011—9/16" .... $.60
No. 36012—3/8" .... .60

SAFETY TERMINAL

Combination high voltage terminal and thru-bushing. Tapered contact pin fits firmly into conical socket providing large area, low resistance connection. Pin is swivel mounted in cap to prevent twisting of lead wire.

No. 37001, Black or Red .... $.40
No. 37501, Low loss .... .55

TERMINAL STRIP

A sturdy four-terminal strip of molded black Textolite. Barriers between contacts. "Non turning" studs, threaded 8/32 each end.

No. 37104 .... $.60

POSTS, PLATES and PLUGS

Designed for Application! Compact, easy to use. Made in black and red regular bakelite as well as low loss brown mica filled bakelite or steatite for R.F. uses. Posts have captive head.

No. 37202 Plates (pr.) .... $.30
No. 37212 Plugs .... .70
No. 37222 Posts (pr.) .... .40

STEATITE TERMINAL STRIPS

Terminal and lug are one piece. Lugs are Navy turret type and are free floating so as not to strain steatite during wide temperature variations. Easy to mount with series of round holes for integral chassis bushings.

No. 37302 .... $ .60
No. 37303 .... .70
No. 37304 .... .80
No. 37305 .... .90
No. 37306 .... 1.00

MIDGET COIL FORMS

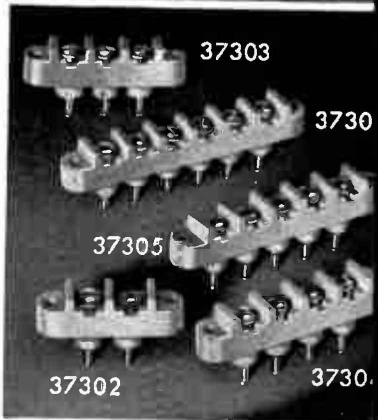
Made of low loss mica filled brown bakelite. Guide funnel makes for easy threading of leads through pins.

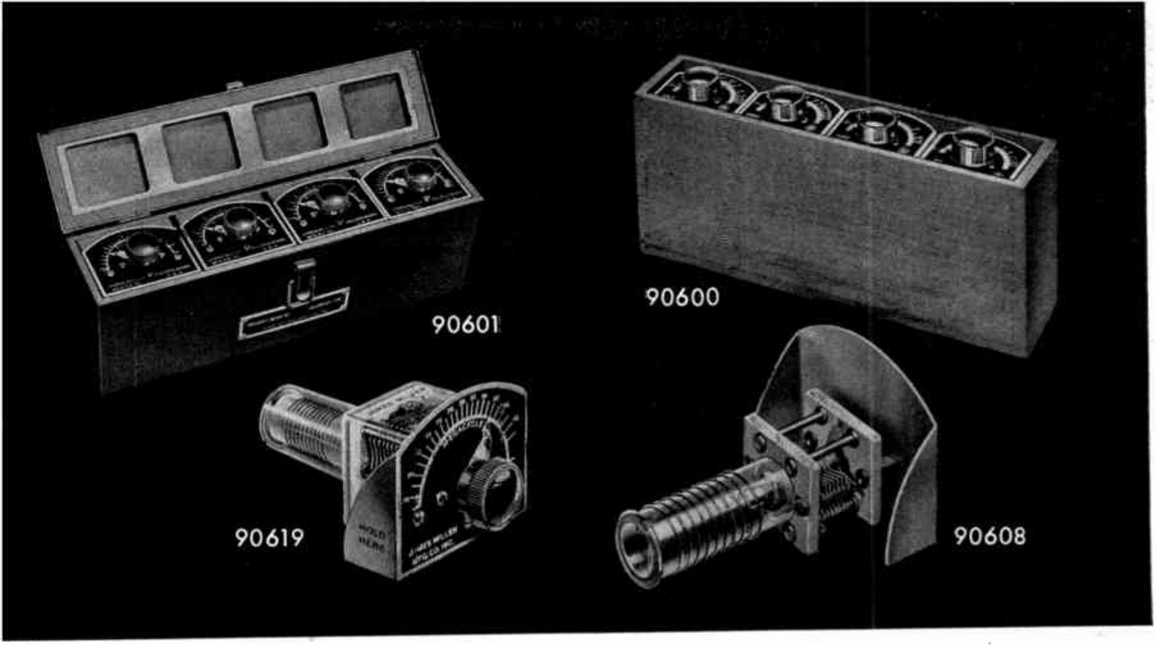
No. 45000 .... $.35
No. 45004 .... .45
No. 45005 .... .45

TUNABLE COIL FORM

Standard octal base of low loss mica-filled bakelite, polystyrene 1/2" diameter coil form, heavy aluminum shield, iron tuning slug of high frequency type, suitable for use up to 35 mc. Adjusting screw protrudes through center hole of standard octal socket.

No. 74001, with iron core $1.85
No. 74002, less iron core 1.50





Midget Absorption Frequency Meters

Many amateurs and experimenters do not realize that one of the most useful "tools" of the commercial transmitter designer is a series of very small absorption type frequency meters. These handy instruments can be poked into small shield compartments, coil cans, corners of chassis, etc., to check harmonics; parasitics; oscillator-doubler, etc., tank tuning; and a host of other such applications. Quickly enables the design engineer to find out what is really "going on" in a circuit.

Types 90605 thru 90609 are extremely small and designed primarily for engineering laboratory use where they will be handled with reasonable care. The most useful combination being the group of four under code No. 90600 and covering the total range of from 3.0 to 140 megacycles. When purchased in sets of four under code No. 90600 a convenient carrying and storage case is included. Series 90601 are slightly larger and very much more rugged. They are further protected by a contour fitting transparent polystyrene case to protect against damage and dirt. This latter series is designed primarily for field use and are not quite as convenient for laboratory use as the 90605 thru 90608 types. All types have dials directly calibrated in frequency.

Code	Description	Net Price
90604	Range 160 to 210 mc.	$ 6.00
90605	Range 3.0 to 10 mc.	4.50
90606	Range 9.0 to 23 mc.	4.50
90607	Range 23 to 60 mc.	4.50
90608	Range 50 to 140 mc.	4.50
90609	Range 130 to 170 mc.	6.00
90610	Range 105 to 150 mc.	6.00
90619	Range 350 to 1000 kc.—Neon Indicator	15.00
90620	Range 150 to 350 kc.—Neon Indicator	15.00
90625	Range 2 to 6 mc.—Neon Indicator	15.00
90626	Range 5.5 to 15 mc.—Neon Indicator	15.00
90600	Complete set of 90605 thru 90608, in case	18.00
90601	Complete set Field type Frequency Meters in metal carrying case 1.5 to 40 mc.	29.00

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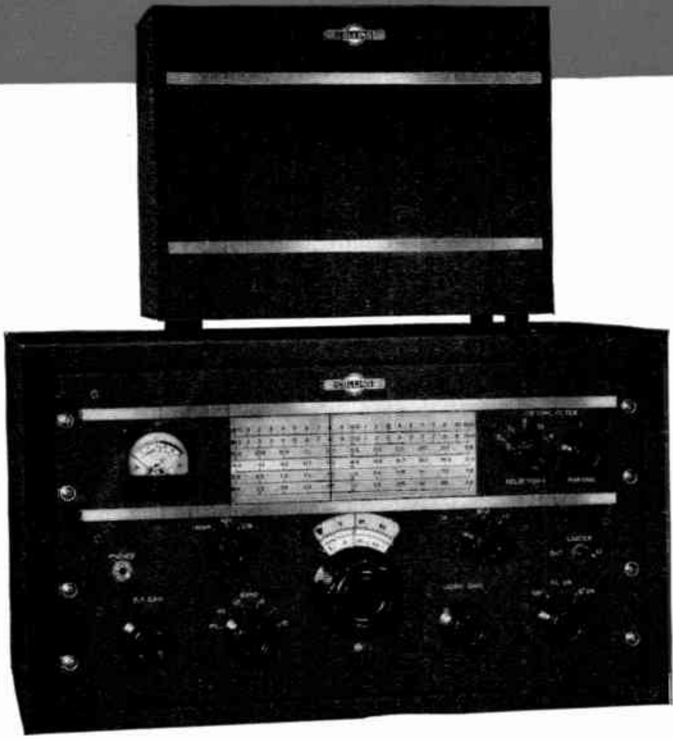
ENJOY COLLINS PERFORMANCE



COLLINS 75A-1 RECEIVER

The 75A-1 receiver is a product of the most advanced engineering, and was designed specifically to give the radio amateur the best possible performance in the 80, 40, 20, 15, 11 and 10 meter bands.

Double conversion and crystal filter controls, with a high frequency first i-f and a low frequency second i-f, provide approximately 50 db image rejection in all bands. The received band width is variable in 5 steps from 4 kc to 200 cycles at 6 db down from the peak of the resonant frequency. The 6AK5 r-f stage makes possible a threshold sensitivity far better than can be realized in normal installations.

Very high accuracy and stability result from the use of precision quartz crystals in the first conversion circuit, the extreme accuracy and stability of the v.f.o. in the second conversion circuit, and linearity and absence of backlash in the tuning mechanism. The band-lighted slide rule dial roughly indicates frequency, while the vernier dial provides a direct reading in kilocycles. Panel controls include tuning, bandswitch, r-f gain, audio gain, c-w pitch, on-off-standby, crystal selectivity, crystal phasing, avc-manual-c-w, noise limiter on-off.

Dimensions: 21⅛" wide, 12¼" high, 13⅞" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 93 pounds including speaker.
Net domestic price, complete with 13 tubes and rectifier, speaker and cabinet assembly, and instruction book (exclusive of state tax but including excise tax)...$375.00

For results in amateur radio, it's . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

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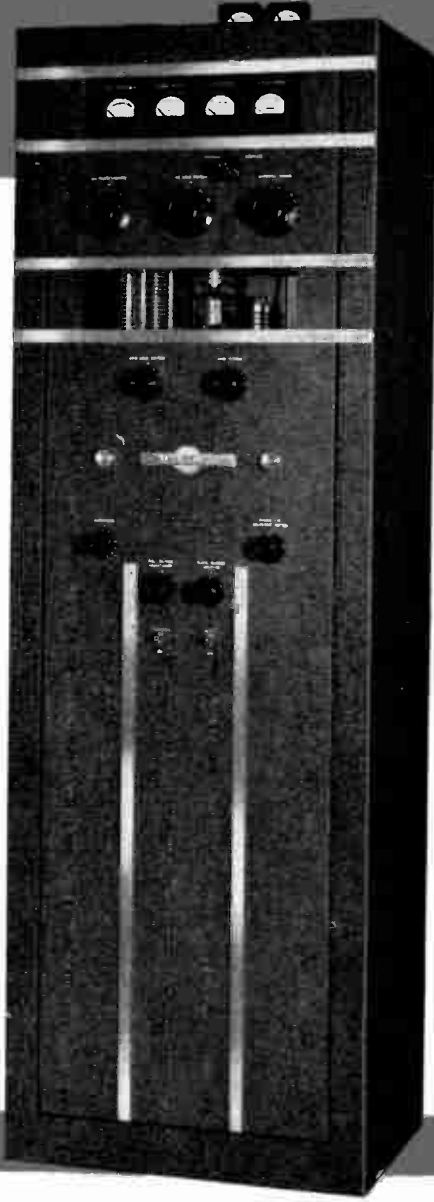


COLLINS 310A-3 EXCITER

The 310A-3 exciter for the 30K-1 has a Collins 70E-8A PTO, and an r-f output of 10 watts. All circuits are ganged together and controlled by a single tuning knob. The slide rule dial gives a rough indication of operating frequency; the vernier dial provides direct reading in kilocycles. A c-w sidetone oscillator is controlled at the front panel for pitch and volume. Grid block keying is provided. A keying filter eliminates key clicks. Removal of the key from the jack closes the circuit for phone operation.

Dimensions: 17¼" wide, 10½" high, 12½" deep.
Power source: 30K-1 power supply.

TUBE LINE-UP:

1—6SJ7 PTO	1—807 multiplier
1—6AG7 buffer amplifier	1—6SL7GT c-w side tone oscillator
1—6AG7 doubler	2—VR105 voltage regulators
1—807 multiplier	2—VR150 voltage regulators

COLLINS 30K-1 TRANSMITTER—500 watts input c-w, 375 watts phone

The owner of a 30K-1 has the best performing half-kilowatt rig money can buy. Operating on the 80, 40, 20, 15, 11 or 10 meter bands, he can run 500 watts of stable c-w or 375 watts of clean, intelligible phone into his PA amplifier. He has bandswitching in all transmitting circuits except the antenna tuning network, where one plug-in coil covers 80 and 40 meters, and another covers 20, 15, 11 and 10. He has very accurate Collins PTO control right on his desk, in the 310A-3 exciter.

It is often said that you can spot a Collins 30K-1 transmitter on phone as soon as you hear it, and that it seems to have more sock than its rated power. One reason for this is well engineered speech clipping, which permits running the audio gain at high level, with 100% modulation. Another reason lies in good audio design and fine components, providing remarkable clarity of voice transmission.

Dimensions: 22" wide, 66½" high, 16½" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 601 pounds including 310A-3 exciter.
Net domestic price, complete with 310A-3 exciter, tubes, microphone cord, r-f cable, power cable, and instruction book $1450.00

TUBE LINE-UP:

1—4-125A r-f power amplifier	1—6B4G modulator driver
1—6SJ7 speech amplifier	2—75TH Class B modulators
1—6SN7 audio amplifier	1—5R4GY bias rectifier
1—6H6 speech clipper	1—5R4GY low voltage rectifier

2—866A high voltage rectifiers

32V-1 TRANSMITTER

150 watts input c-w, 120 watts phone

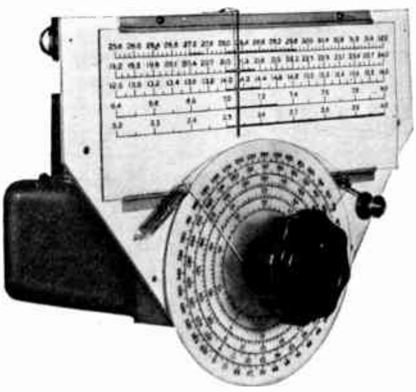
The 32V-1 is a high performance band-switching, gang-tuned transmitter covering the 80, 40, 20, 15, 11 and 10 meter bands. Its heart is the 70E-8A permeability tuned oscillator, used as the VFO. Accuracy and stability are outstanding. The entire transmitter is built into a cabinet 21⅛" wide, 12⁷⁄₁₆" high, and 13⅞" deep—identical in size and styling with the 75A-1 receiver cabinet.

The r-f tube line-up: a 6SJ7 VFO, 6AK6 buffer, 6AG7, 7C5 and 7C5 frequency multipliers and 4D32 final amplifier. Speech line-up: a 6SL7 in cascade to a 6SN7 to a pair of 807 modulators, which furnish 60 watts of audio power to modulate the final amplifier. The power supply contains a 5Z4 (low voltage) and two 5R4GY (high voltage) rectifiers, a VR-75 bias regulator, and two 0A2 screen voltage limiters.

The 32V-1 can be operated by a push-to-talk switch on the microphone, a key, or a separate switch. Terminals are provided for supplying the energizing voltage to the coil of an antenna change-over relay. There are also terminals, paralleled with the operate switch, with which to disable the receiver when the transmitter is in the SEND position. Grid-block keying is utilized on three stages following the VFO. The back-wave of the VFO as heard in a receiver placed beside the 32V-1 is negligible; thus break-in operation is accomplished without difficulty. Keying is very clean, without chirp or clicks. The keyer circuit also includes a side tone oscillator, utilizing a 6SL7, which is used as a c-w keying monitor. The output network consists of a single-ended pi and will load the transmitter into a wide variety of antennas.

Dimensions: 21⅛" wide, 12⁷⁄₁₆" high, 13⅞" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 128 pounds.
Net domestic price, complete with tubes and instruction book $475.00

70E-8A PTO WITH DIAL



The 70E-8A is a versatile and extremely accurate, stable variable frequency oscillator. It has a linear* range of 1600-2000 kc. Sixteen turns of the vernier dial are required to cover the 400 kc range. In the factory calibration of the 70E-8A we use a secondary frequency standard, continually checked against WWV. A special corrector mechanism in the oscillator produces a linear* calibration curve. To assure operation free from humidity effects this oscillator is baked until thoroughly dry, then completely sealed and moisture proofed. As an added protection against leakage, a silica gel capsule is factory inserted. The 70E-8A is employed in the 32V-1 transmitter and in all of the Collins exciters, but may be purchased separately.

Dimensions: Oscillator, 2¾" wide, 5¼" high, 4⅝" deep.
Dial, 10" wide, 8" high, 1¾" behind panel.
Shipping weight: 8 pounds.
Net domestic price, complete with tube, slide rule and vernier dial assembly, and instruction book $40.00
***.015% under normal conditions, which is ½ kc on the 80 meter band.**

COLLINS PTO EXCITER UNITS

Collins PTO (permeability tuned oscillator) exciters give you not only the flexibility and convenience of variable frequency, but also the accurate calibration and high stability inherent in the Collins 70E-8A PTO. These units provide a precision frequency control usually found only in laboratory instruments. Yet they are built for continuous operation under all normal fluctuations in operating conditions.

The slide rule dial roughly indicates operating frequency, while the vernier dial provides a direct reading in kilocycles. There are no reference charts or curves to interpolate. Accuracy and stability are so high as to be surprising to one who has not previously operated Collins PTO controlled equipment. Actual coverage of the PTO is from 1600 kc to 2000 kc. Sixteen complete turns of the vernier dial are required to cover this 400 kc range.

310B-1 and 310B-3 EXCITERS

The 310B-1 and 310B-3 are identical except for their output circuits. The 310B-1 has a link circuit output to work into a final of higher power. The 310B-3 has a series-parallel tunable matching antenna network of the universal type, and is a highly satisfying low power transmitter with many uses around any shack. It is excellent for standby, spot frequency network and emergency work, and is a natural for the beginner, who needs only add a final amplifier when more power is called for.

The 6SJ7 PTO of the 310B's is followed by three 6AG7 multipliers and a 2E26 r-f amplifier. The multiplier circuits are gang tuned and employ bandswitching. Plug-in coils are utilized in the 2E26 plate circuit. The self-contained power supply uses a 5R4GY high voltage rectifier, a 5Z4 low voltage rectifier, a 6H6 bias rectifier, a VR105 voltage regulator, and a VR150 voltage regulator.

Keying is accomplished by applying blocking bias to grid circuits of multiplier stages. A keying filter is included, to remove clicks. The keyer circuit also includes a side tone oscillator, used as a c-w keying monitor.

Dimensions: 19″ wide, 9³⁄₁₆″ high, 10⅜″ deep.
Power source: 115 volts a-c ±10%.
Shipping weight: 310B-1, 50 pounds.
310B-3, 54 pounds.

Net domestic price, complete with tubes and instruction book:
310B-1 Exciter $190.00
310B-3 Exciter with antenna network 215.00

Dimensions: 15³⁄₁₆″ wide, 8¾″ high, 8⅜″ deep.
Power requirements of 310C-1: 6.3 v. a-c @ 1.0 amp; 300 v. d-c @ 40 ma.
Power source for 310C-2: 115 v. a-c ±10%.
Shipping weight, 310C-1: 23 pounds.
310C-2: 27 pounds.

Net domestic price, complete with tubes and instruction book:
310C-1, without power supply .... $ 85.00
310C-2, with power supply 100.00

310C-1 and 310C-2 EXCITERS

The 310C-1 exciter is a straightforward unit consisting of the accurate, stable 70E-8A PTO and a multiplier, with an r-f output of approximately 80 volts rms across 40,000 ohms. Its output frequency range is from 3.2 mc to 4.0 mc. The output can be plugged into the crystal socket, or applied to the grid of an 807 buffer stage, thus giving your present rig a versatility far greater than a large number of crystals could provide. Yet you retain crystal accuracy and stability.

The 310C-2 has a self-contained power supply. Otherwise it is identical with the 310C-1. It employs a 6SJ7 oscillator, 6AG7 multiplier, 6X5 rectifier, VR105 regulator, and VR150 regulator.

THE **EASIEST-TO-USE** RADIO KEYS IN THE WORLD!

AMAZING NEW VIBROPLEX

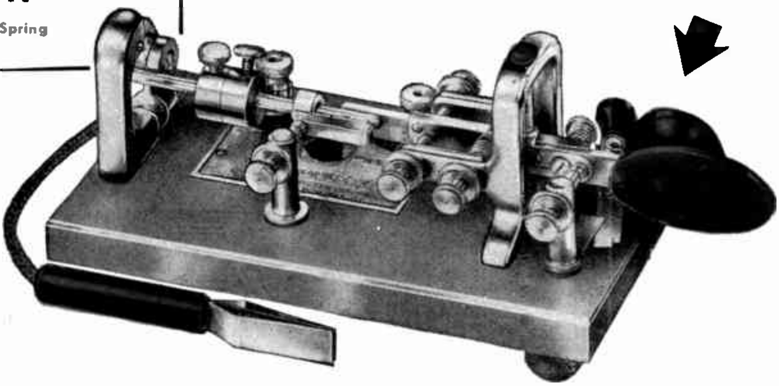
Super DeLuxe Model

"PRESENTATION"

With Super-Speed Control Main Spring and Other Great New Features

VIBROPLEX

Reg. Trade Marks: Vibroplex, Lightning Bug, Bug



- **Touch Control** . . . instant adjustment to personal touch.
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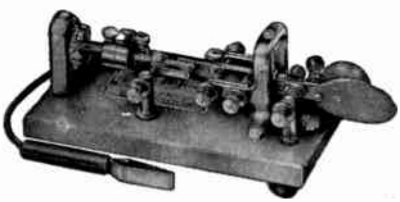
- **Non-Slip Rubber Feet** . . . hold bug firmly to table.
- **Ultra Modern Design** . . . polished chromium parts, 24-karat gold-plated base top.

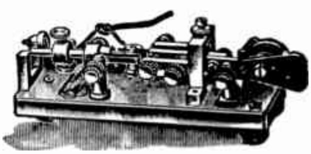
Illustrated above is the New, Super DeLuxe Vibroplex Model PRESENTATION, polished chromium parts, 24-K Gold-Plated Base Top **$27.50**

The New, Super DeLuxe Vibroplex Model "PRESENTATION" is the latest word in sending ease and enjoyment. With this amazing New Vibroplex key you'll be able to send better, faster and easier than ever before. So smooth and easy in action . . . strong, firm signals every time . . . suits any hand or any style of sending . . . sets firmly on table . . . built to last a lifetime.

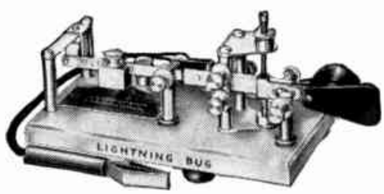
OTHER VIBROPLEX MODELS



Vibroplex Model ORIGINAL, famous the world over for high-class sending performance and ease of operation. Polished chromium parts and base, **$19.50**. Black base, **$15.95**.

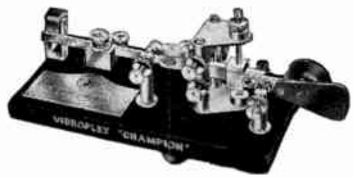


Vibroplex Model BLUE RACER, patterned after the Original Model but smaller in size. Capable of the same high-class sending performance for which the original is famous and just as easy to use. Polished chromium parts and base, **$19.50**. Black base, **$15.95**.



Vibroplex Model LIGHTNING BUG, advanced design, many new features contributing to exceptional sending efficiency and operating ease. Polished chromium parts and base, **$17.50**. Black base, **$13.95**.

Vibroplex Model CHAMPION, a full size, efficient key designed for radio use only. Modelled after the Lightning Bug and easy to use. An ideal key for the beginner. Chrome finished parts and black base, **$9.95**.



Vibroplex CARRYING CASE, plush-lined, finished in handsome simulated black morocco. Reinforced corners, flexible leather handle. Lock and key. Protects key from dust, dirt and moisture. Insures safe keeping when not in use, **$5.50**.

Write for FREE illustrated catalog and name of nearest dealer.

The "BUG" Trade Mark identifies the Genuine Vibroplex key.

It's your guaranty of complete satisfaction. Accept no substitute.

THE VIBROPLEX CO., INC.

833 Broadway New York 3, N. Y.

W. W. ALBRIGHT, President

THERE'S AN RCA TUBE

RCA Power Tube Chart for Amateur Transmitters

CW, FM AND PHONE TO 30 Mc.

This table of representative tube types has been set up to give suitable choice of tubes for the final and for a preceding stage to drive the final. A choice of buffer, doubler, or oscillator driver stage is provided. The tubes listed have been chosen conservatively to provide ample driving power at 30 Mc. even in circuits having higher than usual losses.

Final Amplifier			Tube Types for Driving Final Amplifier (CW, FM, and Phone)			Class B Modulator
Input Power Watts CW & FM	Phone	Tube Type	As Buffer	As Doubler	As Oscillator	Tube Type
40	27	1-2E26	2E26 6AG7 6AK6 6F6	2E26 6AG7 6F6 6N7 6V6GT	2E26 6AG7 6F6 6V6GT	2-6L6 (AB_1) 2-6F6 (AB_2)
75 75 75	54 60 60	2-2E26 1-815 1-807	2E26 6AG7 6F6 802 807	2E26 6AG7 6F6 6L6 802 807	2E26 6AG7 6F6 6L6 802 807	2-2E26 2-807 1-815
150	120	2-807	2E26 6F6 802 807	2E26 6F6 6L6 6N7 802 807	2E26 6F6 6L6 6V6GT 802 807	2-807 2-811
225	150	1-811	2E26 802 807	807 809 811 814	807 814	2-807 2-811
260	175	1-812-A	2E26 802 807	2E26 6L6 802 807	2E26 6L6 802 807	2-807 2-811
300 300	200 240	1-808 1-8005	2E26 802 807	807 809 811 814	807 814	2-808 2-811 2-8005
450	300	2-811	2-2E26 2-802 807 809 812-A 815	2-807 809 811 814	2-807 814 828	2-808 2-811 2-8005
500 500	375 400	1-4-125A/4D21 1-813	2E26 802 807	2E26 6L6 802 807	2E26 6L6 802 807	4-807 2-811 2-8005
520	350	2-812-A	2E26 802 807 809 812-A 815	807 809 811 814 815	807 814 815	2-808 2-811 2-8005
600 600	400 480	2-808 2-8005	2-2E26 807 809 811 812-A 815	2-807 809 811 812-A 814	2-807 814	2-808 2-811 2-8005
750	500	1-8000	807 809 811 812-A 814	807 809 811 814	807 814	4-811 2-8005
750 1000	500 600	1-810 1-806	809 811 812-A 814	808 811 814 828	not recommended	4-811 2-8005
1000 1000 1000	675 750 800	1-4-250A/5D22 2-4-125A/4D21 2-813	2E26 802 807 815	2E26 2-6L6 802 807 815	2E26 2-6L6 802 807 815	2-810 2-8000 4-8005
1000	1000	1-833A	808 809 811 812-A 814 8005	808 811 814 828	not recommended	2-810 2-8000 4-8005
1000	1000	2-8000	2-807 2-809 811 812-A 814	2-809 808 811 814	not recommended	2-810 2-8000 4-8005
1000	1000	2-810	2-809 808 811 812-A 814 8005	808 811 813 828	not recommended	2-810 2-8000 4-8005

FOR EVERY AMATEUR SERVICE

Ratings and Characteristics—Amateur Transmitting Types

RCA Type	Filament or Heater (H) Volts	Filament or Heater (H) Amps.	Amplification Factor	Max. Frequency for Full Input Mc.	Max. ICAS □ Ratings (Class C Telegraphy) Screen Input Watts	Max. ICAS □ Ratings (Class C Telegraphy) Plate Input Watts
RCA POWER TRIODES						
2C43	6.3 (H)	0.9	48	1500	—	16.7 §
6C24	11.0	12.1	30	160	—	1500 §
806	5.0	9.5	12.6	30	—	1000
808	7.5	4	47	30	—	300
810	10.0	4.5	36	30	—	750
811	6.3	4	160	60	—	225
812-A	6.3	4	29	30	—	260
826	7.5	4	31	250	—	130
833-A	10.0	10	35	30	—	1500
8005	10.0	3.25	20	60	—	300
8025-A	6.3	1.92	18	500	—	50
RCA BEAM POWER TUBES						
2E26	6.3 (H)	0.8	6.5•	125	2.5	40
807	6.3 (H)	0.9	8•	60	3.5	75
813	10.0	5	8.5•	30	22	500
815	6.3 / 12.6 (H)	1.6 / 0.8	6.5•	125	4.5*	75*
829-B	6.3 / 12.6 (H)	2.25 / 1.125	9•	200	7*	120*
RCA TETRODES AND PENTODES						
4-125A/ 4D21	5.0	6.5	6.2•	120	20	675
802	6.3 (H)	0.9	7.3•	30	6	33

RCA RECTIFIERS AND THYRATRONS

RCA Type	Volts	Amps.	Description
2D21	6.3 (H)	0.6	Gas thyratron, miniature type. Two tubes in grid-controlled, full-wave circuit, up to 80 watts at 400 volts.
5R4-GY	5.0	2	Full-wave, vacuum rectifier, with choke input, 175 ma. at 750 volts.
816	2.5	2	Half-wave, mercury-vapor rectifier. Two tubes in full-wave with choke input, 250 ma. up to 2380 volts.
866-A	2.5	5	Half-wave, mercury-vapor rectifier. Two tubes in full-wave with choke input, 500 ma. up to 3180 volts.
2050	6.3 (H)	0.6	Gas thyratron. Two tubes in grid-controlled, full-wave circuit, up to 80 watts at 400 volts.
5557	2.5	5	Mercury-vapor thyratron. Two tubes in grid-controlled, full-wave circuit, up to 1500 watts at 1500 volts.

RCA VOLTAGE REGULATORS

RCA Type	Operating Voltage	Current Range	
OA2	150 Volts	5 to 30 ma.	Glow-discharge types for regulating voltages to oscillators (ECO or XTAL types), oscillator power supplies, to stabilize bias voltages, and for spark-over protection. OA2 and OB2 are miniature types.
OB2	108 Volts	5 to 30 ma.	
OA3/ VR75	75 Volts	5 to 40 ma.	
OC3/ VR105	105 Volts	5 to 40 ma.	
OD3/ VR150	150 Volts	5 to 40 ma.	

□ ICAS—Intermittent Commercial and Amateur Service
§ Values shown are for Continuous Commercial Service
• Control grid-screen grid mu-factor
* Total for tube

● RCA has a popular tube for every Amateur service, every power, and every active band. A few of the best-known types in each classification are listed.

In addition, there are special-application types, such as *phototubes*, *acorns*, *kinescopes*, *iconoscopes*, and the well-known receiving types in metal, glass, and miniature.

For additional technical data on these RCA tube types, see your local RCA Tube Supplier, or write RCA, Commercial Engineering, Section AM35, Harrison, N. J.

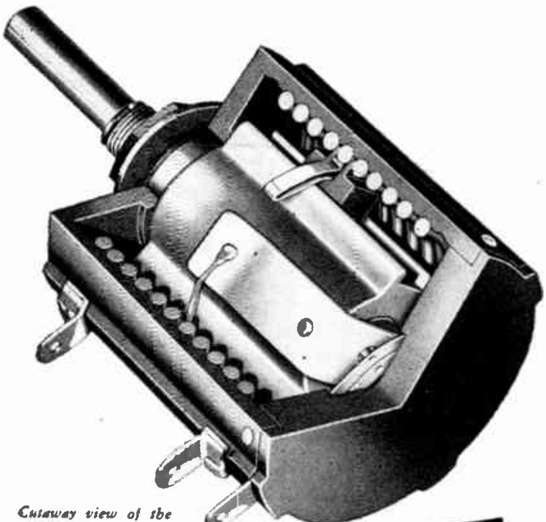
Are you getting RCA HAM TIPS? There's a copy waiting for you at your RCA Tube Supplier.

TUBE DEPARTMENT

RADIO CORPORATION of AMERICA

HARRISON, N. J.

For new simplicity, wide range, and high accuracy in the control of modern electronic circui



Cutaway view of the HELIPOT (Model A—10 Turn—1¾" Diameter)

THE BECKMAN Helipot

(Trademark of the HELical POTentiometer)

Provides *many times* greater resistance control in *same* panel space as conventional potentiometers!

IF YOU are designing or manufacturing *any type* of precision electronic equipment be sure to investigate the greater convenience, utility, range and compactness that can be incorporated into your equipment by using the revolutionary HELIPOT for rheostat-potentiometer control applications... and by using the new DUODIAL turns-indicating knob described at right.

Briefly, here is the HELIPOT principle... whereas a conventional potentiometer consists of a *single* coil of resistance winding, the HELIPOT has a resistance element *many times* longer coiled *helically* into a case which requires no more panel space than the conventional unit. A simple, foolproof guide controls the slider contact so that it follows the helical path of the resistance winding from end to end as a single knob is rotated. Result... *with no increase in panel space requirements*, the HELIPOT gives you as much as 12 times* the control surface. You get far greater accuracy, finer settings, increased range—with maximum compactness and operating simplicity!

COMPLETE RANGE OF TYPES AND SIZES

The HELIPOT is available in a complete range of types and sizes to meet a wide variety of control applications . . .

MODEL A: *5 watts, 10 turns, 46" slide wire length, 1¾" case dia., resistances 10 to 50,000 ohms, 3600° rotation.*

MODEL B: *10 watts, 15 turns, 140" slide wire length, 3¼" case dia., resistances 50 to 200,000 ohms, 5400° rotation.*

MODEL C: *3 watts, 3 turns, 13½" slide wire length, 1¾" case dia., resistances 5 to 15,000 ohms, 1080° rotation.*

MODEL D: *15 watts, 25 turns, 234" slide wire length, 3¼ case dia., resistances 100 to 300,000 ohms, 9000° rotation.*

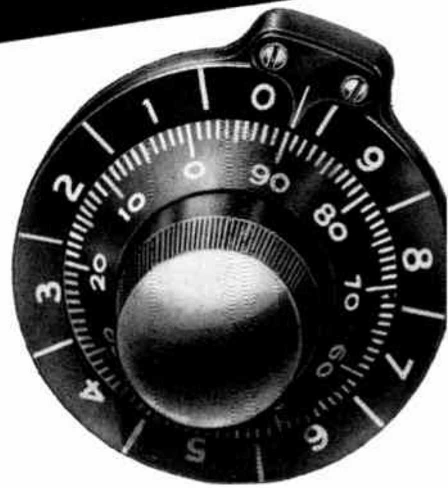
MODEL E: *20 watts, 40 turns, 373" slide wire length, 3¼" case dia., resistances 150 to 500,000 ohms, 14,400° rotation.*

Also, the HELIPOT is available in various special designs . . . with double shaft extensions, in multiple assemblies, integral dual units, etc.

Let us study *your* potentiometer problems and suggest how the HELIPOT can be used—possibly is already being used by others in your industry—to increase the accuracy, convenience and simplicity of modern electronic equipment. No obligation, of course. Write today outlining your problem.

**Data for Model A, 1¾" dia. Helipot. Other models give even greater control range in 3" case diameters.*

THE BECKMAN Duodial



The inner, or Primary dial of the DUODIAL shows exact angular posi tion of shaft during each revolution. The outer, or Secondary dia shows number of complete revolutions made by the Primary dial

A multi-turn rotational-indicating knob dial for us with the HELIPOT and other multiple turn devices

THE DUODIAL is a unique advancement in knob dial design It consists essentially of a primary knob dial geared to a concentric turns-indicating secondary dial—and the entire unit is so compact it requires only a 2" diameter panel space!

The DUODIAL is so designed that—as the primary dial rotate through each complete revolution—the secondary dial moves one division on its scale. Thus, the secondary dial counts the number of complete revolutions made by the primary dial. When used with the HELIPOT, the DUODIAL registers *both* the angular position of the slider contact on any given helix as well as the particular helix on which the slider is positioned.

Besides its use on the HELIPOT, the DUODIAL is readily adaptable to other helically wound devices as well as to many conventiona gear-driven controls where extra dial length is desired without wasting panel space. It is compact, simple and rugged. It contains only two moving parts, both made entirely of metal. It cannot be damaged through jamming of the driven unit, or by forcing beyond any mechanical stop. It is not subject to error from backlash of internal gears.

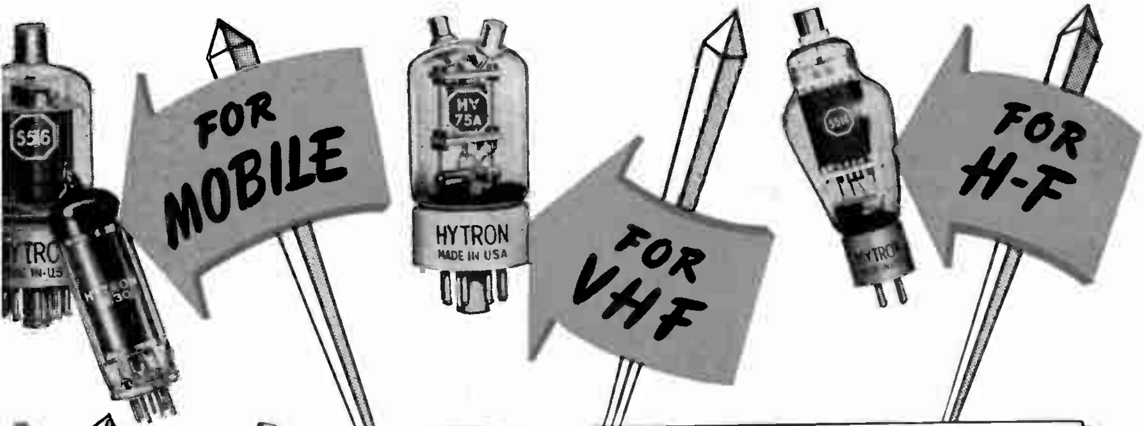
TWO SIZES—MANY RATIOS

The DUODIAL is now available in two types—the Model "R" (illustrated above) which is 2" in diameter, and the new Model "W" which is 4¾" in diameter and is ideal for main control applications. Standard turns-ratios include 10:1, 15:1, 25:1 and 40:1 (ratio between primary and secondary dials). Other ratios can be provided on special order. The 10:1 ratio DUODIAL can be readily employed with devices operating *fewer* than 10 revolutions and is recommended for the 3-turn HELIPOT. In all types, the primary dial and shaft operate with a 1:1 ratio, and all types mount directly on a ¼" round shaft.

Send for this HELIPOT AND DUODIAL CATALOG!

Contains complete data, construction details, etc., on the many sizes and types of HELIPOTS... and on the many unique features of the DUODIAL. Send for your free copy today!

THE Helipot CORPORATION, SOUTH PASADENA 10, CALIFORN



HYTRON TRANSMITTING AND SPECIAL PURPOSE TUBES
CONTINUOUS COMMERCIAL SERVICE RATINGS

Description	Type No.	Filament Ratings Volts	Filament Ratings Amps	Type	Max Plate Volts	Max Plate Ma	Max Plate Dis	Amateur Net Price
LOW AND MEDIUM MU TRIODES	10Y	7.5	1.25	Thor	450	65	15	$1.95
	HY24	2	0.13	Oxide	180	20	2	1.50
	801A/801	7.5	1.25	Thor	600	70	20	3.75
	864	1.1	0.25	Oxide	135	5	—	1.75
	1626	12.6	0.25	Cath	250	25	5	1.85
HIGH-MU TRIODES	HY31Z§	6	2.55	Thor	500	150*	30*	5.50
	HY1231Z§	6 12	3.2 1.6	Thor	500	150*	30*	5.50
	5514#	7.5	3	Thor	1500	175	65	4.95
VHF TRIODES	2C26A	6.3	1.15	Cath	3500	NOTE	10	7.75
	HY75A#§	6.3	2.6	Thor	450	90	15	4.70
	HY1148§	1.4	0.155	Oxide	180	12	1.8	2.25
	HY615	6.3	0.175	Cath	300	20	3.5	2.25
	955	6.3	0.15	Cath	200	8	1.8	3.60
	9002	6.3	0.15	Cath	200	8	1.8	2.50
BEAM PENTODES AND PENTODES	2E25#§	6	0.8	Thor	450	75	15	5.50
	2E30§	6	0.65	Oxide	250	60	10	2.45
	384	1.25 2.5	.330 .165	Oxide	150	25	3	3.60
	3D21A	6.3	1.7	Cath	3500	NOTE	15	7.50
	HY69§	6	1.6	Thor	600	100	30	5.50
	807	6.3	0.9	Cath	600	120	25	2.50
	837	12.6	0.7	Cath	500	80	12	4.75
	HY1269§	6 12	3.2 1.6	Thor	750	120	30	5.50
	1625	12.6	0.45	Cath	600	120	25	2.65
	5516§	6	0.7	Oxide	600	90	15	5.95
ACORNS MINIATURES	954	6.3	0.15	Cath	Sharp cutoff pentode			5.65
	9001	6.3	0.15	Cath	Sharp cutoff pentode			3.10

Description	Type No.	Filament Ratings Volts	Filament Ratings Amps	Type Rect	Peak Plate Ma	Max D-C Ma†	Inv Peak Pot.	Amateur Net Price
RECTIFIERS	816	2.5	2.0	Mer	500	250	5000	$1.30
	866A/866	2.5	5.0	Mer	1000	500	10000	1.95
	1616	2.5	5.0	Vac	800	260	6000	8.65

Description	Type No.	Average Operating Voltage	Operating Ma Min	Operating Ma Max	Av Volts Reg	Min Starting Voltage	List Price
GASEOUS VOLTAGE REGULATORS	OA2	150	5	30	2	185	$4.35
	O82	108	5	30	1	133	4.55
	OC3/VR105	108	5	40	2	133	2.65
	OD3/VR150	150	5	40	3.5	185	2.65

*Both sections of twin triode. NOTE: Special pulse tube, not recommended for c-w; consult Hytron Commercial Engineering Dept. for data. #5514 supplants the HY30Z, HY40, HY40Z, HY51A, HY51B, and HY51Z; the HY75A the HY75; and the 2E25 the HY65. †Current for full wave. §Instant-heating.

For better reception, it's also Hytron — GT, G, lock-in, metal, or miniature.

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PUT YOURSELF ON YOUR CARRIER
Reproduce the Individuality of Your Voice...
Transmit Your Own Personality
...with an E-V Microphone
The CARDAX
World's favorite premium crystal microphone. The only high level crystal cardioid with dual frequency response.
The "630"
Super dynamic. Long proved in service. Ideal frequency response. High output. Acoustalloy diaphragm.
The "910"
Rugged, handsome, crystal microphone. Outstanding performer, at low cost.
With an E-V microphone, you assure accurate reproduction of your own speaking voice. The shading and warmth of your speech arrive at the other end of the QSO undistorted and undiminished. Your carrier is modulated with your exact speech . . . the individuality of your voice is clearly retained . . . your personality is on your carrier.
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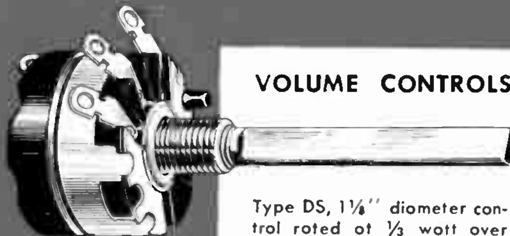
WHEREVER THE CIRCUIT SAYS

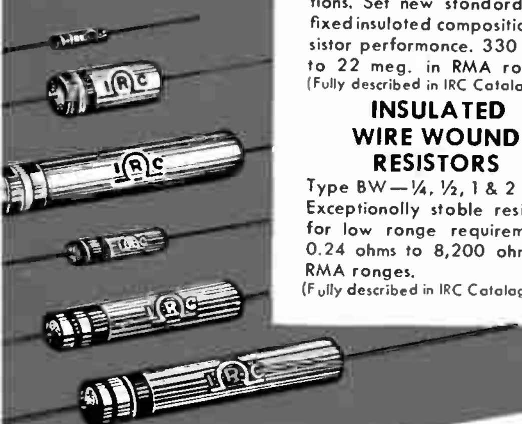
VOLUME CONTROLS

Type DS, 1⅛" diameter control rated at ⅓ watt over entire element.

Type D all-purpose control with IRC Top-In Shaft. Accommodates any one of 11 shafts. Both types feature exclusive Spiral Spring Connector and Five Finger Contactor.

(Fully described in IRC Catalog A-3.)





ADVANCED TYPE BT RESISTORS

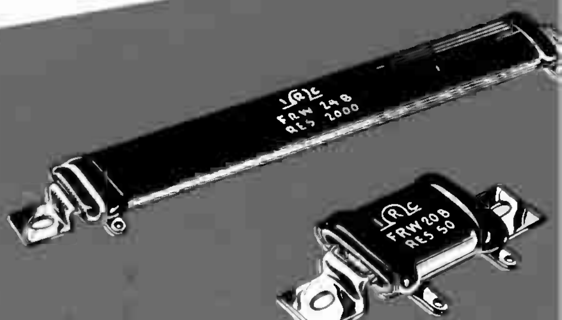
New Type BT Insulated Composition Resistors—⅓, ½, 1 & 2 watt. Meet JAN-R-11 specifications. Set new standards for fixed insulated composition resistor performance. 330 ohms to 22 meg. in RMA ranges.

(Fully described in IRC Catalog B-1)

INSULATED WIRE WOUND RESISTORS

Type BW—¼, ½, 1 & 2 watt. Exceptionally stable resistors for low range requirements. 0.24 ohms to 8,200 ohms in RMA ranges.

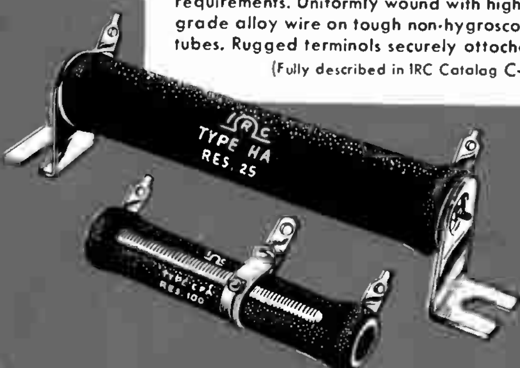
(Fully described in IRC Catalog B-5)



POWER WIRE WOUNDS

Available in full range of sizes, types and terminals. Two types of special cement coating to meet varied types of service requirements. Uniformly wound with highest grade alloy wire on tough non-hygroscopic tubes. Rugged terminals securely attached.

(Fully described in IRC Catalog C-2.)



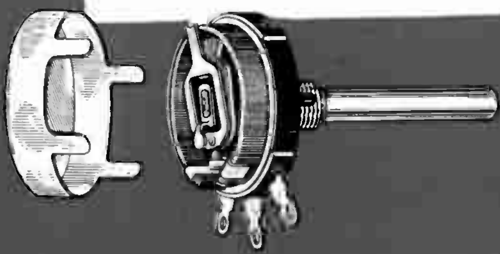
FLAT POWER WIRE WOUNDS

Designed for vertical or horizontal mounting, singly or in stacks. Higher space-power ratio than standard tubular wire wounds. Lightweight construction with extreme mechanical strength. Fixed and adjustable types.

(Fully described in IRC Catalog C-1.)

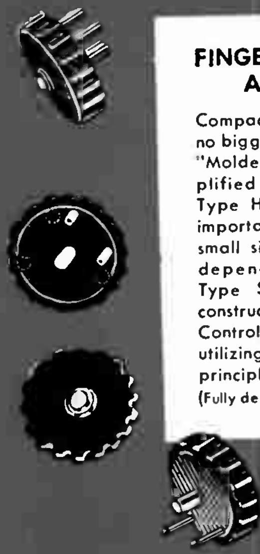
2 WATT WIRE WOUND POTENTIOMETER

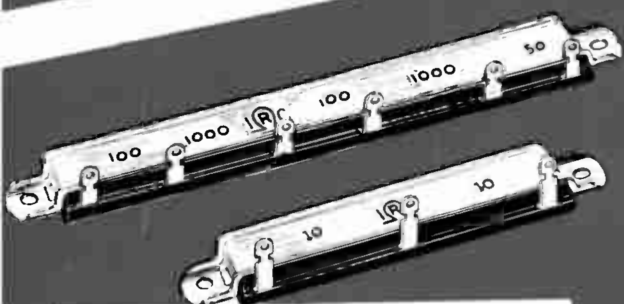
A fully dependable wire wound control providing maximum adaptability to most rheostat and potentiometer applications within its power rating. 1¼" diameter featuring IRC Spirol Spring Connector, long wearing alloy contactor and welded terminals between resistance element and terminals.
(Fully described in IRC Catalog A-2.)



FINGER-TIP CONTROL AND SWITCH

Compact, wafer-thin control, no bigger 'round than a nickel. "Molded-In" element and simplified construction enable Type H Control to fill many important applications where small size must combine with dependable performance. Type SH Switch, similar in construction to the Fingertip Control, is a four point switch utilizing the rotating cover principle.
(Fully described in IRC Catalog A-1.)





PRECISION RESISTORS

A scientifically designed resistor combining highest quality materials with maximum in accuracy and dependability. Used extensively by leading instrument makers. 1% accuracy is standard; closer tolerances available at slightly increased cost.
(Described in IRC Catalog D-1.)

WIRE WOUND RESISTORS

Type MW is a flat wire wound resistor of radically different design. Completely insulated and protected. Offers many opportunities in cost reduction by low initial cost, lower mounting cost, flexibility in providing taps at low cost, and saving in space. Multi-section feature permits exceptional flexibility for voltage dividing applications.
(Fully described in IRC Catalog B-2.)

other produc
in IRC's com
plete resist
line are describ
on the follo
ing pag

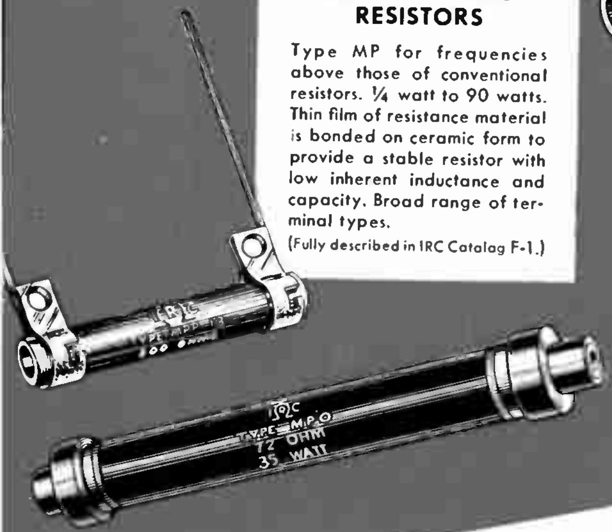
IRC

INTERNATIONAL RESISTANCE COMPANY

401 N. Broad Street Philadelphia 8, Pa.

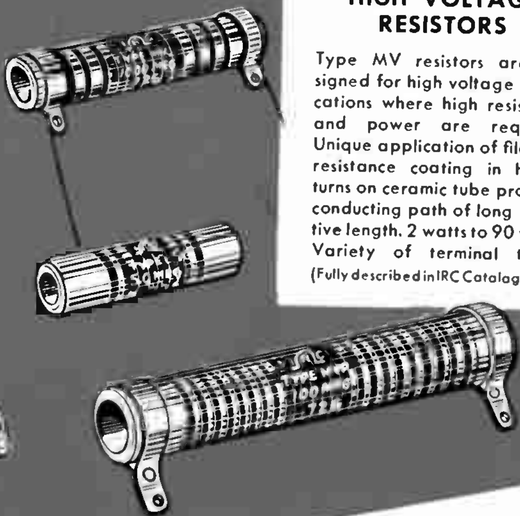
In Canada: International Resistance Co., Ltd., Toronto, Licensee

WHEREVER THE CIRCUIT SAYS —ΛΛ/Λ—



HIGH FREQUENCY RESISTORS

Type MP for frequencies above those of conventional resistors. ¼ watt to 90 watts. Thin film of resistance material is bonded on ceramic form to provide a stable resistor with low inherent inductance and capacity. Broad range of terminal types.
(Fully described in IRC Catalog F-1.)

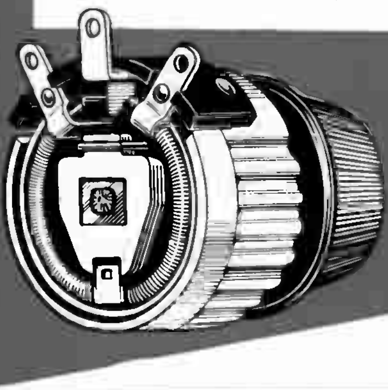


HIGH VOLTAGE RESISTORS

Type MV resistors are designed for high voltage applications where high resistance and power are required. Unique application of filament resistance coating in helical turns on ceramic tube provides conducting path of long effective length. 2 watts to 90 watts. Variety of terminal types.
(Fully described in IRC Catalog G-1.)

POWER RHEOSTAT

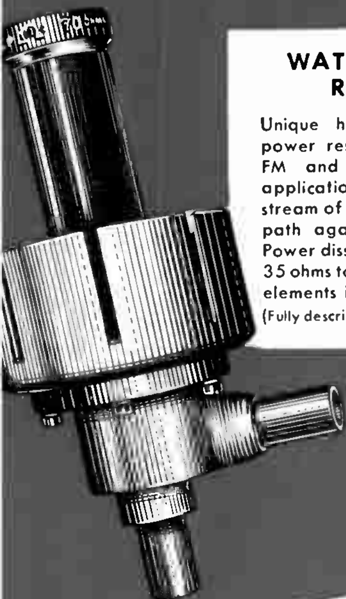
Type PR 25 and 50 watt. All-metal construction. Heat dissipating qualities of aluminum fully utilized. Operate at full rating at approximately half the temperature rise of equivalent units. Can be operated at full power in as low as 25% of rotation without appreciable difference in temperature rise.
(Fully described in IRC Catalog E-2.)



RHEOSTAT AN3155

Type PRT 25 and 50 watt. Developed to meet rigid Army-Navy specifications. Totally enclosed for protection against dirt and damage. All-metal construction. Can be operated down to 25% of full rotation with only minor increase in temperature rise.
(Fully described in IRC Catalog E-1.)

other products in IRC's complete resistor line are described on the preceding pages

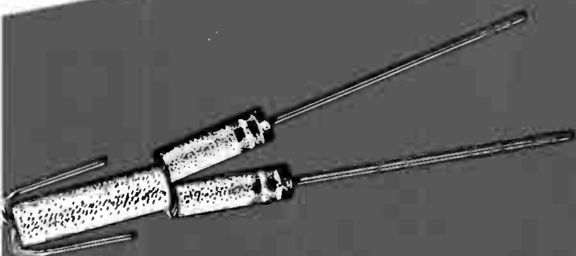


WATER-COOLED RESISTOR

Unique high frequency-high power resistor for television, FM and dielectric heating applications. High velocity stream of water flows in spiral path against resistance film. Power dissipation up to 5 K.W. 35 ohms to 1500 ohms. Resistor elements interchangeable.
(Fully described in IRC Catalog F-2.)

VOLTMETER MULTIPLIERS

Type MF consists of a number of IRC Precisions interconnected and encased in a glazed ceramic tube. Tube is hermetically sealed. Completely impervious to humidity. Maximum current: 1.0 M.A.; 0.5 megohms to 6 megohms.
(Fully described in IRC Catalog D-2.)



MATCHED PAIR RESISTORS

Two resistors matched in series or parallel to as close as 1% initial accuracy. Dependable low cost solution to close tolerance requirements. Both IRC Type BT and BW resistors are available in Matched Pairs.
(Fully described in IRC Catalog B-3.)

IRC RESIST-O-GUIDE

New aid in resistor range identification. Turn three wheels to correspond with color code and standard RMA Range is automatically indicated. 10¢ at all IRC Distributors. When ordering direct send stamps or coin.

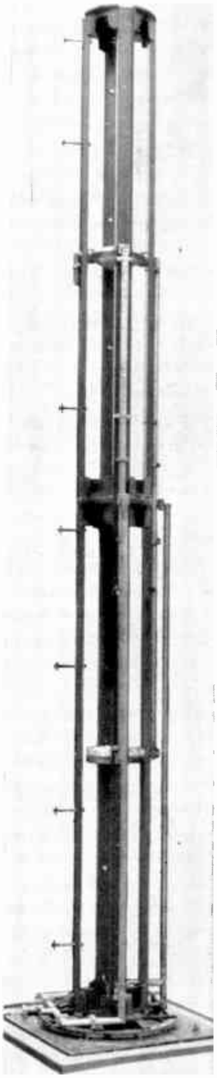
For detailed information on any of IRC's many resistor types write for catalog data bulletins specifying the product in which you are interested.

All standard IRC resistors are readily available in nominal quantities right from distributors' well-stocked shelves. These stock units are listed in Catalog 50 ... write for your copy and the name of your nearest IRC distributor.

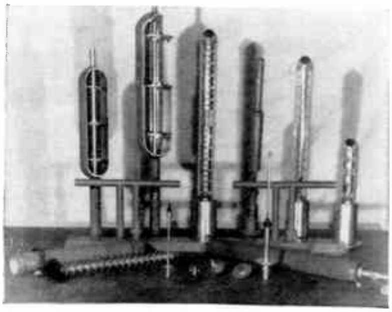
INTERNATIONAL RESISTANCE COMPANY
401 N. Broad Street — Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

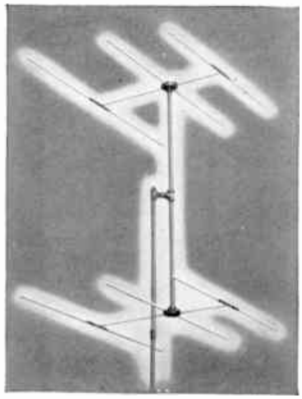
ANTENNAS
by
The Workshop Associates



The name WORKSHOP has become synonymous with fine high-frequency antennas — amateur, television, FM, parabolic, tower, beacon, and many other highly specialized types. During the war and ever since, the WORKSHOP research and test laboratories — the finest in the industry — have pioneered many new types of antenna equipment that today are widely copied and referred to as the criterion of design and performance. When you buy a WORKSHOP antenna, you *know* you are getting the best.



A representative group of high-precision radar antennas made by the WORKSHOP during the war.



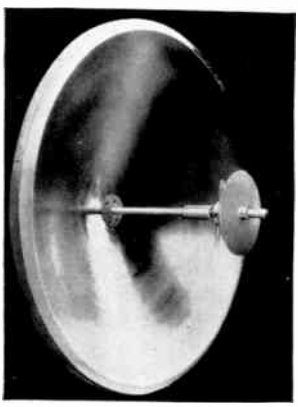
FM "Tower" Broadcast Antenna. Manufactured by Raytheon under exclusive WORKSHOP license.

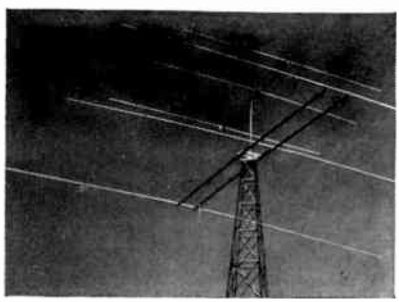
6-Element Super High Gain Television Antenna for sharp, clear reception at 100 miles and over.

Standard Parabolic Relay Antenna for Studio-to-Transmitter link on 920-960 mc. band.

High Gain Beacon Antenna. Recommended by all 152-162 mc. equipment manufacturers. Hundreds are in use throughout the country.

10 over 20 Stacked Array with Rotator and Indicator — the last word in amateur antenna equipment.





THE WORKSHOP ASSOCIATES, Inc.

Specialists in High-Frequency Antennas

54 NEEDHAM ST., NEWTON HIGHLANDS 61, MASSACHUSETTS

Be Right with OHMITE

BROWN DEVIL RESISTORS

Small, extra sturdy, wire wound, vitreous enameled resistors for voltage dropping, bias units, bleeders, etc. Proved right in vital installations the world over. In 5, 10 and 20-watt sizes in values to 100,000 ohms.

DUMMY ANTENNA RESISTORS

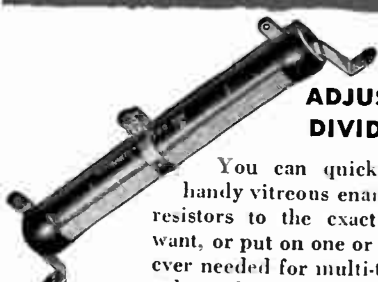
For loading transmit ters other r.f. sources. New rugg vitreous enameled units are "no reactive" within their recon mended frequency range. Tw types of terminals. In vario resistances, 100 and 250 watt.

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.

NEW HIGH FREQUENCY CHOKES

Single layer wound on lo power factor steatite or bak lite cores, with moisture proof coating. Seven stoc sizes for all frequencies fro 3 to 520 mc. Two units rated 600 ma all others are rated 1000 ma.



ADJUSTABLE DIVIDOHMS

You can quickly adjust these handy vitreous enameled Dividohm resistors to the exact resistance you want, or put on one or more taps wherever needed for multi-tap-resistors and voltage dividers. In 7 sizes from 10 to 200 watts. Resistances to 100,000 ohms.

PARASITIC SUPPRESSOR

Small, light, compact nonin ductive resistor and choke in parallel, 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 over-all and ⅝" in diameter.

FIXED RESISTORS

Resistance wire is wound over a ceramic core, permanently locked in place, insulated and protected by Ohmite vitreous enamel. Terminated by lugs. 25, 50, 100, 160 and 200-watt stock sizes, in resistances from 1 to 250,000 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. Wound on a ceramic core and has a moistureproof coating. Three stock sizes, rated 5, 10, and 20 amps.

HEOSTATS...RESISTORS...CHOKES...
OTENTIOMETERS...SWITCHES

Accurate ★ Dependable ★ Long-lived

CLOSE-CONTROL RHEOSTATS

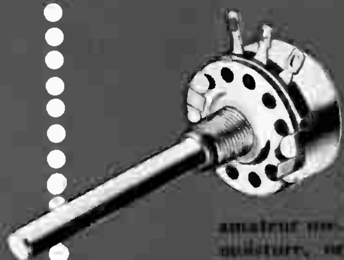
Insure permanently smooth, close control in communication, electronic and electrical devices. Widely used in industry. All ceramic, vitreous enameled. 25, 50, 75, 100, 150, 225, 300, 500, 750 and 1000-watt sizes.

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.

RB-2 DIRECTION INDICATOR POTENTIOMETER

A compact, low cost unit used in a simple potentiometer circuit as a transmitting element to indicate, remotely, the position of a rotary-beam antenna. Used with a 0-1 milliammeter and 6-v. battery.

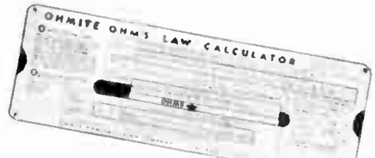


MOLDED COMPOSITION POTENTIOMETER

A high quality 2-watt unit with a good margin of safety, for industrial and amateur use. Unaffected by heat, cold, moisture, or length of service. *Sold through Ohmite distributors.*

LITTLE DEVIL INDIVIDUALLY MARKED INSULATED COMPOSITION RESISTORS

New, tiny, molded fixed resistors each marked with resistance and wattage rating. ½ Watt, 1 watt, and 2 watt sizes, ±10% tolerance. Also ±5%. 10 Ohms to 22 megohms. *Sold through Ohmite distributors.*



HANDY OHM'S LAW CALCULATOR

Figures ohms, watts, volts, amperes—quickly, easily. Solves any Ohm's Law problem with one setting of the slide. New pocket size—9"x3" has all computing scales on one side. Resistor color code on back. Send 25¢ in coin to cover handling cost.

OHMITE

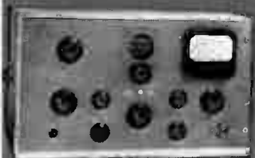
OHMITE MANUFACTURING COMPANY
4834 Flournoy Street, Chicago, U. S. A. Cable "Ohmiteco"

SEND FOR FREE CATALOG

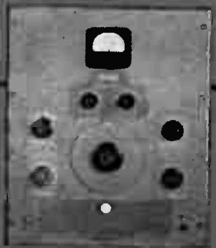
Stock catalog lists hundreds of units, gives helpful information.

MEASURING SPEED AND ACCURA

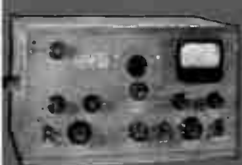
-hp- Model 200C
Resistance-Tuned Oscillator



-hp- Model 206A
Audio Signal Generator



-hp- Model 300A
Wave Analyzer



-hp- Model 330B
Distortion Analyzer

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
LOW FREQUENCY STANDARDS	100A	100 kc, 10 kc, 1 kc, 100 cps	Accuracy 3 cps per mc per degree Centi
	100B	100 kc, 10 kc, 1 kc, 100 cps	Temperature controlled; accuracy 0.00
FREQUENCY DIVIDER	110	100 to 10 cps	Controlled by 100A or 100B. Multipliers available up to 1 mc
RESISTANCE-TUNED OSCILLATORS	200A	35 to 35,000 cps	Output 1 watt into 500 ohms; 1% distor
	200B	20 to 20,000 cps	Output 1 watt into 500 ohms; 1% distor
	200C	20 to 200,000 cps	Output 10 volts into 1,000 ohms; 1% dist
	200D	7 to 70,000 cps	Output 10 volts into 1,000 ohms; 1% dist
	200I	6 to 6,000 cps	Frequency setting closer than 1%; output 1 into 1,000 ohms; 1% distortion
	201B	20 to 20,000 cps	Output 3 watts at 1% and 1 watt at ½ distortion into 600 ohms
	202B	⅓ to 1,000 cps	For low frequency studies. Output 10 vo into 1,000 ohms; 1% distortion
	202D	2 to 70,000 cps	Output 10 volts into 1,000 ohms; 1% disto
AUDIO SIGNAL GENERATORS	205A	20 to 20,000 cps	Output 5 watts, 1% distortion into impedan 50, 200, 600, 5,000 ohms. Output VTVM an db attenuator, 1 db steps
	205AG	20 to 20,000 cps	Same as 205A, plus separate VTVM for com gain measurements
	205AH	1 to 100 kc	Output 5 watts, 3% distortion into 50, 200, 5,000 ohm impedances. Output VTVM and 1 attenuator, 1 db steps
	206A	20 to 20,000 cps	Output +15 dbm with less than 0.1% dist into 50, 150, 600 ohm impedances. Output V and 111 db attenuator in 0.1 db steps
SQUARE WAVE GENERATOR	210A	20 to 10,000 cps	Output 50 volts peak to peak; 1,000 ohm int impedance; 70 db attenuator, 5 db steps
WAVE ANALYZER	300A	30 to 16,000 cps	Variable selectivity; measurement range 1 m 500 volts; 5% accuracy

Precision accuracy, ease of operation, fast measurements, compact size, rugged durability—those are family characteristics of *-hp-* laboratory instruments. For electronic measurements anywhere—laboratory, production line, radio station or amateur's shack.

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
DISTORTION ANALYZERS	320A	400 cps and 5 kc	Measures total distortion as low as 0.1%. 70 db attenuator, 1 db steps for comparison
	320B	50, 100, 400 cps and 1, 5 and 7.5 kc	Same as 320A
	325B	30, 50, 100, 400, 1,000 cps; 5, 7.5, 10 and 15 kc	Measures total distortion as low as 0.1%. Input amplifier and complete VTVM each usable separately
	330B	Any frequency 20 to 20,000 cps	Similar to 325B but measures at any frequency and includes AM detector
	330C	Any frequency 20 to 20,000 cps	Similar to 330B, no AM detector. Meter has VU characteristics to meet FCC requirements for FM broadcasting
FM BROADCAST MONITOR	335B	88 to 108 mc	FCC approved. Monitors carrier frequency and modulation. High fidelity output for aural monitoring
ATTENUATORS	350A	Max 100 kc	110 db, 1 db steps; 5 watts, 500 ohm level. Bridged T type. Accuracy 1 db in 50 db at 100 kc
	350B	Max 100 kc	Same as 350B but 600 ohm level
VACUUM TUBE VOLTMETERS	400A	10 cps to 1 mc	Nine ranges 0.03 to 300 volts full scale. Accuracy ±3% to 100 kc, ±5% to 1 mc. Average reading. Calibrated in rms.
	410A	20 cps to 700 mc	AC: six ranges 1 to 300 volts. DC: seven ranges 1 to 1,000 volts. Resistance: seven ranges 0.2 ohm to 500 megohms
AMPLIFIERS	450A	10 to 1,000,000 cps	40 db and 20 db stabilized gain. Input impedance 1 megohm shunted by approximately 15 uuf.
ELECTRONIC REQUENCY METER	500A	5 cps to 50 kc	Ten ranges, ±2% accuracy. Input 0.5 to 200 volts
ELECTRONIC TACHOMETER	505A	300 to 3,000,000 rpm	Ten ranges, ±2% accuracy
	505B	5 to 50,000 rps	Same as 505A except calibrated in rps
NAL GENERATORS	610A	500 to 1,350 mc	Calibrated output 0.1 microvolt to 0.1 volt. Internal pulse modulation. Direct calibration
	616A	1,800 to 4,000 mc	Direct reading. Pulse modulation, CW and FM. Calibrated output 0.1 microvolt to 0.2 volts
	650A	10 cps to 10 mc	Direct reading. Six bands. Output 3 volts to 600 ohm load. VTVM and output attenuator
POWER SUPPLY	710A		Any dc voltage 180 to 360 for 0 to 75 ma load approximately 1% regulation. Also 6.3 volts, 5 amps ac.

-hp- Model 400A Vacuum Tube Voltmeter

-hp- Model 410A Vacuum Tube Voltmeter

-hp- Model 610A UHF Signal Generator

-hp- Model 650A Audio Signal Generator

Brief specifications of internationally known -hp- instruments appear here. For full details see -hp- Catalog 19-A, available on request.

hp laboratory instruments

FOR SPEED AND ACCURACY

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Export: Frazar & Hansen, 301 Clay Street, San Francisco 11, California

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Most HAM RIGS ARE EQUIPPED WITH *Astatic* MICROPHONES

THE FINAL TEST of product quality and performance is the acceptance of users. Small wonder that Astatic takes extreme pride in the preference among amateurs for Astatic Crystal, Dynamic and Ceramic Microphones. The Astatic Corporation guards this acceptance zealously, spares no effort to produce superior products that will warrant your continued approval. Illustrated are the most popular models, new and old. If you would like to have more information, write us.

CONNEAUT CRYSTAL MICROPHONE—the ultimate in streamlining, bright chrome finish, shown with Type "S" Off-On Switch.

CARDINAL CRYSTAL, DYNAMIC OR CERAMIC MICROPHONE—a major new accomplishment in terms of quality performance at modest cost, has die-cast case in bright gold finish, compact to fit the hand, rests in streamlined CB base (as shown) or lies flat on felt-covered back.

VELVET VOICE CRYSTAL, DYNAMIC OR CERAMIC MICROPHONE—sparkling beauties in gold-finished case and handle, with dark brown, detachable base for convertibility to desk stand, floor stand, hand use.

D-104 CRYSTAL MICROPHONE—first practical crystal microphone developed, with few changes still the top favorite of amateurs.

Astatic Crystal Devices manufactured under Brush Development Co. patents

THE *Astatic* CORPORATION
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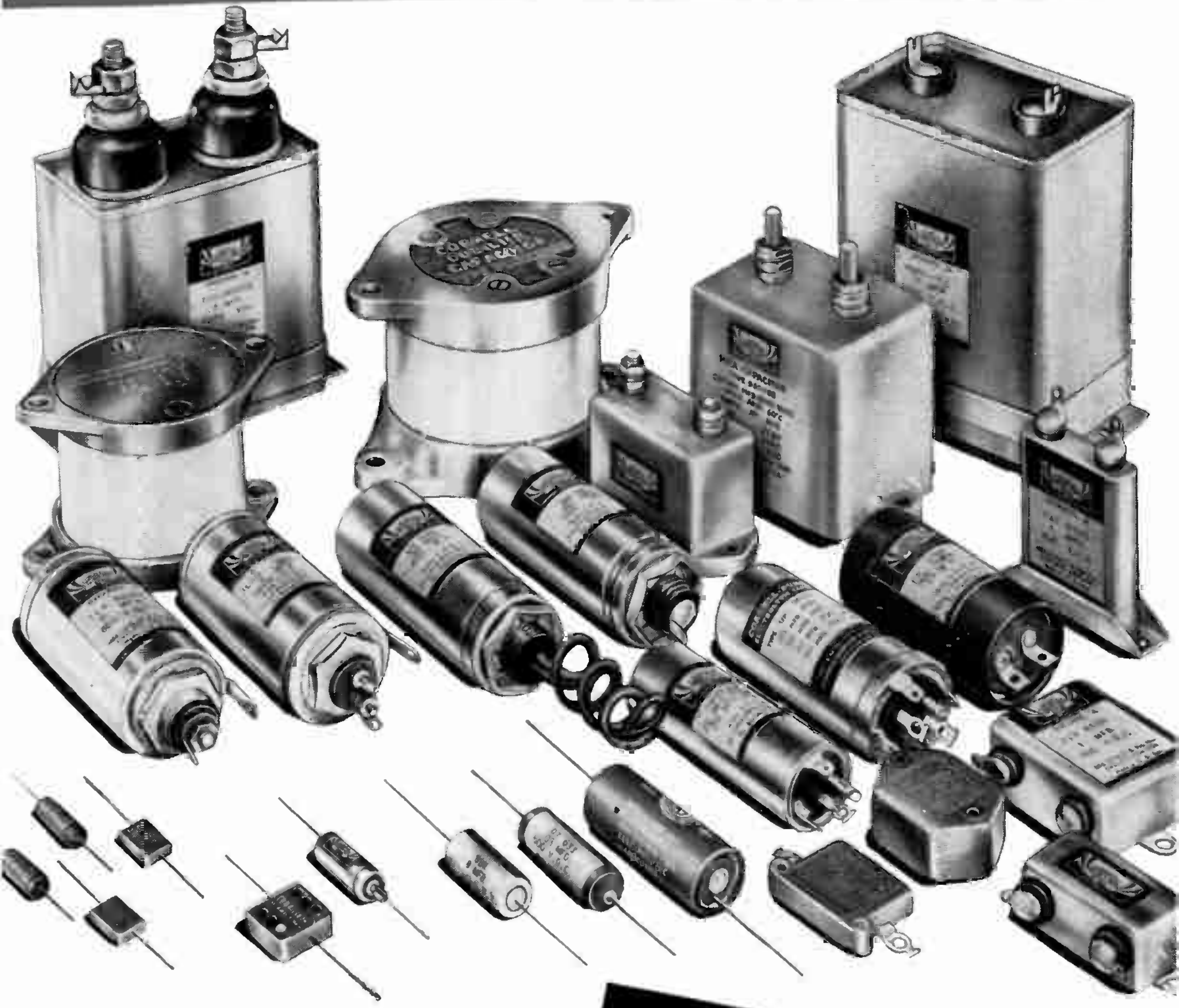

Learn Code the EASY Way

Beginners, Amateurs and Experts alike recommend the **INSTRUCTOGRAPH,** *to learn code and increase speed.*

Learning the INSTRUCTOGRAPH way will give you a decided advantage in qualifying for Amateur or Commercial examinations, and to increase your words per minute to the standard of an expert. The Government uses a machine in giving examinations.

Motor with adjustable speed and spacing of characters on tapes permit a speed range of from 3 to 40 words per minute. A large variety of tapes are available — elementary, words, messages, plain language and coded groups. Also an "Airways" series for those interested in Aviation.

MAY BE PURCHASED OR RENTED

The INSTRUCTOGRAPH is made in several models to suit your purse and all may be purchased on convenient monthly payments if desired. These machines may also be rented on very reasonable terms and if when renting should you decide to buy the equipment the first three months rental may be applied in full on the purchase price.

ACQUIRING THE CODE

It is a well-known fact that practice and practice alone constitutes ninety per cent of the entire effort necessary to "Acquire the Code," or, in other words, learn telegraphy either wire or wireless. The Instructograph supplies this ninety per cent. It takes the place of an expert operator in teaching the student. It will send slowly at first, and gradually faster and faster, until one is just naturally copying the fastest sending without conscious effort.

BOOK OF INSTRUCTIONS

Other than the practice afforded by the Instructograph, all that is required is well directed practice instruction, and that is just what the Instructograph's "Book of Instructions" does. It supplies the remaining ten per cent necessary to acquire the code. It directs one how to practice to the best advantage, and how to take advantage of the few "short cuts" known to experienced operators, that so materially assists in acquiring the code in the quickest possible time. Therefore, the Instructograph, the tapes, and the book of instructions is everything needed to acquire the code as well as it is possible to acquire it.

MACHINES FOR RENT OR SALE

The Instructograph

ACCOMPLISHES THESE PURPOSES:

FIRST: *It teaches you to receive telegraph symbols, words and messages.*

SECOND: *It teaches you to send perfectly.*

THIRD: *It increases your speed of sending and receiving after you have learned the code.*

With the Instructograph it is not necessary to impose on your friends. It is always ready and waiting for you. You are also free from Q.R.M. experienced in listening through your receiver. This machine is just as valuable to the licensed amateur for increasing his speed as to the beginner who wishes to obtain his amateur license.

Postal Card WILL BRING FULL PARTICULARS IMMEDIATELY

THE INSTRUCTOGRAPH CO.

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Henry Radio stores in Butler, Missouri and 11240 West Olympic Blvd., Los Angeles, California have complete stocks of amateur, FM, Television, Short Wave, Communications, Recording, and other radio equipment. I promise you lowest prices, complete stocks, quick delivery, easy terms, generous trade-ins. I promise that you will be satisfied on every detail. Write, wire, phone or visit either store today.

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Collins 75A	$ 375.00
Collins 32V	475.00
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National HRO-7T	292.50
National HRO-7C	372.45
National HFS	142.00
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Hallicrafters S38	49.95
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Hallicrafters S51	149.50
Hallicrafters S58	59.50
Hallicrafters S55	129.50
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Hallicrafters TS4	189.50
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RME HF-10-20	77.00
RME VHF-152A	86.60
RME DB22A	71.00
Hammarlund HQ129X	177.30
Gon-Set 10-11 converter	39.95
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Hunter Cyclemaster	169.50

Millen, Sonar, Bud, Gon-Set, Silver, Premax, Workshop, Amphenol-Mims; Jensen, Meissner, Browning; I have everything.

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LOW PRICES

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Hallicrafters, National, Hammarlund, Collins, Millen, RME, Meissner, Meck, Gordon, Amphenol-Mims, RCA, Vibroplexs, Sonar, all other amateur receivers, transmitters, beams, parts, etc. If it is amateur or communications equipment—I can supply it.

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Mail, phone, or wire your order. *Shipment at once.*

EASY TERMS

I have the world's best time sale plan because I finance the terms myself. I save you time and money. I cooperate with you. Write for details.

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TEN DAY FREE TRIAL

Try any receiver ten days, return it for full refund if not satisfied.

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I service everything I sell free for 90 days. At a reasonable price after 90 days.

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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. 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.

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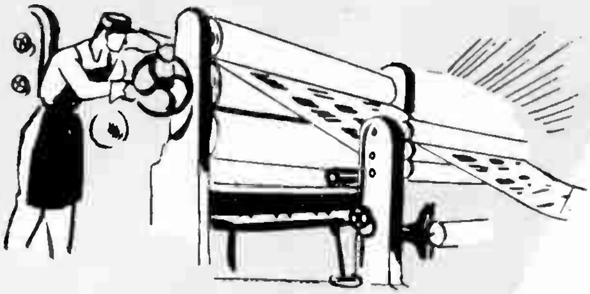


THE OFFICIAL MAGAZINE OF THE AMERICAN RADIO RELAY LEAGUE

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 $4.00 in the United States and Possessions, $4.50 in the Dominion of Canada, $5.00 in all other countries. Elsewhere in this book will be found an application blank for A.R.R.L. membership.

For thirty-four years (and thereby the oldest American radio magazine) ***QST*** has been the "bible" of Amateur Radio.

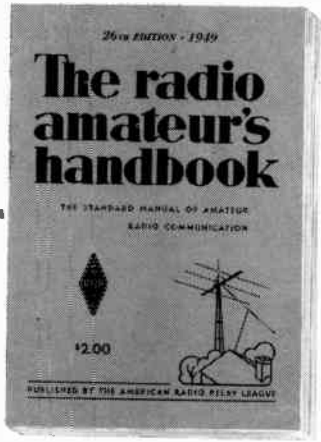
QST BINDERS

Price $2.50 each, postpaid

Available only in U. S. and Possessions

Those who take pride in the appearance of their bookshelf and wish to keep their reference file of *QST*'s in a presentable manner, appreciate the *QST* binder. It is stiff-covered, finished in beautiful and practical fabrikoid. Cleverly designed to take each issue as received and hold it firmly without mutilation, it permits removal of any desired issue without disturbing the rest of the file. It accommodates 12 copies of *QST*. Opens flat at any page of any issue.

With each Binder is furnished a sheet of gold and black gummed labels for years 1929 through 1949. The proper one can be cut from the sheet and pasted in the space provided for it on the back of the binder. The back copies of *QST* contain the record of development of modern amateur technique. They are invaluable as technical references. Some back copies are available—list will be sent upon request.



The *Handbook* tells the things which are needed for a comprehensive understanding of Amateur Radio. From the story of how Amateur Radio started through an outline of its wide scope of the present—from suggestions on how to learn the code through explanations of traffic-handling procedure and good operating practices—from electrical and radio fundamentals through the design, construction, and operation of amateur equipment—this book covers the subject thoroughly. It includes the latest and the best information on everything in Amateur Radio.

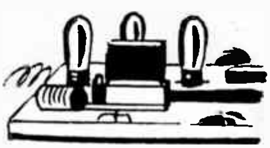
$2.00

U.S., Its Possessions and Canada

$2.50 Elsewhere

Buckram Bound
$3.00

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.



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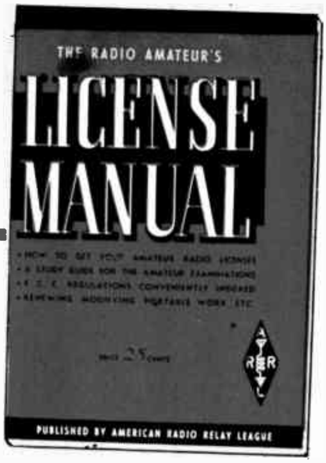
Postpaid Anywhere



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.

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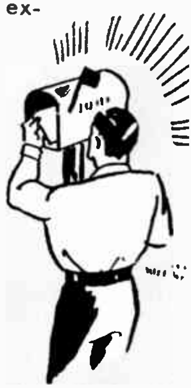

This booklet is designed to train students to handle code skillfully and with precision, both in sending and in receiving. Employing a novel system of code-learning based on the accepted method of *sound* conception, it is particularly excellent for the student who does not have the continuous help of 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

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Lightning Calculators

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.

TYPE A

Radio Calculator

; calculator is useful for the problems that confront the ateur every time he builds a new rig or rebuilds an old one winds a coil or designs a circuit. It has two scales for physical iensions of coils from one-half inch to five and one-half inches diameter and from one-quorter to ten inches in length; a quency scale from 400 kilocycles through 150 megacycles; vavelength scale from two to 600 meters; a capacity scale m 3 to 1,000 micro-microfarads; two inductance scales with ange of from one microhenry through 1,500; a turns-per-inch le to cover enameled or single silk covered wire from 12 to gauge, double silk or cotton covered from 0 to 36 and ible cotton covered from 2 to 36. Using these scales in the ple manner outlined in the instructions on the back of the culator, it is possible to solve problems involving frequency kilocycles, wavelength in meters, inductance in microhenrys d capacity in microfarads, for practically all problems that amateur will have in designing—from high-powered transters down to simple receivers. Gives the direct reading swers for these problems with accuracy well within the erances of practical construction. **$1.00** *Postpaid.*

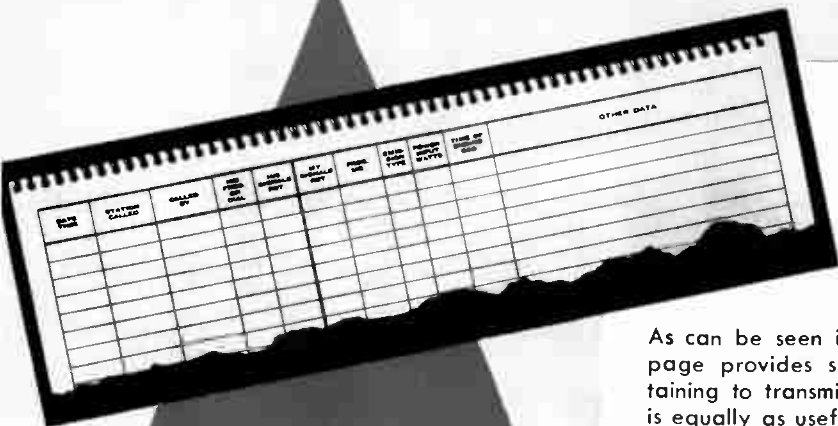
TYPE B

Ohm's Law Calculator

This calculator has four scales: a power scale from 10 microwatts through 10 kilowatts, a resistance scale from .01 ohm 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 scales, 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.*

STATION OPERATING *Sup*



LOG BOOK

As can be seen in the illustration, the log page provides spaces for all facts pertaining to transmission and reception, and is equally as useful for portable or mobile operation as it is for fixed. The log pages with an equal number of blank pages for notes and a sheet of graph paper are spiral bound, permitting the book to be folded back flat at any page, requiring only the page size of 8½ x 11 on the operating table. In addition, a number sheet, with A.R.R.L. Numbered Texts printed on back, for traffic handlers, is included with each book.

50c per book

Official Radiogram Forms

The radiogram blank is designed to comply with the proper order of transmission. All blocks for fill-in are properly spaced for use in typewriter. It has a heading that you will like. Radiogram blanks, 8½ x 7¼, lithographed in green ink, and padded 100 blanks to the pad, 25c per pad, postpaid.

Message Delivery Cards

Radiogram delivery cards embody the same design as the radiogram blank and are available in two styles—on stamped government postcard, 2c each; unstamped, 1c each.

The operating supplies shown on this page have been designed by the A.R.R.L. Communications Department.

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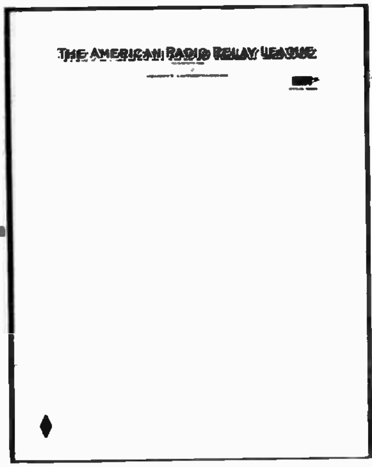
MEMBERSHIP



Members' Stationery

Members' stationery is lithographed on standard 8½ x 11 bond paper which every member should be proud to use for his radio correspondence.

100 Sheets, **$1.00** 250 Sheets, **$1.50**

500 Sheets

$2.50

postpaid

In the January, 1920 issue of ***QST*** there appeared an editorial requesting suggestions for the design of an A.R.R.L. emblem—a device whereby every amateur could know his brother amateur when they met. In the July, 1920 issue the design was announced—the familiar diamond that greets you everywhere in Ham Radio. For years it has been the unchallenged emblem of amateur radio.

The League Emblem, with gold border and lettering, and with black enamel background, is available in either pin (with safety clasp) or screw-back type. In addition, there are special colors for Communications Department appointees. • Red enameled background for the SCM. • Blue enameled background for the ORS or OPS.

50c each *postpaid*

The Emblem Cut. A mounted printing electrotype, ⅝″ high, for use by members on amateur printed matter, letterheads, cards, etc.

$1.00

each

postpaid

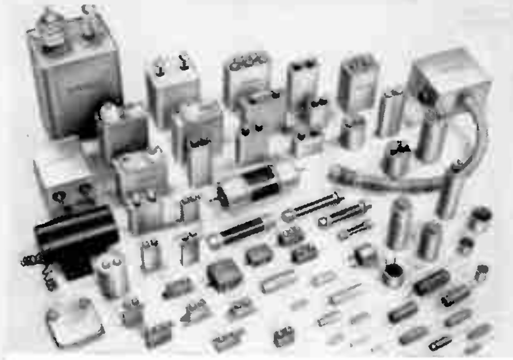
Here's Capacitor Dependability!

Shown here are a few of more than 9675 capacitor and *Koolohm resistor types that Sprague produces every year. Many of these are for critical industrial applications, others for national defense and ultra-exacting scientific needs.

From this vast array come the capacitors that are carefully selected for amateur radio uses—types that mean more for your money because they're better engineered, built more dependably. Catalog on request.

DRY ELECTROLYTICS

Sprague offers the most diversified dry electrolytic capacitor line ever presented for standard distributor stock. Tiny "Atom" midgets; self-mounting multi-section units; high-capacity, low-voltage tubulars; rectangular and cylindrical shapes; lug, bracket and self-mounting types; terminals and lead connections and many others!

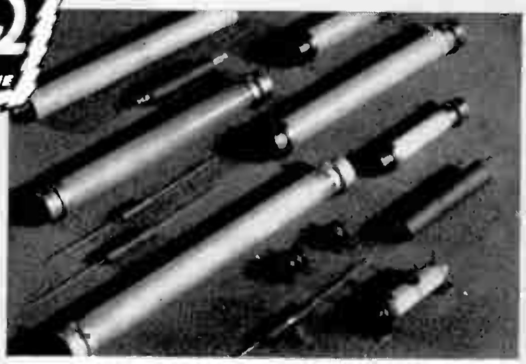


PAPER DIELECTRICS

Standard Sprague paper dielectric capacitors for amateur use include 15 types and over 250 items. Chief among them are three small, popularly priced transmitting types that are both filled and impregnated with *KVO, the exclusive Sprague dielectric. And don't forget the TC Tubular By-pass types—"Not a failure in a million!"

MICA DIELECTRICS

Sprague distributors carry complete stocks of popular mica capacitors including all needed capacities and voltage ratings—in sizes from "postage stamp" silvered micas to high-voltage ceramic-jacketed units. All provide maximum quality for R-F applications where low power factor and high insulation resistance at high frequencies are essential.



*KOOLOHM RESISTORS

Sprague Koolohm Resistors are wound with wire insulated before winding with a flexible ceramic coating that is impervious to heat as high as 1000° C. Doubly protected by glazed ceramic shells and moisture resistant seals. Insulated for 10,000 volts resistance breakdown to ground. Larger, sturdier wire sizes in smaller resistors. Use Koolohms at full wattage ratings—anywhere!

*Trademarks Reg. U. S. Pat. Off.

SPRAGUE

PIONEERS OF ELECTRONIC AND ELECTRICAL PROGRESS

SPRAGUE PRODUCTS COMPANY, NORTH ADAMS, MASS.

Jobbing distributing organization for products of Sprague Electric Co.

THE "GRID-DIP" METER

BECOMES A

Laboratory Standard

MEASUREMENTS CORPORATION

MEGACYCLE METER

Model 59

- For the determination of the resonant frequency of tuned circuits, antennas, transmission lines, by-pass condensers, chokes or any resonant circuit.
- For measuring capacitance, inductance, Q, mutual inductance.
- For preliminary tracking and alignment of receivers.
- As an auxiliary signal generator; modulated or unmodulated.
- For antenna tuning and transmitter neutralizing, power off.
- For locating parasitic circuits and spurious resonances.
- As a low sensitivity receiver for signal tracing.

The Model 59 provides the amateur, service man, technician or engineer with a versatile instrument that embodies the same expert engineering skill and precision manufacturing that has made Measurements Corporation instruments a standard of accuracy in laboratories the world over.

TELEVISION

The Megacycle Meter has many applications in the construction and servicing of television receivers. It can be used for aligning video amplifiers, peaking coils, sound traps, filters, stagger tuned i.f.s, stagger tuned amplifiers, sound i.f.s, local oscillators, carrier circuits, grid mixing circuits, etc.

SPECIFICATIONS:

FREQUENCY:
2.2 Mc. to 400 Mc.; seven plug-in coils.

MODULATION:
CW or 120 cycles; or external.

DIMENSIONS:
Power Unit, 5⅛" wide; 6⅛" high; 7½" deep. Oscillator Unit, 3¾" diameter; 2" deep.

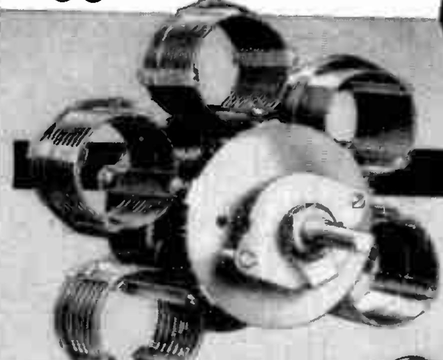
POWER SUPPLY:
110–120 volts, 50–60 cycles; 20 watts.

CIRCULAR ON REQUEST

MANUFACTURERS OF
Standard Signal Generators
Pulse Generators
FM Signal Generators
Square Wave Generators
Vacuum Tube Voltmeters
UHF Radio Noise & Field Strength Meters
Capacity Bridges
Megohm Meters
Phase Sequence Indicators
Television and FM Test Equipment

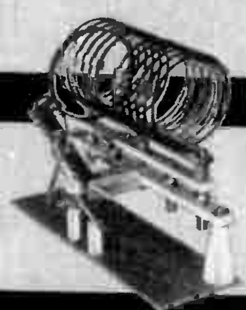
MEASUREMENTS CORPORATION
BOONTON NEW JERSEY

B&W PARTS an



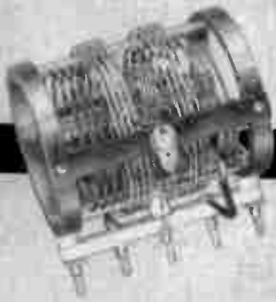
B & W BAND SWITCHING TURRETS

Pioneered by B & W—These compact 5-band switching Turrets are available from 35-watt ratings up through units that may be operated at voltages up to 1000 volts and input powers as high as 150 watts. They provide instant band switching, covering the 80 to 10 meter bands, and are regularly stocked as center-linked, center-tapped coils or end-linked, untapped.



B & W TYPE HD INDUCTORS

Rated up to 1000 Watts Input—Three general types available: without link, fixed center link with center tap, and variable center link with center tap. Type HD coils are ruggedly built, reasonably priced and are typical of the many B & W coils available for specific applications.

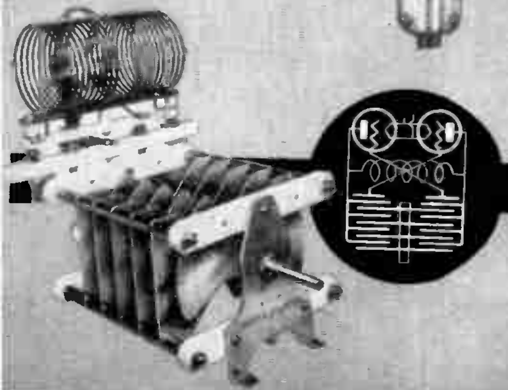


B & W 3400 SERIES INDUCTORS

With Fixed and Adjustable Coupling—Designed for those who want the utmost in sturdy construction and electrical flexibility in coils handling up to 500 watts. These famous coils pioneered individual adjustable link construction, thus providing precise impedance matching up to 600 ohms.

B & W COAXIAL CONNECTOR CC-50

Provides efficient, watertight coaxial cable connections for amateur and commercial use. Also serves as a center insulator. Made of aluminum with steatite insulation, this unit comes complete with weatherproof cement and assembly screws. Weight 12 oz., pull strength 500 lbs.



B & W HEAVY-DUTY VARIABLE CAPACITORS

Type CX is a radically designed split stator, butterfly rotor variable capacitor that permits mounting the tank coil assembly directly on the capacitor frame as illustrated. Opposed stator sections provide short R-F paths desirable in high power rigs. Built-in neutralizing capacitors are provided for, on rear end plate.

B & W PLUG-IN LINKS

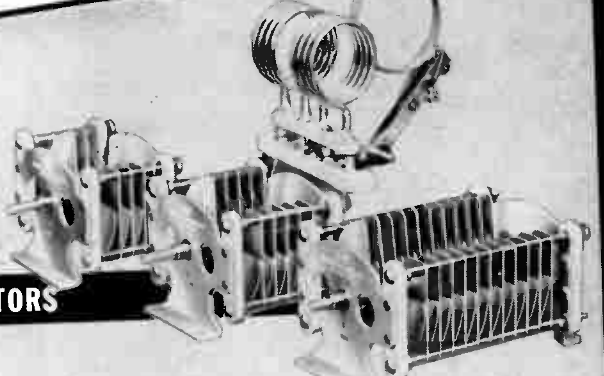
For impedance matching, just plug in the proper link. These new B & W plug-in links make your rig adaptable to practically any impedance as quickly as you can plug in a link with the correct number of turns. Type 3750 mounting bar for HDV coils, Type 3550 for TVH-TVL-BVL coils. Links available in 1-3-6 and 10 turns.

BARKER & WILLIAMSON, Inc.

EQUIPMENT

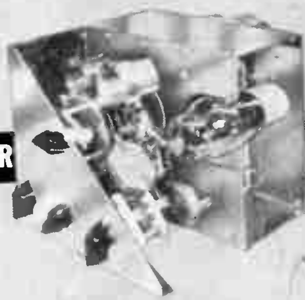
PIONEERED, DESIGNED AND BUILT FOR EXACTING ELECTRONIC ENGINEERS AND EXPERIMENTERS

In addition to the B & W products shown here, there are dozens of others in our general catalog. All are made under the direct supervision of men who know amateur radio requirements personally. And all are produced to the high quality and design standards that are characteristic of B & W equipment.



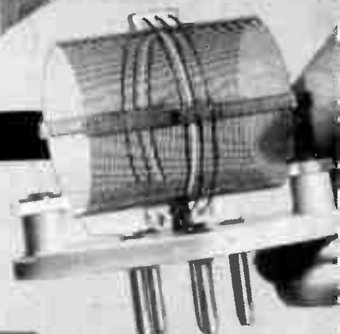
B & W BUTTERFLY VARIABLE CAPACITORS

Type JCX small butterfly variable capacitors offer all the electrical and mechanical features of the larger B & W units, and are ideally small in size for general ham and other uses. They accommodate B & W "B" or "BX" series coils and offer many advantages over conventional type units.



B&W ALL-BRAND FREQUENCY MULTIPLIER

Model 504—A fixed-tuned, broad-band frequency multiplier designed for use with either a V.F.O. or Crystal input. Makes transmission on any band available at the flip of a switch.



B & W "BABY" AIR INDUCTORS

25 Watts Rating—Ideal for crowded layouts, portables and any other application where space is at a premium and high efficiency a "must." Many other types and sizes available. All offer famous B & W "air wound" construction.

B & W TEST INSTRUMENTS

Accurate—Inexpensive—Reliable

AUDIO OSCILLATOR—*Model 200*—An extremely low distortion source of frequencies between 30 and 30,000 cycles.

DISTORTION METER—*Model 400*—Measures total harmonic distortion for the range of 50 to 15,000 cycles.

SINE WAVE CLIPPER—*Model 250*—Provides test signal particularly useful in examining the phase angle, transient and frequency response of audio circuits.

FREQUENCY METER—*Model 300*—An accurate and convenient means of making direct measurements of unknown audio frequencies up to 30,000 cycles.

WRITE FOR COMPLETE B & W INSTRUMENT CATALOG!

Get this Handy Catalog

for full details on inductors, variable capacitors and accessories every ham needs for his rig.

... FIELD AVE. UPPER DARBY, PA.

For Long Life and Dependability

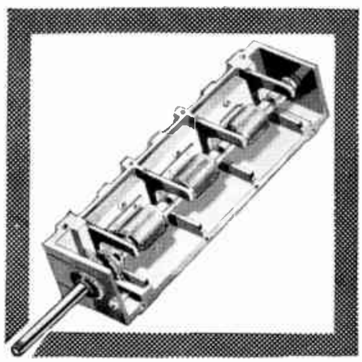
USE MALLORY *APPROVED* PRECISION PRODUCTS

SWITCHES . . . The Mallory line of rotary selector switches, lever action switches, and push button switches provides a complete selection to meet every demand.

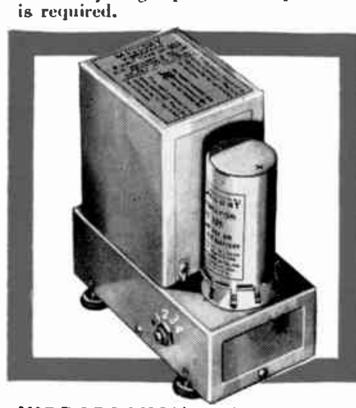
CAPACITORS . . . Proved long life is the well earned reputation of Mallory capacitors. Mallory offers a full line of electrolytic, oil, and wax-and-oil impregnated capacitors.

RESISTORS . . . Mallory Vitreous enamel resistors are used as voltage dividers, and as dropping, load, and shunt resistors in all circuits where relatively high power dissipation is required.

CONTROLS . . . Mallory controls, including the famous 15/16" Mallory Midgetrol are built for hard use and have characteristics that insure complete satisfaction and dependable performance over long life.



INDUCTUNERS* . . . To provide infinitely variable inductance tuning over all frequencies within the range of 44 to 216 megacycles.



VIBRAPACKS** . . . A completely dependable source of high voltage where commercial AC is not available.

After putting lots of time and expense into your rig, you still have the final and sometimes difficult problem of adjusting it. The fewer variables in the component parts used, the smaller the problem. You can depend on consistent performance from Mallory Approved Precision Parts. And you can depend on good service when you see your Mallory Distributor.

**Registered trademark of P. R. Mallory & Co., Inc., for inductance tuning devices covered by Mallory-Ware patents.*

"INSIST ON MALLORY"

***Reg. U. S. Pat. Off.*

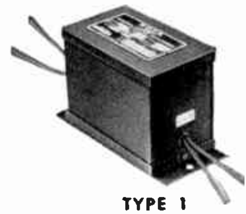
P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA

TYPE 11

SMALL, LOW-COST, SOLA CONSTANT VOLTAGE TRANSFORMERS FOR CHASSIS MOUNTING

Reliable communications equipment must have stabilized voltage—and the right place to provide for it is in the equipment itself. These three types of SOLA Constant Voltage Transformers have been specifically designed for "built-in" applications. They are low in cost and their use will often permit the elimination of other components. For complete information consult Bulletin 34CV-102, available on request.

TYPE 12



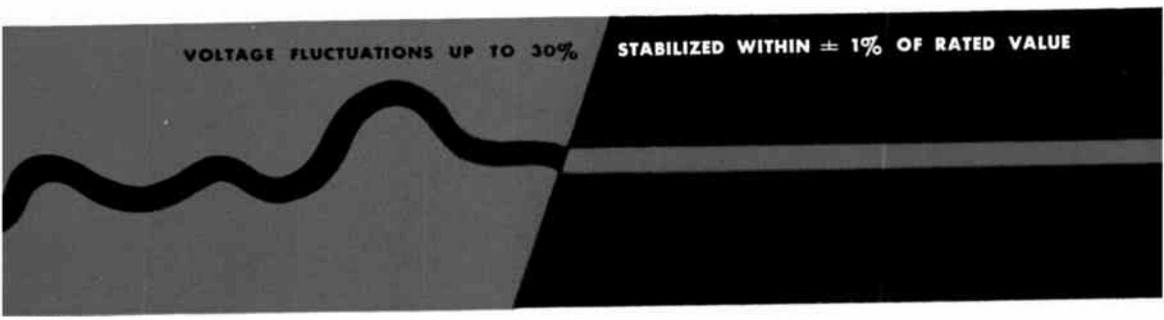
TYPE 1

	Catalog Number	Output Capacity in VA	Input Volts	Output Volts	Dimensions in Inches A	B	C	E	F	Approx. Shipping Weight	List Price Each
YPE 1	30488	15	95-125	6.0	5 11/16	2 5/8	3 7/16	5 1/16		6	$15.00
	30492	15	95-125	6.3	5 11/16	2 5/8	3 7/16	5 1/16		6	15.00
	30498	15	95-125	115.0	5 11/16	2 5/8	3 7/16	5 1/16		6	15.00
YPE 11	30785	17	95-125	6.3	5 13/16	3 21/32	2 19/32	3	2	5½	20.00
	30955	17	95-125	115.0	5 13/16	3 21/32	2 19/32	3	2	5½	20.00
YPE 12	301002	15	95-125	6.3	5 5/16	3½	2¼	3	1½	2½	18.50
	301003	15	95-125	115.0	5 5/16	3½	2¼	3	1½	2½	18.50

*Condenser supplied as separate unit.

DIMENSIONS:
A: Overall Length
B: Overall Width
C: Overall Height
E & F: Mounting Dimensions

Prices subject to chang without notice.

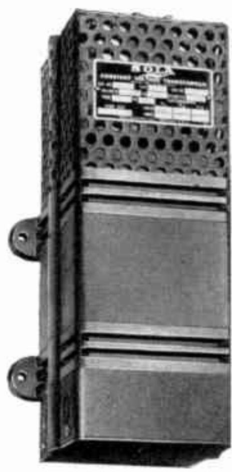


FOR COMMUNICATIONS EQUIPMENT NOW IN SERVICE

Where provision for constant voltage protection has not been made within the equipment itself, these standard SOLA Constant Voltage Transformers can be easily installed. They require no supervision or maintenance, are instantaneous in operation and they protect both themselves and the equipment against short-circuit. Other capacities ranging from 10VA to 15KVA fully described in Bulletin 34CV-102, available on request.

TYPE 2

	Catalog Number	Output Capacity in VA	Input Volts	Output Volts	Dimensions in Inches A	B	C	E	F	Approx. Shipping Weight	List Price Each
YPE 2	30804	30	95-125	115.0	8 9/16	4 3/16	4 3/8	7 15/16	2 3/8	12	$17.00
	30805	60	95-125	115.0	8 13/16	4 3/16	4 3/8	8 1/16	2 3/8	13	24.00
	30806	120	95-125	115.0	9 11/16	4 3/16	4 3/8	8 15/16	2 3/8	17	32.00
YPE 3	30807	250	95-125	115.0	11 5/8	6 15/16	5 5/8	3¼	6 1/8	30	52.00
	30M807	250	190-250	115.0	11 5/8	6 15/16	5 5/8	3¼	6 1/8	30	52.00
	30808	500	95-125	115.0	14½	6 15/16	5 5/8	5	6 1/8	40	75.00
	30M808	500	190-250	115.0	14½	6 15/16	5 5/8	5	6 1/8	40	75.00



TYPE 3

SOLA Constant Voltage Transformers

SOLA ELECTRIC COMPANY, 4633 WEST 16TH STREET, CHICAGO 50, ILLINOIS

PAR-METAL *Standard* RACKS, CABINETS CHASSIS, PANELS

ADAPTABLE FOR EVERY REQUIREMENT . . .

. . . from a Small Receiver to a DeLuxe Transmitter

FREE 28-PAGE CATALOG — No. 48

A 28-page descriptive catalog is available which has **a complete listing** of: Type "C" Cabinet Racks for **19″** Rack Panels, Type "C" Transmitter Racks for **19″**, **24″** and **30″** Rack Panels. Type "A" Relay Racks for **19″** Rack Panels.

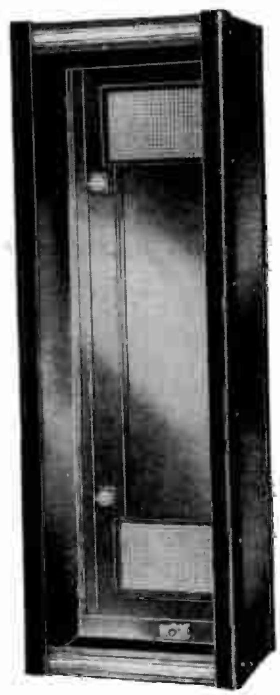
Also—Utility Desk Cabinets, Desk Panel Cabinet Racks, **Amplifier Foundation Chassis, Chassis Bases, 19″** Rack Panels, Rack Panel Assemblies, Cabinets, Cases, Shelves and Roller Trucks.

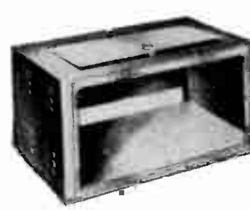
WRITE TODAY for your copy of Catalog No. 48.

Years of experience in building metal equipment for electronic products has enabled us to develop the units listed in our Catalog No. 48.

Par-Metal products represent a **combination** of **skill** and **material**—a **fusion** of **careful** workmanship and mechanical **efficiency** with furniture steel. Such products are dependable and will give you years of satisfactory service at a **reasonable cost.**

The use of **PAR-METAL STANDARDIZED** units enables you to build electronic equipment that is professional both in construction and appearance. These units are being extensively used for **Amateur** equipment (also for Commercial, Marine, Airline, and Broadcast equipment both here and abroad.





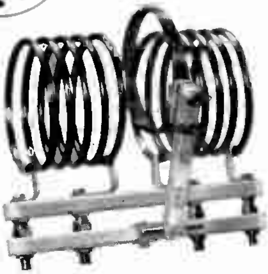
PAR-METAL PRODUCTS CORPORATION

32-62 49th Street, Long Island City 3, N. Y.

EXPORT DEPT.: ROCKE INT. CORP., 13 E. 40th St., New York 16, N. Y.

GREATER EFFICIENCY WITH

NEW JOHNSON HAM INDUCTORS



Coil windings are a wire size larger than on most available inductors.

A COIL TO MATCH YOUR TUBE — A LINK TO MATCH YOUR LINE

JOHNSON's new and comprehensive line of inductors and swinging link assemblies now make it possible for the ham to enjoy commercial efficiency. There are two models for each band for use with either high voltage low current, or low voltage high current tubes.

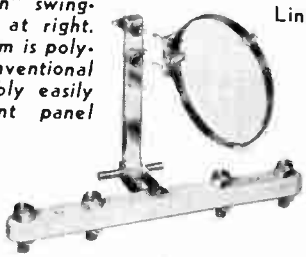
With these new JOHNSON Ham Inductors and "plug-in" Swinging Link Assemblies you can instantly match coil to tube—link to line. These outstanding inductors are also available in semi-fixed models.

HEAVIER WINDINGS ON ALL MODELS

Efficiency is further increased because coil windings are a wire size larger than on most available inductors—resulting in less heating, lower loss and consequently higher efficiency.

Remember, too, that the new JOHNSON Inductors and "plug-in" Link Assemblies fit all conventional inductor assemblies.

Principle of "plug-in" swinging link illustrated at right. The swinging link arm is polystyrene — not conventional plastic. Link assembly easily connected for front panel control.



FREE BOOKLET

A new JOHNSON "Air Wound Ham Inductor Catalog", containing information and tables which will enable you to select the correct inductor, link or links for your individual application, is yours for the asking. Write for it today.

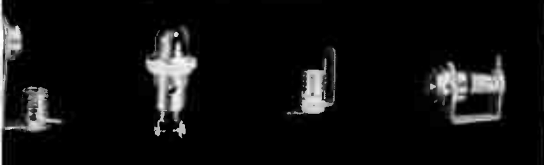
INSULATORS

JOHNSON insulators are specifically designed to handle high RF with low loss. They possess, in addition, logical proportions, clean-cut accurate molding, and high grade nickel plated brass hardware with milled — not stamped — nuts. The Johnson line includes stand-off, cone, thru-panel, antenna, feeder and strain insulators.

SPEED X TRANSMITTING KEYS

The Speed X line, long a leader in its field, is now manufactured by JOHNSON. It includes everything from buzzers to high speed semi-automatic keys. Pictured are the hand key, Model 326, and beautiful chrome finish, new and improved Model 501 semi-automatic. Model 501, Amateur Model 515 and Junior 510 also available in left hand models.

PILOT LIGHTS



To round out its line, JOHNSON purchased the entire Gothard line of fine pilot lights. The Gothard line is a complete line and will be maintained to provide a wide choice and permit selection of a light which will more exactly meet your needs. All metal parts are brass with the exception of hex nuts. Parts are heavily plated and jewel holders are polished chrome or nickel.

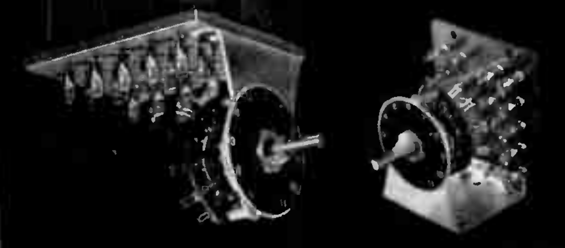
TUBE SOCKETS

JOHNSON Tube Sockets have consistently led the way to better design for better results. Present day demands for ever better radio-electronic circuits and equipment are more than adequately met with JOHNSON Tube Sockets. Superior in mechanical and electrical design, JOHNSON Tube Sockets are available in both standard and "special" sizes.

NEW CABINETS, RADIO PANELS AND CHASSIS

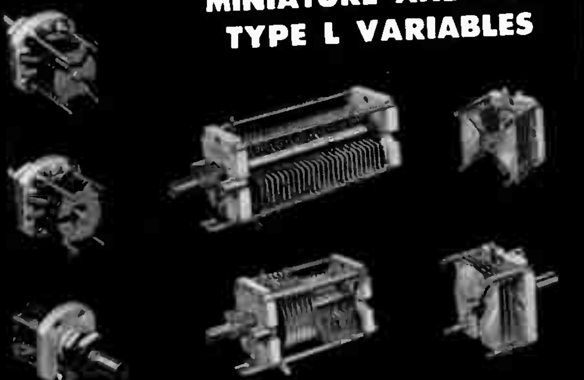
The skill of JOHNSON in building cabinets for its Phasing and Antenna Coupling Equipment is now directed to mass production of cabinets, racks, panels and chassis. They are professional in appearance, characteristically reasonable in price. A unique feature is the ventilation system which permits units to be placed flush side-by-side. Chassis have a new type flush joint which eliminates sharp and protruding edges.

NEW INSTANT CRYSTAL SELECTOR



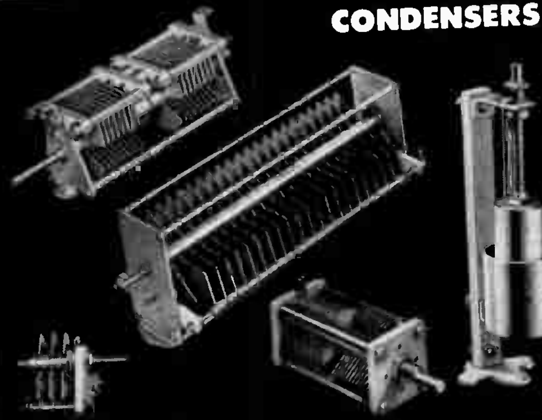
QSY with the speed of an ECO and still enjoy all the advantages of crystal control! The new JOHNSON Instant Crystal Selector gives you ten frequencies with a twist of the knob, accommodates all crystals with ½" spacing. With adaptors you can also use up to six of your upright ¾" spaced crystals plus four with ½" spacing. Extra position on switch for ECO.

MINIATURE AND NEW TYPE L VARIABLES



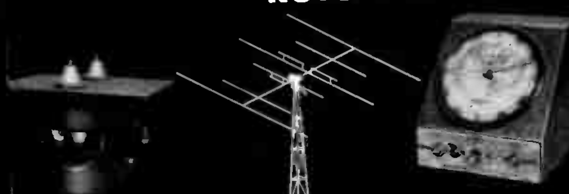
JOHNSON offers you three of the smallest air variables ever built! Pictured above, left, they are available in Single, Butterfly and Differential types. JOHNSON'S new line of Type L Variables, above right, are ceramic soldered for permanent maintenance of capacities. No screws or eyelets to come loose.

CONDENSERS



Precision engineered for peak performance and durability! JOHNSON makes a condenser for every stage of the amateur transmitter. The exacting requirements of amateur, commercial broadcast and industrial operation are fully met by JOHNSON Condensers.

ROTOMATIC BEAM



The JOHNSON Rotomatic Antenna Array is a perfect solution to QRM on the crowded DX bands! Strong, yet light, it has broad band characteristics—provides tremendous increase in signal strength. Heavy duty drive, or may be purchased without motor for hand drive. 360° rotation. The combined direction indicator, with great circle map and beam control, is a marvel of operating efficiency!

JOHNSON Q ANTENNA

High efficiency! That describes JOHNSON'S Q antenna system. Applications include half-wave doublet, either horizontal or vertical, harmonic or "longwire" radiator, radiator reflector, radiator director, "V" beam, JOHNSON Q Beam, and others. Q beam consists of two half-wave Q antennas spaced 1-5 wave.

See JOHNSON products at your jobbers — or write for latest JOHNSON Catalog.

JOHNSON . . . *a famous name in Radio!*

E. F. JOHNSON CO., WASECA, MINNESOTA


UTC
COMPONENTS FOR EVERY APPLICATION
LINEAR STANDARD
High Fidelity Ideal
HIPERM ALLOY
High Fidelity . . . Compact
ULTRA COMPACT
Portable . . . High Fidelity
OUNCER
Wide Range . . . 1 ounce
SUB OUNCER
Weight 1/3 ounce
COMMERCIAL GRADE
Industrial Dependability
SPECIAL SERIES
Quality for the "Ham"
POWER COMPONENTS
Rugged . . . Dependable
VARITRAN
Voltage Adjustors
MODULATION UNITS
One watt to 100KW
VARIABLE INDUCTOR
Adjust like a Trimmer
TOROID HIGH Q COILS
Accuracy . . . Stability
TOROID FILTERS
Any type to 300KC
MU-CORE FILTERS
Any type 1/2 – 10,000 cyc.
EQUALIZERS
Broadcast & Sound
PULSE TRANSFORMERS
For all Services
SATURABLE REACTORS
Power or Phase Control
LARGE UNITS
To 100KW Broadcast
PLUG-IN TYPE
Quick change service
CABLE TYPE
For mike cable line
VERTICAL SHELLS
Husky . . . Inexpensive
REPLACEMENT
Universal Mounting
STEP-DOWN
Up to 2500W . . . Stock
LINE ADJUSTORS
Match any line voltage
CHANNEL FRAME
Simple Low cost
United Transformer Co.
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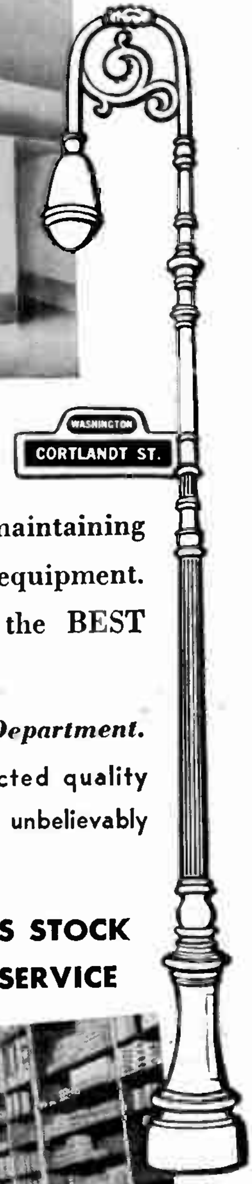
CABLES "ARLAB"

the House on Cortlandt Street

SHOP WITH CONFIDENCE



Terminal has kept its lead in New York by maintaining a complete stock of quality radio parts and equipment. Fair prices and friendly service assure you of the BEST VALUES at TERMINAL.

Visit our "SPECIALS" Department.

Countless values of selected quality parts and equipment at unbelievably low prices!

OUR TREMENDOUS STOCK MEANS FASTER SERVICE

TERMINAL RADIO CORP.

85 CORTLANDT STREET, NEW YORK 7, N. Y.

PHONE: WOrth 4-3311

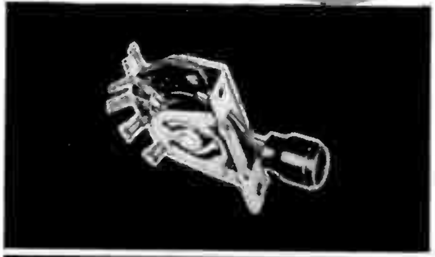
Look to Centralab

CRL

Division of Globe-Union Inc., Milwaukee, Wis.

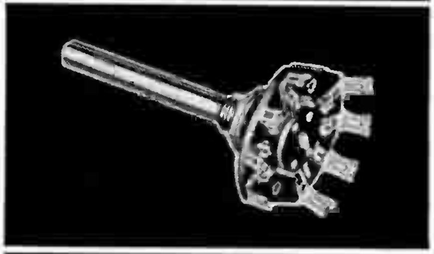
SWITCHES

LEVER SWITCHES: features coil spring mechanism with index spring, replaceable without removal of switch from chassis. Life test — 50,000 cycles. 8 basic combinations of indexing available. Shorting or non-shorting contacts.



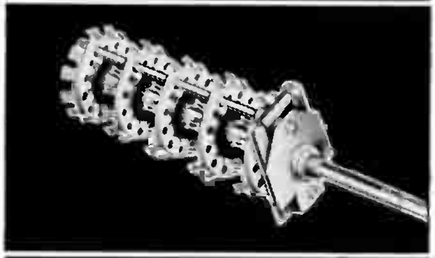
"F" INDEX: an inexpensive, single or two-section type switch for radio-phono operations, simple band change or general switching applications. 30° indexing. 2-12 position fixed stops. Life test through five positions, 10,000 cycles.

TONE SWITCHES: 3-4-6-8-9 or 10 clips available in tone switch group. 8 clip type shown. All rated at 6 watts (1 amp.-6 v.). Shorting or non-shorting contacts. 3-6-10 clip type can be supplied with spring return. Minimum life operation 10,000 cycles.



"H" INDEX: primarily for band change and general tap switch applications. Positive action. Life test through 5 positions, 10,000 cycles. Spring and ball index mechanism. Index springs available in different thicknesses to produce desired turning torque.

"G" INDEX: ideal for industrial applications. Provides positive, stall-proof indexing in 30° and 60° positioning. 2-12 positions, 30° index. 1 to 6 positions, 60° index. Life test through 5 positions, 250,000 cycles.



POWER SWITCH: for transmitters, power supply converters, X-ray equipment, etc. Efficient performance up to 20 megacycles. 7½ amp. at 60 cycles, 115 volts AC. Minimum voltage breakdown between critical points — 3,000 volts RMS, 60 cycles. Life tests — 25,000 cycles.

CON

"R" RADIOHMS: two types: 1) *wire-wound*, rated at 3 watts, wound in linear taper only. 20-10,000 ohms resistance; 2) *composition*, rated at 1 watt with 7 resistance tapers. Non-rubbing contact. Low noise level.

"E" RADIOHMS: composition type only. 6 different resistance tapers available. Resistance from 10,000 ohms to 3 megohms. Current rating ¼ watt. 37/64" in diameter. 17/32" deep behind panel. 25/32" deep behind panel with AC line switch.

"M" RADIOHMS: Composition type. 7 resistance tapers. Resistance from 500 ohms to 10 megohms. Current rating ½ watt. 1 3/64" in diameter. 23/32" deep with AC line switch. Can be twinned, tripled, twinned with switch, or twinned with concentric shafts.

SWITCH COVERS: five types for "R" Radiohms, 4 types for "M" Radiohms, 1 type for "E" Radiohms. Minimum life of 50,000 mechanical and electrical operations. underwriters approved. Rated at 3 amp. 125 volts; 1 amp. 250 volts.

"T" RADIOHMS: no larger than a dime, specially designed for hearing aids, pocket radio receivers, miniature amplifiers. 6 resistance tapers. 500 ohms to 5 megohms resistance. Diameter .625", less knob. Knob diameter 27/32". Rating 1/10 watt. Exceptionally low noise level.

RHEOSTATS: for small motor speed controls, charging rate adjusters, or soldering iron temperature regulators, etc. All metal frame and core, insulated with treated asbestos. Two sizes: 25-watt, ¾" deep behind panel. 50 watt, 1¼" deep behind panel. Diameter of both 2 inches.

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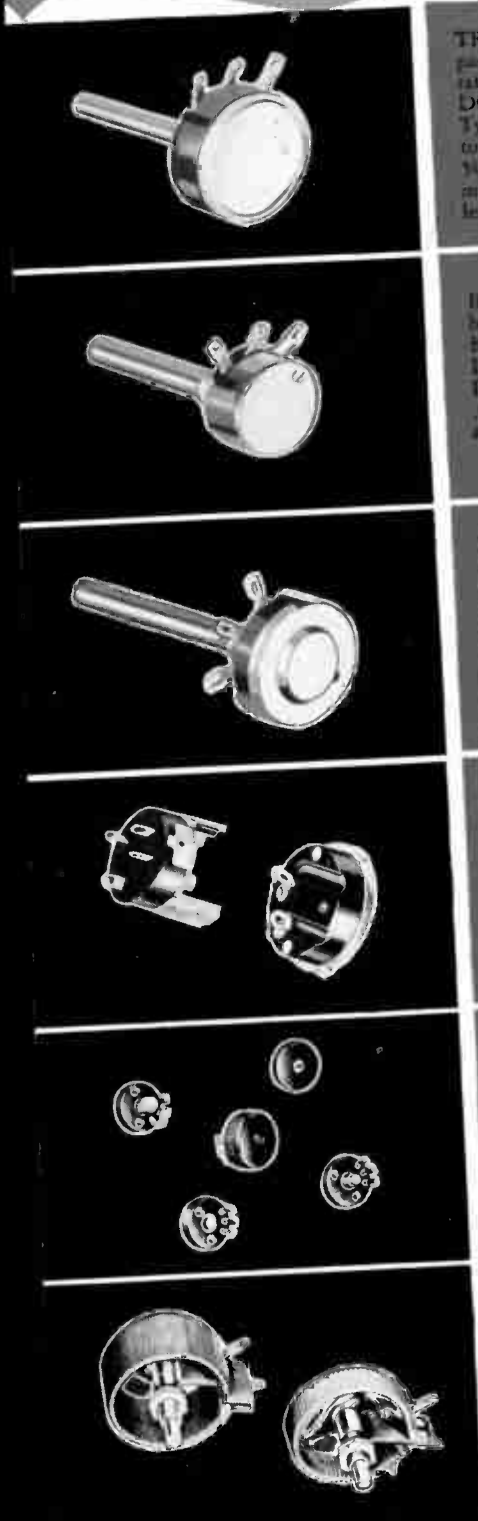
TROLS

CAPACITORS



TRIMMERS: ceramic. Full capacity change within 120° rotation. Working voltages, 500 DC. Flash test, 1100 volts DC. Type 820 — 3 ranges from 2.6 to 35 mmf. Type 822 — 2 to 50 mmf. Type 823 — 5 to 125 mmf. Power factor of all — less than 0.2% at 1 megacycle.

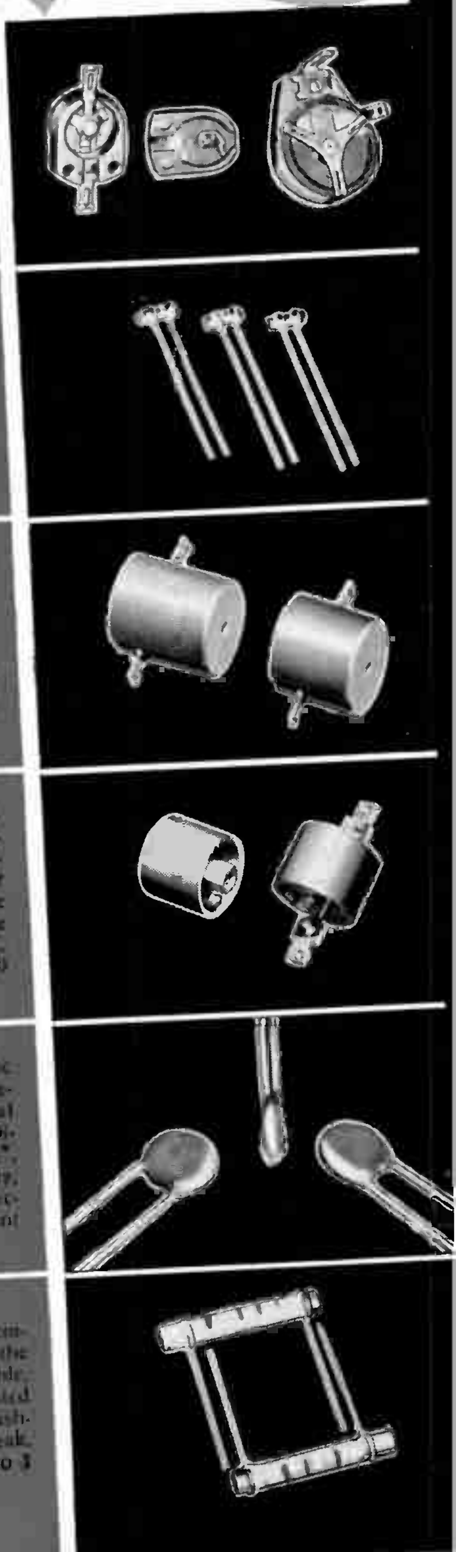
DC TUBULARS: ceramic, for bypass coupling applications. Basic working voltages of 500 DC. Flash tested, 1000 volts DC. Capacity range, from .000010 to .01 mfd. Individually color coded.

HIGH ACCURACY: pure silver electrodes fired on high dielectric constant Ceramic-X. For rigid frequency control applications. Individually tested. Temperature coefficient is maintained with ± 10 parts per million. Standard working voltage, 500 volts DC.

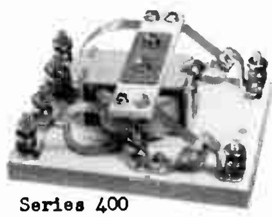
HIGH VOLTAGE: low power factor, stable retrace characteristics and maximum flashover. Plates are pure silver fired to ceramic surface. Five sizes with working voltage from 5000 to 15,000 volts DC. Flashover voltage from 10,000 to 30,000 volts DC.

DISC: miniature ceramic disc capacitor combining utmost reliability with small physical size and low mass weight. Diameter ⅝". Thickness ⅛". Guaranteed minimum capacity, .005 mmf. Pure silver electrodes bonded to permanent Ceramic-X.

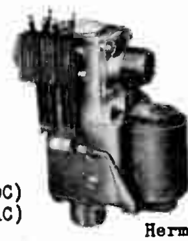
TC TUBULARS: ceramic, temperature compensating — the most stable capacitors available. Individually color coded. Rated at 500 working volts DC. Flash-tested 60 cycles, 1400 volts peak. Capacity range from 860 to 1 mmf.



101-K (AC) 201-K (DC)



Series 400



Types 8200 (DC) 7200 (AC)

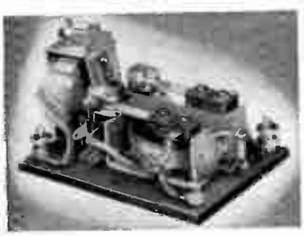
Hermetically sealed

Series 900

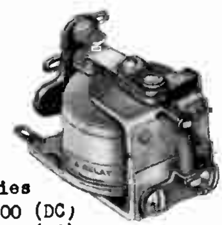
Type 750

Series 300

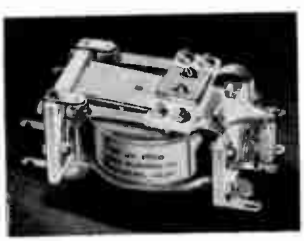
Type 800



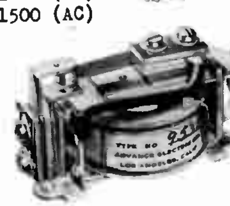
Series 600



Series K1600 (DC) K1500 (AC)



Series 1000



Series 950

Type 1200

Series 5200 (AC) 6200 (DC)

TYPE	USE	CONTACTS	COIL VOLTS	SIZE
101-K (AC) 201-K (DC)	Keying	1/4" Silver DPST N.O.	1 - 440 V.A.C. 1 - 220 V.D.C.	2 x 3 x 1 9/16
Series 300	Time delay	1/4" Silver SPST - DPDT NO or NC	1 - 440 V.A.C. 1 - 220 V.D.C.	3 3/4 x 2 5/8 x 1 1/2
Series 400	Antenna relay	1/4" Silver DPDT	1 - 440 V.A.C. 1 - 220 V.D.C.	3 9/16 x 2 13/16 x 1 5/8
Series 600	Lock-up Elect. reset	1/8" - 3/8" DPDT, DPST N.O., DPST N.C.	1 - 440 V.A.C. 1 - 220 V.D.C.	3 7/16 x 2 1/2 x 2 3/8
Type 750	Overload Elect. reset	1/4" Silver DPST	Type A 250-500 Ma. Type B 500-1000 Ma.	4 x 2 15/16 x 2 5/16
Type 800	Underload	1/4" Silver DPST	Type A 150-300 Ma. Type B 300-500 Ma.	3 x 3 x 1 7/8
Series 900	Impulse	DPST, DPDT 1/8, 3/16, 1/4	1 - 440 V.A.C. 1 - 220 V.D.C.	3 1/4 x 2 1/4 x 2
Series 1000	Midget ceramic insulated rel.	1/4" Silver DPST or DPDT	1 - 440 V.A.C. 1 - 220 V.D.C.	2 3/4 x 1 1/2 x 1 5/16
Series K1600 (DC) Series K1500 (AC)	Midget	1/8" or 3/16" up to 3PDT	1 - 220 V.A.C. 1 - 150 V.D.C.	1 3/4 x 1 3/16 x 1 1/8
Type 1200	Ultra-sensitive DC relay	1 Amp 110 V.A.C.	1 - 30,000 Ohms	2 9/16 x 2 x 1 1/2
Type 8200 (DC) Type 7200 (AC)	52 or 75 Ohm Coaxial rel.	1/4" Silver with DPDT Auxil. contacts	1 - 440 V.A.C. 1 - 220 V.D.C.	3 3/4 x 2 7/8 x 1 3/4
Series 950	Industrial Control	1/8 - 3/8 SPST - 3PDT	1 - 440 V.A.C. 1 - 220 V.D.C.	2 21/32 x 1 1/2 x 1 1/2
Series 5200 (AC) Series 6200 (DC)	Midget Telephone	1/8" or 3/16" SPST - 4PDT	1 - 440 V.A.C. 1 - 220 V.D.C.	1 1/2 x 1 1/8 x 1 5/16
Hermetically sealed	This is a sealed unit available for use with most midget and standard sized relays.			

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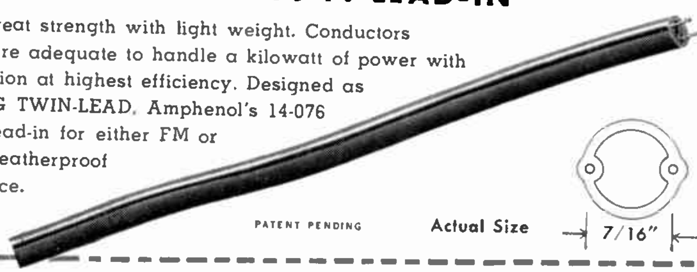
The DX "Champ" on 10 and 20 Meters

Extremely effective for reception as well as transmission, the Deluxe Dual-Three Signal Squirter is the first rotary beam offering full performance on both 10 and 20 meters.
Each of the two three-element arrays is coupled to the line with a separate Inductostub inductive coupling. Match between antenna and line is so simplified that the Signal Squirter can be assembled, installed and operated without tedious, complicated adjustments.

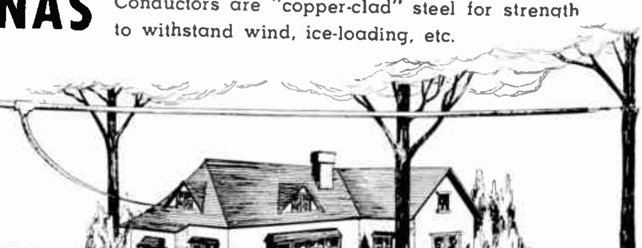
Weatherproof 300 OHM *Tubular* TRANSMITTING TWIN-LEAD *or* TV LEAD-IN

Tubular construction permits great strength with light weight. Conductors are 7 strands No. 26 copper wire adequate to handle a kilowatt of power with low losses and ideal for reception at highest efficiency. Designed as an outstanding TRANSMITTING TWIN-LEAD, Amphenol's 14-076 also proves to be a superior lead-in for either FM or Television. Specify this new weatherproof tubular line for best performance.



TWIN-LEAD *Folded Dipole* AMATEUR ANTENNAS

Ready cut to your choice of four bands—a real inexpensive and rugged DX Antenna. The folded dipole antenna section is Amphenol Twin-Lead No. 14-022. Conductors are "copper-clad" steel for strength to withstand wind, ice-loading, etc.



The 75 foot lead-in is Amphenol 300 ohm Twin-Lead No. 14-056. It affords a perfect match, and is joined to the antenna with a weatherproof molded polyethylene jacket.

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CRYSTAL CONTROLLED OSCILLATORS

	TYPE	RANGE	PIN DIA.	PIN SPACE	CASE DIM.	APPLICATION FEATURES
AMATEUR	AX2	20-40-80 Meter Bands	.093"	.486"	1 1/16"x11/16"x7/16"	High activity, plated xtal for precision frequency control.
AMATEUR	AX3	24-24.33 MC 25-25.55 MC	.093"	.486"	1 1/16"x13/16"x[illegible]/16"	Plated, overtone xtal for use on 2-6-10-11 meters with CCO-2A.
AMATEUR	CF6	455 KC 465 KC 500 KC	SOLDERING LUGS	SOLDERING LUGS	1 [illegible]/16"x1 7/32"x7/16"	Single signal filter with high Q for general communications rec'rs.
AMATEUR	KV3	100 KC	.093"	.486"	1 3/16"x1" DIA.	Low drift, plated xtal for secondary standards.
AMATEUR	CCO-2A	2-6-10-11 METERS	SOCKET .093"	SOCKET .486"	3"x2 1/4"x2 1/4"	Packaged oscillator for VHF crystal control. Nucleus for new construction.
COMMERCIAL	BH6	200-600 KC 1-100 MC	.050"	.486"	25/32"x23/32"x11/32"	Compact hermetically sealed metal case. Supplied as CR-18, CR-23 and other types for military use.
COMMERCIAL	BH7	15-75 MC	.062"	COAXIAL	.555"x.560"DIA.	Cylindrical hermetically sealed metal case. Supplied as CR-9, CR-17 and CR-24 for military use.
COMMERCIAL	SR5	600-15000 KC	.125"	.500"	1 3/8"x1 7/16"x1/2"	Gasket sealed molded phenolic case. Supplied as CR-1 for military use.
COMMERCIAL	SR12	70-200 KC	.125"	.750"	1 3/4"x1 3/4"x3/4"	Precision plated crystal in gasket sealed phenolic case for calibrators and standards.
COMMERCIAL	TCO-1	TEMP. STABILIZER	STD. OCTAL	STD. OCTAL	1 1/2"x1 1/4" DIA.	Constant temperature oven used with type BH6 unit for high stability.

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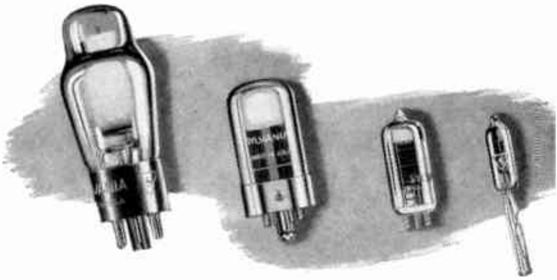
Now you can get the same high quality in transmitting tubes as you've been getting in Sylvania receiving tubes!

Typical of this whole new quality line, is the 3D24 (shown), a 4-electrode amplifier and oscillator with 45 watt anode dissipation. The electronic graphite plate allows high dissipation for a small area and maintains constant inter-electrode relationship and uniform tube characteristics.

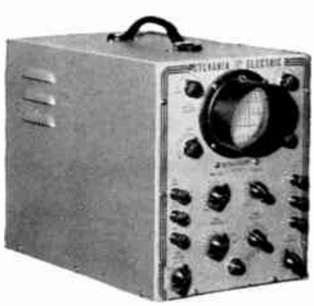


RECEIVING TUBES

From the tiny subminiatures to the famous Lock-I tubes, Sylvania research has brought many importa advances in radio receiving tubes over the past twent years.

You'll want Sylvania tubes for their well-know quality and performance. Sylvania has its own pla for the manufacture of radio tube parts. Every step the way is quality-controlled. You can be sure of sati faction when you buy any tubes in the big Sylvani Receiving Tube line.

Type 132
(Seven-inch Screen)



Type 131
(Three-inch Screen)

OSCILLOSCOPES

These modern Sylvania Oscilloscopes are ac-operated genera purpose cathode ray instruments used to study waveforms measure voltages and currents in various types of circuits. Excellent for audio circuit analysis, transmitter monitoring, filter an hum analysis, frequency comparisons.

Type 131 utilizes a type 3AP1 cathode ray tube having electrostatic deflection and focus. With horizontal and vertical amplifiers 0.5 volts rms gives 1-inch peak-to-peak deflection. Sweep generator variable from 15 to 40,000 cycles.

Two unique push-pull amplifiers give extra sensitive pattern on the 7GP1 screen of the Sylvania Type 132 Oscilloscope. One inch peak-to-peak vertical deflection is obtained with .21 volt rms input. Z-axis input provided for intensity modulation applications.

These are versatile instruments for every progressive well-equipped ham operator.

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POLYMETER

Sylvania Polymeter Type 134Z is a multi-purpose ıum tube voltmeter that greatly simplifies the job pplying many accurate measurements and tests to o equipment.

's beautifully styled, compactly designed; stabilized inst errors due to gas current in tubes. All essential ssories are included.

ests audio, ac and rf voltages from 20 cps to 300 mc ough use of proximity fuze-type tube built into handy be. Full scale range of 3. 10, 30, 100, 300 volts.

leasures dc voltages from .1 to 1,000 volts in full e ranges of 3, 10, 30, 100, 300, 1,000.

leasures dc current from .05 milliampere to 10 am-es in full scale ranges of 3, 10, 30, 100, 300, 1,000 liamperes and 10 amperes.

leasures resistance from ½ ohm to 1,000 megohms full scale ranges of 1,000, 10,000, 100,000 ohms and 0, 1,000 megohms.

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e this new DC Voltage Multiplier on your Sylvania lymeter and the 1,000 vdc range setting will read ,000 vdc full scale; the 300 vdc range setting will read 00 vdc full scale.

This accessory extends your Polymeter applications : Transmitter Plate Voltages. Experimental Power pplies. Television Circuits. Electronic Photographic uipment, and other high dc voltage plications.

MODULATION METER

You can now monitor modulation percentage and speech quality with Modulation Meter Type X-7018. Compact and economical. Helps you keep transmission efficient and at the same time comply with FCC overmodulation regulations. Also indicates carrier shift.

AUDIO OSCILLATOR

The Audio Oscillator Type 145 is one of the most versatile and convenient test instruments made. Its powerful signal of known frequency provides an accurate tone source for distortion checking of radio receivers. It may be used as a modulating signal for radio transmitters or as a simple frequency meter.

Frequency range 20 to 20,000 cycles.

Output power exceeds 1-watt at 500 ohms load; varies less than ± 2 db from 500 cps over entire range of instrument.

Maximum rms distortion less than 2% from 150 to 20,000 cycles, less than 4% from 30 to 150 cycles.

Output impedance variable by 3-position panel selector switch.

Signal-to-hum ratio 1,000 to 1 at 1 watt input.

Frequency drift negligible, even during warm-up periods.

Easily portable; weighs only 27½ pounds. An exceptionally flexible and valuable test instrument. Send for complete information.

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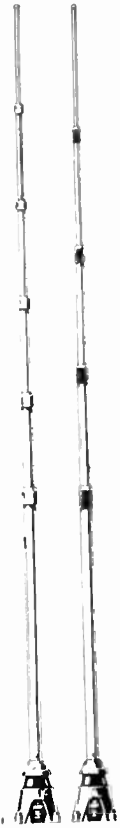


Alumi'm Monel
Insulators extra)

Weather Resistant Steel Antennas

No.	Description	*Extended Length*	*Collapsed Length*	*Base O.D.*	*Base I.D.*	*Weight Each*
112-M	2-sec. telescoping	11'8"	6'1"	.656"	.556"	4 lbs.
318-M	3-sec. telescoping	17'3"	6'2"	.875"	.775"	7 lbs.
224-M	4-sec. telescoping	22'9"	6'3"	1.063"	.963"	11 lbs.
130-M	5-sec. telescoping	28'3"	6'4"	1.250"	1.150"	15 lbs.
136-M	6-sec. telescoping	33'9"	6'5"	1.500"	1.400"	20 lbs.

Light-Weight Aluminum Antennas

No.	Description	*Extended Length*	*Collapsed Length*	*Base O.D.*	*Base I.D.*	*Weight Each*
AL-106	1-pc. tapered rod	6'3"	6'2"	.313"	....	¼ lb.
AL-312	2-sec. telescoping	12'4"	6'4"	.500"	.334"	1½ lbs.
AL-518	3-sec. telescoping	18'5"	6'4"	.750"	.584"	3 lbs.
AL-324	4-sec. telescoping	24'4"	6'4"	1.000"	.834"	5 lbs.
AL-530	5-sec. telescoping	30'0"	6'5"	1.250"	1.084"	7 lbs.
AL-535	6-sec. telescoping	35'8"	6'5"	1.500"	1.310"	12 lbs.

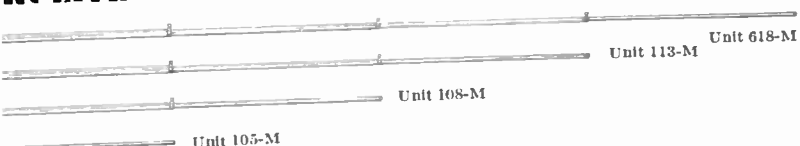
Heavy-Duty Aluminum Masts

No.	Description	*Extended Length*	*Collapsed Length*	*Base O.D.*	*Base I.D.*	*Weight Each*
AM-017	1-pc. tapered tube	17'9"	17'9"	.969"	.689"	5½ lbs.
AM-035	2-sec. tapered	35'0"	17'9"	2.000"	1.732"	19 lbs.

Long-Enduring Monel Antennas

No.	Description	*Extended Length*	*Collapsed Length*	*Base O.D.*	*Base I.D.*	*Weight Each*
MM-313	2-sec. telescoping	about 13'	6'9"	.615"	.545"	2¾ lbs.
MM-419	3-sec. telescoping	about 19'	6'9"	.747"	.667"	5 lbs.
MM-825	5-sec. telescoping	about 25'	5'8"	.893"	.799"	7½ lbs.
MM-430	5-sec. telescoping	about 30'	6'9"	1.065"	.945"	13 lbs.
MM-435	5-sec. telescoping	about 35'	7'8"	1.065"	.945"	15 lbs.
MM-635	5-sec. tele. H.D.	about 35'	7'8"	1.250"	1.120"	21 lbs.

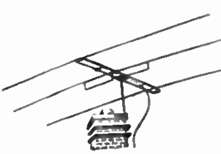
RULITE ELEMENTS for Beam Arrays



'remax Corulite Elements meet the need for light-weight but sturdy elements for hori-ontal arrays and similar applications. Exceptionally light weight yet they provide the needed strength and rigidity so essential in horizontal installations — and at extremely low cost. The special steel tubing used in these elements is a Premax development to insure unusual stiffness and strength. Heavily electroplated to insure corrosion resistance and high electrical conductivity. Fully adjustable to any desired length.

No.	Description	*Extended Length*	*Collapsed Length*	*Base O.D.*	*Recommended For*	*Weight Per Pr.*
105-M	1-section	5'0"	5'0"	.625"	6-meter	1 lb.
108-M	2-section	8'2"	4'7"	.750"	10-meter	2 lbs.
113-M	3-section	12'4"	4'8"	.875"		3½ lbs.
618-M	4-section	17'0"	5'3"	1.000"	20-meter	5½ lbs.

(Sold only in pairs, complete with Premax "Hairpin" Tuning Bar)



ROTARY BEAM KIT

Complete Rotary Beam Kit No. RB-6309 for 6, 10 and 11 meters, includes aluminum frame, 3 pair Elements, T-Match accessories and necessary hardware. Weighs only 30 pounds.

Base Insulator, Type 1: Heavy duty with compression rating up to 10,000 lbs. Rigid or hinged post designs in galvanized malleable iron or bronze to fit ¾" to 1 23/32" I.D. masts.

Type 1

Base Insulator Type 2: Light design for masts up to 18' or higher if guyed or supported by standoff insulators. ¾" top post is standard but with use of adapters will fit other sizes.

Type 2

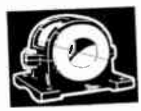
Base Insulator Type 6: for tower platform, rooftops or marine. Lead-thru construction permits antenna connections below roof or deck. Available for ¾" to 1⅜" I.D. tubular masts.

Type 6

Type 3 Standoff Insulator for supporting verticals or for use in pairs as complete antenna or element mountings. Galvanized iron or bronze with porcelain body. Styles to fit ½" to 1⅝" O.D. elements.

Type 3

Type 8-C Insulated Mounting Clamp for horizontal arrays, verticals, etc. Galvanized iron with porcelain split bushing. For ⅝" to 1" O.D. masts.

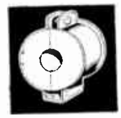


Type 8-C

Type 9-C Insulated Mounting Clamp for horizontal elements, verticals, etc. Galvanized iron with porcelain split bushing. Fits ⅝", ¾", ⅞" or 1" O.D. elements.

Type 9-C

Type 10-C Insulated Mounting Clamp. Electroplated stamped steel with porcelain split bushing; light-weight for rotary and dipole installations. For ⅝" to 1" elements.



Type 10-C

Type 10-S Standoff Insulator. Chrome plated bronze base and head-caps, porcelain insulator. Fits ⅞" to 1¼" O.D. Elements or masts.

No. 5-D Deck Bushing of brown glazed porcelain with galvanized malleable flange which bolts through rubber gasket to roof or deck. I.D. ¾", 1¼" or 1¾".

Bushing

Bronze Mounting Clip for horizontal elements, vertical antennas or for feed and transmission connections. For ¾", ⅞" or 1" O.D.

Mounting Clip

Wall Bracket of heavy steel for mounting vertical antennas on side walls, parapets, etc. Drilled to fit Types 1 and 2 Base Insulators.

Wall Bracket

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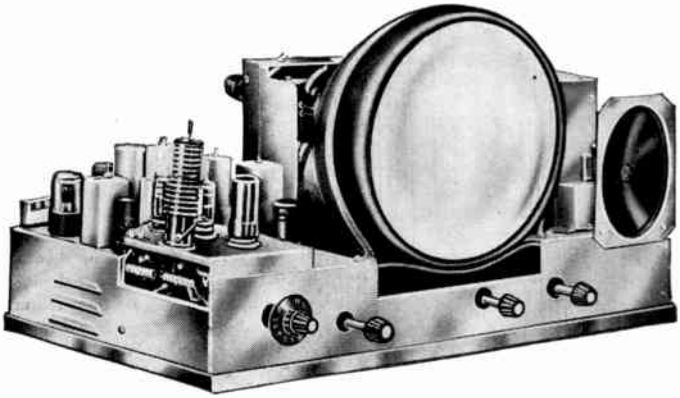

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Transvision **MODEL 10A TV KIT**, less cabinet.................... **Net $199.00**

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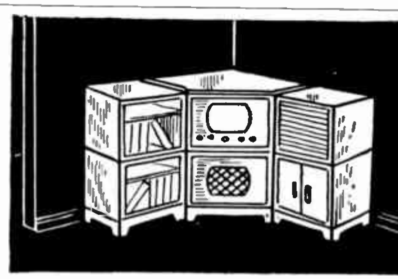
FREE **162 p. Transvision TELEVISION COURSE** ***with purchase of any Transvision TV Kit.***

You don't need this course to assemble a Transvision TV Kit, because the job is easy enough and our instruction sheet is simple and clear. BUT, if you want a *good introduction to television fundamentals* as a basis for further study, the Transvision Television Home-Study Course is ideal. *Remember, you pay nothing extra for this course.* Ask your distributor.



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The Model 770

AN ACCURATE POCKET-SIZE VOLT-OHM MILLIAMMETER

(SENSITIVITY: 1000 Ohms per Volt)

Features

- Compact—measures 3⅛" x 5⅞" x 2¼".
- Uses latest design 2% accurate 1 Mil. D'Arsonval type meter.
- Same zero adjustment holds for both resistance ranges. It is not necessary to readjust when switching from one resistance range to another. This is an important time-saving feature never before included in a V.O.M. in this price range.
- Housed in round-cornered, molded case.
- Beautiful black etched panel. Depressed letters filled with permanent white, insures long-life even with constant use.

Specifications

6 A.C. VOLTAGE RANGES:
0—15/30/150/300/1500/3000 VOLTS

6 D.C. VOLTAGE RANGES:
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4 D.C. CURRENT RANGES:
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2 RESISTANCE RANGES:
0—500 OHMS 0—1 MEGOHM

The Model 770 comes complete with self-contained batteries, test leads and all operating instructions.

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