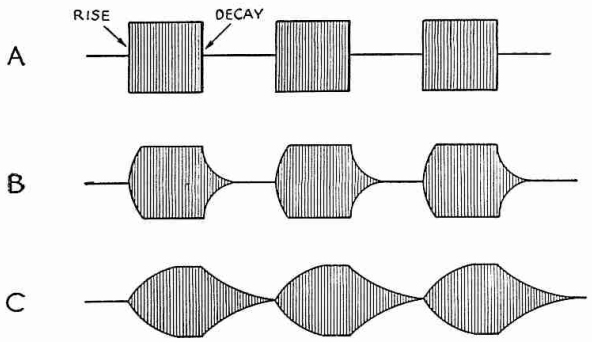


Fig. 8-2—Typical oscilloscope displays of a code transmitter. The rectangular-shaped dots (A) have serious key clicks extending many kc. either side of the transmitter frequency. Using proper shaping circuits increases the rise and decay times to give signals with the envelope form of B. This signal would have practically no key clicks. Carrying the shaping process too far, as in C, results in a signal that is too "soft" and is not easy to copy.

Backwave

The last factor is "backwave," a signal during key-up conditions from some amplifier-keyed transmitters. Some operators listening in the shack to their own signals and hearing a backwave think that the backwave can be heard on the air. It isn't necessarily so, and the best way to check is with an amateur a mile or more away. If he can't hear a backwave on the S9+ signal, you can be sure that it isn't there when the signal is weaker. Backwave is undesirable because it makes a signal harder to copy, even with acceptable shaping and no chirp.

Amplifier Keying

Many two-, three- and even four-stage transmitters are utterly incapable of completely chirp-free amplifier keying because the severe "modulation" of the output stage has an effect on the oscillator frequency and "pulls" through the several stages. This is particularly true when the oscillator stage is on the same frequency as the keyed output stage, but it can also happen when frequency multiplying is involved. Another source of reaction is the variation in oscillator supply voltage under keying conditions, although this can usually be handled by stabilizing the oscillator supply with a VR tube. If the objective is a completely chirp-free transmitter, the very first step is to make sure that keying the contemplated amplifier stage (or stages) has no effect on the oscillator frequency. This can be checked by listening on the oscillator frequency while the amplifier stage is keyed. Listen for chirp on either side of zero beat to eliminate the possible effect of a chirpy receiver caused by line-voltage changes or pulling. If no chirp of the steadily running oscillator can be detected, the transmitter can be keyed without chirp in the stage or stages used for the test. This is no assurance that the transmitter can be keyed without chirp in an earlier stage until the same test is passed by the earlier stage.

An amplifier can be keyed by any method that reduces the output to zero. Neutralized stages can be keyed in the cathode circuit, although where powers over 50 or 75 watts are involved it is often desirable to use a keying relay or vacuum tube keyer, to minimize the chances for electrical shock. Tube keying drops the supply voltages and adds cathode bias, points to be considered where maximum output is required. Blocked-grid keying is applicable to many neutralized stages, but it presents problems in high-powered amplifiers and requires a source of negative voltage. Output stages that aren't neutralized, such as many of the tetrodes and pentodes in widespread use, will usually leak a little and show some backwave regardless of how they are keyed. In a case like this it may be necessary to key two stages to eliminate backwave. They can be keyed in the cathodes, with blocked-grid keying, or in the screens. When screen keying is used, it is not always sufficient to reduce the screen voltage to zero; it may have to be pulled to some negative value to bring the key-up plate current to zero, unless fixed negative control-grid bias is used. It should be apparent that where two stages are keyed, keying the earlier stage must have no effect on the oscillator frequency if completely chirp-free output is the goal.

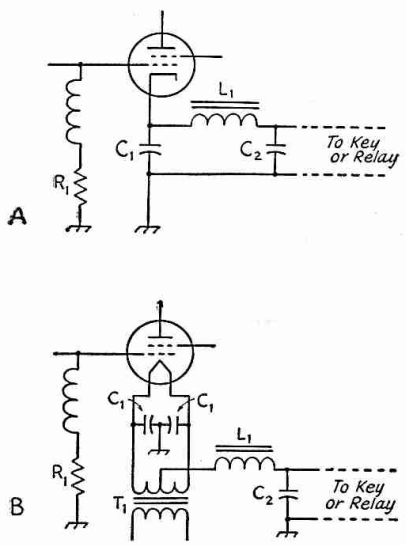


Fig. 8-3—The basic cathode (A) and center-tap (B) keying circuits. In either case C_1 is the r.f. return to ground, shunted by a larger capacitor for shaping. Voltage ratings at least equal to the cut-off voltage of the tube are required. T_1 is the normal filament transformer. C_2 can be about 0.01 μf.

The shaping of the signal is controlled by the values of L_1 and C_1. Increased capacitance at C_1 will make the signal softer on break; increased inductance at L_1 will make the signal softer on make. In many cases the make will be satisfactory without any inductance.

Values at C_1 will range from 0.5 to 4 μf., depending upon the tube type and operating conditions. The value of L_1 will also vary with tube type and conditions, and may range from a fraction of a henry to several henrys. When tetrodes or pentodes are keyed in this manner, a smaller value can sometimes be used at C_1 if the screen-voltage supply is fixed and not obtained from the plate supply through a dropping resistor.

Oscillators keyed in the cathode circuit cannot be softened on break indefinitely by increasing the value of C_1 because the grid-circuit time constant enters into the action.

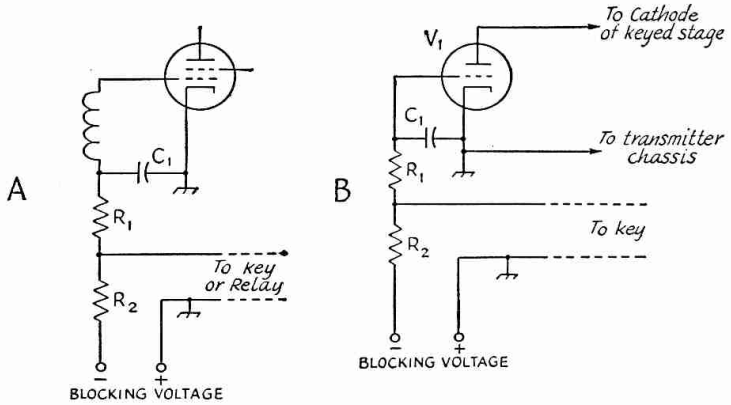


Fig. 8-4—The basic circuit for blocked-grid keying is shown at A. R_1 is the normal grid leak, and the blocking voltage must be at least several times the normal grid bias. The click on make can be reduced by making C_1 larger, and the click on break can be reduced by making R_2 larger. Usually the value of R_2 will be 5 to 20 times the resistance of R_1. The power supply current requirement depends upon the value of R_2, since closing the key circuit places R_2 across the blocking voltage supply.

An allied circuit is the vacuum-tube keyer of B. The tube V_1 is connected in the cathode circuit of the stage to be keyed. The values of C_1, R_1 and R_2 determine the keying envelope in the same way that they do for blocked-grid keying. Values to start with might be 0.47 megohm for R_1, 4.7 megohm for R_2 and 0.0047 μf. for C_1.

The blocking voltage supply must deliver several hundred volts, but the current drain is very low. The 6B4-G or other low plate-resistance triode is suitable for V_1. To increase the current-carrying ability of a tube keyer, several tubes can be connected in parallel.

A vacuum-tube keyer adds cathode bias and drops the supply voltages to the keyed stage and will reduce the output of the stage.

Shaping of the keying is obtained in several ways. Blocked-grid and vacuum-tube keyers get suitable shaping with proper choice of resistor and capacitor values, while cathode and screen-grid keying can be shaped by using inductors and capacitors. Sample circuits are shown in Figs. 8-3, 8-4 and 8-5, together with instructions for their adjustment. There is no "best" adjustment, since this is a matter of personal preference and what you want your signal to sound like. Most operators seem to like the make to be heavier than the break. All of the circuits shown here are capable of a wide range of adjustment.

If the negative supply in a grid-block keyed stage fails, the tube will draw excessive key-up current. To protect against tube damage in this eventuality, an overload relay can be used or, more simply, a fast-acting fuse can be included in the cathode circuit.

Oscillator Keying

The reader may wonder why oscillator keying hasn't been mentioned earlier, since it is widely used. The sad fact of life is that excellent oscillator keying is infinitely more difficult to obtain than is excellent amplifier keying. If the objective is no detectable chirp, it is probably *impossible* to obtain with oscillator keying, particularly on the higher frequencies. The reasons are simple. Any keyed-oscillator transmitter requires shaping at the oscillator, which involves changing the operating conditions of the oscillator over a significant period of time. The output of the oscillator doesn't rise to full value immediately, so the drive on the following stage is changing, which in turn may reflect a variable load on the oscillator. No oscillator has been devised that has no change in frequency over its entire operating voltage range and with a changing load. Furthermore, the shaping of the keyed-oscillator envelope usually has to be exaggerated, because the following stages will tend to sharpen up the keying and introduce clicks unless they are operated as linear amplifiers (as described in detail later).

Acceptable oscillator keying can be obtained on the lower-frequency bands, and the methods used to key amplifiers can be used, but chirp-free clickless oscillator keying is probably not possible at the higher frequencies, unless at some future date a completely voltage-insensitive oscillator circuit is devised. Often some additional shaping of the signal will be introduced on "make" through the use of a clamp tube in the output amplifier stage, because the time constant of the screen bypass capacitor plus screen dropping resistor increases the screen-voltage rise time, but it is of no help on the "break" portion of the signal.

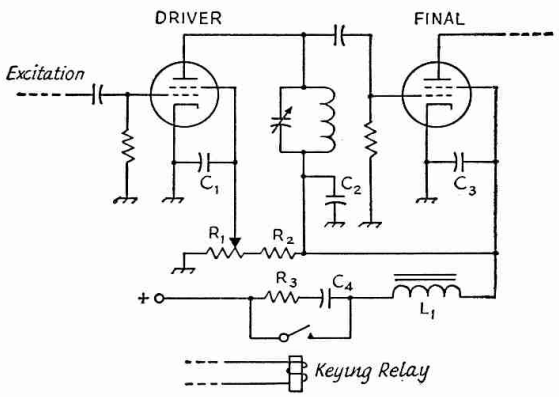


Fig. 8-5—When the driver stage plate voltage is roughly the same as the screen voltage of a tetrode final amplifier, combined screen and driver keying is an excellent system. The envelope shaping is determined by the values of L_1, C_4, and R_3, although the r.f. bypass capacitors C_1, C_2 and C_3 also have a slight effect. R_1 serves as an excitation control for the final amplifier, by controlling the screen voltage of the driver stage. If a triode driver is used, its plate voltage can be varied for excitation control.

The inductor L_1 will not be too critical, and the secondary of a spare filament transformer can be used if a low-inductance choke is not available. The values of C_4 and R_3 will depend upon the inductance and the voltage and current levels, but good starting values are 0.1 μf. and 50 ohms.

To minimize the possibility of electrical shock, it is recommended that a keying relay be used in this circuit, since both sides of the circuit are "hot." As in any transmitter, the signal will be chirp-free only if keying the driver stage has no effect on the oscillator frequency.

Break-In Keying

The usual argument for oscillator keying is that it permits break-in operation, which is true. If break-in operation is not contemplated and as near perfect keying as possible is the objective, then keying an amplifier or two by the methods outlined earlier is the solution. For operating convenience, an automatic transmitter "turner-onner" (see Campbell, *QST*, Aug., 1956), which will turn on the power supplies and switch antenna relays and receiver muting devices, can be used. The station switches over to the complete "transmit" condition where the first dot is sent, and it holds in for a length of time dependent upon the setting of the delay. It is equivalent to voice-operated phone of the type commonly used by s.s.b. stations. It does not permit hearing the other station whenever the key is up, as does full break-in.

Full break-in with excellent keying is not easy to come by, but it is easier than many amateurs think. Many use oscillator keying and put up with a second-best signal.

Three solutions to chirp-free break-in keying have been developed. One is the "silent v.f.o.," which consists of a well-shielded oscillator and buffer stage running continuously at a low frequency. The output is keyed before it gets out of the shielded compartment, and in some applications several subsequent stages are also keyed.

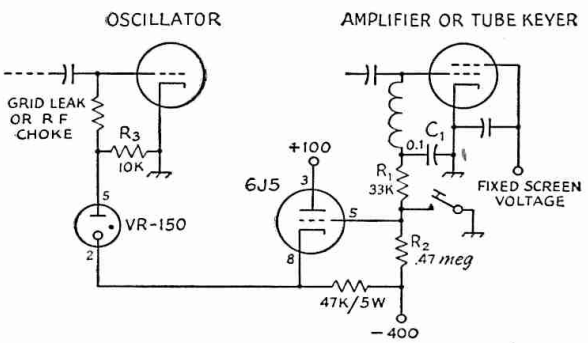


Fig. 8-6—When satisfactory blocked-grid or tube keying of an amplifier stage has been obtained, this VR-tube break-in circuit can be applied to the transmitter to furnish differential keying. The constants shown here are suitable for blocked-grid keying of a 6146 amplifier; with a tube keyer the 6J5 and VR tube circuitry would be the same.

With the key up, sufficient current flows through R_3 to give a voltage that will cut off the oscillator tube. When the key is closed, the cathode voltage of the 6J5 becomes close to ground potential, extinguishing the VR tube and permitting the oscillator to operate. Too much shunt capacity on the leads to the VR tube, and too large a value of grid capacitor in the oscillator, may slow down this action, and best performance will be obtained when the oscillator (turned on and off this way) sounds "clicky." The output envelope shaping is obtained in the amplifier, and it can be made softer by increasing the value of C_1. If the keyed amplifier is a tetrode or pentode, the screen voltage should be obtained from a fixed voltage source or stiff voltage divider, not from the plate supply through a dropping resistor.

A switch connected in series with the VR tube will, when opened, turn on the oscillator for "frequency spotting."

A second approach is to use a conversion exciter, in which two oscillators (one crystal-controlled, one v.f.o.) run continuously and their outputs, with suitable buffer stages intervening, are fed to a mixer stage. The mixer stage output is the sum or difference frequency of the two oscillator frequencies, which have been selected to give a sum or difference in an amateur band. When the mixer stage is turned off by keying, no output appears in the amateur band, and the effect is the same as keying an oscillator stage that cannot possibly chirp. The oscillator frequencies must be selected carefully so that none of their harmonics fall within an amateur band, and sufficient selectivity must be present in stages following the mixer to insure that no spurious signals are amplified.

Differential Keying

A third approach is to turn the oscillator on fast before a keyed amplifier stage can pass any signal and turn off the oscillator fast after the keyed amplifier stage has cut off. The principle is called "differential keying" and a number of circuits have been devised for accomplishing the action. One of the simplest can be applied to any grid-block keyed amplifier or tube-keyed stage by the addition of a triode and a VR tube, as in Fig. 8-6. Using this keying system for break-in, the keying will be chirp-free if it is chirp-free with the VR tube removed from its socket, to permit the oscillator to run all of the time. If the transmitter can't pass this test, it indicates that more isolation is required between keyed stage and oscillator.

Another VR-tube differential keying circuit, useful when the screen-grid circuit of an amplifier is keyed, is shown in Fig. 8-7. The normal screen keying circuit is made up of the shaping capacitor C_1, the keying relay (to remove dangerous volt-

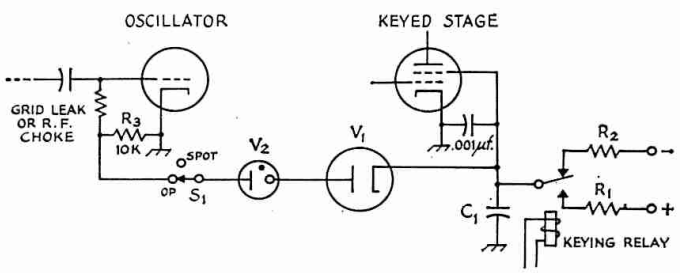


Fig. 8-7—VR-tube differential keying in an amplifier screen circuit.

With key up and current flowing through V_1 and V_2, the oscillator is cut off by the drop through R_3. The keyed stage draws no current because its screen grid is negative. C_1 is charged negatively to the value of the—source. When the relay is energized, C_1 charges through R_1 to a + value. Before reaching zero (on its way +) there is insufficient voltage to maintain ionization in V_2, and the current is broken in R_3, turning on the oscillator stage. As the screen voltage goes positive, the VR tube, V_2, cannot reignite because the diode, V_1, will not conduct in that direction. The oscillator and keyed stage remain on as long as the relay is closed. When the relay opens, the voltage across C_1 must be sufficiently negative for V_2 to ionize before any bleeder current will pass through R_3. By this time the screen of the keyed stage is so far negative that the tube has stopped conducting.

ages from the key), and the resistors R_1 and R_2. The + supply should be 50 to 100 volts higher than the normal screen voltage, and the − voltage should be sufficient to ignite the VR tube, V_2, through the drop in R_2 and R_3. Current through R_2 will be determined by voltage required to cut off oscillator; if 10 volts will do it the current will be 1 ma. For a desirable keying characteristic, R_2 will usually have a higher value than R_1. Increasing the value of C_1 will soften both "make" and "break."

The tube used at V_2 will depend upon the available negative supply voltage. If it is between 120 and 150, a 0A3/VR75 is recommended. Above this a 0C3/VR105 can be used. The diode, V_1, can be any diode operated within ratings. A 6AL5 will suffice with screen voltages under 250 and bleeder currents under 5 ma. For maximum life a separate heater transformer should be used for the diode, with the cathode connected to one side of the heater winding.

Clicks in Later Stages

It was mentioned earlier that key clicks can be generated in amplifier stages following the keyed stage or stages. This is often a puzzling problem to an operator who has spent considerable time adjusting the keying in his exciter unit for clickless keying, only to find that the clicks are bad when the amplifier unit is added. There are two possible causes for the clicks: low-frequency parasitic oscillations and amplifier "clipping."

Under some conditions an amplifier will be momentarily triggered into low-frequency parasitic oscillations, and clicks will be generated when the amplifier is driven by a keyed exciter. If these clicks are the result of low-frequency parasitic oscillations, they will be found in "groups" of clicks occurring at 50- to 150-kc. intervals either side of the transmitter frequency. Of course low-frequency parasitic oscillations can be generated in a keyed stage, and the operator should listen carefully to make sure that the output of the exciter is clean before he blames a later amplifier. Low-frequency parasitic oscillations are usually caused by poor choice in r.f. choke values, and the use of more inductance in the plate choke than in the grid choke for the same stage is recommended. (See Chapter Six and "low-frequency parasitic oscillations.")

When the clicks introduced by the addition of an amplifier stage are found only near the transmitter frequency, amplifier "clipping" is indicated. It is quite common when fixed bias is used on the amplifier and the bias is well past the "cut-off" value. The effect can usually be minimized or eliminated by using a combination of fixed and grid-leak bias for the amplifier stage. The fixed bias should be sufficient to hold the key-up plate current only to a low level and not to zero. In a triode amplifier, overdriving the amplifier can also result in clipping that will add key clicks, and the cure is to reduce the drive. The output won't suffer appreciably.

A linear amplifier (Class AB_1, AB_2 or B) will amplify the excitation without adding any clicks, and if clicks show up a low-frequency parasitic oscillation is probably the reason.

Testing Your Keying

The choice of a keying circuit is not as important as its testing. Any of the circuits shown in this chapter can be made to give satisfactory keying, but must be adjusted properly.

The easiest way to find out what your keyed signal sounds like on the air is to trade stations with a near-by ham friend some evening for a short QSO. If he is a half mile or so away, that's fine, but any distance where the signals are still S9 will be satisfactory.

After you have found out how to work his rig, make contact and then have him send slow dashes, with dash spacing. (The letter "T" at about 5 w.p.m.) With minimum selectivity, cut the r.f. gain back just enough to avoid receiver overloading (the condition where you get crisp signals instead of mushy ones) and tune slowly from out of beat-note range on one side of the signal through to zero and out the other side. Knowing the tempo of the dashes, you can readily identify any clicks in the vicinity as yours or someone else's. A good signal will have a thump on "make" that is perceptible only where you can also hear the beat note, and the click on "break" should be practically negligible at any point. Fig. 8-8A shows how it should sound. If your signal is like that, it will sound good, provided there are no chirps. Then have him run off a string of 35- or 40-w.p.m. dots with the bug — if they are easy to copy, your signal has no "tails" worth worrying about and is a good one for any speed up to the limit of manual keying. Make one last check with the selectivity in (Fig. 8-8B), to see that the clicks off the signal are negligible even at high signal level.

If you don't have any convenient friends with whom to trade stations, you can still check your keying, although you have to be a little more careful. The first step is to get rid of the r.f. click at the key, as described earlier, because if you don't you cannot make further observations.

So far you haven't done a thing for your signal on the air and you still don't know what it sounds like, but you may have cleaned up some clicks in the broadcast set. Now disconnect the antenna from your receiver and short the antenna terminals with a short piece of wire. Tune in your own signal and reduce the r.f. gain to the point where your receiver doesn't overload. Detune any antenna trimmer the receiver may have. If you can't avoid overload within the r.f. gain-control range, pull out the r.f. amplifier tube and try again. If you still can't avoid overload, listen to the second

Keying Tests

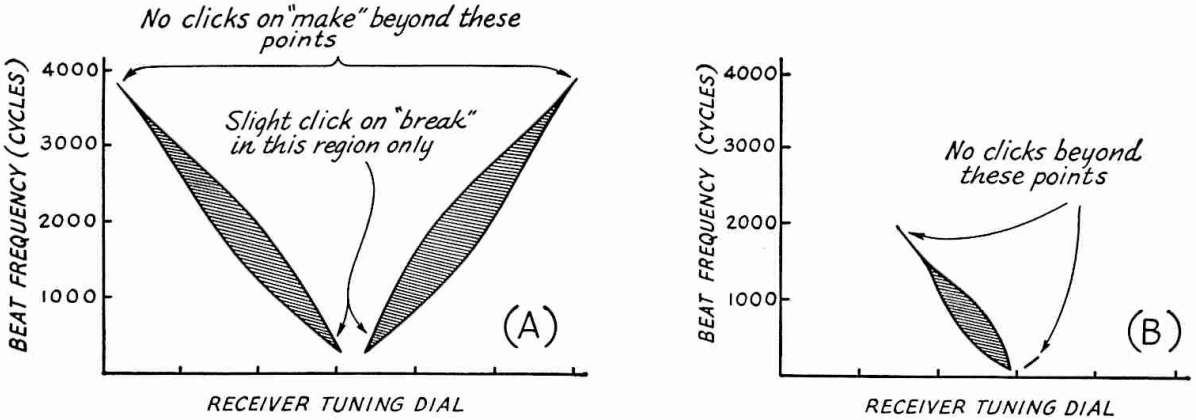


Fig. 8-8—Representations of a clean c.w. signal as a receiver is tuned through it. (A) shows a receiver with no selectivity and the b.f.o. set in the center of the pass band, and (B) shows the selectivity in and the receiver adjusted for single-signal reception. The variation in thickness of the lines represents the relative signal intensity. The audio frequency where the signal disappears will depend upon the receiver selectivity characteristic and the strength of the signal.

harmonic as a last resort. An overloaded receiver can generate clicks.

Describing the volume level at which you should set your receiver for these "shack" tests is a little difficult. The r.f. filter should be effective with the receiver running wide open and with an antenna connected. When you turn on the transmitter and take the other steps mentioned to reduce the signal in the receiver, run the audio up and the r.f. down to the point where you can just hear a little "rushing" sound with the b.f.o. off and the receiver tuned to the signal. This is with the selectivity in. At this level, a properly adjusted keying circuit will show no clicks off the rushing-sound range. With the b.f.o. on and the same gain setting, there should be no clicks outside the beat-note range. When observing clicks, make the slow-dash and fast-dot tests outlined previously.

Now you know how your signal sounds on the air, with one possible exception. If keying your transmitter makes the lights blink, you may not be able to tell too accurately about the chirp on your signal. However, if you are satisfied with the absence of chirp when tuning *either side of zero beat*, it is safe to assume that your receiver isn't chirping with the light flicker and that the observed signal is a true representation. No chirp either side of zero beat is fine. Don't try to make these tests without first getting rid of the r.f. click at the key, because clicks can mask a chirp.

Exchanging stations temporarily with another interested amateur is probably the best way to check your keying. The second-best method is to check it in the shack as outlined above. The least satisfactory way is to ask another ham on the air how your keying sounds. The reason it is the least satisfactory is that most hams are reluctant to be highly critical of another amateur's signal. In a great many cases they don't actually know what to look for or how to describe any aberrations they may observe.

Vacuum-Tube Keyers

The practical tube-keyer circuit of Fig. 8-9 can be used for keying any stage of any transmitter. Depending upon the power level of the keyed stage, more or fewer Type 2A3 tubes can be connected in parallel to handle the necessary current. The voltage drop through a single 2A3 varies from about 70 volts at 50 ma. to 40 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 connected to the transmitter ground. Thus the keyer can be used only in negative-lead or cathode keying.

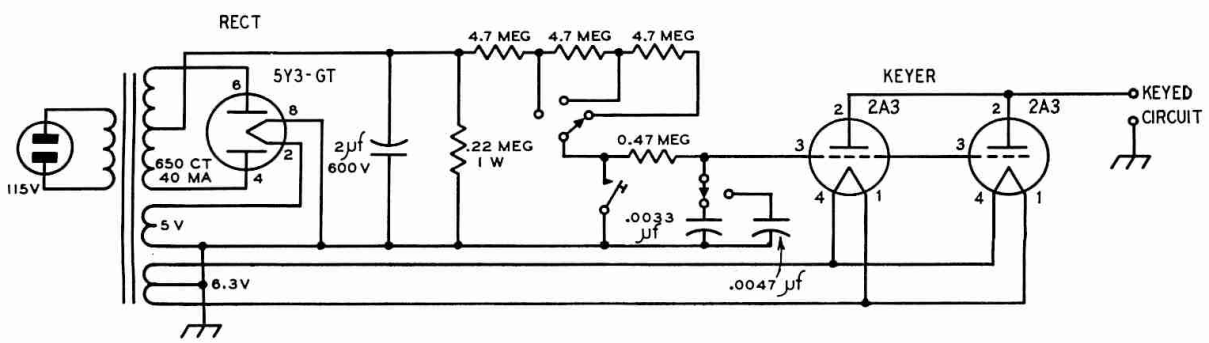


Fig. 8-9—Wiring diagram of a practical vacuum-tube keyer.

When used in cathode keying, it will introduce cathode bias to the stage and reduce the output. This can be compensated for by a reduction in the grid-leak bias of the stage.

The negative-voltage supply can be eliminated if a negative voltage is available from some other source, such as a bias supply. A simplified version of this circuit could eliminate the switches and associated resistors and capacitors, since they are incorporated only to allow the operator to select the combination he prefers. But once the values have been selected, they can be soldered permanently in place. The rule for adjusting the keying characteristic is the same as for blocked-grid keying.

A Low-Power Keyer

If a low-level stage running only a few watts is to be keyed, the tube-keyer circuit of Fig. 8-10 offers a simple solution. By using a 117L7 type tube, which incorporates its own rectifier, it is only necessary to connect to some existing power supply at the point marked "X". The keying characteristic will vary with many factors, so the values of R_1 and R_2 only represent starting points for experimentation.

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 backwave, but it can be cured by better insulation, or by reduced values of resistors and increased values of capacitors.

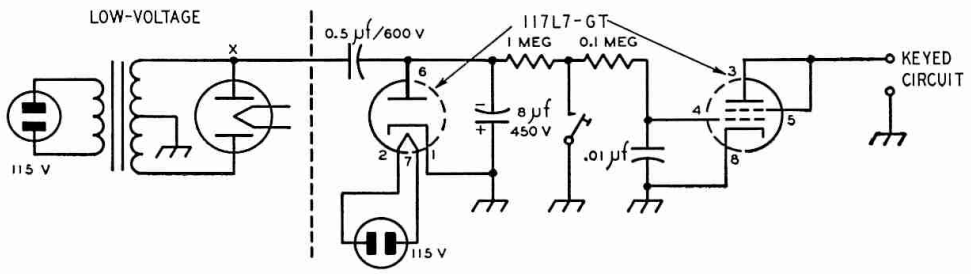


Fig. 8-10—Simple low-power vacuum-tube keyer. Connect keyer to a low-voltage power supply at point "X".

Monitoring of Keying

In general, there are two common methods for monitoring one's "fist" and signal. The first, and perhaps less common type, involves the use of an audio oscillator that is keyed simultaneously with the transmitter.

The second method is one that permits receiving the signal through one's receiver, and this generally requires that the receiver be tuned to the transmitter (not always convenient unless working on the same frequency) and that some method be provided for preventing overloading of the receiver, so that a good replica of the transmitted signal will be received. Except where quite low power is used, this usually involves a relay for simultaneously shorting the receiver input terminals and reducing the receiver gain.

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.

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. 8-11 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, K_1. 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

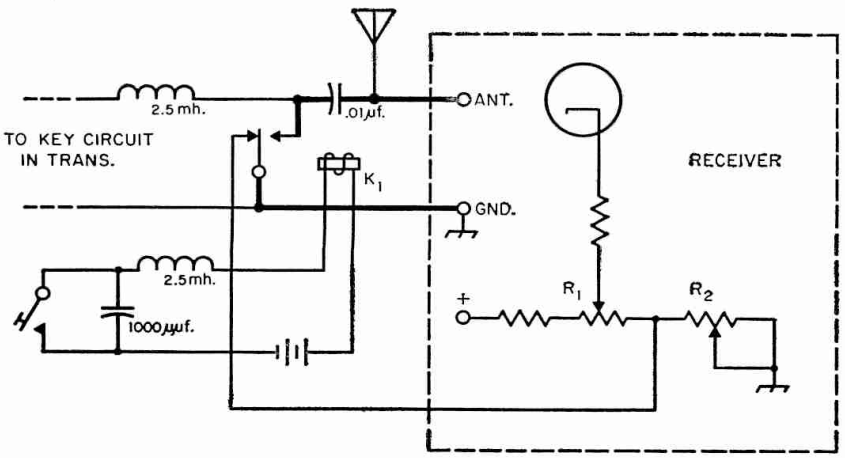


Fig. 8-11—Wiring diagram for smooth break-in operation. The lead shown as a heavy line and the lead from bottom relay contact to ANT post on receiver should be kept as short as possible for minimum pickup of the transmitter signal.
R_1—Receiver manual gain control.
R_2—5000- or 10,000-ohm wire-wound potentiometer.
K_1—S.p.d.t. keying relay. Although battery and d.c. relay are shown, any suitable a.c. or d.c. relay and power source can be used.

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. A filter at the key suppresses 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.

A few of the recent communications receivers bring the return lead from the r.f. gain control to a normally shorted terminal at the rear of the receiver. The preceding break-in system can be readily applied to a receiver of this type, and it will repay the receiver owner to study the instruction book and determine if his receiver already has this connection made in it. Other receivers have provision for reducing the gain or for blanking the receiver; one popular model has provision for bringing in negative bias from a transmitter grid leak to cut off an audio stage during transmit periods.

Full descriptions of systems for break-in operation can be found in the following *QST* articles:

Crawfis, "Simplified 'Break-In with One Antenna,' " Nov., 1954.
Goodman, "VR Break-In Keying," Feb., 1954.
Hays, "Selenium Break-In Keying," July, 1955.
Miller and Meichner, "TVG — An Aid to Break-In," March, 1953.
Puckett, "'De Luxe' Keying Without Relays," September, 1953; Part II, Dec., 1953.
Puckett, "C.W. Man's Control Unit," Feb., 1955.

Receiver Muting and Grid-Block Keying

The muting system shown in Fig. 8-12 can be used with any grid-block or tube-keyed transmitter, and it is particularly applicable to the VR-tube differential keying circuit of Fig. 8-6. Referring to Fig. 8-12, R_1, R_2 and C_1 have the same values and functions that the similarly designated components in Figs. 8-4 and 8-6 have. When the key is open, a small current will flow through R_3, the 0A2 and R_2, and the voltage drop across R_3 will be sufficient to cut off the 6C4. With the 6C4 cut off, there is no current through R_4 and consequently no voltage appearing across R_4. The voltage of the receiver a.v.c. bus is zero with respect to ground.

When the key is closed, there is insufficient voltage across the 0A2 to maintain conduction, and consequently there is no current flow through R_3. With zero voltage between grid and cathode, the 6C4 passes current. The drop across R_4, and thus the negative voltage applied to the a.v.c. line in the receiver, is determined by the value of R_4. Thus the key-down gain of the receiver can be adjusted to permit listening to one's own signal, by increasing the value of R_4 until the receiver output level is a comfortable one. To utilize the same antenna for transmitting and receiving, and thus benefit during receiving from

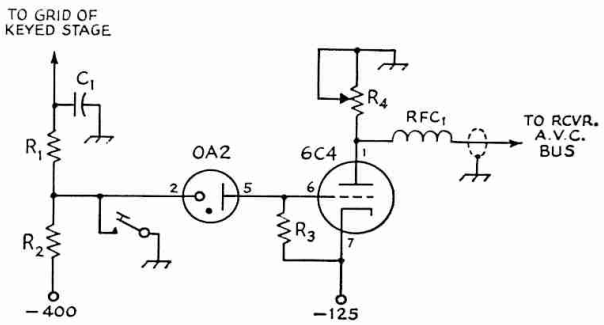


Fig. 8-12—Circuit diagram of a receiver muter for use with grid-block or tube keying.
C_1—Shaping capacitor, see text.
R_1, R_2—Shaping resistors, see text.
R_3—0.1 megohm.
R_4—15,000-ohm 2-watt potentiometer
RFC_1—1 mh. or less.

any directional properties of the antenna, an electronic transmit-receive switch can be used (see later in this chapter).

The receiver a.v.c. bus can be located by reference to the receiver instruction manual, and connection be made to it through a length of shielded wire. The a.v.c. switch in the receiver must be turned to ON for the muter to be effective.

If desired, the muting circuit can be built into the transmitter, or it can be mounted on a shelf or small chassis behind the receiver. The two negative voltages can be furnished by one supply and a reasonably heavy voltage divider; the main requirement of the supply is that the nominal −125 volts remain below the normal voltage drop of the 0A2 (150 volts). Installation of the muting circuits should have little or no effect on the keying characteristic of the transmitter; if it does the characteristic can be restored by proper values for R_1, R_2 and C_1.

The "Matchtone"

The "Matchtone" is a combination of the Monimatch (see Chapter 21) and a c.w. tone-generating monitor. It consists of a transistor audio oscillator which uses the Monimatch as a keyed source of d.c. power. In addition to the usual function it can be used by the sightless amateur as an audible transmitter-antenna tuning indicator.

While direct monitoring of c.w. transmissions via the receiver is a preferred method because it can reveal much about the keying characteristics, transmissions offset from the receiving frequency call for a separate monitor. The self-powered transistorized monitor fills the bill nicely. The use of the r.f. bridge, already connected in the r.f. transmission line, as a source of power for the monitor is a logical choice.

The circuit of the Matchtone and the connections to the Monimatch and the receiver are shown in Fig. 8-13. A small 2- or 3-to-1 push-pull grid-to-plate audio interstage transformer is used for feedback as well as for coupling to the receiver. If a transformer having a p.p. grid winding is not available from the junk box, the audio coupling to the receiver can be obtained by connecting C_2 to the ungrounded end of R_1. While use of a low value of capacitance for C_2 is necessary to avoid excessive shunting of the high-impedance receiver audio circuit, the value shown will provide sufficient coupling for a good audio tone level from the monitor. A third possibility for the audio output connection from the monitor is to substitute the headphones for R_1, together with a single-pole double-throw switch or relay to switch the phones between the monitor and the receiver. The on-off switch, S_1, can be made a part of R_2 by use of a volume control switch attachment.

The value shown for C_1 gives an audio pitch in the 500–1000 cycle range, depending somewhat on the particular transformer, the setting of R_2 and the transmitter output power. Other values of C_1 can be used to adjust the pitch to the operator's individual preference. R_2 may be adjusted to compensate for the changes in the d.c. current from the Monimatch caused by a change in transmitter frequency band or power. Using either a 2N109 or a CK722 transistor, the circuit should oscillate with usable audio level with as little as 0.1 ma. d.c. flowing to ground through the monitor. Other low-cost transistors such as the 2N107 and the 2N170 should work equally well.

Because the pitch of the audio tone is to some degree dependent upon the d.c. voltage obtained from the Monimatch, the pitch gives a reasonably accurate indication of correct final amplifier plate circuit tuning (maximum power output) and, if an antenna tuner is used, will also indicate resonance of the tuner to the transmitter output frequency. This characteristic of the Matchtone should be of considerable aid to sightless amateurs. (From *QST*, January, 1958.)

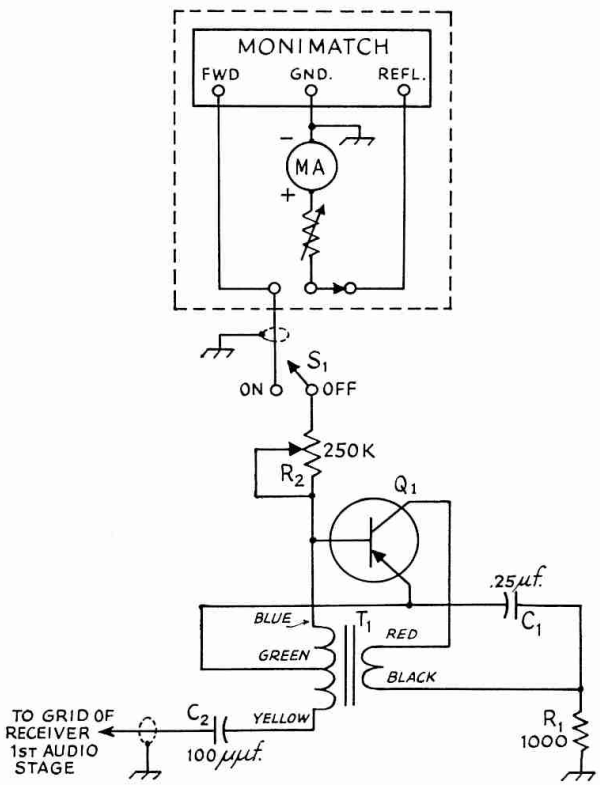


Fig. 8-13—Circuit of the Matchtone. Section enclosed in dashed line is the Monimatch and its indicating circuit. Braid of shielded lead to audio grid should connect to receiver chassis.

C_1—Paper.
C_2—Mica or ceramic.
Q_1—2N109, CK722 or similar.
R_1—1000 ohms, ½ watt.
R_2—0.25-megohm volume control.
S_1—S.p.s.t. toggle.
T_1—Push-pull interstage audio transformer, 2:1 or 3:1 total grid to plate.

Electronic Transmit-Receive Switches

No antenna relay is fast enough to switch an antenna from transmitter to receiver and back at normal keying speeds. As a consequence, when it is desired to use the same antenna for transmitting and receiving (a "must" when directional antennas are used) and to operate c.w. break-in or voice-controlled sideband, an electronic switch is used in the antenna. The word "switch" is a misnomer in this case; the transmitter is connected to the antenna at all times and the t.r. "switch" is a device for preventing burn-out of the receiver by the transmitter.

One of the simplest approaches is the circuit shown in Fig. 8–14. The 6C4 cathode follower couples the incoming signal on the line to the receiver input with only a slight reduction in gain. When the transmitter is "on," the grid of the 6C4 is driven positive and the rectified current biases the 6C4 so that it can pass very little power on to the receiver. The factors that limit the r.f. voltage the circuit can handle are the voltage break-down rating of the 47-$\mu\mu$f. capacitor and the voltage that may be safely applied between the grid and cathode of the tube.

To avoid stray pick-up on the lead between the cathode and the antenna terminal of the receiver, this lead should be kept as short as possible. The entire unit should be shielded and mounted on the receiver near the antenna terminals. In wiring the tube socket, input and output circuit components and wiring should be separated to reduce feed-through by stray coupling.

The t.r. switch of Fig. 8–15 differs in two ways from the preceding example. By using a grounded cathode and a tuned plate circuit, a voltage gain is obtained through the tube. The input is taken from the plate of the transmitter output stage instead of from the transmission line, and as a result the voltage build-up in the transmitter tank is utilized. Unlike the preceding t.r. switch, which permits listening on frequencies or bands to which the transmitter is not tuned, this switch will not permit much receiver response at frequencies removed from the transmitter frequency. Usually this is no problem, since most operation is around one's transmitter frequency. The 2.2K resistor across the plate circuit broadens the frequency response and reduces the need for retuning over a band. In a commercial version of this switch, a broadband output transformer replaces L_1 and the variable capacitor, and no coil changes are required in the range 3.5 to 30 Mc.

The switch of Fig. 8–15 can be built in a small metal box and mounted in the transmitter close to the output stage. The plate and heater power can be "borrowed" from the transmitter; the plate power will be less than 15 ma. at 100 to 150 volts. The coaxial line to the receiver can be any convenient length.

The capacitive voltage divider for feeding the t.r. switch is composed of the t.r. switch input capacitance (about 10 $\mu\mu$f.) and a series capacitor for connection to the plate tank. A conservative value of the series capacitor for an a.m. plate-modulated final can be calculated by the following formula:

$$C_1(\mu\mu\text{f.}) = \frac{2500}{d.c.\ plate\ volts}$$

The series capacitance as calculated above may be doubled in value when the final is not modulated, as in c.w., grid modulation or in a linear power amplifier.

The series capacitance is generally less than 20 $\mu\mu$f. The capacitor should be of the low-loss variety and should be capable of withstanding the tank voltage. For plate voltages of 800 volts or less, the disk type ceramic capacitors have been found to be adequate. For greater voltages, an inexpensive capacitor may be fabricated from RG-8/U coaxial cable. This cable has a rating of approximately 6000 peak r.f. volts, and in the laboratory it withstands in excess of 20,000 volts of d.c. Actually, in normal use it is usually limited by current rather than voltage. The capacitance of the cable is 30 $\mu\mu$f. per foot, so that one may measure off the required capacitance by the inch, and end up with a really low-loss and practical unit.

The t.r. switch input is a high impedance for low frequencies. It is advantageous, therefore, to have the tank circuit at d.c. ground potential so that crosstalk at power-line frequencies will be eliminated. Fortunately, this is the case in practically all modern transmitters. A type of noise customarily picked up with electronic t.r. switches is that caused by plate current flowing in the power amplifier. It is necessary, therefore, to bias the tubes beyond cutoff when receiving.

TVI and T.R. Switches

The preceding t.r. switches generate harmonics when their grid circuits are driven positive, and these harmonics can cause TVI if steps are not

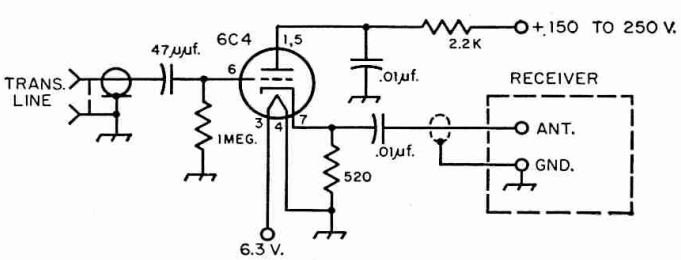


Fig. 8-14—Schematic diagram of cathode-follower t.r. switch. Resistors are ½-watt. The unit should be assembled in a small chassis or shield can and mounted on or very close to the receiver antenna terminals. The transmitter transmission line can be connected at the coaxial jack with an M-358 Tee adapter.

The heater and plate power can be "borrowed" from the receiver in most cases.

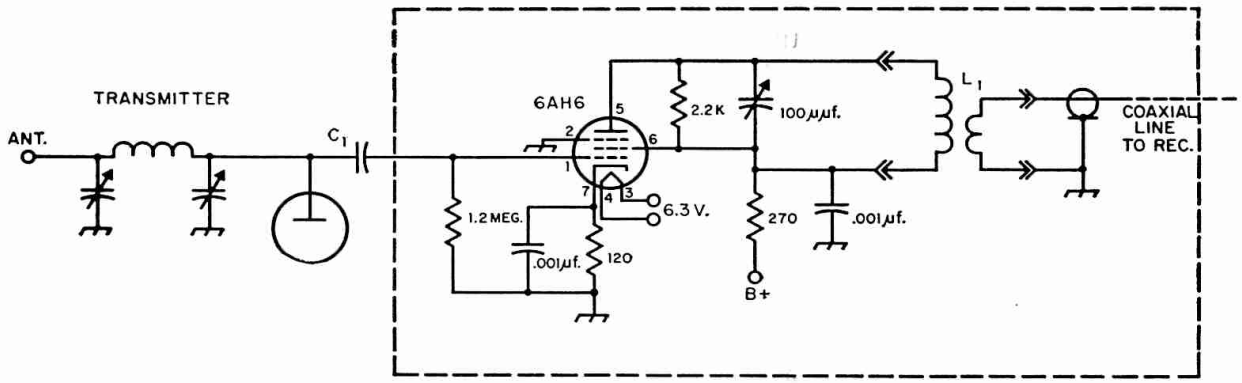


Fig. 8-15—A t.r. switch that mounts in the transmitter. Resistors are ½-watt.

C_1—Depends upon transmitter. See text.
L_1—Plug-in coil to tune to band in use. Coupling coil to receiver, 20 per cent turns in L_1 wound tight over "cold" end of L_1.

taken to prevent it. The switch of Fig. 8–14 should be well-shielded and used in the antenna transmission line between transmitter and low-pass filter. The switch of Fig. 8–15, when mounted in a transmitter that was TVI-free, should not introduce any TVI because the filtering that is successful for the transmitter should be successful for the harmonics generated by the t.r. switch.

Speed Keys

The average operator finds that a speed of 20 to 25 words per minute is the limit of his ability with a straight hand key. However, he can increase his speed to 30 to 40 w.p.m. by the use of a "speed key." The mechanical speed keys, available in most radio stores, give additional speed by making strings of dots when the key lever is pushed to the right; dashes are made manually by closing the key to the left. After practicing with the speed key, the operator obtains the correct "feel" for the key, which allows him to release the dot lever at exactly the right time to make the required number of dots. A speed key can deliver practically perfect code characters when used by an operator who knows what good code sounds like; however, one will not compensate for an operator's poor code ability.

An electronic speed key will not compensate for an operator's poor sending ability, either. However, the electronic speed key has the feature that it makes strings of both dots *and* of dashes, by proper manipulation of the key lever, and in current designs the dashes are *self-completing*. This means that it is impossible to send anything but the correct length of dash when the key lever is closed on the dash side. It is, of course, possible to send an incorrect number of dashes through poor operator timing.

An Electronic Speed Key

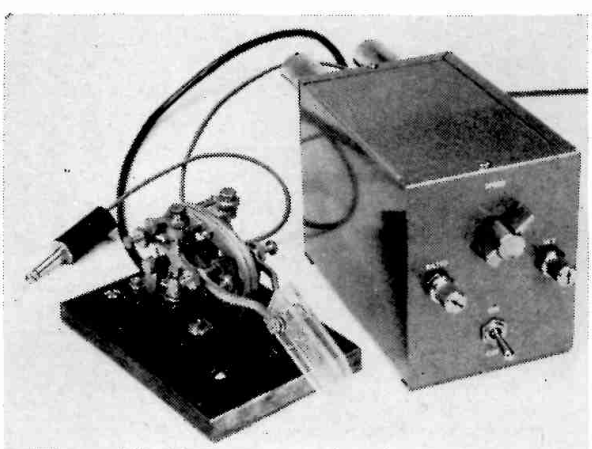


Fig. 8-16—This electronic speed key has a range of approximately 8 to 35 w.p.m., set by the speed control at top center. It has relay output and can be used with any transmitter that can be keyed by a hand key. The key (left) is made from two telegraph keys and a pair of ⅛-inch thick sheet plastic paddles.

The unit shown in Figs. 8-16 and 8-18 represents one of the simpler designs of an electronic key. The total cost of the key, in dollars and construction time, is quite low. The keying lever is made from parts taken from two straight telegraph keys; these are available at less than a dollar each in the war-surplus version (J-38). A more elegant keying lever can be built from a (more-expensive) war-surplus mechanical speed key.

Referring to Fig. 8-17, the timing of the key is provided by the oscillator V_{1A}. When the key is closed, a sawtooth wave is generated by the fast charge and slow discharge of the .25-μf. capacitor in the cathode circuit. The rate of discharge is set by the total resistance across the capacitor, and the voltage to which the capacitor is charged is determined by the setting of R_1. The sawtooth wave, applied to the grid of V_{2A}, cannot drive the grid very positive because the 3.3-megohm resistor limits the current; the effect is to "clip the tops" of the sawtooth cycles. The

An Electronic Speed Key

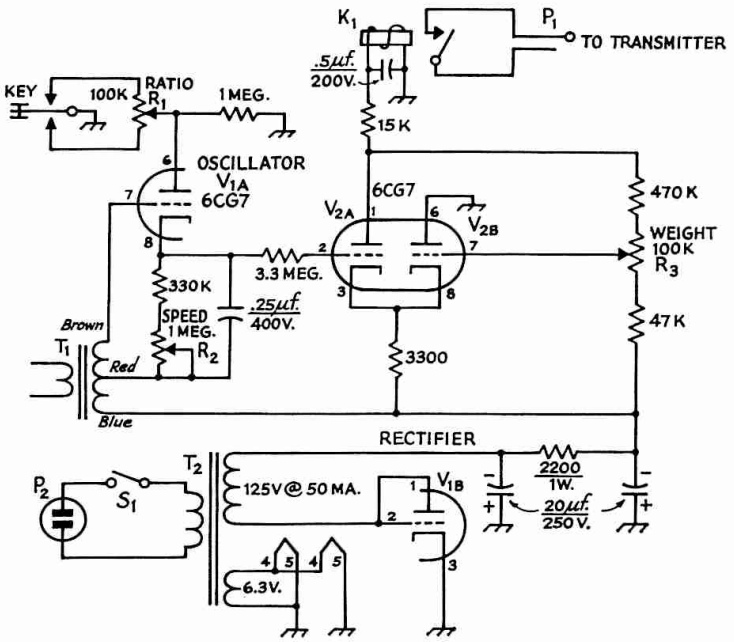


Fig. 8-17—Circuit diagram of the electronic speed key. Unless otherwise specified, resistors are ½ watt. Polarity-marked capacitors are electrolytic, others are tubular paper.

K1—5000-ohm 3-ma. relay (Sigma 41F-5000S-SIL).
P1—Phone plug.
P2—A.c. line plug.
R1, R3—100,000-ohm potentiometer, linear taper.
R2—1-megohm potentiometer, linear taper.
S1—S.p.s.t. toggle.
T1—5-watt 25,000-to-4-ohm output transformer, secondary not used (Stancor A-3857).
T2—125-v. 50-ma. and 6.3-v. 2-amp. transformer (Thordarson 26R38 or similar).

voltage at which V_{2A} passes enough current to close the relay is set by the position of the arm of R_3.

Except for the tubes, the keyer circuit is housed in a grey Hammertone 6 × 5 × 4-inch Minibox (Bud CU-2107), as shown in Fig. 8-16. The tube sockets are mounted so that the two tubes project outside at the rear of the unit. The power transformer is mounted on the rear wall, and the toggle switch and the three controls are mounted on the "front" panel. The power line to P_2, the two-wire cable to P_1, and the three-wire cable to the key leave the cabinet at the rear through individual rubber grommets. Use multiple tie points generously for the support of the fixed resistors and capacitors.

To make the key, first remove the keys from their bases and strip the bases of their remaining hardware. The four support legs for the key are formed from the original tie strips and shorting switch arms. At the front they bolt to the key frame at the countersunk holes; at the rear they make up to the binding posts. The three-wire cable connects to two binding posts and a supporting leg. A heavy base of ½-inch thick steel adds weight to the structure, and rubber or cork feet glued to the steel prevent its scratching the table.

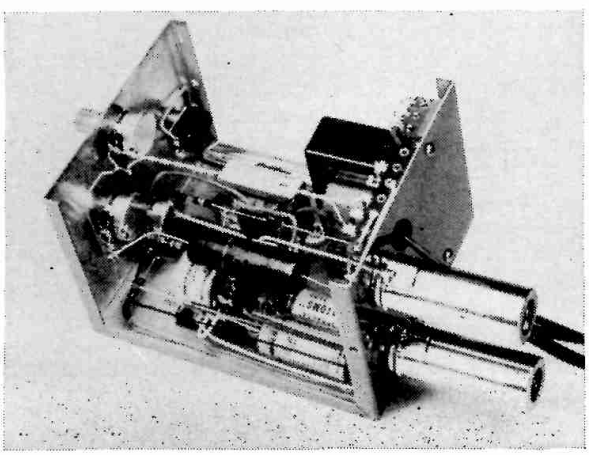


Fig. 8-18—Components for the electronic speed key are mounted on the three walls of a Minibox section, with the tubes projecting out the back. Keep wires away from screw holes, to prevent short circuits when the box is assembled.

Adjustment of Electronic Speed Key

In operation, the three controls will serve as their labels indicate. There is a unique (but not highly critical) combination of settings of the weight and ratio controls that will give automatic dots and dashes at the same speed; this setting can only be determined by ear and will be dependent on how well the operator can recognize good code. If the operator taps his foot to count groups of four dots or two dashes, the dots and dashes will have the same speed when the beat is the same. It is easy to determine whether dots or dashes are too heavy or too light. Connect an ohmmeter to P_1; holding the dot lever closed should make the ohmmeter needle hover around half scale, and holding the dash lever closed should make the ohmmeter hover around 75 per cent of the short-circuit reading. Lacking an ohmmeter, the transmitter plate milliammeter can be used; dots and dashes should give 50 per cent and 75 per cent of the key-down value when the keyer controls have been properly adjusted.

QST articles describing other types of electronic speed keys include:

Brann, "In Search of the Ideal Electronic Key," Feb., 1951
Bartlett, "Compact Automatic Key Design," Dec., 1951
Kaye, "All-Electronic 'Ultimatic' Keyer," April, May, 1955

Speech Amplifiers and Modulators

The audio amplifiers used in radiotelephone transmitters operate on the principles outlined earlier in this book in the chapter on vacuum tubes. The design requirements are determined principally by the type of modulation system to be used and by the type of microphone to be employed. It is necessary to have a clear understanding of modulation principles before the problem of laying out a speech system can be approached successfully. Those principles are discussed under appropriate chapter headings.

The present chapter deals with the design of audio amplifier systems for communication purposes. In voice communication the primary objective is to obtain the most *effective* transmission; i.e., to make the message be understood at the receiving point in spite of adverse conditions created by noise and interference. The methods used to accomplish this do not necessarily coincide with the methods used for other purposes, such as the reproduction of music or other program material. In other words, "naturalness" in reproduction is distinctly secondary to intelligibility.

The fact that satisfactory intelligibility can be maintained in a relatively narrow band of frequencies is particularly fortunate, because the width of the channel occupied by a phone transmitter is directly proportional to the width of the audio-frequency band. If the channel width is reduced, more stations can occupy a given band of frequencies without mutual interference.

In speech transmission, amplitude distortion of the voice wave has very little effect on intelligibility. The importance of such distortion in communication lies almost wholly in the fact that many of the audio-frequency harmonics caused by it lie outside the channel needed for intelligible speech, and thus will create unnecessary interference to other stations.

Speech Equipment

In designing speech equipment it is necessary to know (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 undue distortion anywhere in the system.

MICROPHONES

The **level** of a microphone is its electrical output for a given sound intensity. Level varies greatly with microphones of different types, and depends on the distance of the speaker's lips from the microphone. Only approximate values based on averages of "normal" speaking voices can be given. The values given later 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. 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).** When used with a vacuum-tube amplifier, the microphone is connected in the cathode circuit of a low-μ triode, as shown in Fig. 9-1A.

Sound waves striking the diaphragm cause it to vibrate in accordance with the sound, and the pressure on the granules alternately increases and decreases, causing a corresponding decrease and increase in the electrical resistance of the microphone. The instantaneous value of this resistance determines the instantaneous value of plate current through the tube, and as a consequent the voltage drop across the plate load resistor increases and decreases with the increases and decreases in granule pressure.

The carbon microphone finds its major amateur application in mobile and portable work; a good microphone in the circuit of Fig. 9-1A will deliver 25 to 35 volts peak output.

Piezo-electric 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 a 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 input circuit for the crystal microphone is shown in Fig. 9-1B.

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.

The **ceramic microphone** utilizes the piezoelectric effect in certain types of ceramic materials to achieve performance very similar to that of the crystal microphone. It is less affected by temperature and humidity. Output levels are similar to those of crystal microphones for the same type of frequency response.

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.

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. 9-1C). 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. 9-1D.

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 loud-speaker. A lightweight voice coil is rigidly attached to a diaphragm, the coil being suspended 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 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.

In general, the dynamic microphones have the smoothest peak-free response and widest frequency range, and they are also the least susceptible to damage from shock and extremes of temperature and humidity.

THE SPEECH AMPLIFIER

The audio-frequency amplifier stage that 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 the modulator 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.

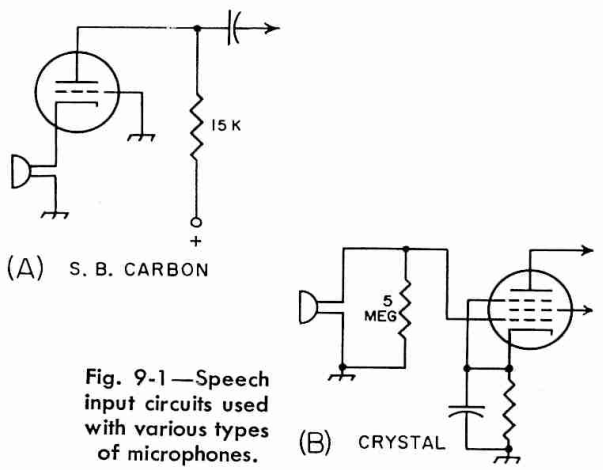
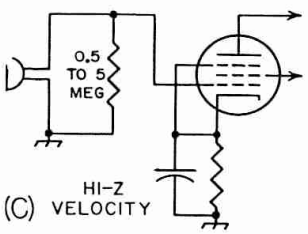
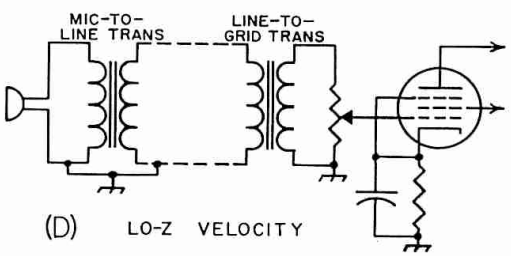


Fig. 9-1—Speech input circuits used with various types of microphones.





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, and this power in turn depends on the type of modulation system selected, as described in Chapter 10. 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 a later chapter. Generally speaking, it is advisable to choose a tube or tubes for the last stage of the speech amplifier that will be capable of

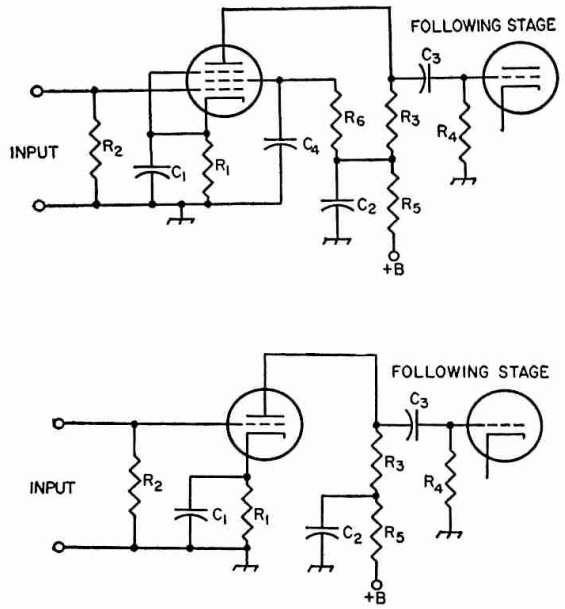


Fig. 9-2—Resistance-coupled voltage-amplifier circuits. A, pentode; B, triode. Designations are as follows:

C_1—Cathode bypass capacitor.
C_2—Plate bypass capacitor.
C_3—Output coupling capacitor (blocking capacitor).
C_4—Screen bypass capacitor.
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-I. Values in the decoupling circuit, C_2R_5, are not critical. R_5 may be about 10% of R_3; an 8- or 10-μf. electrolytic capacitor is usually large enough at C_2.

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.

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.

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. 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 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 most satisfactory type of coupling 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. 9-2 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 at the end of this book.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 9-3. The circuit at A combines resistance and transformer coupling, and may be used for exciting the

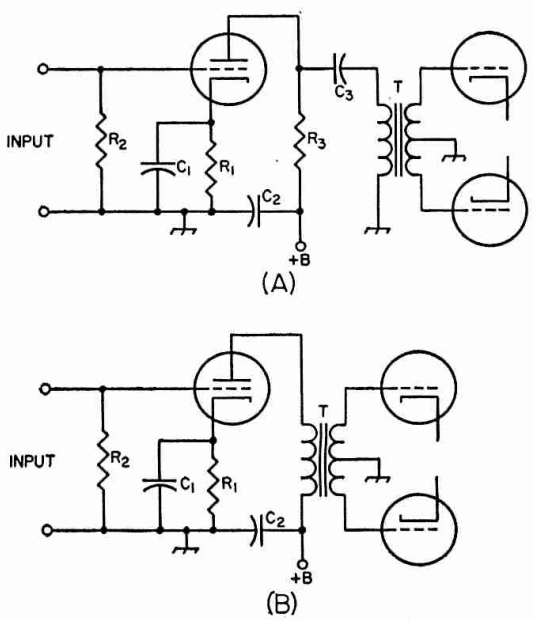


Fig. 9-3—Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B for transformer coupling. Designations correspond to those in Fig. 9-2. In A, values can be taken from Table 9-I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given in the tube tables for the particular type of tube used.

Table 9–I

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 ratio of the new voltage to 300 volts. Voltage gain is measured at 400 cycles. Capacitor values given are based on 100-cycle cutoff. For increased low-frequency response, all capacitors may be made larger than specified (cut-off frequency in inverse proportion to capacitor 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 Bypass μf.	Cathode Bypass μf.	Blocking Capacitor μf.	Output Volts (Peak) [1]	Voltage Gain [2]
6SJ7, 12SJ7	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, 12J7-GT	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, 12AU6, 12SH7	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, 6AQ7, 6AT6, 6Q7, 6SL7GT, 6T8, 12AT6, 12Q7-GT, 12SL7,-GT (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
6AV6, 12AV6, 12AX7 (one triode)	0.1	0.1	——	1300	——	4.6	0.027	43	45
		0.22	——	1500	——	4.0	0.013	57	52
		0.47	——	1700	——	3.6	0.006	66	57
	0.22	0.22	——	2200	——	3.0	0.013	54	59
		0.47	——	2800	——	2.3	0.006	69	65
		1.0	——	3100	——	2.1	0.003	79	68
	0.47	0.47	——	4300	——	1.6	0.006	62	69
		1.0	——	5200	——	1.3	0.003	77	73
		2.2	——	5900	——	1.1	0.002	92	75
6SC7, 12SC7 [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
6CG7, 6J5, 7A4, 7N7, 6SN7-GT, 12J5-GT, 12SN7-GT (one triode)	0.047	0.047	——	1300	——	3.6	0.061	59	14
		0.1	——	1580	——	3.0	0.032	73	15
		0.22	——	1800	——	2.5	0.015	83	16
	0.1	0.1	——	2500	——	1.9	0.031	68	16
		0.22	——	3130	——	1.4	0.014	82	16
		0.47	——	3900	——	1.2	0.0065	96	16
	0.22	0.22	——	4800	——	0.95	0.015	68	16
		0.47	——	6500	——	0.69	0.0065	85	16
		1.0	——	7800	——	0.58	0.0035	96	16
6C4, 12AU7 (one triode)	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	——	8800	——	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

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.

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** or **phase-splitter** circuits as shown in Fig. 9-4.

The circuits shown in Fig. 9-4 are of the "self-balancing" type. In A, 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, thus giving push-pull output. The part that appears across R_7 from V_2 opposes the voltage from V_1 across R_7, thus reducing the signal applied to the grid of V_2. The negative feedback so obtained tends to regulate the voltage applied to the phase-inverter tube so that the output voltages from both tubes are substantially equal. The gain is slightly less than twice the gain of a single-tube amplifier using the same operating conditions.

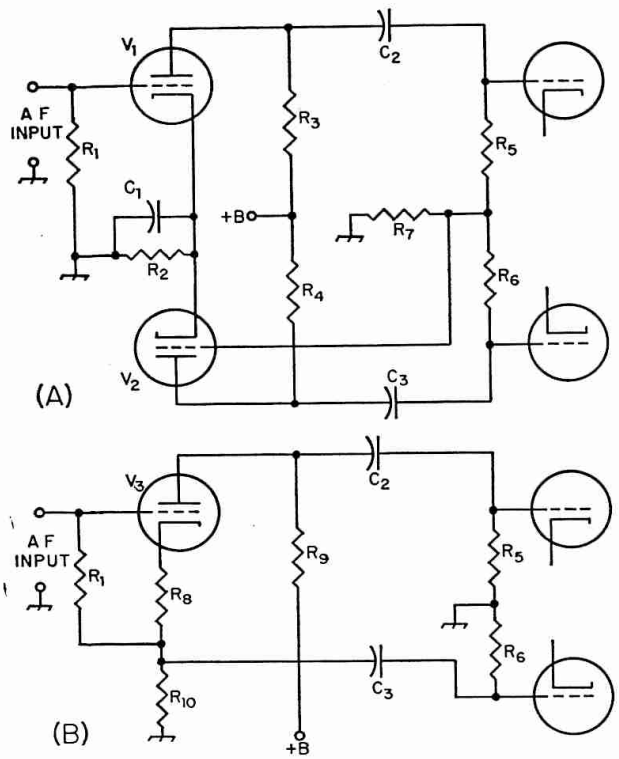


Fig. 9-4—Self-balancing phase-inverter circuits. V_1 and V_2 may be a double triode such as the 12AU7 or 12AX7. V_3 may be any of the triodes listed in Table 9-I, or one section of a double triode.

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-μf. electrolytic.
C_2, C_3—0.01- to 0.1-μf. paper.

In the single-tube circuit shown in Fig. 9-4B 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 necessary for keeping the final output at the proper level for modulating the transmitter. 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 signal voltage level is so low there is no danger that the stages ahead of the gain control will be overloaded by the full microphone output. 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. Also, in a high-gain amplifier it is better to operate the first tube at maximum gain, since this gives the best signal-to-hum ratio. 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, a Class B amplifier may be required. Select a suitable tube type and determine from the tube tables at the end of this book the grid driving power required, if any.

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 or output stage of the speech-amplifier. Receiver-type power tubes can be used (beam tubes such as the 6L6 may be needed in some cases) as determined from the receiving-tube tables. If the speech amplifier is to drive a Class B modulator, use a Class A or AB_1 amplifier.

4) If the speech-amplifier output stage is also the modulator and must operate Class AB_2 to develop the required power output, use a low- or medium-μ triode to drive it. If more power is needed than can be obtained from one tube, use two 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 speech-amplifier output stage operates Class A or AB_1, it may be driven by a voltage amplifier. If the output 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-to-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 maximum-signal cathode 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. 9-4A) 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 described earlier in this chapter.

7) Divide the voltage required to drive the output 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. If high gain is required, a pentode may be used for the first speech-amplifier stage, but it is *not* advisable to use a second pentode because of the possibility of feedback 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 medium-μ triode may be used as a third stage.

A high-μ double triode with the sections in cascade makes a good low-level amplifier, and will give somewhat greater gain than a pentode followed by a medium-μ triode. With resistance-coupled input to the first section the cathode of that section may be grounded (contact potential bias), which is helpful in reducing hum.

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 feedback. For reasonably humless operation, the hum voltage should not exceed about 1 per cent of the maximum audio output voltage — that is, the hum and noise should be at least 40 db. below the output level.

Unwanted feedback, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feedback can be minimized by isolating each stage with decoupling resistors and capacitors, 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 steel 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 electric 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 feedback. It is always safe, although not absolutely necessary, to separate the speech amplifier and its power supply, building them on separate chassis.

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; this 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. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

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.

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 feedback 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 self-oscillation of the amplifier.

When using paper capacitors as bypasses, be sure that the terminal marked "outside foil" is connected to ground. This utilizes the outside foil of the capacitor as a shield around the "hot" foil. When paper capacitors are used for coupling 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.

Modulators and Drivers

● CLASS AB AND B MODULATORS

Class AB or B modulator circuits are basically identical no matter what the power output of the modulator. The diagrams of Fig. 9-5 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

The audio ratings of various types of transmitting tubes are given in the chapter containing the tube tables. 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 15 to 25 per cent greater than the actual power needed for modulation.

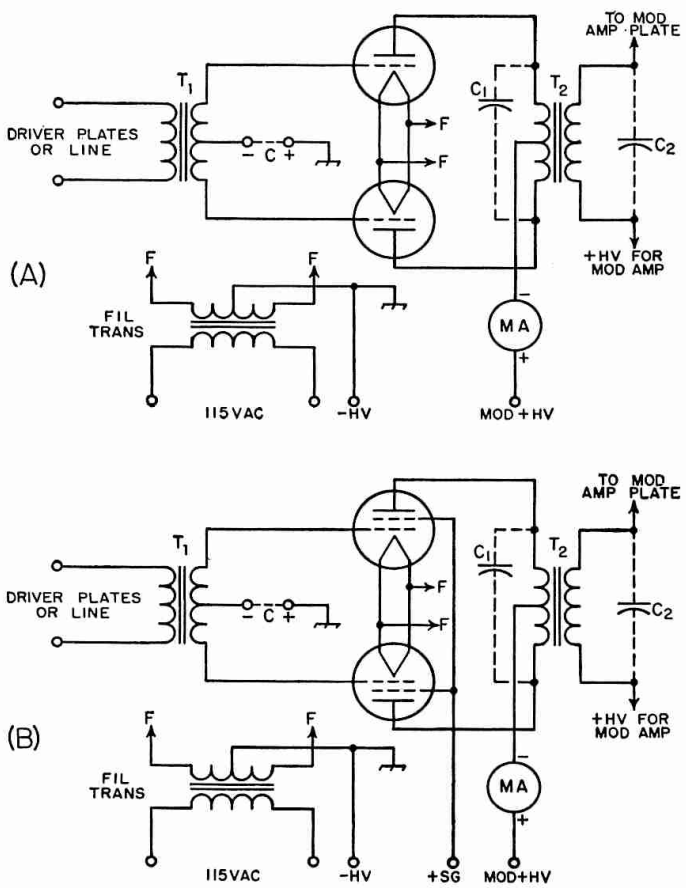


Fig. 9-5—Modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

Matching to Load

In giving audio 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

$$N = \sqrt{\frac{Z_p}{Z_m}} = \sqrt{\frac{6900}{5000}} = \sqrt{1.38} = 1.175{:}1.$$

The required transformer ratios for the ordinary range of impedances are shown graphically in Fig. 9-6.

Many modulation transformers are provided with primary and secondary taps, so that various turns ratios can be obtained to meet the requirements of particular tube combinations. However, it may be that the exact turns ratio required 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 usually 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 very far without exceeding the ratings of the Class C tubes for either plate voltage or plate current, even though the power input is kept at the same figure.

Suppressing Audio Harmonics

Distortion in either the driver or Class B modulator 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 capacitors C_1 and C_2 across the primary and secondary, respectively, of the Class B output transformer in Fig. 9-5 is to reduce the strength of harmonics and unnecessary high-frequency components existing in the modulation. The capacitors 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 0.01 μf. will be required; the larger values are necessary with the lower values of load resistance. The voltage rating of each capacitor 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 will be supplied by the plate bypass or blocking capacitor in the modulated amplifier.

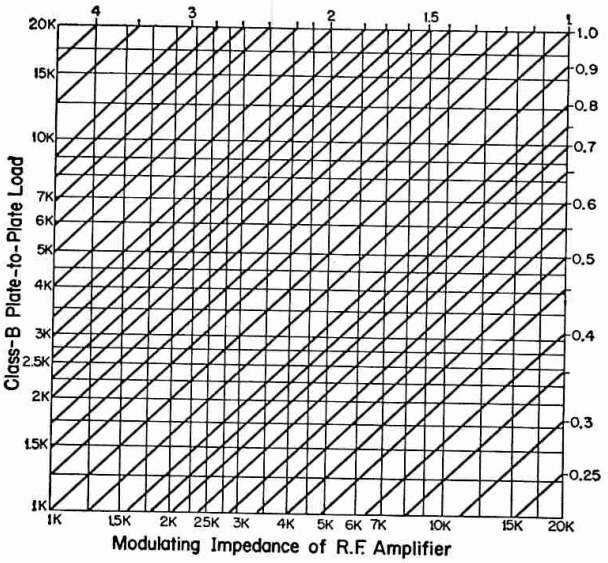


Fig. 9-6—Transformer ratios for matching a Class C modulating impedance to the required plate-to-plate load for the Class B modulator. The ratios given on the curves are from total primary to secondary. Resistance values are in kilohms.

A still better arrangement is to use a low-pass filter as shown later, even though clipping is not deliberately employed.

Grid Bias

Certain triodes designed for Class B audio work can be operated without grid bias. Besides eliminating the grid-bias supply, the fact that grid current flows over the whole audio cycle means that the load resistance for the driver is fairly constant. With these tubes the grid-return lead from the center-tap of the input transformer secondary is simply connected to the filament center-tap or cathode.

When the modulator tubes require bias, it should always be supplied from a *fixed* voltage source. Cathode bias or grid-leak bias cannot 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. For larger bias voltages a heavy-duty "B" battery may be used if the grid current does not exceed 40 or 50 milliamperes on voice peaks. The batteries are charged by the grid current rather than discharged, but a battery nevertheless 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 the power supply chapter.

Plate Supply

In addition to adequate filtering, 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 load current, 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.

Good dynamic regulation — i.e., with suddenly applied loads — is equally as important as good regulation under steady loads, since an instantaneous drop in voltage on voice peaks also will limit the output and cause distortion. The output capacitor of the supply should have as much capacitance as conditions permit. A value of at least 10 μf. should be used, and still larger values are desirable. It is better to use all the available capacitance in a single-section filter rather than to distribute it between two sections.

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. The audio impedance between screen and cathode also must be low.

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 can happen even though the modulation percentage, as defined in the chapter on amplitude modulation, is less than 100 per cent, if the modulator is incapable of delivering the audio power required to modulate the transmitter.)

As shown later, such a condition may be reached by deliberate design, in case the modulator is to be adjusted for peak clipping. But whether it happens by accident or intention, the splatter and spurious sidebands can be eliminated by inserting a low-pass filter (Fig. 9-13) between the modulator and the modulated amplifier, and then taking care to see that the actual modulation of the r.f. amplifier does not exceed 100 per cent.

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 may be developed in the primary — frequently high enough to break down the transformer insulation.

DRIVERS FOR CLASS-B MODULATORS

Class AB_2 and Class B amplifiers are driven into the grid-current region, so power is con-

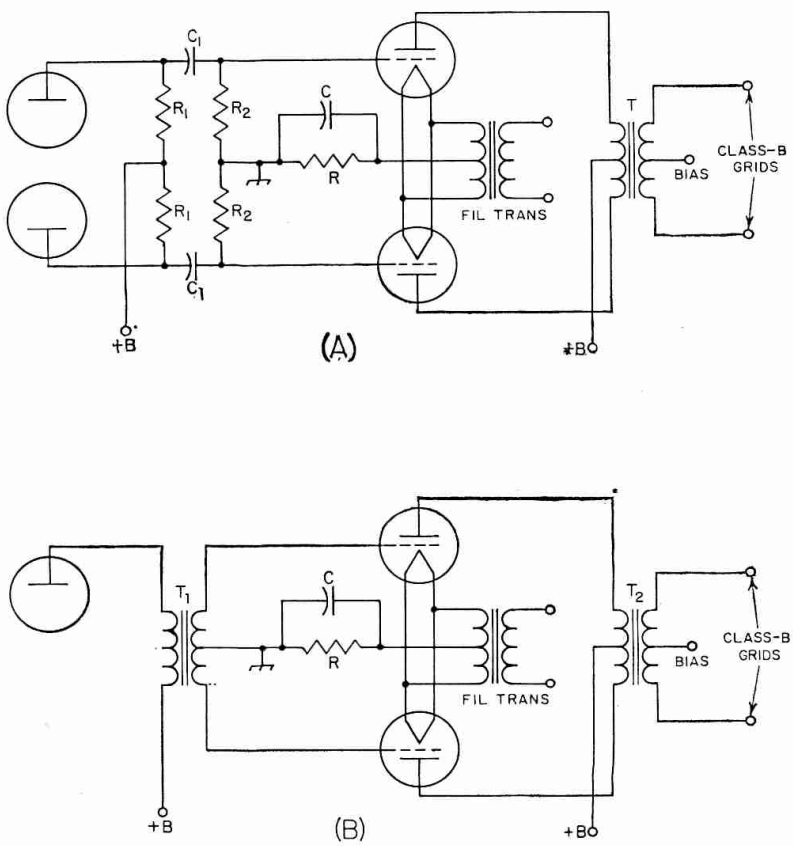


Fig. 9-7—Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling capacitor and grid resistor, respectively; values also may be taken from Table 9-I.

In both circuits the output transfomer, (T_1T_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 bypass, is determined as described in the text.

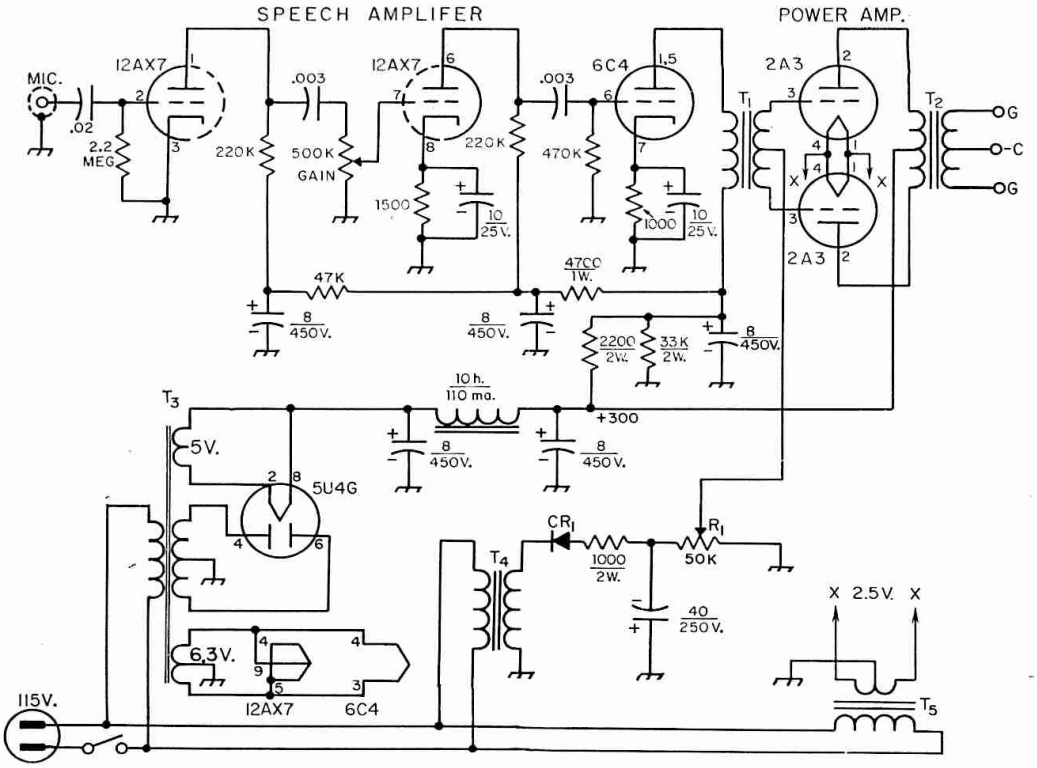


Fig. 9-8—Speech-amplifier driver for 10–15 watts output. Capacitances are in μf. Resistors are ½ watt unless specified otherwise. Capacitors with polarity indicated are electrolytic; others may be paper or ceramic.

CR1—Selenium rectifier, 20 ma.
R1—50,000-ohm potentiometer, preferably wire wound.
T1—Interstage audio transformer, single plate to push-pull grids, turns ratio 2 to 1 or 3 to 1, total secondary to primary.
T2—Class-B driver transformer, 3000 ohms plate-to-plate; secondary impedance as required by Class-B tubes used; 15 watt rating.
T3—Power transformer, 700 volts c.t., 110 ma.; 5 volts, 3 amp.; 6.3 volts, 4 amp.
T4—Power transformer, 125 volts, 20 ma.; 6.3 volts, 0.6 amp.
T5—2.5-volt 5-ampere filament transformer (Thordarson 21F00).

sumed 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 varying 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 wave form 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.

Driver Tubes

To secure good voltage regulation the internal impedance of the driver, as seen by the modulator grids, must be low. The principal component of this impedance is the plate resistance of the driver tube or tubes as reflected through the driver transformer. Hence for low driving-source impedance the effective plate resistance of the driver tubes should be low and the turns ratio of the driver transformer, primary to secondary, should be as large as possible. The maximum turns ratio that can be used is that value which just permits developing the modulator grid-to-grid a.f. voltage required for the desired power output. 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.

Low-μ triodes such as the 2A3 have low plate resistance and are therefore good tubes to use as drivers for Class AB_2 or Class B modulators. Tetrodes such as the 6V6 and 6L6 make very poor drivers in this respect when used without negative feedback, but with such feedback the effective plate resistance can be reduced to a value comparable with low-μ triodes.

Fig. 9-7 shows representative circuits for a push-pull triode driver using cathode bias. If the amplifier operates Class A the cathode resistor need not be bypassed, 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 bypassed. The bypass capacitance required can be calculated by a simple rule: the cathode resistance in ohms multiplied by the bypass capacitance in microfarads should equal at least 25,000. The

voltage rating of the capacitor 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 2A3s 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 100 ma. From Ohm's Law,

$$E = RI = 780 \times 0.10 = 78.6 \text{ volts}$$

From the rule mentioned previously, the bypass capacitance required is

$$C = 25{,}000/R = 25{,}000/780 = 32\ \mu\text{f}.$$

A 40- or 50-μf. 100-volt electrolytic capacitor would be satisfactory

Fig. 9-8 is a typical circuit for a speech amplifier suitable for use as a driver for a Class AB_2 or Class B modulator. An output of about 13 watts can be realized with the power supply circuit shown (or any similar well-filtered supply delivering 300 volts under load). This is sufficient for driving any of the power triodes commonly used as modulators. The 2A3s in the output stage are operated Class AB_1. The circuit provides several times the voltage gain needed for communications-type crystal or ceramic microphones.

The two sections of a 12AX7 tube are used in the first two stages of the amplifier. These are resistance coupled, the gain control being in the grid circuit of the second stage. Although the cathode of the first stage is grounded and there is no separate bias supply for the grid, the grid bias actually is about one volt because of "contact potential."

The third stage uses a medium-μ triode which is coupled to the 2A3 grids through a transformer having a push-pull secondary. The ratio may be of the order of 2 to 1 (total secondary to primary) or higher; it is not critical since the gain is sufficient without a high step-up ratio.

The output transformer, T_2, should be selected to couple between push-pull 2A3s and the grids of the particular modulator tubes used.

The power supply has a capacitor-input filter the output of which is applied to the 2A3 plates through T_2. For the lower-level stages, additional filtering is provided by successive RC filters which also serve to prevent audio feedback through the plate supply.

Grid bias for the 2A3s is furnished by a separate supply using a small selenium rectifier and a TV "booster" transformer, T_4. The bias may be adjusted by means of R_1, and should be set to -62 volts or to obtain a total plate current of 80 ma. (as measured in the lead to the primary center tap of T_2) for the 2A3s.

In building an amplifier of this type the constructional precautions outlined earlier should be observed. The Class AB_1 modulators described subsequently in this chapter are representative of good constructional practice.

Negative Feedback

Whenever tetrodes or pentodes are used as drivers for Class B modulators, negative feedback should be used in the driver stage, for the reason already discussed.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-9. Fig. 9-9A 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 total resistance of R_4 and R_5 in series should be ten or more times the rated load resistance of V_2. Instead of the voltage divider, a tap on the transformer primary can be used to supply the feedback voltage, if such a tap is available.

The amount of feedback 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. Circuit values for typical tube combinations are given in detail in Fig. 9-9.

The push-pull circuit in Fig. 9-9B requires an audio transformer with a split secondary. The feedback

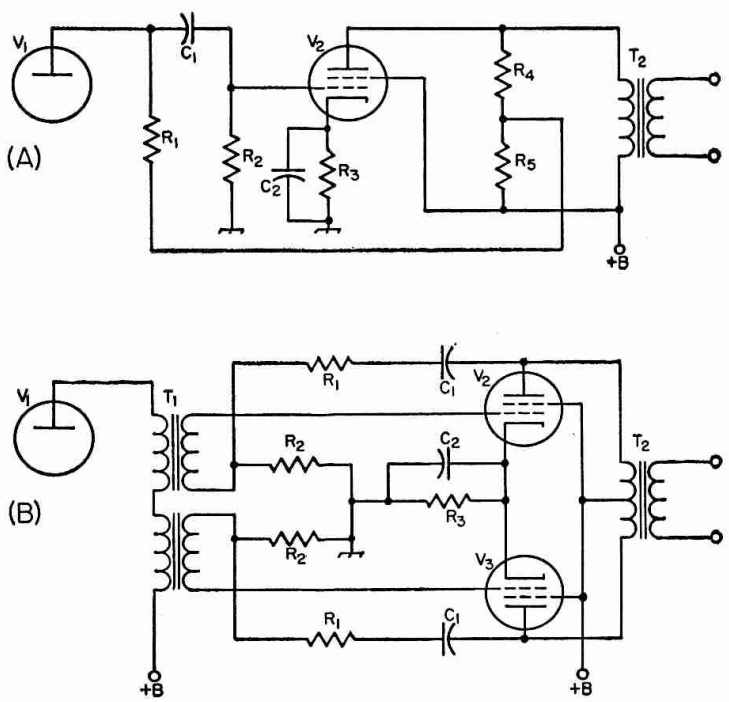


Fig. 9-9—Negative-feedback circuits for drivers for Class B modulators. A—Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, or one section of a 6CG7 and a 6AQ5, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μf.; C_2, 50 μf. B—Push-pull beam-tetrode driver. If V_1 is a 6J5 or 6CG7 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 μf.; C_2, 100 μf.

voltage is obtained from the plate of each output tube by means of the voltage divider, R_1, R_2. The blocking capacitor, C_1, prevents the d.c. plate voltage from being applied to R_1R_2; the reactance of this capacitor 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 (ten times or more) compared with the rated load resistance for V_2 and V_3.

In this circuit the feedback voltage that is developed across R_2 appears at the grid of V_2 (or V_3) through the transformer secondary and grid-cathode circuit of the tube, provided the tubes are not driven to grid current. The per cent feedback is

$$n = \frac{R_2}{R_1 + R_2} \times 100$$

where n is the feedback percentage, and R_1 and R_2 are connected as shown in the diagram. The higher the feedback 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 feedback 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 or similar triode, and T_1 has a turns ratio of 2-to-1, total secondary to primary, it is possible to use over 30 per cent feedback without going beyond the output-voltage capabilities of the triode. Twenty per cent feedback will reduce the effective plate resistance to the point where the output voltage regulation is better than that of 2A3s without feedback. The power output under these conditions is about 20 watts.

Increasing the Effectiveness of the Phone Transmitter

The effectiveness of an amateur phone transmitter can be increased to a considerable extent by taking advantage of speech characteristics. 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, a large portion of speech power is normally found below 500 cycles. If these low frequencies are attenuated, 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.

One simple way to reduce low-frequency response is to use small values of coupling capacitance between resistance-coupled stages, as shown in Fig. 9-10A. A time constant of 0.0005 second for the coupling capacitor 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 capacitor of 0.001 μf. will give the required time constant.

The high-frequency response can be reduced by using "tone control" methods, utilizing a capacitor 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. 9-10B. The capacitor 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 without unduly sacrificing intelligibility.

Restricting the frequency response 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

Although it is obviously desirable to modulate

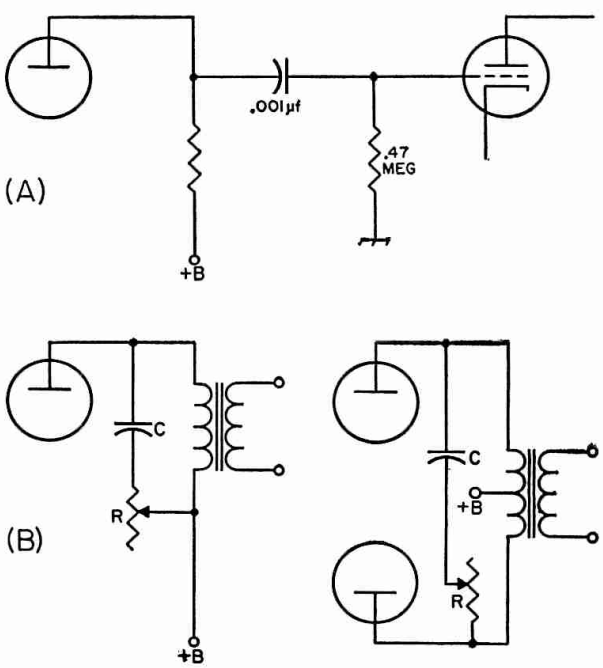


Fig. 9-10—A, use of a small coupling capacitor to reduce low-frequency response; B, tone-control circuits for reducing high-frequency response. Values for C and R are discussed in the text; 0.01 μf. and 25,000 ohms are typical.

the transmitter as completely as possible, 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. 9-11. V_1, a medium-μ triode, has its grid connected in parallel with the grid of the last speech amplifier tube (the stage preceding the power stage) through the gain control R_1. The amplified output is coupled to a full-wave rectifier, V_2. The rectified audio output develops a negative d.c. voltage across C_1R_3, which has a sufficiently long time constant to hold the voltage at a reasonably steady value between syllables and words. The negative d.c. voltage is applied as control bias to the suppressor grid of the first tube in the speech amplifier (this circuit requires a pentode first stage), effecting a reduction in gain. The gain reduction is substantially proportional to the average microphone output and thus tends to hold the amplifier output at a constant level.

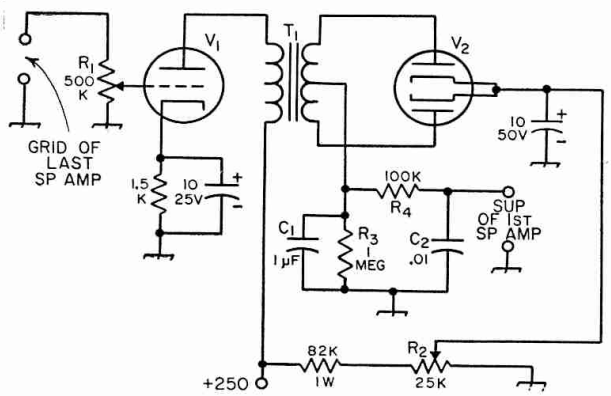


Fig. 9-11—Speech-amplifier output limiting circuit.
V_1—6C4, 6C5, 6CG7, 6J5, 12AU7, etc.
V_2—6H6, 6AL5, etc.
T_1 — Interstage audio, single plate to p. p. grids.

An adjustable bias is applied to the cathodes of V_2 to cut off the tube at low levels and thus prevent rectification until a desired output level is reached. R_2 is the "threshold control" which sets this level. R_1, the gain control, determines the rate at which the gain is reduced with increasing signal level.

The hold-in time can be increased by increasing the resistance of R_3. C_2 and R_4 may not be necessary in all cases; their function is to prevent too-rapid gain reduction on a sudden voice peak. The "rise time" of this circuit can be increased by increasing C_2 or R_4, or both.

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

In speech wave forms the average power content is considerably less than in a sine wave of the same peak amplitude. Since modulation percentage is based on peak values, the modulation or sideband power in a transmitter modulated 100 per cent by an ordinary voice wave form will be considerably less than the sideband power in the same transmitter modulated 100 per cent by a sine wave. In other words, the modulation percentage with voice wave forms is determined by peaks having relatively low average power content.

If the low-energy peaks are clipped off, the remaining wave form will have a considerably higher ratio of average power to peak amplitude. More sideband power will result, therefore, when such a clipped wave is used to modulate the transmitter 100 per cent. Although clipping distorts the wave form and the result therefore does not sound exactly like the original, it is possible to secure a worth-while increase in modulation power without sacrificing intelligibility. Once the system is properly adjusted *it will be impossible to overmodulate the transmitter* because the maximum output amplitude is fixed.

By itself, clipping generates the same high-order harmonics that overmodulation does, and therefore will cause splatter. To prevent this, the audio 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 high attenuation for all frequencies above 3000 cycles.

It is possible to use as much as 25 db. of clipping before intelligibility suffers; 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.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 to 12 db.) there is almost no change in "quality" but the voice power is increased considerably.

Before drastic clipping can be used, the speech signal must be amplified several 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.

One type of clipper-filter system is shown in block form in Fig. 9-12A. The clipper 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 or clipping 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" or modulation control. This control is set initially so that the

Speech Clipping

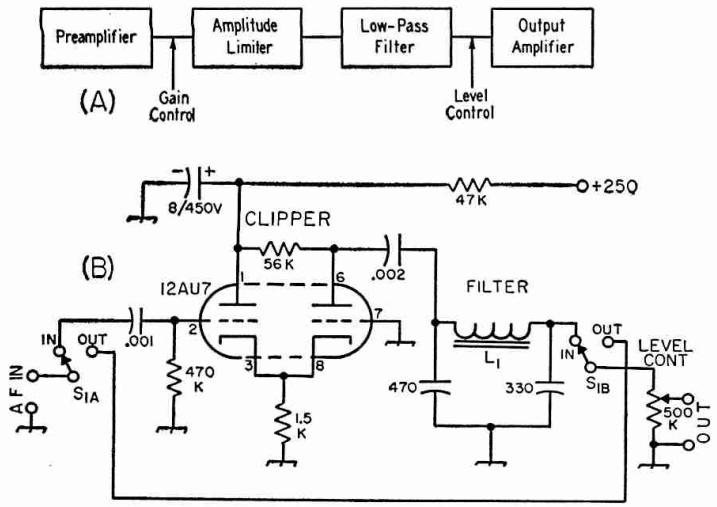


Fig. 9-12—(A) Block diagram of speech-clipping and filtering amplifier. (B) Practical speech clipper circuit with low-pass filter. Capacitances below 0.001 μf. are in μμf. Resistors are ½ watt.

L_1—20 henrys, 900 ohms (Stancor C-1515).
S_1—D.p.d.t. toggle or rotary.

amplitude-limited output of the clipper-filter cannot cause more than 100 per cent modulation.

It should be noted that the peak amplitude of the audio wave form actually applied to the modulated stage in the transmitter is not necessarily held at the same relative level as the peak amplitude of the signal coming out of the clipper stage. When the clipped signal goes through the filter, the relative phases of the various frequency components that pass through the filter are shifted, particularly those components near the cut-off frequency. This may cause the peak amplitude out of the filter to exceed the peak amplitude of the clipped signal applied to the filter input terminals. Similar phase shifts can occur in amplifiers following the filter, especially if these amplifiers, including the modulator, do not have good low-frequency response. With poor low-frequency response the more-or-less "square" waves resulting from clipping tend to be changed into triangular waves having higher peak amplitude. Best practice is to cut the low-frequency response *before* clipping and to make all amplifiers following the clipper-filter as flat and distortion-free as possible.

The best way to set the modulation control in such a system is to check the actual modulation percentage with an oscilloscope connected as described in the section on modulation. With the gain control set to give a desired clipping level with normal voice intensity, the level control should be adjusted so that the maximum modulation does not exceed 100 per cent no matter how much sound is applied to the microphone.

A practical clipper-filter circuit is shown in Fig. 9-12B. It may be inserted between two speech-amplifier stages (but after the one having the gain control) where the level is normally a few volts. The cathode-coupled clipper circuit gives some over-all voltage gain in addition to performing the clipping function. The filter constants are such as to give a cut-off characteristic that combines reasonably good fidelity with adequate high-frequency suppression.

High-Level Clipping and Filtering

Clipping and filtering also can be done at high level — that is, at the point where the modulation is applied to the r.f. amplifier — instead of in the low-level stages of the speech amplifier. In one rather simple but effective arrangement of this type the clipping takes place in the Class-B modulator itself. This is accomplished by carefully adjusting the plate-to-plate load resistance for the modulator tubes so that they saturate or clip peaks at the amplitude level that represents 100 per cent modulation. The load adjustment can be made by choice of output transformer ratio or by adjusting the plate-voltage/plate-current ratio of the modulated r.f. amplifier. It is best done by examining the output wave form with an oscilloscope.

The filter for such a system consists of a choke coil and capacitors as shown in Fig. 9-13. The values of L and C should be chosen to form a low-pass filter section having a cut-off frequency of about 2500 cycles, using the modulating impedance of the r.f. amplifier as the load resistance. For this cut-off frequency the formulas are

$$L_1 = \frac{R}{7850} \quad \text{and} \quad C_1 = C_2 = \frac{63.6}{R}$$

where R is in ohms, L_1 in henrys, and C_1 and C_2 in microfarads. For example, with a plate-modulated amplifier operating at 1500 volts and 200 ma. (modulating impedance 7500 ohms) L_1 would be 7500/7850 = 0.96 henry and C_1 or C_2 would be 63.6/7500 = 0.0085 μf. By-pass capacitors in the plate circuit of the r.f. amplifier should be included in C_2. Voltage ratings for C_1 and C_2 when connected as shown must be the same as for the plate blocking capacitor — i.e., at least twice the d.c. voltage applied to the plate of the modulated amplifier. L and C values can vary 10 per cent or so without seriously affecting the operation of the filter.

Besides simplicity, the high-level system has the advantage that high-frequency components

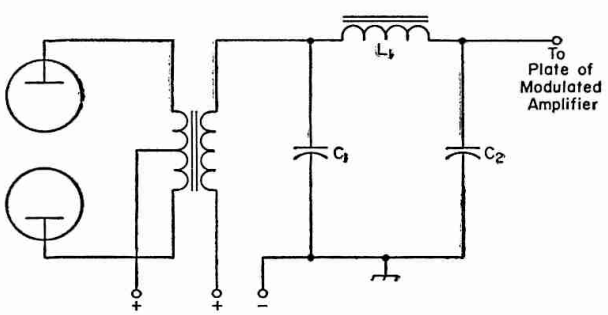


Fig. 9-13—Splatter-suppression filter for use at high level, shown here connected between a Class B modulator and plate-modulated r.f. amplifier. Values for L_1, C_1 and C_2 are determined as described in the text.

of the audio signal fed to the modulator grids, whether present legitimately or as a result of amplitude distortion in lower-level stages, are suppressed along with the distortion components that arise in clipping. Also, the undesirable effects of poor low-frequency response following clipping and filtering, mentioned in the preceding section, are avoided. Phase shifts can still occur in the high-level filter, however, so adjustments preferably should be made by using an oscilloscope to check the actual modulation percentage under all conditions of speech intensity. (For further discussion see Bruene, "High-Level Clipping and Filtering", *QST*, November, 1951.)

Low-Power Modulator

A modulator suitable for plate modulation of low-power transmitters or for screen or control-grid modulation of high-power amplifiers is pictured in Figs. 9-14 and 9-16. As shown in Fig. 9-15, it uses a pair of 6AQ5's in push-pull in the output stage. These are driven by a 6C4 phase inverter. A two-stage preamplifier using a 12AX7 brings the output voltage of a crystal or ceramic microphone up to the proper level for the 6C4 grid. A power supply is included on the same chassis.

The undistorted audio output of the amplifier is 7-8 watts. This is sufficient for modulating the plate of an r.f. amplifier running 10 to 15 watts input, or for modulating the control grids or screens of r.f. amplifiers using tubes having plate-dissipation ratings up to 250 watts. When screen modulation is used the screen power for the modulated amplifier (up to 250 volts) can be taken from the modulator power supply. The wiring shown in Fig. 9-15 provides for this, through an adjustable tap on the 25,000-ohm bleeder resistor, R_5, in the power supply. If a separate screen supply is used, or if the modulator is used for grid-bias or plate modulation of an r.f. amplifier, the d.c. circuit should be opened at point "X" in Fig. 9-15.

The amplifier uses resistance coupling up to the output-stage grids. The first section, V_{1A}, of the 12AX7 has "contact-potential" bias. The gain control, R_1, is in the grid circuit of the second section, V_{1B}, of the 12AX7. Negative feedback from the secondary of the output transformer, T_1, is introduced at the cathode of this tube section. The feedback voltage is dependent on the ratio of R_2 to R_3, approximately, and with the constants given is sufficient to result in a considerable reduction in distortion along with improved regulation of the audio output voltage. The latter is important when the unit is used for modulating a screen or control grid, as described in the chapter on amplitude modulation.

The phase inverter is of the split-load type described earlier in this chapter. It drives the push-pull 6AQ5's in the power amplifier. The output transformer used in the power stage is a multitap modulation transformer suitable for any of the types of modulation mentioned above.

Capacitor C_1 across the secondary of the output transformer, T_1, is used to reduce the high-frequency response of the amplifier. Without it, self-oscillation is likely to occur at a high audio frequency (usually above audibility) because phase shift in the output transformer at the end of its useful frequency range causes the feedback to become positive.

The power supply uses a replacement-type transformer and choke with a capacitor-input filter. Voltage under the modulator and speech-amplifier load is 250. The decoupling resistance-capacitance networks in the plate circuits of V_{1A} and V_{1B} contribute additional smoothing of the d.c. for these low-level stages.

The unit includes provision for send-receive switching, S_1 being used for that purpose. S_{1B} can be used to control the r.f. section — for example, by being connected in parallel with the key used for c.w. operation. Simultaneously S_{1A} short-circuits the secondary of T_1 so the transformer will not be damaged by being left

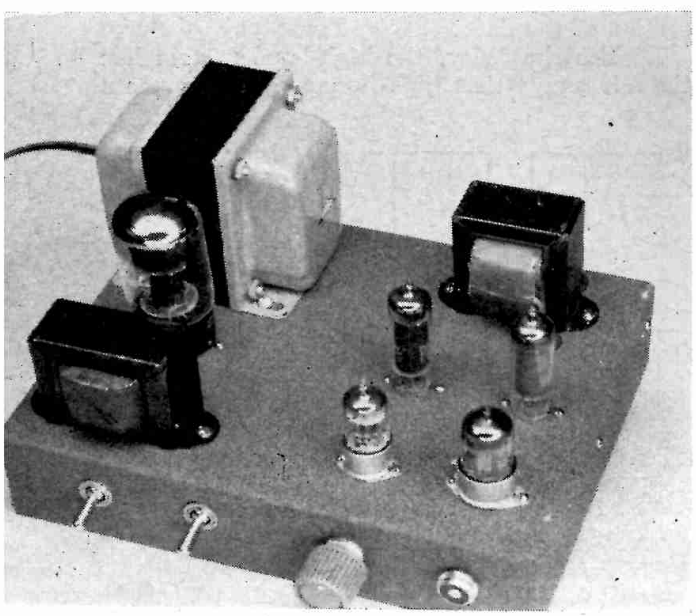


Fig. 9-14—Speech amplifier and low-power modulator suitable for screen or control-grid modulation of high-power amplifiers, or for plate modulation of an r.f. stage with up to 15 watts plate input. It is assembled on a 7 × 9 × 2-inch steel chassis, with the power supply occupying the left-hand section and the audio circuits the right. The 12AX7 preamplifier is at the lower right-hand corner, the 6C4 phase inverter is to its left, and the 6AQ5 power amplifiers are behind the two. Controls along the chassis edge are, left to right, the power switch, send-receive switch, gain control, and microphone jack.

A Low-Power Modulator

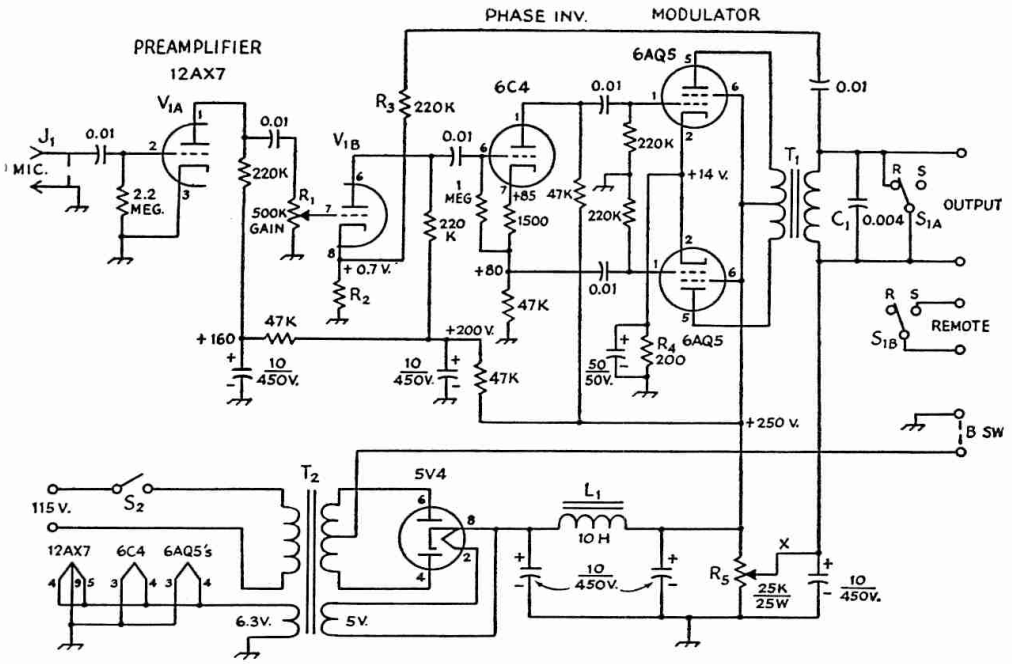


Fig. 9-15—Circuit of the speech amplifier and modulator. All capacitances are in μf.; capacitors with polarities marked are electrolytic, others are ceramic. Resistors are ½ watt except as noted below. Voltages measured to chassis with v.t. voltmeter.

J_1—Microphone connector (Amphenol 75-PC1M).
L_1—10 henrys, 90 ma. (Triad C-7X).
S_1—D.p.d.t. toggle.
S_2—S.p.s.t. toggle.
T_1—Modulation transformer, tapped secondary, primary 10,000 ohms plate to plate (Thordarson 21M68).
T_2—Power transformer, 525 v.c.t., 90 ma.; 6.3 v., 5 amp.; 5 v., 2 amp. (Triad R-10A).
R_2—1500 ohms, ½ watt.
R_4—App. 200 ohms, 2 watts (two 390-ohm 1-watt resistors in parallel).

without load. If S_{1B} is connected across the transmitter key, S_1 also can be used as a phone-c.w. switch, being left in the "R" position for c.w. operation.

The terminals marked "B Switch" should be short circuited (indicated by the dashed line) if S_1 is used as a send-receive switch. If a switch on the transmitter is used for send-receive, these terminals may be used for turning the plate voltage in the modulator on and off through an extra pair of contacts on the transmitter send-receive switch. In that case S_1 should be left in the "send" position for phone operation.

The proper secondary taps to use on T_1 will depend on the impedance of the load to which the amplifier is connected. Methods for determining the modulating impedance with various types of modulation are given in the section on amplitude modulation, together with information on connecting the modulator to the r.f. stage.

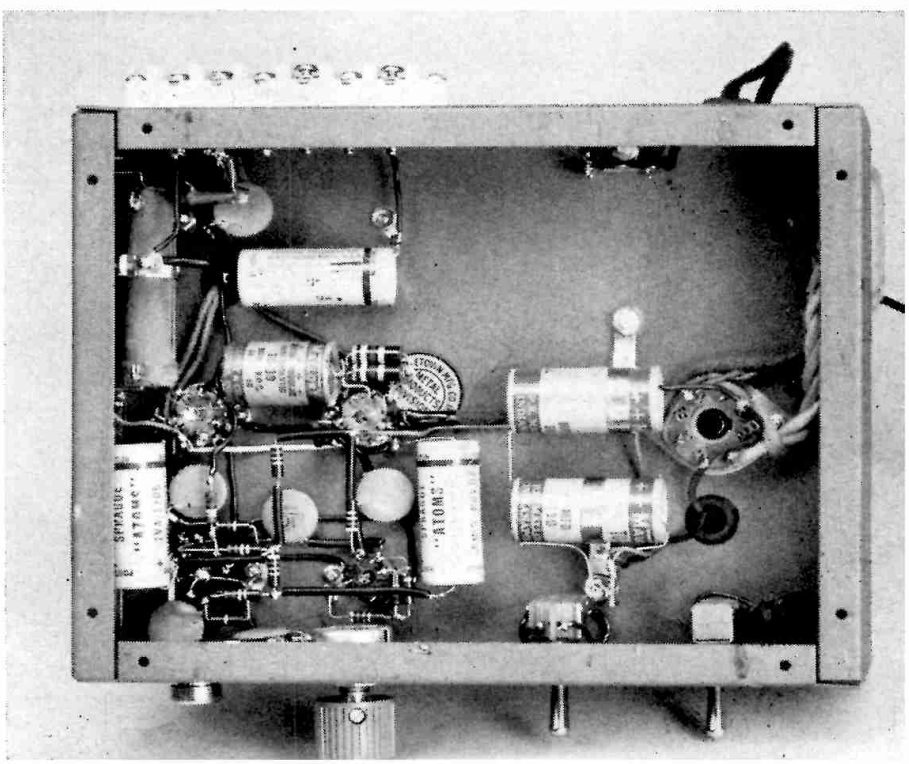


Fig. 9-16—Below-chassis view of the modulator. The rectifier-tube socket and electrolytic filter capacitors are at the right in this view. The 12AX7 socket is at the lower left. Bleeder resistor R_5 is at the upper left, near the 6-terminal connection strip on the rear edge of the chassis. Placement of components is not critical, but the leads in the first two stages should be kept short and close to the chassis to minimize hum troubles.

25-Watt Modulator using Push-Pull 6BQ6GTs

The speech amplifier-modulator shown in Figs. 9-17 to 9-19, inclusive, can be used for plate modulation of low-power transmitters running 25 to 50 watts input to the final stage. The circuit as shown is capable of an audio output of 25 watts, but this can be increased to 30 watts by a simple modification. The 6BQ6s in the output stage are operated in Class AB_1. Inexpensive receiver-type replacement components are used throughout, except for the modulation transformer.

Circuit

The speech amplifier uses a pentode first stage resistance-coupled to a triode second stage. This combination gives sufficient gain for a crystal microphone. The pentode and triode are the two sections of a dual tube, the 6AN8. Transformer coupling is used between the triode and the modulator tubes, in order to get push-pull voltage for the 6BQ6GT grids. Cathode bias is used on the final stage.

The coupling capacitance between the first and second stages is purposely made small to reduce the low-frequency response, and the primary of the output transformer is shunted by C_2 to reduce the amplification at the high-frequency end. C_1, on the first stage, also tends to reduce high-frequency response in addition to bypassing any r.f. that might be picked up on the microphone cord. These measures confine the frequency response to the most useful portion of the voice range.

S_2 is the "send-receive" switch. One section opens the power transformer center tap, thus cutting off the plate voltage during receiving periods. The other section can be connected to the key terminals on the transmitter, as indicated in the circuit diagram, to turn the transmitter on and off along with the modulator. If the transmitter is one in which the oscillator is not keyed, S_{2B} may be used to control the transmitter plate voltage, usually by being connected in the 115-volt circuit to the plate-supply transformer.

The "phone-c.w." switch, S_3, short-circuits the secondary of the modulation transformer, T_3, when the transmitter is to be keyed, and also opens the center-tap of T_1 so plate voltage cannot be applied to the modulator.

The power supply uses a receiver replacement-type transformer with a capacitor-input filter. Additional filtering for the speech-amplifier stages is provided by the 10-μf. capacitors and the series resistors in the plate circuits. Hum is also reduced by the VR-150 used to regulate the modulator screen voltage. Note that the regulator tube is connected between the screens and cathodes so that the actual screen voltage is 150 and is not reduced by the drop in the cathode bias resistor. Maintaining full screen voltage is important if the rated output is to be secured.

Operating

The 6BQ6GT amplifier requires a plate-to-plate load of 4000 ohms, and the output transformer ratio must be chosen to reflect this load to the plates (see later section on matching a modulator to its load). For most small transmitters running 30 to 50 watts input to the final stage a 1-to-1 transformer ratio will be satisfactory, since the modulating impedance of such transmitters usually is in the neighborhood of 4000 ohms. The secondary of T_3 is connected in series with the d.c. lead to the plate (and screen, if a screen-grid tube) of the Class C amplifier to be modulated. For further details, see the chapter on amplitude modulation.

For checking the modulator operation a milliammeter (0–200 range satisfactory) may be connected in the lead to the center-tap of the

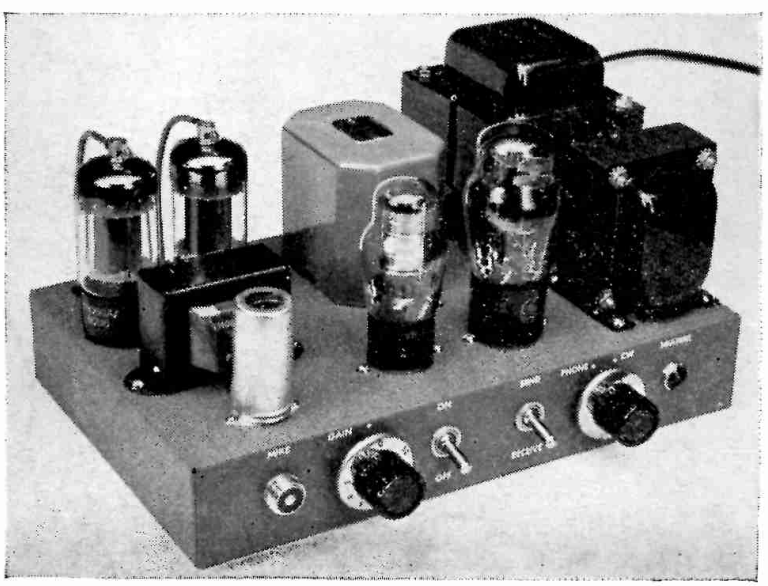


Fig. 9-17—A modulator for transmitters operating at plate inputs up to 50 watts. The speech amplifier and modulator are at the left in this view; power supply components are at the right. The chassis is 7 × 11 × 2 inches.

25-Watt Modulator

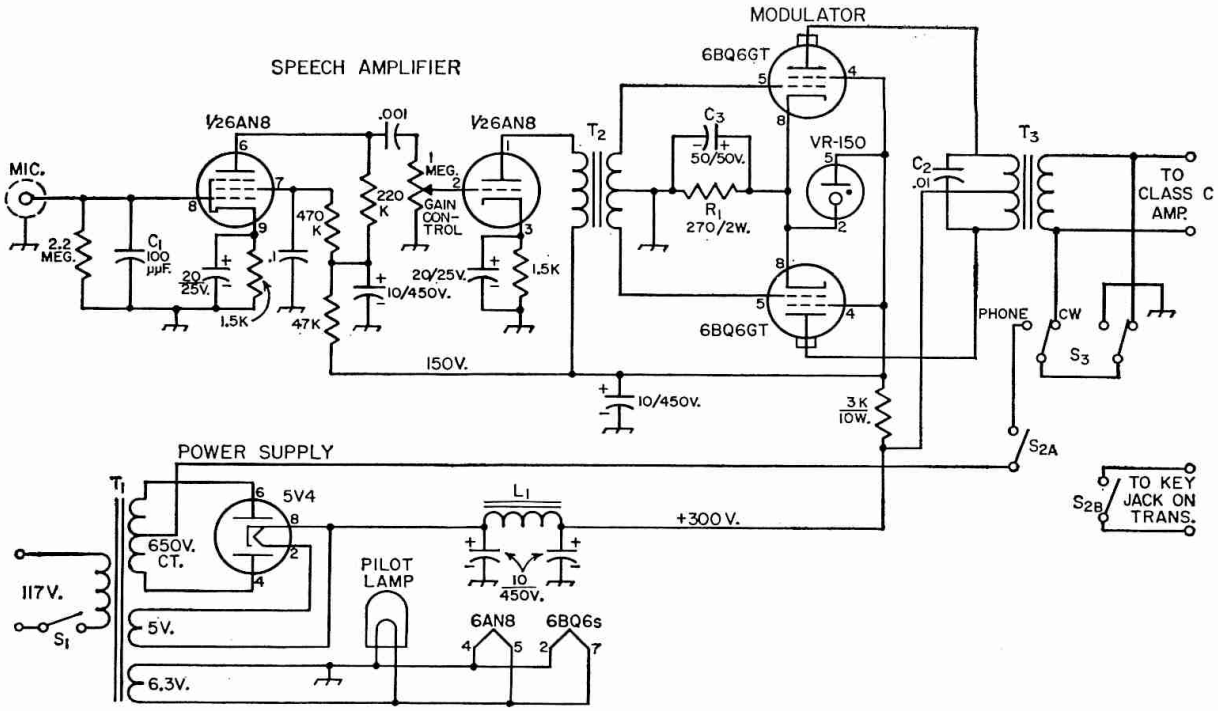


Fig. 9-18—Circuit diagram of the 25-watt modulator. Capacitances below 0.001 μf. are in μμf. Capacitors up to 0.01 μf. are ceramic. Resistors are ½ watt unless otherwise specified.

L1—8 henrys, 150 ma.
S1—S.p.s.t. toggle.
S2—D.p.d.t. toggle.
S3—2-pole 2-position rotary (Centralab PA-2003).
T1—Power transformer, 650 volts c.t., 150 ma. 5 volts 3 amp.; 6.3 volts, 5 amp.
T2—Interstage audio, single plate to p.p. grids, pri, to total sec. ratio 1 to 3.
T3—Modulation transformer, multimatch type (UTC S-19).

primary of T_3. Without voice input to the microphone the plate current should be approximately 50 ma. When modulating the transmitter, the current should "kick" to 60 or 70 ma.; this will usually represent 100 per cent modulation. If the amplifier can be tested with a single-tone signal replacing the microphone, the plate current will be about 165 ma. at full output.

The audio power output can be increased to about 30 watts, sufficient for modulating an 807 at its full phone rating, if the 6BQ6GT cathodes are grounded and bias of about 30 volts from a fixed source such as a small battery is applied to the grids. The battery may be substituted for the cathode resistor if the ground connection is moved from the center tap of the secondary of T_2 to the cathodes of the 6BQ6GTs.

(From *QST*, December, 1955.)

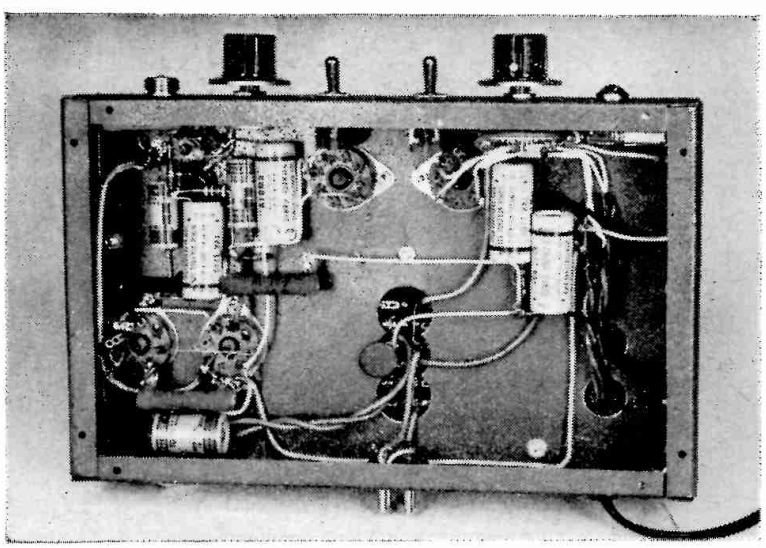


Fig. 9-19—Under-chassis view of the 6BQ6GT modulator. The two large capacitors at the right are the filter capacitors in the power supply. The modulator bias resistor and bypass capacitor (R_1C_3) are at lower left. Leads from the modulation transformer go through the three holes in the chassis. Shielded wire is used for heater, microphone input, and gain-control leads.

Class AB_1 Modulator Using 807s

The modulator unit shown in Figs. 9-20 to 9-22, inclusive, uses a pair of 807s as Class AB_1 power amplifiers. Its audio power output depends on the plate voltage applied to the 807s; approximate values and optimum plate-to-plate load resistances are as follows:

Plate Voltage	*Plate-to-Plate Load*	*Power Output*
400	6200 ohms	30 watts
500	8000 ohms	40 watts
600	9800 ohms	45 watts
750	12,500 ohms	60 watts

The power-output figures are conservative, and will vary somewhat with the losses in the output transformer. These in turn may vary with different combinations of tap connections. The nominal tube output (without transformer losses) is 20 to 25 per cent higher than the figures given.

The modulator is intended for use with an external plate and screen supply for the 807s, but includes a screen regulator circuit. The unit has a built-in power supply for the speech amplifier section. Fixed bias for the 807s is taken from this supply.

Speech Circuit

The speech amplifier uses a high-μ dual triode as a two-stage resistance-coupled amplifier, followed by a medium-μ triode. The latter is transformer-coupled to the modulator grids. The gain from the microphone input to the 807 grids is more than ample for crystal microphones and others of similar output level.

The frequency response of the amplifier is adjusted to put maximum energy in the range where it contributes most to speech intelligibility; that is, the output is highest between 500 and 1200 cycles and drops off gradually on either side. The lower frequencies are reduced by using low values of coupling capacitance between the resistance-coupled stages, and the high-frequency end is attenuated by C_1. Further high-frequency attenuation, particularly for such components generated in the modulator itself, is provided by capacitor C_2, connected across the output terminals of the modulation transformer.

Power Supply

The plate-supply requirements of the 12AX7 and 6C4 in the speech amplifier are quite small and easily can be supplied by a small "TV booster" type transformer, T_3. As shown in the diagram, a half-wave selenium rectifier works into a capacitor-input filter from this transformer. Bias for the 807s is obtained from this supply by making the output current flow through R_2 and R_3 in series, these resistors being connected between the negative output terminal of the supply and ground so that a negative voltage is developed with respect to chassis. The bias is adjustable by varying R_2. A single variable resistor having a total resistance of 10,000 ohms can be used instead of the two 5000-ohm units in series; the adjustment becomes somewhat more critical with the larger resistor but the operation is otherwise the same.

Heater power for the speech amplifier and modulator tubes is supplied by a separate filament transformer, T_4.

Plate power for the 807s is intended to be taken from an external source at a voltage level suitable for the output power desired. Screen voltage for the 807s comes from the same source, but is regulated at 300 volts by means of two 0A2 voltage-regulator tubes in series. Such regulation is essential for proper operation of the modulator tubes. The current through the 0A2s should be adjusted to 25 to 30 ma., with no signal on the

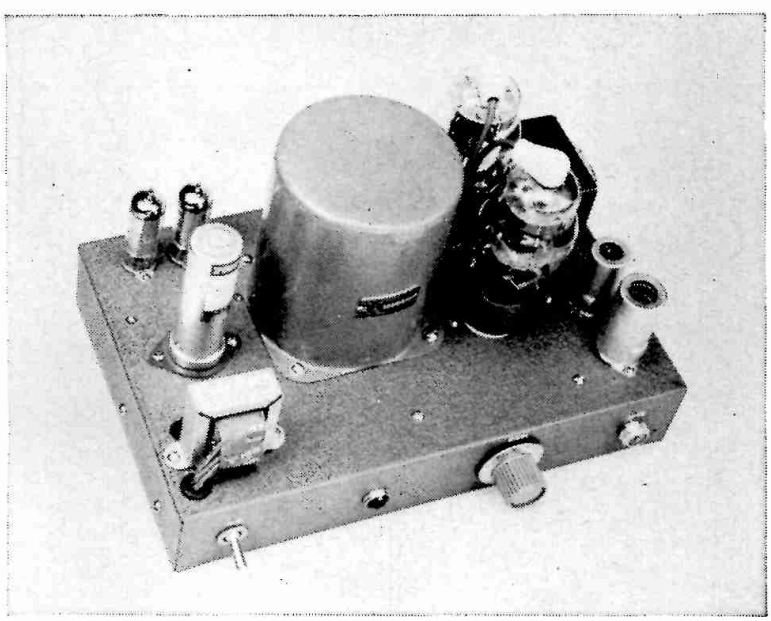


Fig. 9-20—Speech amplifier and modulator using Class AB_1 807s. Depending on plate voltage used, audio power outputs up to at least 60 watts may be obtained.

807 Modulator

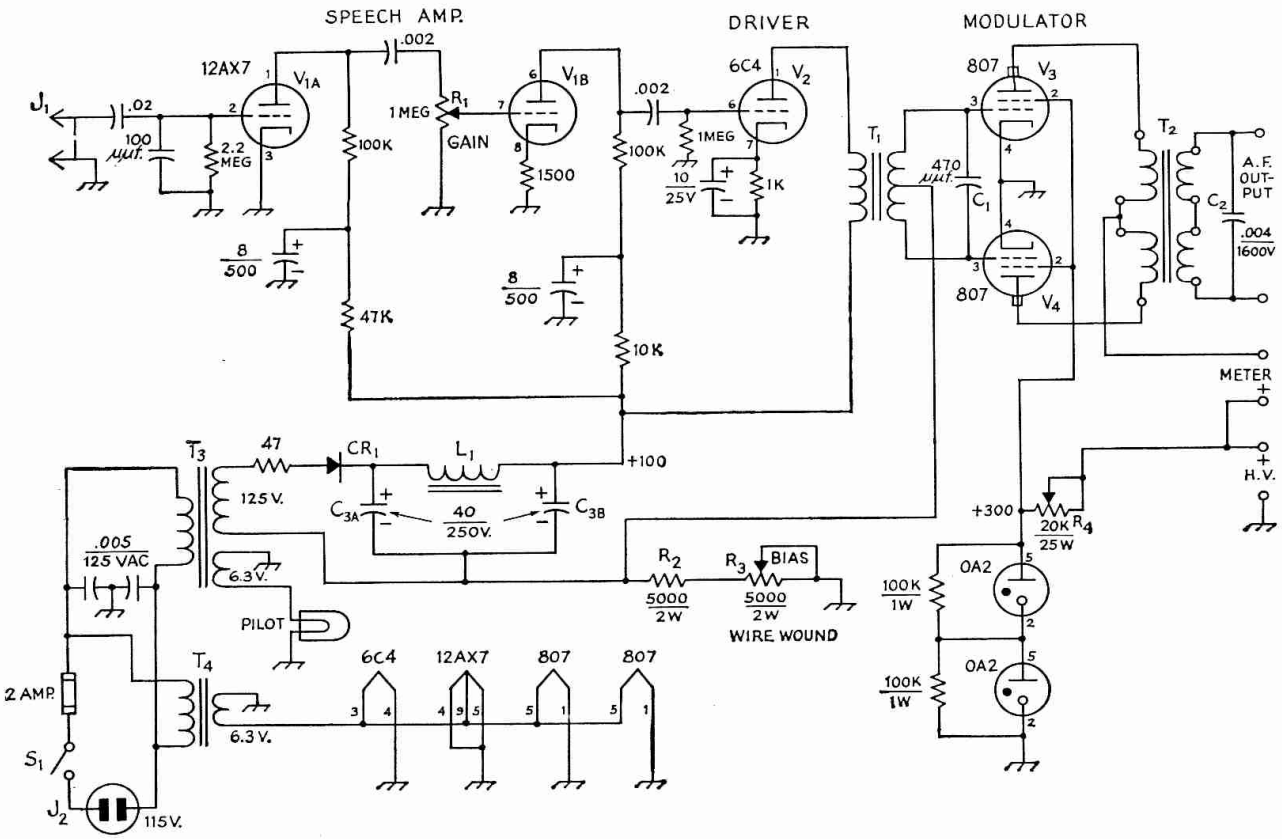


Fig. 9-21—Circuit diagram of the 807 modulator. Capacitances are in μf. unless otherwise specified; electrolytics are marked with polarity; others may be either ceramic or paper. Resistors are ½ watt except as indicated.

C_1—470-μμf. mica or ceramic.
C_2—App. 0.005 μf., 1600 volts (see discussion on modulators earlier in this chapter).
C_3—Dual 40-μf. electrolytic, 250 volts. Must be type that can be insulated from chassis.
CR_1—Selenium rectifier, 20-ma. or higher rating, 130 volts.
J_1—Chassis-type microphone connector (Amphenol 75-PC1M).
L_1—10 henrys, 50 ma. (Triad C-3X).
R_1—1-megohm control, audio taper.
R_2—5000 ohms, 2 watts.
R_3—5000-ohm wire-wound control, 2 watts.
R_4—20,000-ohm adjustable wire-wound, 25 watts.
S_1—S.p.s.t. toggle.
T_1—Interstage audio transformer, single plate to push-pull grids; 10-ma. primary; 3-to-1 turns ratio, total secondary to primary (Merit A-2914).
T_2—Multimatch modulation transformer, 30-watt rating adequate for voice work (UTC CVM-1).
T_3—Power transformer, 125 volts at 15 ma.; 6.3 volts at 0.6 amp. (Stancor PS8415).
T_4—Filament transformer, 6.3 volts at 3 amp.

grids of the 807s, by setting the slider on the 20,000-ohm adjustable resistor, R_4.

A pair of terminals is provided for connecting a d.c. milliammeter (0–200 ma. range is suitable) in series with the 807 plates for measuring plate current. Such a meter is useful as a check on the operation of the modulator during initial testing, and as a modulation indicator during actual operation. If a meter is not used the meter terminals should be connected together through a jumper.

Construction

The modulator shown is built on a 7 × 11 × 2-inch steel chassis, but other chassis sizes and layouts may be used if the builder prefers. The principal constructional precaution to be observed is to keep the modulation transformer, T_2, reasonably well separated from the low-level speech components so stray coupling between the wiring of these stages is minimized. The interstage transformer, T_1, should not be mounted too close to the power transformer, T_3, since there is a possibility of hum pickup in T_1 if these two units are close together.

It is necessary to cut a large hole — about 3 inches in diameter — for mounting the particular type of modulation transformer used in the unit shown. The connection terminals on this transformer are lugs on the bottom of the case, so the chassis opening must be large enough to permit making connections without danger of a short-circuit to chassis.

In wiring the speech-amplifier section, the leads to grids and plates should be kept short and separated as much as possible from heater wiring. The heater leads should be run along a fold in the edge of the chassis except where they must be brought out to reach the tube sockets. In this unit shielded wire was used for the heater wiring, but this is not necessary as a hum-reducing precaution. The principal reason here was mechanical; the shielded wire stays in place better and the shields can be "tacked" together with a spot of solder as a simple method of cabling.

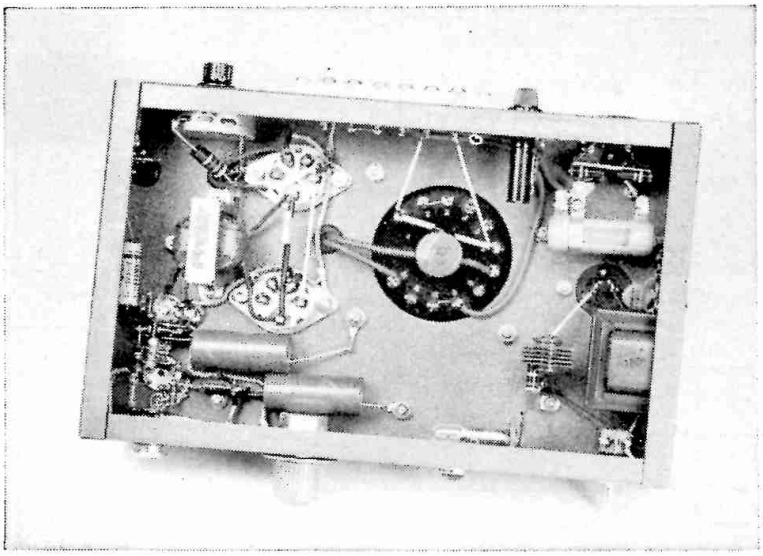


Fig. 9-22—Below-chassis view of the 807 modulator.

In the top view, Fig. 9-20, the speech-amplifier section is along the right-hand edge of the chassis. The tube near the front is the 12AX7 dual-triode amplifier. The 6C4 driver is just behind it, and the filament transformer, T_4, is on the rear right-hand corner. The modulation transformer is in the left center alongside the 807 modulator tubes. Along the left edge of the chassis are the power transformer, T_3, the dual filter capacitor, C_3, and the two gas regulator tubes. The negative terminal of C_3 must be insulated from the chassis; the capacitor shown is a "twist-lok" type with a bakelite socket.

In the below-chassis view, Fig. 9-22, the power-supply components are at the right. L_1 is mounted on the right-hand wall of the chassis, with the selenium rectifier, CR_1, just to its left. The dropping resistor for the VR-regulated screen circuit is near the upper right corner, close to the 0A2 sockets. The 115-volt socket and fuse are on the chassis wall near the regulator tubes. The speech-amplifier section is at the lower left in this view, with components laid in as convenient. The interstage audio transformer, T_1, is mounted between the 807 sockets at the left. The control on the top wall is R_3, for setting the grid bias on the 807s. Audio output, high voltage, and meter connections are made through the terminal strip (Millen 37306) between the fuse and bias control.

Operating Notes

The speech amplifier section may be tested independently of the modulator, since it has its own power supply. Testing may be done as described later in the chapter, preferably with an audio oscillator and oscilloscope to check wave form.

The modulation transformer taps to be used will depend on the plate-to-plate load resistance required for the desired power output and on the modulating impedance of the r.f. amplifier. The chart furnished with the transformer should be consulted for this information.

If the Class-C amplifier plate supply has the proper voltage and has sufficient excess capacity to furnish an average current of 70 to 100 ma. in addition to its normal Class C load, it may be used for this modulator as well. If not, a separate supply of conventional design (see chapter on power supply) may be used. It should have a choke-input filter and should have a minimum output capacitance of about 10 μf. for good dynamic regulation.

Before attempting to test the modulator, remove the 807s from their sockets and adjust R_4 (shut off the voltage before making each adjustment) for a current of 25 to 30 ma. through the VR tubes. The current may be measured by connecting a milliammeter of suitable range in series with the positive high-voltage lead between the external power supply and this unit, since with the 807s out of their sockets the only current is that through R_4 and the VR tubes.

After R_4 is properly adjusted, replace the 807s and with R_3 at maximum resistance (maximum bias) connect the plate milliammeter to the meter terminals. Then apply plate power and adjust R_3 for a plate current of 40 to 50 ma.; the value is not especially critical, but should not be too near cutoff and should not be so large as to cause the rated plate dissipation of the tubes to be exceeded. With the Class-C load connected the plate current should rise to approximately 140 ma. at full output, using a sine-wave signal. With voice input the current should kick to 65–75 ma. on peaks. These figures for plate current are the same regardless of the plate voltage used, so long as the screen is maintained at 300 volts.

If c.w. as well as phone operation is to be employed, provision should be made either in the modulator or the r.f. unit for short-circuiting the secondary of the modulation transformer when the transmitter is being keyed.

6146 Modulator and Speech Amplifier

The modulator shown in Figs. 9-23 to 9-25, inclusive, uses a pair of 6146s in AB_1, and is complete with power and bias supplies on a 10 × 17 × 3-inch chassis. The modulator also is equipped with an audio take-off for scope monitoring.

The audio power that can be obtained (based on measurements) is as follows:

Nominal Plate Voltage	*Power Output*	*Plate-to-Plate Load Resistance*
500 volts	75 watts	4200 ohms
600 volts	95 watts	5200 ohms
750 volts	120 watts	6700 ohms

Suitable sets of components for all three of the voltages listed above are readily available, so the power level can be selected to suit the Class C amplifier to be modulated. The modulator shown in the photographs is set up for 750-volt operation, but aside from the power and modulation transformers all components are the same regardless of the voltage level.

Audio Circuits

As shown in the circuit diagram, Fig. 9-24, the audio system consists of a 12AX7 preamplifier with the two tube sections in cascade, followed by a 6C4 voltage amplifier which is transformer-coupled to the grids of the Class AB_1 modulator tubes. The combination provides ample gain for a communications-type crystal, ceramic, or dynamic microphone.

The first stage of the amplifier is "contact-potential" biased, and is resistance-coupled to the second stage. The gain control, R_1, is in the grid circuit of the second stage. Decoupling resistors and capacitors are included in the plate-supply circuits of these two stages; these decoupling circuits also provide additional plate-supply hum filtering for the two low-level stages.

The secondary of T_1, the transformer coupling the third speech stage to the modulator grids, is shunted by a 470-$\mu\mu$f. capacitor to reduce high-frequency response. The optimum value of capacitance will depend on the particular type of audio transformer selected, as well as on the high-frequency characteristics of the microphone employed. Different values should be tried with the object of cutting the high-frequency response as much as possible, consistent with intelligibility.

The modulation transformer is of the multimatch type, and the taps should be selected to reflect the proper plate-to-plate load impedance, as given earlier, for the desired power output. The impedance ratio, secondary to primary, will depend on the modulating impedance of the modulated r.f. amplifier, as described earlier in this chapter. The secondary of the modulation transformer is shunted by C_1 to reduce output at the higher audio frequencies, particularly for attenuating high-frequency harmonics that might be generated in the modulator at high output levels. The value suggested (0.005 μf.) is an average figure and should be modified according to the modulating impedance of the Class-C stage as discussed earlier in this chapter.

Power Supply

Plate power for all tubes in the unit is supplied by a single power transformer. Mercury-vapor rectifiers are used because good voltage regulation is desirable. The filter is a single section with choke input and a large (over 25 μf.) output capacitance. The filter capacitor consists of three 80-μf. 450-volt electrolytic capacitors in series for 750-volt d.c. output. If the output voltage is 600 or less only two capacitors in series will be needed. These capacitors are shunted by 0.1-megohm resistors to help equalize the d.c. voltages across them.

The 200-volt (approximately) supply for the 6146 screens and the plates of the speech-amplifier tubes is taken from the main supply through a dropping resistor, and is regulated by two 0B2 voltage-regulator tubes in series. A 20-μf. ca-

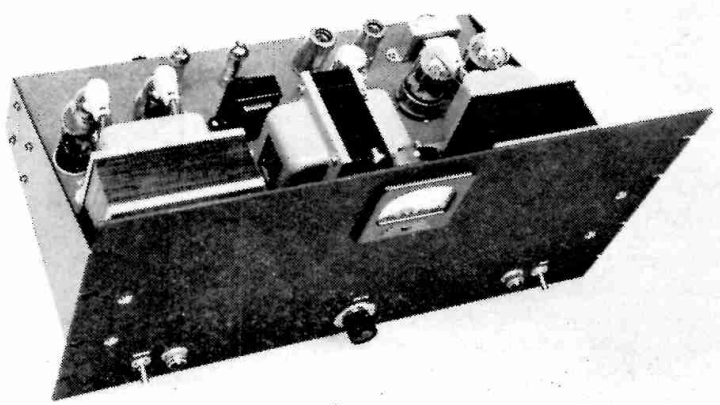


Fig. 9-23—Class-AB_1 modulator using 6146s, complete with speech amplifier and power supply. The relay-rack panel is 10½-inches high. Plate- and filament-supply primary switches, each with its own pilot lamp, are near the lower edge of the panel. The gain control is at lower center. Along the front of the chassis, just behind the panel, are the plate power transformer, filter choke, and modulation transformer, going from left to right. The tubes at the left are the 816 rectifiers, with the 6146s at the right. Along the rear edge are the two voltage-regulator tubes, the 12AX7 and 6C4 speech amplifier tubes, and the interstage audio transformer, T_1.

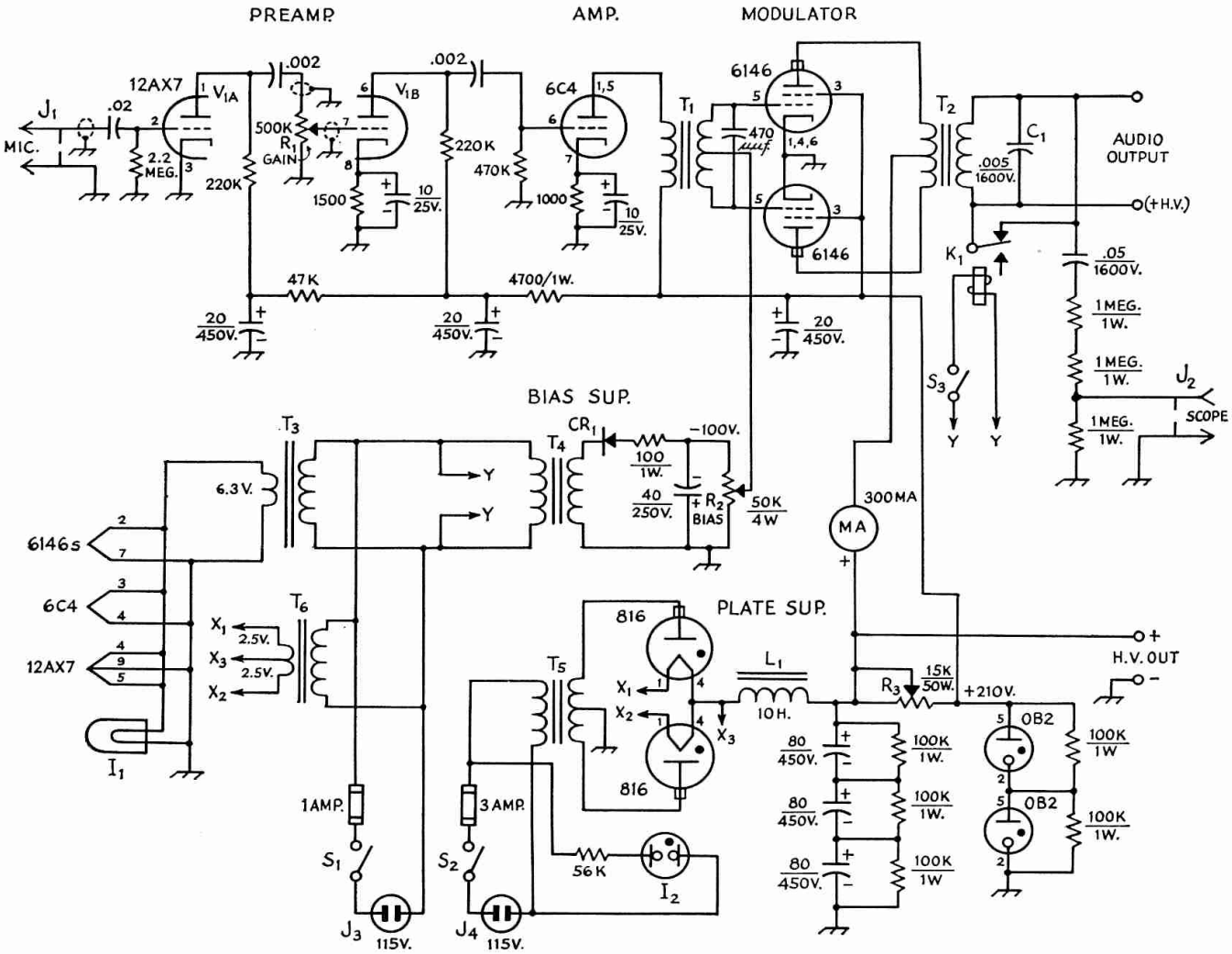


Fig. 9-24—Circuit diagram of the 6146 modulator and power supply. Capacitances are in μf. unless indicated otherwise; capacitors marked with polarity are electrolytic, others may be paper or ceramic as convenient. Resistances are in ohms; resistors are ½ watt except as indicated.

C_1—See text.
CR_1—Selenium rectifier, 20 ma. or higher rating, 130 volts.
I_1—6.3-volt pilot lamp.
I_2—Neon lamp, NE-51.
J_1—Microphone connector (Amphenol 75-PC1M).
J_2—Phono jack.
J_3, J_4—115-volt chassis-mounting plug (Amphenol 61-M1).
K_1—Antenna changeover relay, 115-volt coil (Advance AH/2C/115VA; type AM also suitable).
L_1—Filter choke, 10 henrys, 300 ma. (Triad C-19A).
R_1—0.5-megohm control, audio taper.
R_2—50,000-ohm wire-wound control, 4 watts.
R_3—15,000-ohm adjustable, 50 watts.
S_1, S_2—S.p.s.t. toggle.
S_3—S.p.s.t., mounted on R_1.
T_1—Interstage audio, single plate to p.p. grids, 3-to-1 secondary-to-primary ratio (Stancor A-63-C).
T_2—Multimatch modulation transformer, 125 watts (Triad M-12AL).
T_3—Filament transformer, 6.3 volts at 4 amp. (Triad F-53X).
T_4—Power transformer, 117 volts at 20 ma.; 6.3-volt winding unused (Thordarson 26R32).
T_5—Plate transformer. For 500 volts d.c.: 1235 volts c.t., 310 ma. (Triad P-7A); for 600 volts d.c.: 1455 volts c.t., 310 ma. (Triad P-11A). Transformer shown is for either 600 or 750 volts d.c. output at 310 ma.; sec. voltage 1780 c.t. for 750 volts (Triad P-14A).
T_6—Filament transformer, 5 volts at 3 amp., 2500-volt insulation (Stancor P-4088).

pacitor is connected across the VR tubes to improve the dynamic regulation in the 6146 screen circuit, since the peak instantaneous screen current exceeds the regulating capacity (30 ma.) of the VR tubes when the modulator is driven to maximum output.

Fixed bias for the 6146 grids is taken from a built-in bias supply using a TV "booster" transformer with a selenium rectifier. This bias is adjustable by means of R_2. The bias supply and filament transformer are on the same a.c. circuit so that bias is applied to the modulator grids whenever the tube heaters are energized.

Control and Auxiliary Circuits

The modulator includes an oscilloscope takeoff circuit consisting of the 0.05-μf. capacitor and three 1-megohm resistors in series. This can be

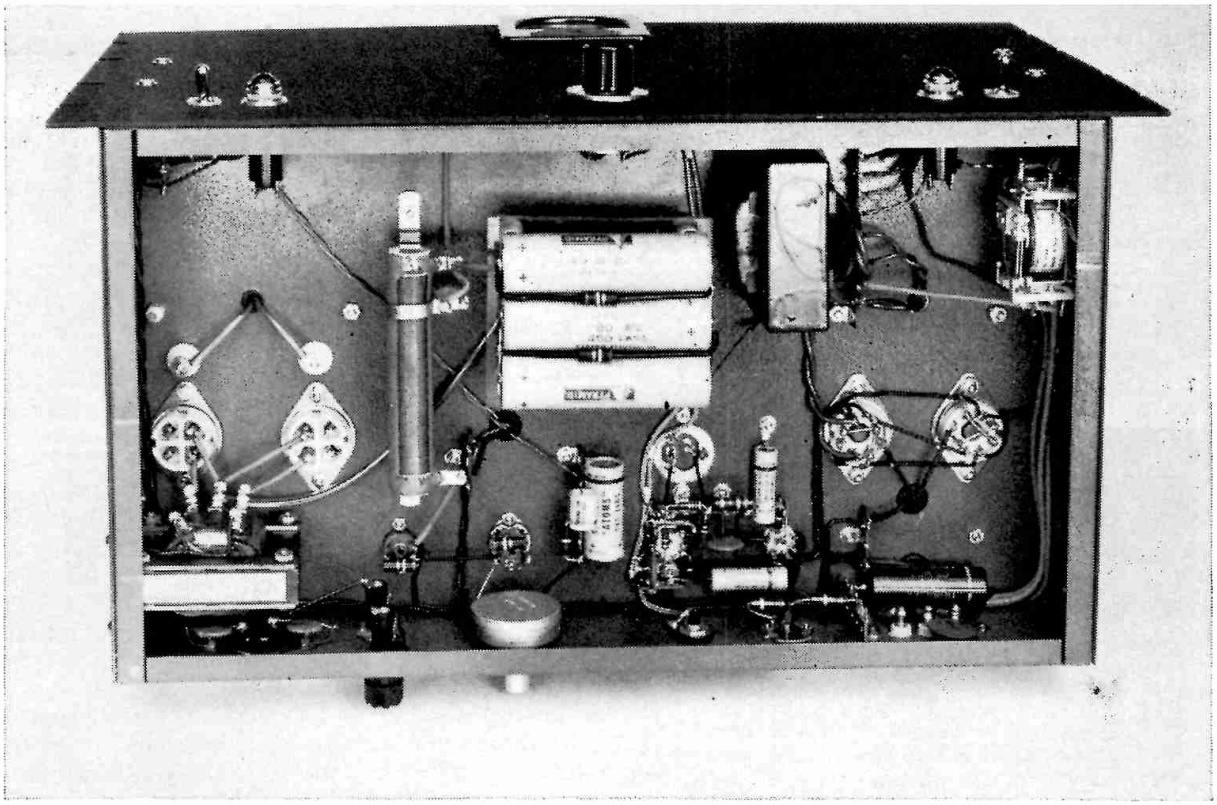


Fig. 9-25—Below-chassis view of the 6146 modulator. The 816 sockets and filament transformer (T_6) are at the lower left. The chassis wall at the bottom has on it, left to right, the 115-volt a.c. plugs, fuse holders, bias control (R_2), microphone input connector (J_1), scope take-off connector (J_2) and a three-terminal strip (Millen 37303) for audio output and positive high voltage connections. The high-voltage filter capacitor bank is in the center, mounted on a plate of plastic insulation which is supported away from the chassis on small pillars. The 6.3-volt transformer (T_3) is to the right of the capacitors. The antenna changeover relay used for shorting the modulation-transformer secondary is on the right-hand chassis wall.

used for horizontal deflection of a c.r. tube to give the trapezoidal modulation pattern (see chapter on amplitude modulation). Usually, it will be necessary to use an external control for adjusting the amplitude of the sweep voltage so obtained. If desired, a 1-megohm control can be substituted for the fixed resistor at the bottom of the string, thus avoiding the necessity for an external control.

The normally closed contacts of an antenna-type relay, K_1, are used to short-circuit the secondary of the modulation transformer when the transmitter is to be used for c.w. work. The switch, S_3, that controls the relay is mounted on the gain control, R_1, so that when the gain is turned all the way off, thus opening the switch, the relay contacts close. This insures that the modulator is inoperative and cannot be driven by accidental voice input (which would result in excessive plate current) when the transformer secondary is short-circuited.

Separate a.c. inputs are provided for the filament-bias and plate power circuits. The plate supply can thus be controlled by an external switch without disturbing the operation of the filament circuits or requiring a modification of the 115-volt wiring.

Terminals are provided for taking out high-voltage d.c. for an external unit. The power-supply equipment has more capacity than is needed by the modulator unit itself (the rating for amateur-type service is somewhat over 300 ma.) and may in some cases be sufficient for operation of the modulated r.f. amplifier as well. At least 200 ma. should be available for this purpose, since the average plate-supply current in the modulator unit alone is less than 100 ma., including the speech-amplifier and VR-tube drain.

Operating Data

The dropping resistor in the screen-supply circuit should be adjusted so that the current through 0B2s is 30 ma. with the bias on the 6146 grids adjusted so that the no-signal plate current is approximately 50 ma. The current through the VR tubes may be measured by temporarily opening the lead to the upper 0B2 at pin 5 and inserting a milliammeter of appropriate range.

If a sine-wave signal is used for testing the modulator, full output should be secured with a modulator plate current of approximately 240 ma. This value will be the same for all plate voltages, provided the screen voltage is maintained at approximately 200 volts and the values of plate-to-plate load resistance as specified earlier are used. With voice input the plate current will kick up to about 100 ma. on peaks, depending on the characteristics of the speaker's voice and those of the microphone used. This peak value should be determined under actual operating conditions with an oscilloscope, after which the plate milliammeter can be used as a modulation indicator.

Class B Modulator with Filter

Representative Class B modulator construction is illustrated by the unit shown in Figs. 9-26 and 9-28. This modulator includes a splatter

Fig. 9-26—A typical Class B modulator arrangement. This unit uses a pair of 811As, capable of an audio power output of 340 watts, and includes a splatter filter. The modulation transformer is at the left and the splatter choke at the right. All high-voltage terminals are covered so they cannot be touched accidentally.

filter, $C_1C_2L_1$ in the circuit diagram, Fig. 9-27, and also has provision for short-circuiting the modulation transformer secondary when c.w. is to be used.

The audio input transformer is not built into this unit, it being assumed that this transformer

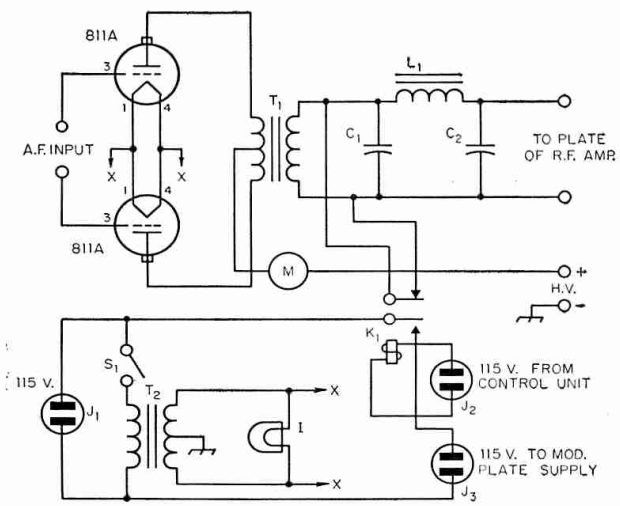


Fig. 9-27—Circuit diagram of the Class B modulator.
C_1, C_2, L_1—See text. (L_1 is Chicago Transformer type SR-300).
K_1—D.p.d.t. relay, high-voltage insulation (Advance type 400).
M—0–500 d.c. milliammeter, bakelite case.
T_1—Variable-ratio modulation transformer (Chicago Transformer type CMS-1).
T_2—Filament transformer, 6.3 v., 8 amp.
I_1—6.3-volt pilot light.
X_1, X_2—Chassis-type 115-volt plugs, male.
X_3—Chassis-type 115-volt receptacle, female.
S_1—S.p.s.t. toggle.

will be included in the driver assembly as is customary. If the modulator and speech amplifier-driver are mounted in the same rack or cabinet, the length of leads from the driver to the modulator grids presents no problem. The bias required by the modulator tubes at their higher plate-voltage ratings should be fed through the center tap on the secondary of the driver transformer. At a plate voltage of 1250 or less no bias is needed and the center-tap connection on the transformer can be grounded.

The values of C_1, C_2 and L_1 depend on the modulating impedance of the Class C r.f. amplifier. They can be determined from the formulas given in this chapter in the section on high-level clipping and filtering. The splatter filter will be effective regardless of whether the modulator operating conditions are chosen to give high-level clipping, but it is worth while to design the system for clipping at 100 per cent modulation if the tube curves are available for that purpose. The voltage ratings for C_1 and C_2 should at least equal the d.c. voltage applied to the modulated r.f. amplifier.

A relay with high-voltage insulation is used to short-circuit the secondary of T_1 when the

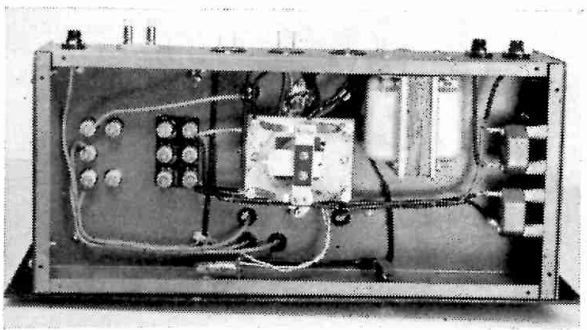


Fig. 9-28—The relay and filament transformer are mounted below the chassis. C_1, C_2 and K_1 are mounted on small stand-off insulators.

relay coil is not energized. A normally closed contact is used for this purpose. The other arm is used to close the primary circuit of the modulator plate supply when the relay is energized. Shorting the transformer secondary is necessary when the r.f. amplifier is keyed, to prevent an inductive discharge from the transformer winding that would put "tails" on the keyed characters and, with cathode keying of the amplifier, would cause excessive sparking at the key contacts. The control circuit should be arranged in such a way that K_1 is not energized during c.w. operation but is energized by the send-receive switch during phone operation.

Careful attention should be paid to insulation since the instantaneous voltages in the secondary circuit of the transformer will be at least twice the d.c. voltage on the r.f. amplifier. If a "hi-fi" amplifier of 10 watts or more output is available, it can be used as the driver for the 811As by coupling as shown in Fig. 9-29.

Checking Speech Equipment

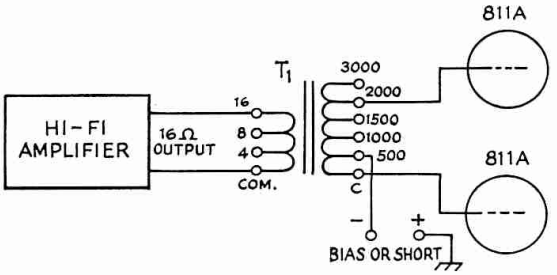


Fig. 9-29—A "hi-fi" audio amplifier will drive a Class-B modulator; a suitable coupling transformer is required. The connections shown here are for a pair of 811As. The amplifier should have an output rating of at least 10 watts.

T_1—10-watt line-to-voice-coil transformer (Stancor A-8104).

Checking Amplifier Operation

An adequate job of checking speech equipment can be done with equipment that is neither elaborate nor expensive. A typical setup is shown in Fig. 9-30. The construction of a simple audio oscillator is described in the chapter on measurements. The audio-frequency voltmeter can be either a vacuum-tube voltmeter or a multirange volt-ohm-milliammeter that has a rectifier-type a.c. range. The headset is included for aural checking of the amplifier performance.

An audio oscillator usually will have an output control, but if the maximum output voltage is in excess of a volt or so the output setting may be rather critical when a high-gain speech amplifier is being tested. In such cases an attenuator such as is shown in Fig. 9-30 is a convenience.

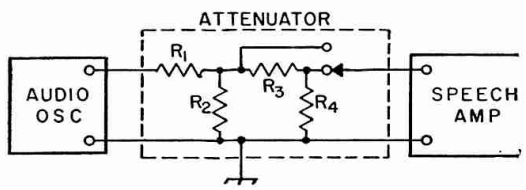


Fig. 9-30—Simple oscillator-attenuator test setup for checking a speech amplifier. It is not necessary that the frequency range of the audio oscillator be continuously variable; one or more "spot frequencies" will be satisfactory. Suitable resistor values are: R_1 and R_3, 10,000 ohms; R_2 and R_4, 1000 ohms.

Each of the two voltage dividers reduces the voltage by a factor of roughly 10 to 1, so that the over-all attenuation is about 100 to 1. The relatively low value of resistance, R_4, connected across the input terminals of the amplifier also will minimize stray hum pickup on the connecting leads.

The output of a power amplifier such as a modulator or driver for a Class B stage may be checked by using a resistance load of the rated value for the amplifier. A useful circuit arrangement is shown in Fig. 9-31. The load resistance, R_1, may be a single adjustable unit of appropriate power rating or may be made up of several resistors in series or parallel to give the required resistance. If measurement of the resistance is necessary an ohmmeter will be sufficiently accurate. In the case of a multimatch output transformer the taps should be those that will actually be used with the Class C amplifier with which the modulator is intended to work. R_1 then should have a value equal to the modulating impedance of the r.f. amplifier.

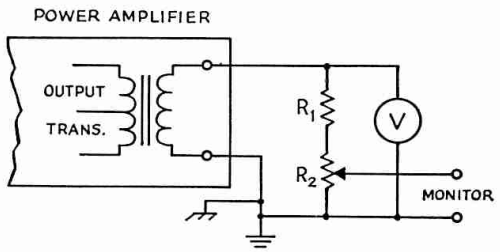


Fig. 9-31—Circuit for measuring power and making qualitative checks of the amplifier output. Values to be used for R_1 and R_2 are discussed in the text. The secondary winding of the output transformer in the amplifier should be disconnected from any d.c. source in the unit and one end connected to chassis as shown. An earth ground should be used on the system.

If an audio oscillator generating a good sine wave is used as the signal source the output power of the amplifier may be measured by an audio-frequency voltmeter as indicated by V. Either a vacuum-tube voltmeter on its a.c. scale or a rectifier-type a.c. voltmeter will be satisfactory, the principal requirements being relatively high impedance (1000 ohms per volt or more) and a reasonably accurate calibration. The power output will be equal to E^2/R_1, where E is the r.m.s. value of the voltage across the resistor (a.c. instruments usually are calibrated in r.m.s. values). This assumes that the distortion generated in the amplifier is small; if distortion is high, the voltmeter reading will be inaccurate.

If the amplifier is a driver for a Class B modulator, the value of R_1 should be calculated from R/N^2, where N is the turns ratio, primary to total secondary, of the class B input transformer, and R is the rated plate-to-plate load for the driver tube or tubes. R_1 should of course be connected across the total secondary in this case.

For a qualitative check on distortion, provision is made in Fig. 9-31 for monitoring the output of the amplifier. R_2 should be a wire-wound potentiometer having a resistance of 10 or 20 ohms. A headset may be connected to the "Monitor" terminals. Using the audio oscillator as a signal source, start with the gain control at minimum and then advance it slowly while listening carefully to the tone signal in the headset. When it begins to sound like a musical octave instead of a single tone, or when higher harmonically related tones can be heard along with the

desired one, distortion is starting to become appreciable. This effect usually will be detectable, but not serious, at full output of the amplifier as indicated by the voltmeter reading. Keep the signal in the headset at a moderate level by adjusting R_2 when necessary. If the amplifier passes the distortion test satisfactorily, reduce the audio input to zero and note whether any hum is audible in the headset. There should be none, if the tone level in the headset at full sine-wave output was no more than moderately high.

After completing these checks with satisfactory results, substitute the microphone for the oscillator input to the amplifier and have someone speak into it at a moderate level. The headset will serve to indicate the speech quality at various output levels. A tape recorder, if available, is useful at this stage since it can be substituted for the headset and will provide a means for comparing the effect of changes and adjustments in the amplifier as well as giving a better over-all check on speech quality than the average headset. The effect of measures taken to attenuate high- or low-frequency response in the amplifier is readily observed by comparing recordings made before and after changes. The output quality of the amplifier also can be compared with the original output of the microphone as registered on the recorder. In using a recorder care must be taken to set R_2 so that the first stage in the recorder amplifier is not overloaded. Use the normal gain setting of the recorder and adjust R_2 to give normal level indications.

Amplifier Troubles

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 to ground, starting with the output amplifier. 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, a defective tube, or inadequate plate-supply filtering, 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 and the amplifier tube. 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.

Using the Oscilloscope

Speech-amplifier checking is facilitated considerably if an oscilloscope of the type having amplifiers and a linear sweep circuit is available. A typical setup for using the oscilloscope is shown in Fig. 9-32. With the connections shown, the sweep circuit is not required but horizontal and vertical amplifiers are necessary. Audio voltage from the oscillator is fed directly to one oscilloscope amplifier (horizontal in this case) and the output of the speech amplifier is connected to the other. The scope amplifier gains should be adjusted so that each signal gives the same line length with the other signal shut off.

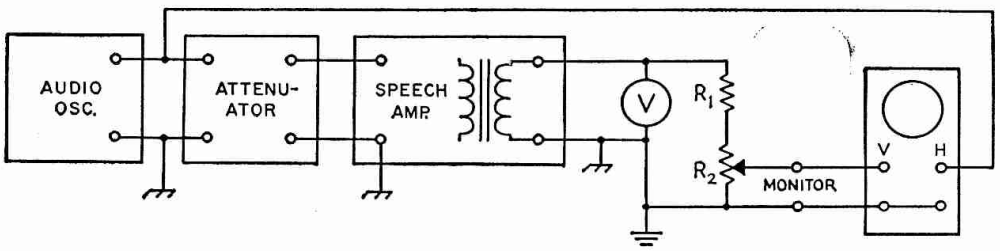


Fig. 9-32—Test setup using the oscilloscope to check for distortion. These connections will result in the type of pattern shown in Fig. 9-33, the horizontal sweep being provided by the audio input signal. For wave-form patterns, omit the connection between the audio oscillator and the horizontal amplifier in the scope, and use the horizontal linear sweep.

Under these conditions, when the input and output signals are applied simultaneously they are compared directly. If the speech amplifier is distortion-free and introduces no phase shift, the resulting pattern is simply a straight line, as shown at the upper left in Fig. 9-33, making an angle of about 45 degrees with the horizontal and vertical axes. If there is no distortion but there is phase shift, the pattern will be a smooth ellipse, as shown at the upper right. The greater the phase shift the greater the tendency of the ellipse to grow into a circle. When there is even-harmonic distortion in the amplifier one end of the line or ellipse becomes curved, as shown in the second row in Fig. 9-33. With odd-harmonic distortion such as is characteristic of overdriven push-pull stages, the line or ellipse is curved at both ends.

Patterns such as these will be obtained when the input signal is a fairly good sine wave. They will tend to become complicated if the input

wave form is complex and the speech amplifier introduces appreciable phase shifts. It is therefore advisable to test for distortion with an input signal that is as nearly as possible a sine wave. Also, it is best to use a frequency in the 500–1000 cycle range, since improper phase shift in the amplifier is usually least in this region. Phase shift in itself is not of great importance in an audio amplifier of ordinary design because it does not change the character of speech so far as the ear is concerned. However, if a complex signal is used for testing, phase shift may make it difficult to detect distortion in the oscilloscope pattern.

Since the oscilloscope amplifiers themselves may introduce phase shift and possibly distortion as well, it is advisable to check the scope before attempting to make checks on the speech amplifier. Apply the signal from the audio oscillator simultaneously to the horizontal and vertical amplifier input terminals. If both amplifiers have the same phase characteristics and negligible distortion the pattern, after suitable adjustment of the gains, will be a straight line as shown at the upper left in Fig. 9-33. If distortion is visible, note whether it changes when the scope gain controls are reduced; if not, the signal voltage from the audio oscillator is too great and should be reduced to the point where the input amplifiers are not overloaded. After finding the proper settings for signal input and scope gains, leave the latter alone in making checks on the speech equipment and adjust the input to the scope by means of R_2 and the output of the audio oscillator. Phase shift in the scope itself is not serious since the presence of distortion in the speech amplifier can be detected by the patterns shown at the right in Fig. 9-33.

In amplifiers having negative feedback, excessive phase shift within the feed-back loop may cause self-oscillation, since the signal fed back may arrive at the grid in phase with the applied signal voltage instead of out of phase with it. Such a phase shift is most likely to be associated with the output transformer. Oscillation usually occurs at some frequency above 10,000 cycles, although occasionally it will occur at a very low frequency. If the pass band in the stage in which the phase shift occurs is deliberately restricted to the optimum voice range, as described earlier, the gain at both very high and very low frequencies will be so low that self-oscillation is unlikely, even with large amounts of feedback.

Generally speaking, it is easier to detect small amounts of distortion with the type of pattern shown in Fig. 9-33 than it is with the wave-form pattern obtained by feeding the output signal to the vertical plates and making use of the linear sweep in the scope. However, the wave-form pattern can be used satisfactorily if the signal from the audio oscillator is a reasonably good sine wave. One simple method is to examine the output of the oscillator alone and trace the pattern on a sheet of transparent paper. The pattern given by the output of the amplifier can then be

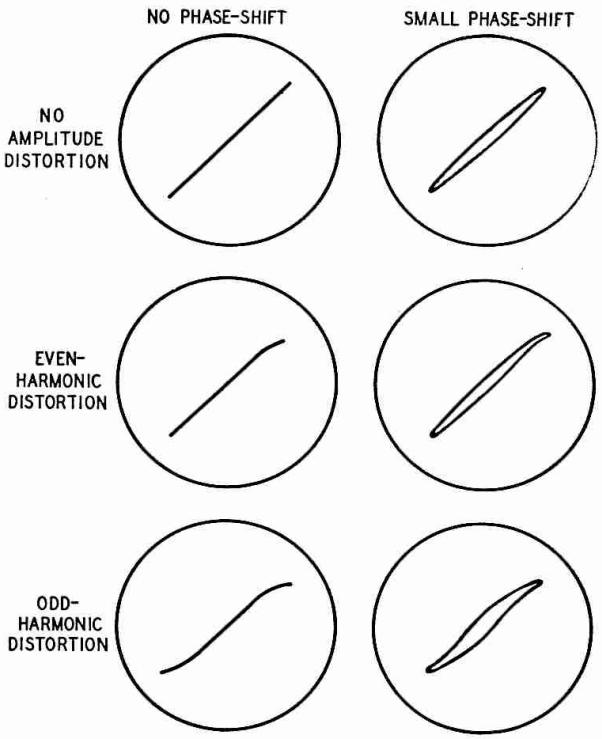


Fig. 9-33—Typical patterns obtained with the connections shown in Fig. 9-32. Depending on the number of stages in the amplifier, the pattern may slope upward to the right, as shown, or upward to the left. Also, depending on where the distortion originates, the curvature in the second row may appear either at the top or bottom of the line or ellipse.

compared with the "standard" pattern by adjusting the oscilloscope gains to make the two patterns coincide as closely as possible. The pattern discrepancies are a measure of the distortion.

In using the oscilloscope care must be taken to avoid introducing hum voltages that will upset the measurements. Hum pickup on the scope leads or other exposed parts such as the amplifier load resistor or the voltmeter can be detected by shutting off the audio oscillator and speech amplifier and connecting first one and then the other to the vertical plates of the scope, setting the internal horizontal sweep to an appropriate width. The trace should be a straight horizontal line when the vertical gain control is set at the position used in the actual measurements. Waviness in the line indicates hum. If the hum is not in the scope itself (check by disconnecting the leads at the instrument) make sure that there is a good ground connection on all the equipment and, if necessary, shield the hot leads.

The oscilloscope can be used to good advantage in stage-by-stage testing to check wave forms at the grid and plate of each stage and thus to determine rapidly where a source of trouble may be located. When the scope is connected to circuits that are not at ground potential for d.c., a capacitor of about 0.1 μf. should be connected in series with the hot oscilloscope lead. The probe lead should be shielded to prevent hum pickup.

Amplitude Modulation

As described in the chapter on circuit fundamentals, the process of modulation sets up groups of frequencies called **sidebands,** which appear symmetrically above and below the frequency of the unmodulated signal or **carrier.** If the instantaneous values of the amplitudes of all these separate frequencies are added together, the result is called the **modulation envelope.** In **amplitude modulation (a.m.)** the modulation envelope follows the amplitude variations of the audio-frequency signal that is being used to modulate the wave.

For example, modulation by a 1000-cycle tone will result in a modulation envelope that varies in amplitude at a 1000-cycle rate. The actual r.f. signal that produces such an envelope consists of three frequencies — the carrier, a side frequency 1000 cycles higher, and a side frequency 1000 cycles lower than the carrier. These three frequencies easily can be separated by a receiver having high selectivity. In order to reproduce the original modulation the receiver must have enough bandwidth to accept the carrier and the sidebands simultaneously. This is because an a.m. detector responds to the modulation envelope rather than to the individual signal components, and the envelope will be distorted in the receiver unless all the frequency components in the signal go through without change in their relative amplitudes.

In the simple case of tone modulation the two side frequencies and the carrier are constant in amplitude — it is only the envelope amplitude that varies at the modulation rate. With more complex modulation such as voice or music the amplitudes and frequencies of the side frequencies vary from instant to instant. The amplitude of the modulation envelope varies from instant to instant in the same way as the complex audio-frequency signal causing the modulation. Nevertheless, even in this case the *carrier* amplitude is constant if the transmitter is properly modulated.

A.M. Sidebands and Channel Width

Speech can be electrically reproduced, with high intelligibility, in a band of frequencies lying between approximately 100 and 3000 cycles. When these frequencies are combined with a radio-frequency carrier, the sidebands occupy the frequency spectrum from about 3000 cycles below the carrier frequency to 3000 cycles above — a total band or **channel** of about 6 kilocycles.

Actual speech frequencies extend up to 10,000 cycles or more, so it is possible to occupy a 20-kc. channel if no provision is made for reducing its width. For communication purposes such a channel width represents a waste of valuable spectrum space, since a 6-kc. channel is fully adequate for intelligibility. Occupying more than the minimum channel creates unnecessary interference. Thus speech equipment design and transmitter adjustment and operation should be pointed toward maintaining the channel width at the minimum.

● THE MODULATION ENVELOPE

In Fig. 10-1, the drawing at A shows the unmodulated r.f. signal, assumed to be a sine wave of the desired radio frequency. The graph can be taken to represent either voltage or current.

In B, the signal is assumed to be modulated by the audio frequency shown in the small drawing above. This frequency is much lower than the carrier frequency, a necessary condition for good modulation. When the modulating voltage is "positive" (above its axis) the envelope amplitude is increased *above* its unmodulated amplitude; when the modulating voltage is "negative" the envelope amplitude is *decreased.* Thus the envelope grows larger and smaller with the polarity and amplitude of the modulating voltage.

The drawings at C shows what happens with stronger modulation. The envelope amplitude is doubled at the instant the modulating voltage reaches its positive peak. On the negative peak of the modulating voltage the envelope amplitude just reaches zero; in other words, the signal is completely modulated.

Percentage of Modulation

When a modulated signal is detected in a receiver, the detector output follows the modulation envelope. 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. 10-1C would produce considerably more useful audio output 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. 10-1, X represents the unmodulated carrier amplitude, Y is the maximum envelope amplitude on the modulation up-peak, and Z is the minimum envelope amplitude on the modulation downpeak.

In a properly operating modulation system the modulation envelope is an accurate reproduction of the modulating wave, as can be seen in Fig. 10-1 at B and C by comparing one side of the outline with the shape of the modulating wave. (The lower outline duplicates the upper, but simply appears upside down in the drawing.)

The **percentage of modulation** is

$$\%\,\text{Mod.} = \frac{Y - X}{X} \times 100 \text{ (upward modulation), or}$$

$$\%\,\text{Mod.} = \frac{X - Z}{X} \times 100 \text{ (downward modulation)}$$

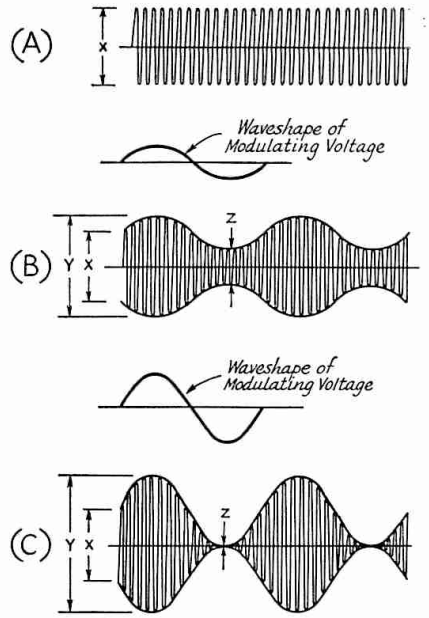


Fig. 10-1—Graphical representation of (A) r.f. output unmodulated, (B) modulated 50%, (C) modulated 100%. The modulation envelope is shown by the thin outline on the modulated wave.

If the wave shape of the modulation is such that its peak positive and negative amplitudes are equal, then the modulation percentage will be the same both up and down. If the two percentages differ, the larger of the two is customarily specified.

Power in Modulated Wave

The amplitude values shown in Fig. 10-1 correspond to current or voltage, so the drawings may be taken to represent instantaneous values of either. The power in the wave varies as the *square* of either the current or voltage, so at the peak of the modulation up-swing the instantaneous power in the envelope of Fig. 10-1C is four times the unmodulated carrier power (because the current and voltage both are doubled). At the peak of the down-swing the power is zero, since the amplitude is zero. These statements are true of 100 per cent modulation no matter what the wave form of the modulation. The instantaneous envelope power in the modulated signal is proportional to the square of its envelope amplitude at every instant. This fact is highly important in the operation of every method of amplitude modulation.

It is convenient, and customary, to describe the operation of modulation systems in terms of sine-wave modulation. Although this wave shape is seldom actually used in practice (voice wave shapes depart very considerably from the sine form) it lends itself to simple calculations and its use as a standard permits comparison between systems on a common basis. With sine-wave modulation the *average* power in the modulated signal over any number of full cycles of the modulation frequency is found to be 1½ times the power in the unmodulated carrier. In other words, the power output increases 50 per cent with 100 per cent modulation by a sine wave.

This relationship is very useful in the design of modulation systems and modulators, because any such system that is capable of increasing the *average* power output by 50 per cent with sine-wave modulation automatically fulfills the requirement that the *instantaneous* power at the modulation up-peak be four times the carrier power. Consequently, systems in which the additional power is supplied from outside the modulated r.f. stage (e.g., plate modulation) usually are designed on a sine-wave basis as a matter of convenience. Modulation systems in which the additional power is secured from the modulated r.f. amplifier (e.g., grid modulation) usually are more conveniently designed on the basis of peak envelope power rather than average power.

The extra power that is contained in a modulated signal goes entirely into the sidebands, half in the upper sideband and half in the lower. As a numerical example, full modulation of a 100-watt carrier by a sine wave will add 50 watts of sideband power, 25 in the lower and 25 in the upper sideband. Supplying this additional power for the sidebands is the object of all of the various systems devised for amplitude modulation.

No such simple relationship exists with complex wave forms. Complex wave forms such as speech do not, as a rule, contain as much average power as a sine wave. Ordinary speech wave forms have about half as much average power as a sine wave, for the same peak amplitude in both wave forms. Thus for the same modulation percentage, the sideband power with ordinary speech will average only about half the power with sine-wave modulation, since it is the peak envelope amplitude, not the average power, that determines the percentage of modulation.

Unsymmetrical Modulation

In an ordinary electric circuit it is possible to increase the amplitude of current flow indefinitely, up to the limit of the power-handling capability of the components, but it cannot very well be decreased to less than zero. The same thing is true of the amplitude of an r.f. signal; it can be modulated *upward* to any desired extent, but it cannot be modulated *downward* more than 100 per cent.

When the modulating wave form is unsymmetrical it is possible for the upward and downward modulation percentages to be different. A simple case is shown in Fig. 10-2. The positive peak of the modulating signal is about 3 times the amplitude of the negative peak. If, as shown in the drawing, the modulating amplitude is adjusted so that the peak downward modulation is just 100 per cent ($Z = 0$) the peak upward modulation is 300 per cent ($Y = 4X$). The carrier amplitude is represented by X, as in Fig. 10-1. The modulation envelope reproduces the wave form of the modulating signal accurately, hence there is no distortion. In such a modulated signal the increase in power output with modulation is considerably greater than it is when the modulation is symmetrical and therefore has to be limited to 100 per cent both up and down.

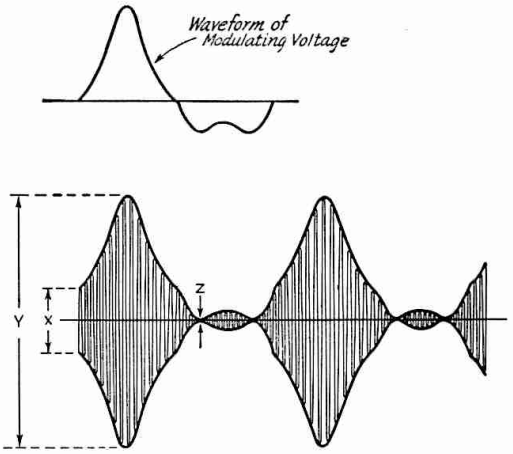


Fig. 10-2—Modulation by an unsymmetrical wave form. This drawing shows 100% downward modulation along with 300% upward modulation. There is no distortion, since the modulation envelope is an accurate reproduction of the wave form of the modulating voltage.

In Fig. 10-2 the peak envelope amplitude, Y, is four times the carrier amplitude, X, so the peak-envelope power is 16 times the carrier power. When the upward modulation is more than 100 per cent the power capacity of the modulating system obviously must be increased sufficiently to take care of the much larger peak amplitudes.

Overmodulation

If the amplitude of the modulation on the downward swing becomes too great, there will be a period of time during which the r.f. output is entirely cut off. This is shown in Fig. 10-3. The shape of the downward half of the modulating wave is no longer accurately reproduced by the modulation envelope, consequently the modulation is distorted. Operation of this type is called **overmodulation.** The distortion of the modulation envelope causes new frequencies (harmonics of the modulating frequency) to be generated. These combine with the carrier to form new side frequencies that widen the channel occupied by the modulated signal. These spurious frequencies are commonly called "splatter."

It is important to realize that the channel occupied by an amplitude-modulated signal is dependent *on the shape of the modulation envelope.* If this wave shape is complex and can be resolved into a wide band of audio frequencies, then the channel occupied will be correspondingly large. An overmodulated signal splatters and occupies a much wider channel than is necessary because the "clipping" of the modulating wave that occurs at the zero axis changes the envelope wave shape to one that contains high-order harmonics of the original modulating frequency. These harmonics appear as side frequencies separated by, in some cases, many kilocycles from the carrier frequency.

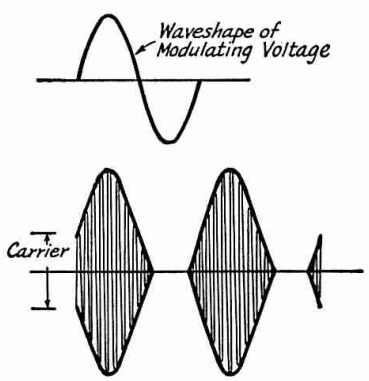


Fig. 10-3—An overmodulated signal. The modulation envelope is not an accurate reproduction of the wave form of the modulating voltage. This or any type of distortion occurring during the modulation process generates spurious sidebands or "splatter."

Because of this clipping action at the zero axis, it is important that care be taken to prevent applying too large a modulating signal in the downward direction. Overmodulation downward results in more splatter than is caused by most other types of distortion in a phone transmitter.

GENERAL REQUIREMENTS

For proper operation of an amplitude-modulated transmitter there are a few general requirements that must be met no matter what particular method of modulation may be used. Failure to meet these requirements is accompanied by distortion of the modulation envelope. This in turn increases the channel width as compared with that required by the legitimate frequencies contained in the original modulating wave.

Frequency Stability

For satisfactory amplitude modulation, the carrier *frequency* must be entirely unaffected by modulation. If the application of modulation 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 FCC 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.

Linearity

At least up to the limit of 100 per cent upward modulation, the amplitude of the r.f. output should be directly proportional to the amplitude of the modulating wave. Fig. 10-4 is a graph of an ideal **modulation characteristic,** or curve showing the relationship between r.f. output amplitude and instantaneous modulation amplitude. The modulation swings the r.f. ampli-

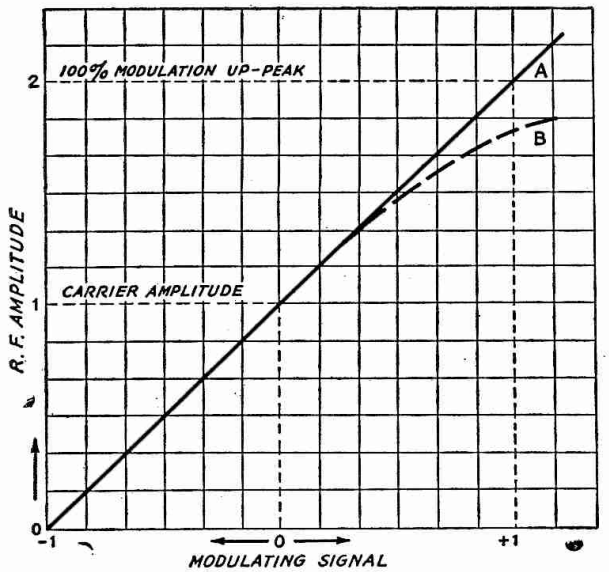


Fig. 10-4—The modulation characteristic shows the relationship between the instantaneous envelope amplitude of the r.f. output current (or voltage) and the instantaneous amplitude of the modulating voltage. The ideal characteristic is a straight line, as shown by curve *A*.

tude back and forth along the curve *A*, as the modulating voltage alternately swings positive and negative. Assuming that the negative peak of the modulating wave is just sufficient to reduce the r.f. output to zero (modulating voltage equal to −1 in the drawing), the same modulating voltage peak in the *positive* direction (+1) should cause the r.f. amplitude to reach twice its unmodulated value. The ideal 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 voltage 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 wave. 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.

The **modulation capability** of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from nonlinearity. The maximum capability can never exceed 100 per cent on the down-peak, but it is possible for it to be higher on the up-peak. The modulation capability should be as close to 100 per cent as possible, so that the most effective signal can be transmitted.

Plate Power Supply

The d.c. power supply for the plate or plates of the modulated amplifier should be well filtered; if it is not, plate-supply ripple will modulate the carrier and cause annoying hum. The ripple voltage should not be more than about 1 per cent of the d.c. output voltage.

In amplitude modulation the plate current of the modulated r.f. amplifier varies at an audio-frequency rate; in other words, an alternating current is superimposed on the d.c. plate current. The output filter capacitor 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 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 capacitor is at least equal to

$$C = 25\,\frac{I}{E}$$

where C = Capacitance of output capacitor in μf.
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 capacitor in the plate-supply filter should be at least

$$C = 25\,\frac{I}{E} = 25 \times \frac{275}{1250} = 25 \times 0.22 = 5.5\ \mu\text{f.}$$

Amplitude Modulation Methods

● MODULATION SYSTEMS

As explained in the preceding section, amplitude modulation of a carrier is accompanied by an increase in power output, the additional power being the "useful" or "talk power" in the sidebands. This additional power may be supplied from an external source in the form of audio-frequency power. It is then added to the unmodulated power input to the amplifier to be modulated, after which the combined power is converted to r.f. This is the method used in plate modulation. It has the advantage that the r.f. power is generated at the high efficiency characteristic of Class C amplifiers — of the order of 65 to 75 per cent — but has the accompanying disadvantage that generating the audio-frequency power is rather expensive.

An alternative that does not require relatively large amounts of audio-frequency power makes use of the fact that the power output of an amplifier can be controlled by varying the potential of a tube element — such as a control grid or a screen grid — that *does not, in itself,* consume appreciable power. In this case the additional power during modulation is secured by sacrificing carrier power; in other words, a tube is capable of delivering only so much total power

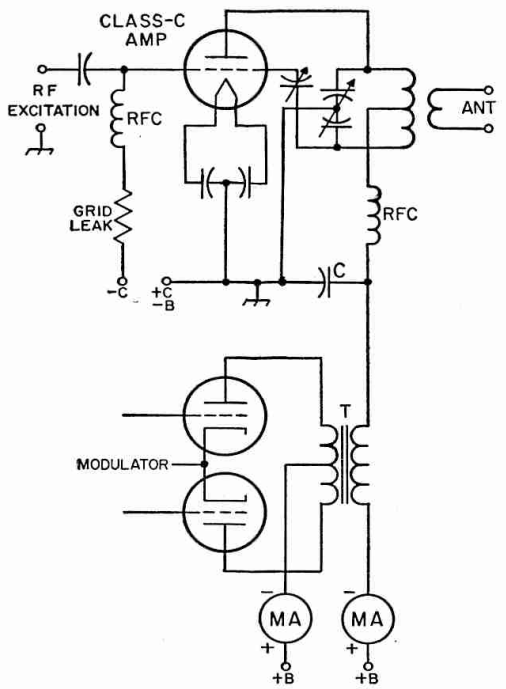


Fig. 10-5—Plate modulation of a Class C r.f. amplifier. The r.f. plate bypass capacitor, C, in the amplifier stage should have reasonably high reactance at audio frequencies. A value of the order of 0.001 μf. to 0.005 μf. is satisfactory in practically all cases. (See chapter on modulators.)

within its ratings, and if more must be delivered at full modulation, then less is available for the unmodulated carrier. Systems of this type must of necessity work at rather low efficiency at the unmodulated carrier level. As a practical working rule, the efficiency of the modulated r.f. amplifier is of the order of 30 to 35 per cent, and the unmodulated carrier power output obtainable with such a system is only about one-fourth to one-third that obtainable from the same amplifier with plate modulation.

It is well to appreciate that no simple modulation scheme that purports to get around this limitation of grid modulation ever has actually done so. Methods have been devised that have resulted in modulation at high over-all efficiency, without requiring audio power, by obtaining the necessary additional power from an auxiliary r.f. amplifier. This leads to circuit and operating complexities that make the systems unsuitable for amateur work, where rapid frequency change and simplicity of operation are almost always essential.

The methods discussed in this section are the basic ones. Variants that from time to time attain passing popularity can readily be appraised on the basis of the preceding paragraphs. A simple grid modulation system that claims high efficiency should be looked upon with suspicion, since it is almost certain that the high efficiency, if actually achieved, is obtained by sacrificing the linear relationship between modulating signal and modulation envelope that is the first essential of a good modulation method.

PLATE MODULATION

Fig. 10-5 shows the most widely used system of plate modulation, in this case with a triode r.f. tube. 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 power 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.

Audio Power

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

$$Z_m = \frac{E_b}{I_p} \times 1000 \text{ ohms}$$

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 instantaneous plate voltage (the r.f. output voltage must be proportional to the plate voltage) for the modulation to be linear. This will be the case when the amplifier operates under Class C conditions. The linearity 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 are described in the chapter on transmitters. 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 the d.c. plate voltage used, 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 from a fixed-bias source of about the cut-off value, supplemented by enough grid-leak bias to bring the total up to the required operating bias.

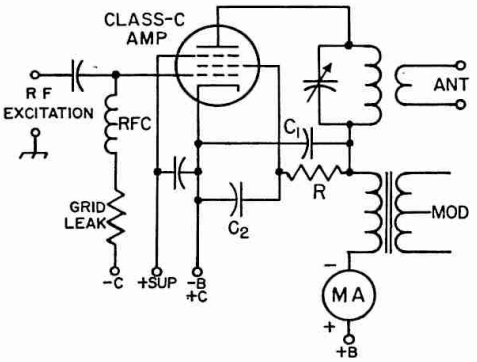


Fig. 10-6—Plate and screen modulation of a Class C r.f. amplifier using a screen-grid tube. The plate r.f. bypass capacitor, C_1, should have reasonably high reactance at all audio frequencies; a value of 0.001 to 0.005 μf. is generally satisfactory. The screen bypass, C_2, should not exceed 0.002 μf. in the usual case.

When the modulated amplifier is a beam tetrode the suppressor connection shown in this diagram may be ignored. If a base terminal is provided on the tube for the beam-forming plates, it should be connected as recommended by the tube manufacturer.

The maximum permissible d.c. plate power input for 100 per cent modulation is twice the sine-wave audio-frequency power output available from 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 in the chapter on 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 total power input (d.c. plus audio-frequency a.c.) increases with modulation, the d.c. plate current of a plate-modulated amplifier should not change when the stage is modulated. 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 wave. D.c. instruments cannot follow the a.f. variations, and since the average d.c. plate current and plate voltage of a properly operated amplifier do not change, neither do the meter readings. A change in plate current with modulation indicates nonlinearity. On the other hand, a thermocouple r.f. ammeter connected in the antenna or transmission line will show an increase in r.f. current with modulation, because instruments of this type respond to power rather than to current or voltage.

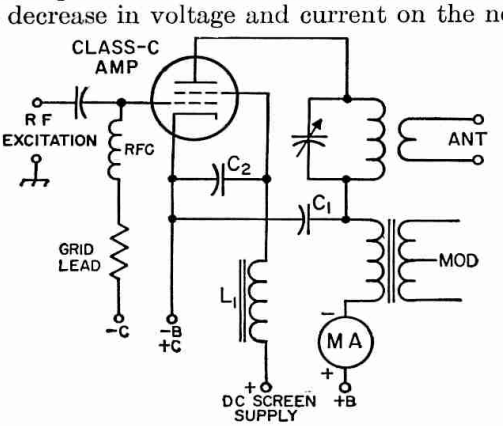


Fig. 10-7—Plate modulation of a beam tetrode, using an audio impedance in the screen circuit. The value of L_1 is discussed in the text. See Fig. 10-6 for data on bypass capacitors C_1 and C_2.

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 voltages is shown in Fig. 10-6. The dropping resistor, R, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

The modulating impedance is found by dividing the d.c. plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents 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 the screen voltage has a much greater effect on the plate current than the plate voltage does. The modulation characteristic is nonlinear if the plate alone is modulated. However, some beam tetrodes can be modulated satisfactorily by applying the modulating power to the plate circuit alone, provided the screen is connected to its d.c. supply through an audio impedance. Under these conditions the screen becomes self-modulating, because of the variations in screen current that occur when the plate voltage is varied. The circuit is shown in Fig. 10-7. 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 screen impedance can be taken to be approximately equal to the d.c. screen voltage divided by the d.c. screen current in amperes.

Choke-Coupled Modulator

The choke-coupled Class A modulator is shown in Fig. 10-8. Because of the relatively low power output and plate efficiency of a Class A amplifier, this method is seldom used except for a few special applications. The audio power output of the modulator is combined with the d.c. power in the plate circuit, as in the case of the transformer-coupled modulator. But there is considerably less freedom in adjustment, since no transformer is available for matching impedances.

The modulating impedance of the r.f. amplifier must be adjusted to the value of load impedance

required by the particular modulator tube used, and the power input to the r.f. stage should not exceed twice the rated a.f. power output of the modulator for 100 per cent modulation. A complication is the fact that the plate voltage on the

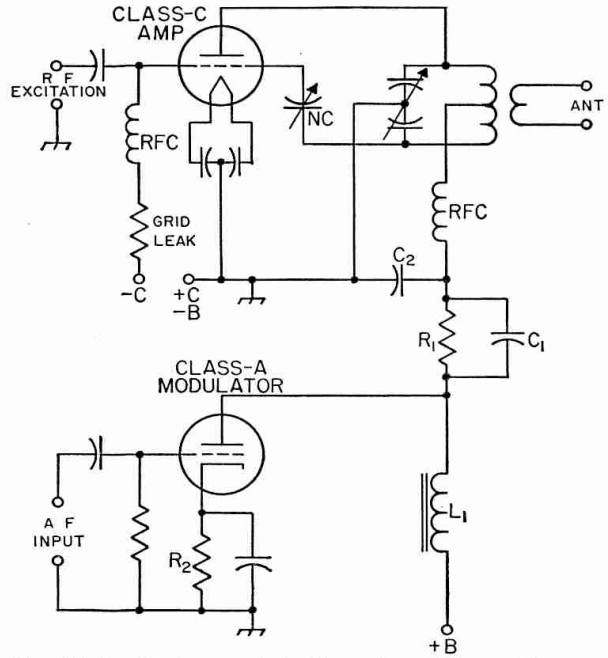


Fig. 10-8—Choke-coupled Class A modulator. The cathode resistor, R_2, should have the normal value for operation of the modulator tube as a Class A power amplifier. The modulation choke, L_1, should be 5 henrys or more. A value of 0.001 to 0.005 μf. is satisfactory at C_2, the r.f. amplifier plate bypass capacitor. See text for discussion of C_1 and R_1.

modulator must be higher than the plate voltage on the r.f. amplifier, for 100 per cent modulation. This is because the a.f. voltage developed by the modulator cannot swing to zero without a great deal of distortion. R_1 provides the necessary d.c. voltage drop between the modulator and r.f. amplifier, but its value cannot be calculated without using the published plate family of curves for the modulator tube used. The d.c. voltage drop through R_1 must equal the minimum instantaneous plate voltage on the modulator tube under normal operating conditions. C_1, an audio-frequency bypass across R_1, should have a capacitance such that its reactance at 100 cycles is not more than about one-tenth the resistance of R_1. Without R_1C_1 the percentage of modulation is limited to 70 to 80 per cent in the average case.

GRID MODULATION

The principal disadvantage of plate modulation is that a considerable amount of audio power is necessary. This requirement can be avoided by applying the modulation to a grid element in the modulated amplifier. However, the convenience and economy of the low-power modulator must be paid for, since no modulation system gives something for nothing. The increased power output that accompanies modulation is paid for, in the case of grid modulation, by a reduction in the carrier power output obtainable from a given r.f. amplifier tube, and by more rigorous operating requirements and more complicated adjustment.

The term "grid modulation" as used here applies to all types — control grid, screen, or suppressor — since the operating principles are exactly the same no matter which grid is actually modulated. With grid modulation 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 as shown in Fig. 10-9. For

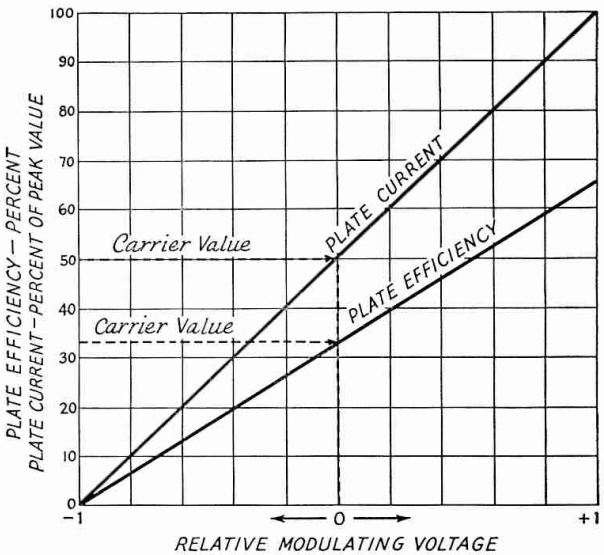


Fig. 10-9—In a perfect grid-modulated amplifier both plate current and plate efficiency would vary with the instantaneous modulating voltage as shown. When this is so the modulation characteristic is as given by curve A in Fig. 10-4, and the peak envelope output power is four times the unmodulated carrier power. The variations in plate current with modulation, indicated above, do not register on a d.c. meter, so the plate meter shows no change when the signal is modulated.

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-envelope peak the power input is doubled, and since the plate efficiency also is doubled at the same instant the peak envelope output power will be four times the carrier power. The efficiency obtainable at the envelope peak depends on how carefully the modulated amplifier is adjusted, and sometimes can be as high as 80 per cent. It is generally less when the amplifier is adjusted for good linearity, and under average conditions a round figure of ⅔, or 66 per cent, is representative. The efficiency without modulation is only half the peak efficiency, or about 33 per cent. This low average efficiency reduces the permissible carrier output to about one-fourth the power obtainable from the same tube in c.w. operation, and to about one-third the carrier output obtainable from the tube with plate modulation.

The modulator is required to furnish only the audio power dissipated in the modulated grid under the operating conditions chosen. A speech amplifier capable of delivering 3 to 10 watts is usually sufficient.

Grid Modulation

Generally speaking, grid modulation does not give quite as linear a modulation characteristic as plate modulation, even under optimum operating conditions. When misadjusted the nonlinearity may be severe, resulting in bad distortion and splatter. However, with careful adjustment it is capable of satisfactory results.

Plate-Circuit Operating Conditions

The d.c. plate power 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. It is generally best to use the maximum plate voltage permitted by the manufacturer's ratings, because the optimum operating conditions are more easily achieved with high plate voltage and the linearity also is improved.

Example: Two tubes having plate dissipation ratings of 55 watts each are to be used with grid modulation.

The maximum permissible 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 average plate current for the two tubes will be

$$I = \frac{P}{E} = \frac{165}{1500} = 0.11 \text{ amp.} = 110 \text{ ma.}$$

At 33% efficiency, the carrier output to be expected is 55 watts.

The plate-voltage/plate-current ratio at *twice* carrier plate current is

$$\frac{1500}{220} = 6.8$$

The tank-circuit L/C ratio should be chosen on the basis of *twice* the average or carrier plate current. If the L/C ratio is based on the plate voltage/plate current ratio under carrier conditions the Q may be too low for good coupling to the output circuit.

Screen Grid Modulation

Screen modulation is probably the simplest form of grid modulation and the least critical of adjustment. The most satisfactory way to apply the modulating voltage to the screen is through a transformer, as shown in Fig. 10-10. With practical tubes it is necessary to drive the screen somewhat negative with respect to the cathode to get complete cut-off of r.f. output. For this reason the peak modulating voltage required for 100 per cent modulation is usually 10 per cent or so greater than the d.c. screen voltage. The latter, in turn, is approximately half the rated screen voltage recommended under maximum ratings for c.w. operation.

The audio power required for 100 per cent modulation is approximately one-fourth the d.c. power input to the screen in c.w. operation, but varies somewhat with the operating conditions. A receiving-type audio power amplifier will suffice as the modulator for most transmitting tubes. The relationship between screen voltage and screen current is not linear, which means that the load on the modulator varies over the

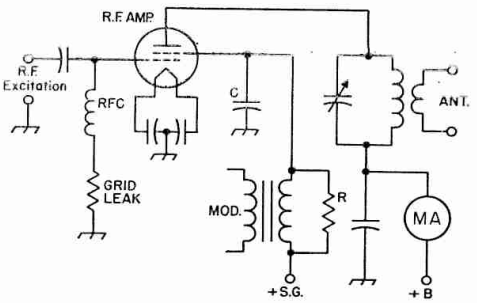


Fig. 10-10—Screen-grid modulation of beam tetrode. Capacitor C is an r.f. bypass capacitor and should have high reactance at audio frequencies. A value of 0.002 μf. is satisfactory. The grid leak can have the same value that is used for c.w. operation of the tube.

audio-frequency cycle. It is therefore highly advisable to use negative feedback in the modulator circuit. If excess audio power is available, it is also advisable to load the modulator with a resistance (R in Fig. 10-10) its value being adjusted to dissipate the excess power. Unfortunately, there is no simple way to determine the proper resistance except experimentally, by observing its effect on the modulation envelope with the aid of an oscilloscope.

On the assumption that the modulator will be fully loaded by the screen plus the additional load resistor R, the turns ratio required in the coupling transformer may be calculated as follows:

$$N = \frac{E_d}{2.5\sqrt{PR_L}}$$

where N is the turns ratio, secondary to primary; E_d is the rated screen voltage for c.w. operation; P is the rated audio power output of the modulator; and R_L is the rated load resistance for the modulator.

Adjustment

A screen-modulated amplifier should be adjusted with the aid of an oscilloscope connected as shown in Fig. 10-11. 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 d.c. plate and screen voltages. 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 voltage until the modulation characteristic shows curvature (see later in this chapter for use of the oscilloscope). If curvature occurs well below 100 per cent modulation, the plate efficiency is too high at the carrier level. Increase the plate loading slightly and readjust the r.f. grid excitation to maintain the same plate current; then apply modulation and check the characteristic again. Continue until the characteristic is as linear as possible from zero to twice the carrier amplitude.

In general, the amplifier should be heavily

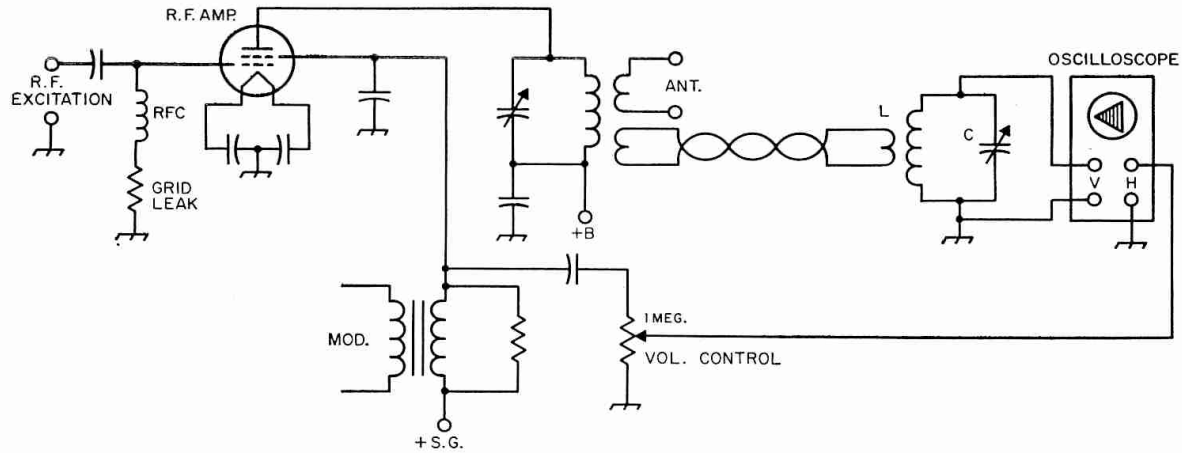


Fig. 10-11—Using the oscilloscope for adjustment of a screen-modulated amplifier. *L* and C should tune to the operating frequency, and may be coupled to the transmitter tank circuit through a twisted pair or coax, using single-turn links at each end. The blocking capacitor (0.05 μf.) that couples the audio voltage from the screen grid to the horizontal plates of the oscilloscope should have a voltage rating equal to at least twice the d.c. voltage on the grid that is being modulated. The r.f. and audio voltages should be fed directly to the deflection plates of the scope tube (through blocking capacitors if necessary or desirable), not through any vertical or horizontal amplifiers that may be in the instrument.

loaded. Under proper operating conditions the plate-current dip as the amplifier plate circuit is tuned through resonance will be little more than just discernible. It is desirable to operate with the grid current as low as possible, since this reduces the screen current and thus reduces the amount of power required from the modulator.

With proper adjustment the linearity is good up to about 90 per cent modulation. When the screen is driven negative for 100 per cent modulation there is a kink in the modulation characteristic at the zero-voltage point. This introduces a small amount of envelope distortion. The kink can be removed and the over-all linearity improved by applying a small amount of modulating voltage to the control grid simultaneously with screen modulation.

In an alternative adjustment method not requiring an oscilloscope the r.f. amplifier is first tuned up for maximum output without modulation and the rated d.c. screen voltage (from a fixed-voltage supply) for c.w. operation applied. Use heavy loading and reduce the grid excitation until the output just starts to fall off, at which point the resonance dip in plate current should be small. Note the plate current and, if possible, the r.f. antenna or feeder current, and then reduce the d.c. screen voltage until the plate current is one-half its previous value. The r.f. output current should also be one-half its previous value at this screen voltage. The amplifier is then ready for modulation, and the modulating voltage may be increased until the plate current just starts to shift upward, which indicates that the amplifier is modulated 100 per cent. With voice modulation the plate current should remain steady, or show just an occasional small upward kick on intermittent peaks.

"Clamp-Tube" Modulation

A method of screen-grid modulation that is convenient in transmitters provided with a screen protective tube ("clamp" tube) is shown in Fig. 10-12. An audio-frequency signal is applied to the grid of the clamp tube, which then becomes a modulator. The simplicity of the circuit is somewhat deceptive, since it is considerably more difficult from a design standpoint than the transformer-coupled arrangement of Fig. 10-10.

For proper modulation the clamp tube must be operated as a triode Class A amplifier, and it will be recognized that the method is essentially identical with the choke-coupled Class A plate modulator of Fig. 10-8 except that a resistance, R_2, is substituted for the choke. R_2 in the usual case is the screen dropping resistor normally used for c.w. operation. Its value should be at least two or three times the load resistance required by the Class A modulator tube for optimum audio-frequency output. Unfortunately, relatively little

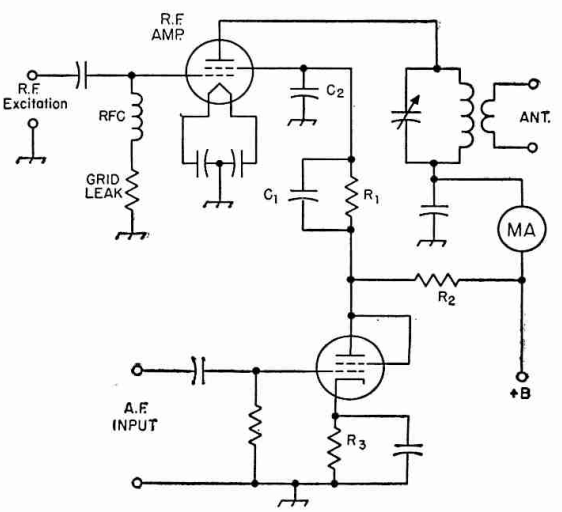


Fig. 10-12—Screen modulation by a "clamp" tube. The grid leak is the normal value for c.w. operation and C_2 should be 0.002 μf. or less. See text for discussion of C_1, R_1, R_2 and R_3. R_3 should have the proper value for Class A operation of the modulator tube, but cannot be calculated unless triode curves for the tube are available.

information is available on the triode operation of the tubes most frequently used for screen-protective purposes.

Like the choke-coupled modulator, the clamp-tube modulator is incapable of modulating the r.f. stage 100 per cent unless the dropping resistor, R_1, and audio bypass, C_1, are incorporated in the circuit. The same design considerations hold, with the addition of the fact that the screen must be driven negative, not just to zero voltage, for 100 per cent modulation. The modulator tube must thus be operated at a voltage ranging from 20 to 40 per cent higher than the screen that it modulates. Proper design requires knowledge of the screen characteristics of the r.f. amplifier and a set of plate-voltage plate-current curves on the modulator tube as a triode.

Adjustment with this system, once the design voltages have been determined, is carried out in the same way as with transformer-coupled screen modulation, preferably with the oscilloscope. Without the oscilloscope, the amplifier may first be adjusted for c.w. operation as described earlier, but with the modulator tube removed from its socket. The modulator is then replaced, and the cathode resistance, R_3, adjusted to reduce the amplifier plate current to one-half its c.w. value. The amplifier plate current should remain constant with modulation, or show just a small upward flicker on occasional voice peaks.

Controlled Carrier

As explained earlier, a limit is placed on the output obtainable from a grid-modulation system by the low r.f. amplifier plate efficiency (approximately 33 per cent) under unmodulated carrier conditions. The plate efficiency increases with modulation, since the output increases while the d.c. input remains constant, and reaches a maximum in the neighborhood of 50 per cent with 100 per cent sine-wave modulation. If the power input to the amplifier can be reduced during periods when there is little or no modulation, thus reducing the plate loss, advantage can be taken of the higher efficiency at full modulation to obtain higher effective output. This can be done by varying the d.c. power input to the modulated stage in accordance with *average* variations in voice intensity, in such a way as to maintain just sufficient carrier power to keep the modulation high, but not exceeding 100 per cent, under all conditions. Thus the carrier amplitude is controlled by the average voice intensity. Properly utilized, controlled carrier permits increasing the effective carrier output at maximum level to a value about equal to the rated plate dissipation of the tube, or twice the output obtainable with constant carrier.

It is desirable to control the power input just enough so that the plate loss, without modulation, is safely below the tube rating. Excessive control is disadvantageous because the distant receiver's a.v.c. system must continually follow the variations in average signal level. The circuit of Fig. 10-13 permits adjustment of both the maximum and minimum power input, and although somewhat more complicated than some circuits that have been used is actually simpler to operate because it separates the functions of modulation and carrier control. A portion of the audio voltage at the modulator grid is applied to a Class A "control amplifier" which drives a rectifier circuit to produce a d.c. voltage negative with respect to ground. C_1 filters out the audio variations, leaving a d.c. voltage proportional to the average voice level. This voltage is applied to the grid of a "clamp" tube to control the d.c. screen voltage and thus the r.f. carrier level. Maximum output is obtained when the carrier-control tube grid is driven to cut-off, the voice level at which this occurs being determined by the setting of R_4. The input without modulation is set to the desired level (usually about equal to the plate dissipation rating of the modulated stage) by adjusting R_2. R_3 may be the normal screen-dropping resistor for the modulated beam tetrode, but in case a separate screen supply is used the resistance need be just large enough to give sufficient voltage drop to reduce the no-modulation power input to the desired value.

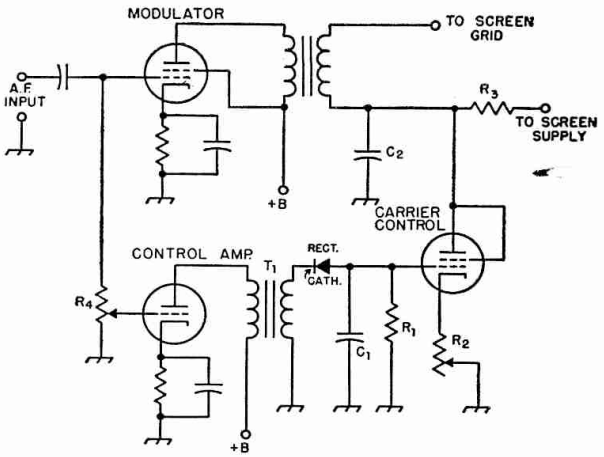


Fig. 10-13—Circuit for carrier control with screen modulation. A small triode such as the 6C4 can be used as the control amplifier and a 6Y6G is suitable as a carrier-control tube. T_1 is an interstage audio transformer having a 1-to-1 or larger turns ratio. R_4 is a 0.5-megohm volume control and also serves as the grid resistor for the modulator. A germanium crystal may be used as the rectifier. Other values are discussed in the text.

C_1R_1 should have a time constant of about 0.1 second. The time constant of C_2R_3 should be no larger. Further details may be found in *QST* for April, 1951, page 64. An oscilloscope is required for proper adjustment.

Suppressor Modulation

Pentode-type tubes do not, in general, modulate well when the modulating voltage is applied to the screen grid. However, a satisfactory modulation characteristic can be obtained by applying the modulation to the suppressor grid. The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 10-14.

The method of adjustment closely resembles that used with screen-grid modulation. If an oscilloscope is not available, the amplifier is first adjusted for optimum c.w. output with zero bias

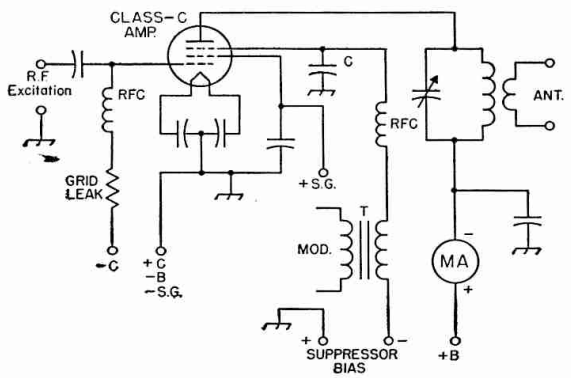


Fig. 10-14—Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. bypass capacitor, C, should be the same as the grid bypass capacitor in control-grid modulation.

on the suppressor grid. Negative bias is then applied to the suppressor and increased in value until the plate current and r.f. output current drop to half their original values. When this condition has been reached the amplifier is ready for modulation.

Since the suppressor is always negatively biased, the modulator is not required to furnish any power and a voltage amplifier can be used. The suppressor bias will vary with the type of pentode and the operating conditions, but usually will be of the order of −100 volts. The peak a.f. voltage required from the modulator is equal to the suppressor bias.

Control-Grid Modulation

Although control-grid modulation may be used with any type of r.f. amplifier tube, it is seldom used with tetrodes and pentodes because screen or suppressor modulation is generally simpler to adjust. However, control-grid modulation is the only form of grid modulation that is applicable to triode amplifiers. A typical triode circuit is given in Fig. 10-15.

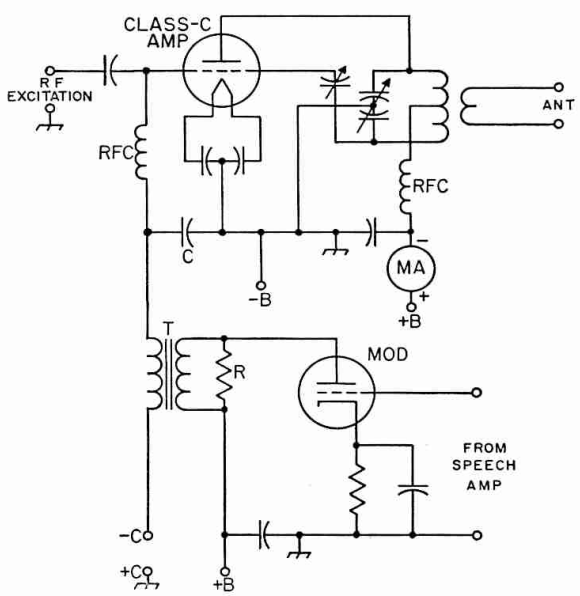


Fig. 10–15—Control-grid modulation of a Class C amplifier. The r.f. grid bypass capacitor, C, should have high reactance at audio frequencies (0.005 μf. or less).

In control-grid modulation the d.c. grid bias is the same as in normal Class C amplifier service, but the r.f. grid excitation is somewhat smaller. The audio voltage superimposed on the d.c. bias changes the instantaneous grid bias at an audio rate, thus varying the operating conditions in the grid circuit and controlling the output and efficiency of the amplifier.

The change in instantaneous bias voltage with modulation causes the rectified grid current of the amplifier to vary, which places a variable load on the modulator. To reduce distortion, resistor R in Fig. 10-15 is connected in the output circuit of the modulator as a constant load, so that the over-all load variations will be minimized. This resistor should be equal to or somewhat higher than the load into which the modulator tube is rated to work at normal audio output. It is also recommended that the modulator circuit incorporate as much negative feedback as possible, as a further aid in reducing the internal resistance of the modulator and thus improving the "regulation" — that is, reducing the effect of load variations on the audio output voltage. The turns ratio of transformer T should be about 1 to 1 in most cases.

The load on the r.f. driving stage also varies with modulation. This in turn will cause the excitation voltage to vary and may cause the modulation characteristic to be nonlinear. To overcome it, the driver should be capable of two or three times the r.f. power output actually required to drive the amplifier. The excess power may be dissipated in a dummy load (such as an incandescent lamp of appropriate power rating) that then performs the same function in the r.f. circuit that resistor R does in the audio circuit.

The d.c. bias source in this system should have low internal resistance. Batteries or a voltage-regulated supply are suitable. Grid-leak bias should not be used.

Satisfactory adjustment of a control-grid modulated amplifier requires an oscilloscope. The scope connections are similar to those shown for screen-grid modulation in Fig. 10-11, with audio from the modulator's output transformer secondary applied to the horizontal plates through a blocking capacitor and volume control, and with r.f. from the plate tank circuits coupled to the vertical plates. The adjustment procedure follows that for screen modulation as previously described.

● CATHODE MODULATION

Circuit

The fundamental circuit for cathode modulation is shown in Fig. 10-16. It is a combination of the plate and grid 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 are modulated.

Because part of the modulation is by the

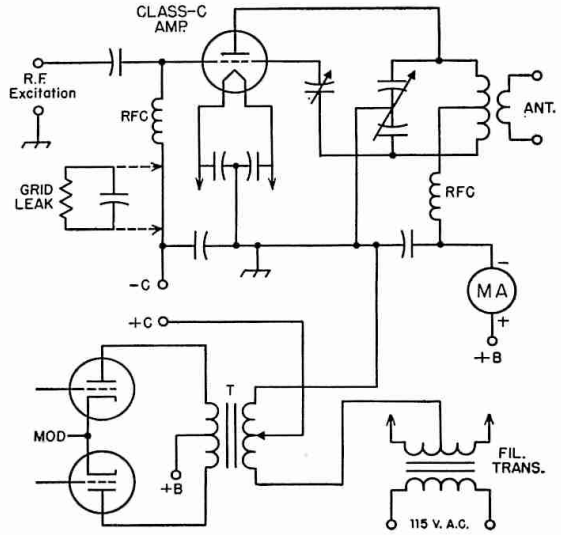


Fig. 10-16—Circuit arrangement for cathode modulation of a Class C r.f. amplifier. Values of bypass capacitors in the r.f. circuits should be the same as for other modulation methods.

control-grid 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. 10-17. In these curves the performance of the cathode-modulated r.f. amplifier is plotted in terms

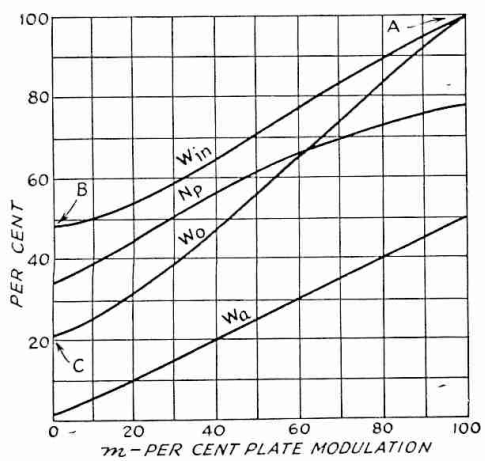


Fig. 10-17—Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class C telephony tube ratings.

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.

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 modulation is increased to make the over-all modulation reach 100 per cent. The limiting condition, 100 per cent plate modulation and no grid modulation, is at the right (A); pure grid 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. 10-17, 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 $= 250 \times 0.65 = 162.5$ watts
Power output $= 190 \times 0.48 = 91.2$ watts

The required audio power, from the chart, is equal to 20% of the d.c. input to the modulated amplifier. Therefore

Audio power $= 162.5 \times 0.2 = 32.5$ 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 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, as described in the chapter on speech equipment.

Conditions for Linearity

R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and control-grid 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 per-

centages 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 bypassed 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. When directly heated tubes are modulated their filaments must be supplied from a separate transformer. The filament bypass capacitors should not be larger than about 0.002 μf., to avoid bypassing 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.

● LINEAR AMPLIFIERS

If a signal is to be amplified after modulation has taken place, the shape of the modulation envelope must be preserved if distortion is to be avoided. This requires the use of a **linear amplifier** — that is, one that will reproduce, in its output circuit, the exact form of the signal envelope applied to its grid.

Linear amplifiers for amplitude-modulated r.f. signals cannot be operated with the grid bias beyond cut-off. To do so would mean that the part of the modulation envelope near the zero axis (see Fig. 10-1C) would be clipped, since there would be times when the instantaneous signal voltage would be below the minimum value that would cause plate-current flow. The result would be overmodulation of the type shown in Fig. 10-3.

However, the grid bias may be set at any value less than cutoff. Usually, such amplifiers are operated at or near the Class B condition — that is, with the grid bias at or somewhat less than cutoff. Although Class B operation results in considerable distortion of the individual r.f. cycles applied to the grid, the modulation *envelope* is not distorted if the operating conditions are chosen properly. The r.f. distortion produces only r.f. harmonics, and these can be eliminated by the selectivity of the output tank circuit.

A linear amplifier used for a.m. has the same disadvantages with respect to efficiency that grid modulation does. The reason also is much the same: since the amplifier must handle a peak-envelope power four times as great as the unmodulated carrier power, it cannot be operated at its full capabilities when it is amplifying only the unmodulated carrier. The plate efficiency of the amplifier varies with the instantaneous value of the modulation envelope in the same way that it varies with the instantaneous modulating voltage in grid modulation (Fig. 10-9). Hence the efficiency at the unmodulated carrier level is only of the order of 30-35 per cent.

Because of this low efficiency, linear amplifiers have not had much application in amateur transmitters, especially since equivalent efficiency can be obtained with grid modulation, along with a less critical adjustment procedure. Recently there has been some increase in use of a.m. linears, particularly at v.h.f., as a means of stepping up the modulated power output of very low power transmitters with a minimum of complication in over-all equipment and operation. To obtain a useful increase in power output by this means the linear amplifier must use a tube or tubes capable of relatively large plate dissipation, since about two-thirds of the d.c. power input to the amplifier is consumed in heating the plate and only about one-third is converted to useful carrier output.

Checking A.M. Phone Operation

● USING THE OSCILLOSCOPE

Proper adjustment of a phone transmitter is aided immeasurably by the oscilloscope. The scope 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.

In the simplest scope circuit, radio-frequency voltage from the modulated amplifier is applied to the vertical deflection plates of the tube, usually through blocking capacitors as shown in the oscilloscopecir cuit in the chapter on measurements, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the instantaneous amplitude of the audio signal varies, the r.f. output of the transmitter likewise varies, and this produces a wedge-shaped pattern or **trapezoid** on the screen. If the oscilloscope has a built-in horizontal sweep, the

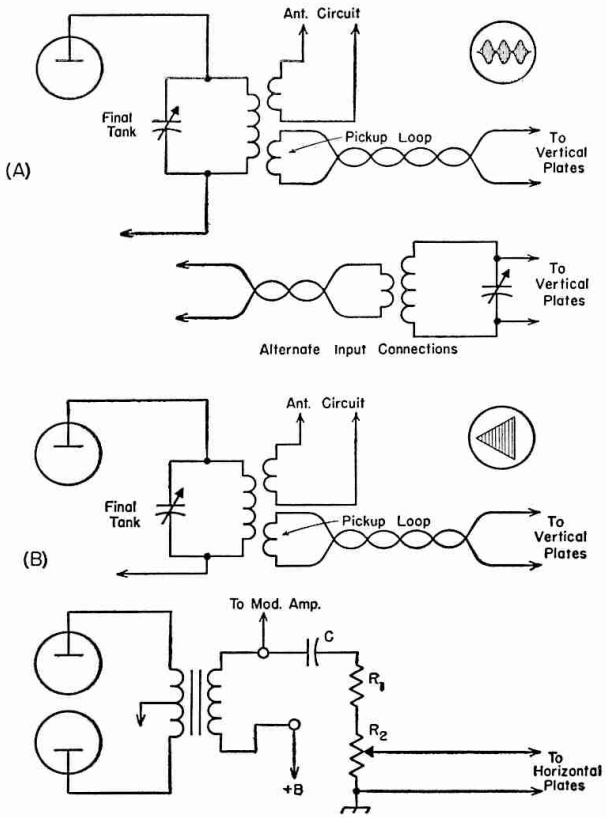


Fig. 10-18—Methods of connecting the oscilloscope for modulation checking. A—connections for wave-envelope pattern with any modulation method; B—connections for trapezoidal pattern with plate modulation. See Fig. 10-11 for scope connections for trapezoidal pattern with screen modulation.

r.f. voltage can be applied to the vertical plates as before (never through an amplifier) and the sweep will produce 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.

The Wave-Envelope Pattern

The connections for the wave-envelope pattern are shown in Fig. 10-18A. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a low-impedance (coax, twisted pair, etc.) 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 convenient means for adjustment of the pattern height.

If it is inconvenient to couple to the final tank coil, as may be the case if the transmitter is tightly shielded to prevent TVI, the pick-up loop may be coupled to the tuned tank of a matching circuit or antenna coupler. Any method (even a short antenna coupled to the tuned circuit shown in the "alternate input connections" of Fig. 10-18A) that will pick up enough r.f. to give a suitable pattern height may be used.

The position of the pick-up coil should be varied until an unmodulated carrier pattern, Fig. 10-19B, 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. 10-19D, where the point X represents the horizontal 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.

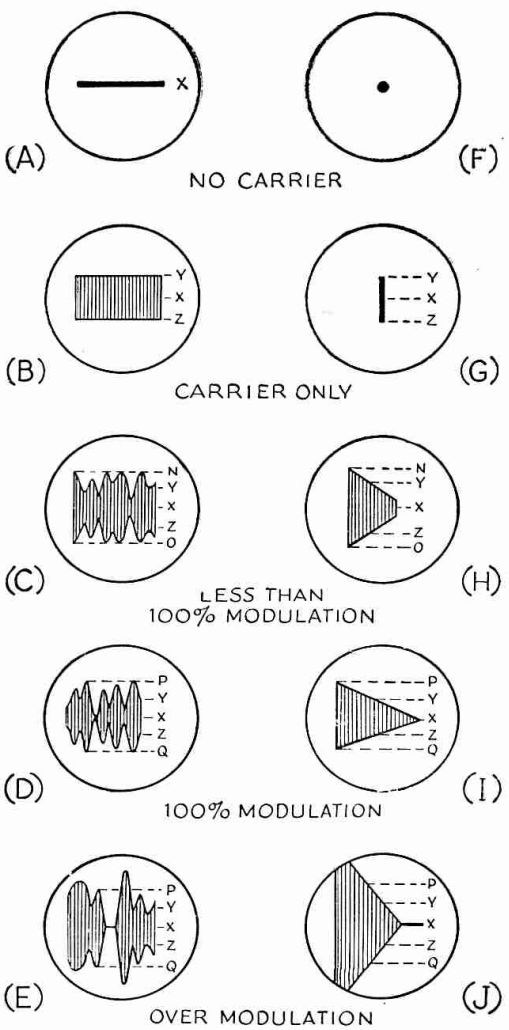


Fig. 10-19—Wave-envelope and trapezoidal patterns representing different conditions of modulation.

The Trapezoidal Pattern

Connections for the trapezoid or wedge pattern as used for checking plate modulation are shown in Fig. 10-18B. The vertical plates of the c.r. tube are coupled to the transmitter tank through a pick-up loop, preferably using a tuned circuit, as shown in the upper drawing, adjustable to the operating frequency. Audio voltage from the modulator is applied to the horizontal plates through a voltage divider, R_1R_2. This voltage should be adjustable so a suitable pattern width can be obtained; a 0.25-megohm volume control can be used at R_2 for this purpose.

The resistance required at R_1 will depend on the d.c. plate voltage on the modulated amplifier. The total resistance of R_1 and R_2 in series should be about 0.25 megohm for each 100 volts of d.c. plate voltage. For example, if the modulated amplifier operates at 1500 volts, the total resistance should be 3.75 megohms, 0.25 megohm at R_2 and the remainder, 3.5 megohms, in R_1. R_1 should be composed of individual resistors not larger than 0.5 megohm each, in which case 1-watt resistors will be satisfactory.

For adequate coupling at 100 cycles the capacitance, in microfarads, of the blocking capacitor, C, should be at least $0.05/R$, where R is the total resistance ($R_1 + R_2$) in megohms. In the example above, where R is 3.75 megohms, the capacitance should be $0.05/3.75 = 0.013$ μf. or more. The voltage rating of the capacitor should be at least twice the d.c. voltage applied to the modulated amplifier. The capacitance can be made up of two or more similar units in series, so long as the total capacitance is equal to that required, in case a single unit of sufficient voltage rating is not available. Two or more units may be used in parallel if capacitors having adequate voltage rating but insufficient capacitance are available.

The corresponding scope connections for screen modulation were given in Fig. 10-11. This circuit will be satisfactory for d.c. screen voltages up to 200 volts or so, which will include most beam tetrodes. If the d.c. screen voltage, adjusted for proper modulation, exceeds 200 volts a voltage divider similar to that shown in Fig. 10-18 should be used, the values being calculated as described above using the screen voltage instead of the plate voltage.

Trapezoidal patterns for various conditions of modulation are shown in Fig. 10-19 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. Over-

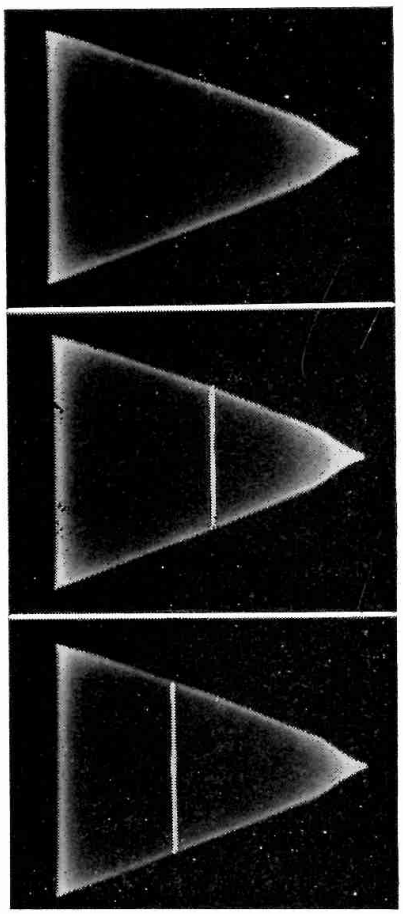


Fig. 10-20—*Top*—A typical trapezoidal pattern obtained with screen modulation adjusted for optimum conditions. The sudden change in slope near the point of the wedge occurs when the screen voltage passes through zero. *Center*—If there is no audio distortion, the unmodulated carrier will have the height and position shown by the white line superimposed on the sine-wave modulation pattern. *Bottom*—Even-harmonic distortion in the audio system, when the audio signal applied to the speech amplifier is a sine wave, is indicated by the fact that the modulation pattern does not extend equal horizontal distances on both sides of the unmodulated carrier.

modulation in the upward direction is indicated by increased height over PQ, and downward by an extension along the axis X at the pointed end.

● CHECKING TRANSMITTER PERFORMANCE

The trapezoidal pattern is generally more useful than the wave-envelope pattern for checking the operation of a phone transmitter. However, both types of patterns have their special virtues, and the best test setup is one that makes both available. The trapezoidal pattern is better adapted to showing the performance of a modulated amplifier from the standpoint of inherent linearity, without regard to the wave form of the audio modulating signal, than is the wave-envelope pattern. Distortion in the audio signal also can be detected in the trapezoidal pattern, although considerable experience in analyzing scope patterns is sometimes required to recognize it.

If the wave-envelope pattern is used with a

sine-wave audio modulating signal, distortion in the modulation envelope is easily recognizable; however, it is difficult to determine whether the distortion is caused by lack of linearity of the r.f. stage or by a.f. distortion in the modulator. If the trapezoidal pattern shows good linearity in such a case the trouble obviously is in the audio system. It is possible, of course, for both defects to be present simultaneously. If they are, the r.f. amplifier should be made linear first; then any distortion in the modulation envelope will be the result of some type of improper operation in the speech amplifier or modulator, or in coupling the modulator to the modulated r.f. stage.

R. F. Linearity

The trapezoidal pattern is actually a graph of the modulation characteristic of the modulated amplifier. The sloping sides of the wedge show the r.f. amplitude for every value of instantaneous modulating voltage, exactly the type of curve plotted in Fig. 10-4. If these sides are perfectly straight lines, as drawn in Fig. 10-19 at H and I, the modulation characteristic is linear. If the sides show curvature, the characteristic is nonlinear to an extent that is shown by the degree to which the sides depart from perfect straightness. This is true regardless of the wave form of the modulating voltage.

Audio Distortion

If the speech system can be driven by a good audio sine-wave signal instead of a microphone, the trapezoidal pattern also will show the presence of even-harmonic distortion (the most common type, especially when the modulator is overloaded) in the speech amplifier or modulator. If there is no distortion in the audio system, the trapezoid will extend horizontally equal distances on each side of the vertical line representing the unmodulated carrier. If there is even-harmonic distortion the trapezoid will extend farther to one side of the unmodulated-carrier position than to the other. This is shown in Fig. 10-20. The probable cause is inadequate power output from the modulator, or incorrect load on the modulator.

An audio oscillator having reasonably good sine-wave output is highly desirable for testing both speech equipment and the phone transmitter as a whole. A very simple audio oscillator such as is shown in Chapter 21 on measurements is quite adequate. With such an oscillator and the scope, the pattern is steady and can be studied closely to determine the effects of various operating adjustments.

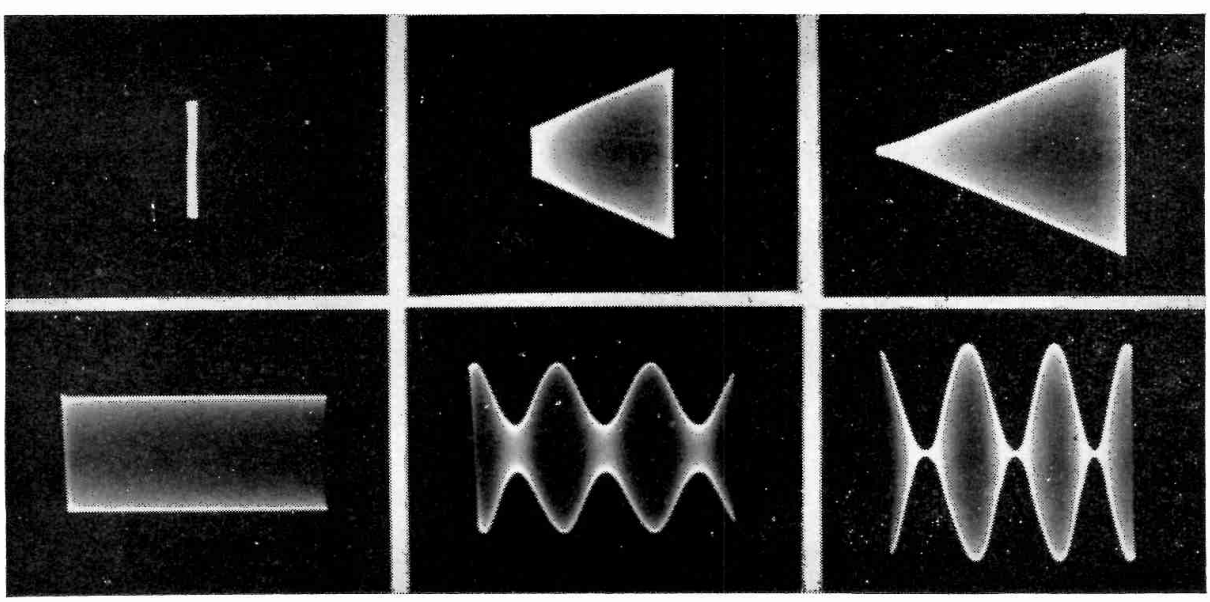
In the case of the wave-envelope pattern, distortion in the audio system will show up in the modulation envelope (with a sine-wave input signal) as a departure from the sine-wave form, and may be checked by comparing the envelope with a drawing of a sine wave. Attributing any such distortion to the audio system assumes, of course, that a check has been made on the linearity of the modulated r.f. amplifier, preferably by use of the trapezoidal pattern.

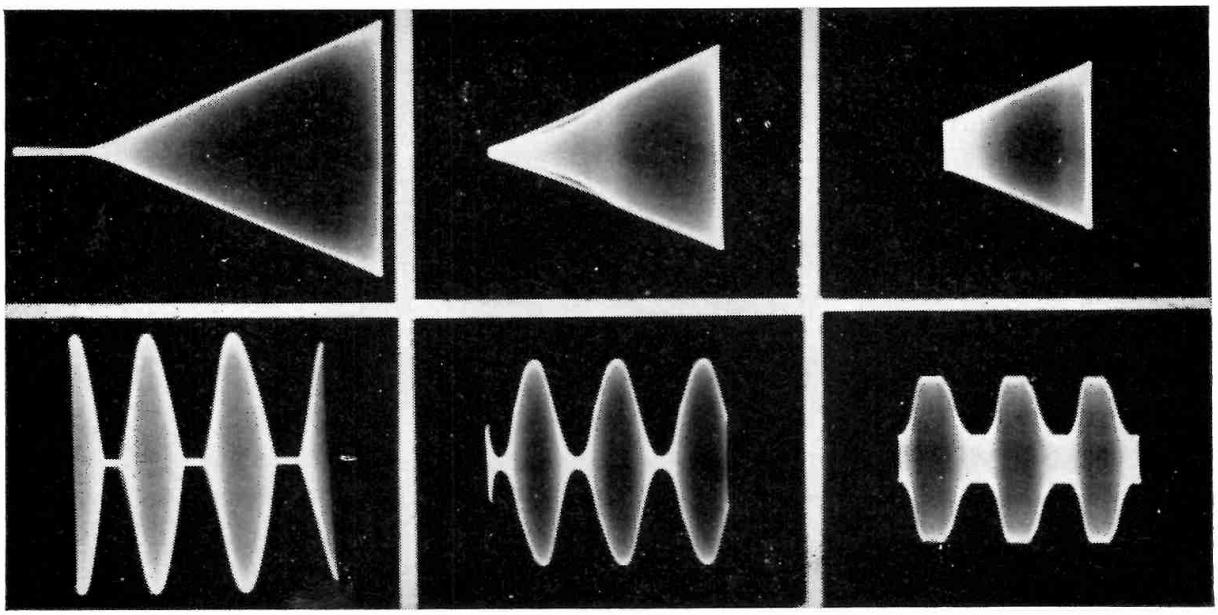
Typical Patterns

Figs. 10-20, 10-21 and 10-22 show some typical scope patterns of modulated signals for different conditions of operation. The screen-modulation patterns, Fig. 10-20, also show how the presence of even-harmonic audio distortion can be detected in the trapezoidal pattern. The pattern

Fig. 10-21—Oscilloscope patterns showing proper modulation of a plate-and-screen modulated tetrode r.f. amplifier. Upper row, trapezoidal patterns; lower row, corresponding wave-envelope patterns. In the latter a linear sweep having a frequency one-third that of the sine-wave audio modulating frequency was used, so that three cycles of the modulation envelope show in the pattern.

Unmodulated carrier. Approximately 50 per cent modulation. 100 per cent modulation.





Modulation over 100 per cent. Improper screen-circuit time constant. Insufficient audio power.

Fig. 10-22—Improper operation or design. These pictures are to the same scale as those in Fig. 10-21, on the same transmitter and with the same test setup.

to be sought in adjusting the transmitter is the one at the top in Fig. 10-20, where the top and bottom edges of the pattern continue in straight lines up to the point representing 100 per cent modulation. If these edges tend to bend over toward the horizontal at the maximum height of the wedge the amplifier is "flattening" on the modulation up-peaks. This is usually caused by attempting to get too large a carrier output, and can be corrected by tighter coupling to the antenna or by reducing the d.c. screen voltage.

Fig. 10-21 shows patterns indicating proper operation of a plate-and-screen modulated tetrode r.f. amplifier. The corresponding wave-envelope pattern is shown with each trapezoidal pattern. The slight "tailing off" at the modulation down peak (point of the wedge) can be minimized by careful adjustment of r.f. grid excitation and plate loading.

Several types of improper operation are shown in Fig. 10-22. In the photos at the left the linearity of the r.f. stage is good but the amplifier is being modulated over 100 per cent. This is shown by the maximum height of the pattern (compare with the unmodulated carrier of Fig. 10-21) and by the bright line extending from the point of the wedge (or between sections of the envelope).

The patterns in the center, Fig. 10-22, show the effect of a too-long time constant in the screen circuit, in an amplifier getting its screen voltage through a dropping resistor, both plate and screen being modulated. The "double-edged" pattern is the result of audio phase shift in the screen circuit combined with varying screen-to-cathode resistance during modulation. The over-all effect is to delay the rise in output amplitude during the up-sweep of the modulation cycle, slightly distorting the modulation envelope as shown in the wave-envelope pattern. This effect, which becomes more pronounced as the audio modulating frequency is increased, is usually absent at low modulation percentages but develops rapidly as the modulation approaches 100 per cent. It can be reduced by reducing the screen bypass capacitance, and also by connecting resistance (to be determined experimentally, but of the same order as the screen dropping resistance) between screen and cathode.

The right-hand pictures in Fig. 10-22 show the effect of insufficient audio power. Although the trapezoidal pattern shows good linearity in the r.f. amplifier, the wave-envelope pattern shows flattened peaks (both positive and negative) in the modulation envelope even though the audio signal applied to the amplifier was a sine wave. More speech-amplifier gain merely increases the flattening without increasing the modulation percentage in such a case. The remedy is to use a larger modulator or less input to the modulated r.f. stage. In some cases the trouble may be caused by an incorrect modulation-transformer turns ratio, causing the modulator to be overloaded before its maximum power output capabilities are reached.

Faulty Patterns

The pattern defects shown in Fig. 10-22 are only a few out of many that might be observed in the testing of a phone transmitter, all capable of being interpreted in terms of improper operation in some part of the transmitter. It is well to keep in mind, however, that it is not always the transmitter that is at fault when the scope shows an unusual pattern The trouble may be in some defect in the test setup.

Patterns representative of two common faults

of this nature are shown in Fig. 10-23. The upper picture shows what happens to the trapezoidal pattern when the audio voltage applied to the horizontal plates of the c.r. tube is not exactly in phase with the modulation envelope. The normal straight edges of the wedge are transformed into ellipses which in the case of 100 per cent modulation (shown) touch at the horizontal axis and reach maximum heights equal to the height of the normal wedge at the modulation up-peak. Such a phase shift can occur (and usually will) if the audio voltage applied to the c.r. tube deflection plates is taken from any point in the audio system other than where it is applied to the modulated r.f. stage. The coupling capacitor shown in the recommended circuit of Fig. 10-18 must have very low reactance compared with the resistance of R_1 and R_2 in series — not larger than a few per cent of the resistance.

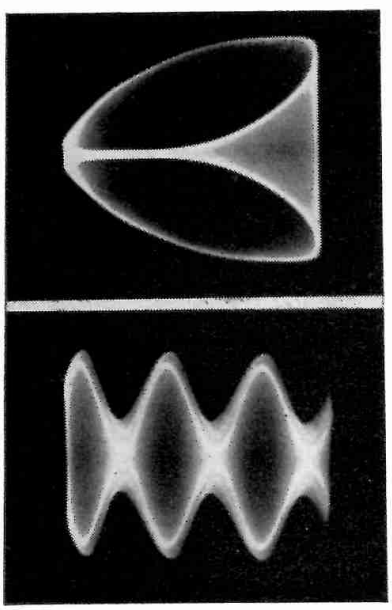


Fig. 10-23—Upper photo—Audio phase shift in coupling circuit between transmitter and horizontal deflection plates. Lower photo—Hum on vertical deflection plates.

The wave-envelope pattern in Fig. 10-23 shows the effect of hum on the vertical deflection plates. This may actually be on the carrier (poor power-supply filtering) or may be introduced in some way from the a.c. line through stray coupling between the scope and the line or because of poor grounding of the scope, transmitter or modulator.

It is important that r.f. from the *modulated stage only* be coupled to the oscilloscope, and then only to the vertical plates. 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 position where the unwanted pick-up disappears, a small bypass capacitor (10 $\mu\mu$f. or more) 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.

MODULATION CHECKING WITH THE PLATE METER

The plate milliammeter of the modulated amplifier provides a simple and fairly reliable means for checking the performance of a phone transmitter, although it does not give nearly as definite information as the oscilloscope does. If the modulated amplifier is perfectly linear, its plate current will not change when modulation is applied if

1) the upward modulation percentage does not exceed the modulation capability of the amplifier,

2) the downward modulation does not exceed 100 per cent, and

3) there is no change in the d.c. operating voltages on the transmitter when modulation is applied.

The plate current should be constant, ideally, with any of the methods of modulation discussed in this chapter, with the single exception of the controlled-carrier system. The plate meter cannot give a reliable check on the performance of the latter system because the plate current increases with the intensity of modulation. With this system the plate-current variations should be correlated with the transmitter performance as observed on an oscilloscope, if the plate meter is to be used for checking modulation.

Plate 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) R.f. amplifier not loaded properly to present the required value of modulating impedance to the modulator.
4) Insufficient output capacitance in the filter of the modulated-amplifier plate supply.
5) D.c. input to the r.f. amplifier, under carrier conditions, is in excess of the manufacturer's ratings for plate modulation. Alternatively, the cathode emission of the amplifier tubes may be low.
6) In plate-and-screen modulation of tetrodes or pentodes, the screen is not being sufficiently modulated along with the plate. In systems in which the d.c. screen voltage is obtained through a dropping resistor, a downward dip in plate current may occur if the screen bypass capacitance is large enough to bypass audio frequencies.
7) Poor voltage regulation of the modulated-amplifier plate supply. This may be caused by voltage drop in the supply itself, when the modulated amplifier and a Class B amplifier are operated from the same supply, or may be caused by voltage drop in the primary supply from the power line when the modulator load is thrown on. It is readily

checked by measuring the voltage with and without modulation. Poor line regulation will be shown by a drop in filament voltage with modulation.

Any of the following may cause an upward shift in plate current:

1) Overmodulation (excessive audio power, audio gain too high).
2) Incomplete neutralization of the modulated amplifier.
3) Parasitic oscillation in the modulated amplifier.

Grid Modulation

With any type of grid modulation, any of the following may cause a downward shift in modulated-amplifier plate current:

1) Too much r.f. excitation.
2) Insufficient grid bias, particularly with control-grid modulation. Grid bias is usually not critical with screen and suppressor modulation, the value of grid leak recommended for c.w. operation being satisfactory.
3) With control-grid modulation, excessive resistance in the bias supply.
4) Insufficient output capacitance in plate-supply filter.
5) Plate efficiency too high under carrier conditions; amplifier is not loaded heavily enough.

Because grid modulation is not perfectly linear (always less so than plate modulation) an amplifier that is properly designed and operated may show a small upward plate-current shift with modulation, 10 per cent or less with sine-wave modulation and amounting to an occasional upward flicker with voice. An upward plate current shift in excess of this may be caused by

1) Overmodulation (excessive modulating voltage).
2) Regeneration (incomplete neutralization).
3) With control-grid or suppressor modulation, bias too great.
4) With screen modulation, d.c. screen voltage too low.
5) Audio distortion in modulator.

In grid-modulation systems the modulator is not *necessarily* operating linearly if the plate current stays constant with or without modulation. It is readily possible to arrive at a set of operating conditions in which flattening of the up-peaks is just balanced by overmodulation downward, resulting in practically the same plate current as when the transmitter is unmodulated. The oscilloscope provides the only certain check on grid modulation.

COMMON TROUBLES IN THE PHONE TRANSMITTER

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 with the voice at 100 per cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing. 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 (to a cold water pipe, for example) on the microphone and speech system usually is essential to hum-free operation.

Spurious Sidebands

A superheterodyne receiver having a variable-selectivity 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. An "S"-meter reading of about half scale is satisfactory. With the crystal filter in its sharpest position 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 "clicks" or crackles well away from the carrier frequency. Sidebands more than 3 to 4 kilocycles from the carrier should be of negligible strength, compared with the carrier, in a properly modulated phone transmitter. The causes are overmodulation or nonlinear operation.

With sine-wave modulation the relative intensities of sidebands can be observed if a tone of 1000 cycles or so is used, since the crystal filter readily can separate frequencies of this order. The "S"-meter will show how the spurious side frequencies (those spaced more than the modulating frequency from the carrier) compare with the carrier itself. Without an "S"-meter, the a.v.c. should be turned off and the b.f.o. turned on; then the r.f. gain should be set to give a moderately strong beat note with the carrier. The intensity of side frequencies can be estimated from the relative strength of the beats as the receiver is tuned through the spectrum adjacent to the carrier.

As an alternative to the sharp crystal filter, a Q-multiplier adjusted for sharpest selectivity can give equivalent results in analyzing the

spectrum of the signal if the same care is used to prevent overloading and spurious receiver responses. This generally requires keeping the r.f. and i.f. gain low.

Receivers having steep-sided band-pass filters for single-sideband reception can be used, but the technique is more difficult. If the band pass is, say, 3 kc., the signal should first be tuned in with the carrier placed at one edge of the pass band. If it is placed at the low edge, for example, the receiver should then be tuned 3 kc. *higher* so its response will be in the region just outside the normal spectrum space occupied by one sideband. Any "crackles" heard in this region represent the results of nonlinearity or overmodulation. This assumes that the precautions mentioned above with respect to receiver overloading have been carefully observed.

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. pickup — 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. pickup, and a ground connection separate from that to which the transmitter is connected is advisable.

If the transmitter is "hot" with r.f., the cause usually is to be found in the method of coupling to the antenna. Any form of coupling that involves either a direct or capacitive connection between the transmitter and the transmission line is likely to cause the transmitter chassis to assume an r.f. potential above ground because of "parallel" type currents on the line. An earth connection to the transmitter does not always help in such a case. The best remedy is to use inductive coupling between the transmitter and line, a matching circuit such as is described in the chapter on transmission lines being suitable.

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 generated 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 the plate current of the modulator 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 high-resistance (1000-ohms-per-volt or more) rectifier-type voltmeter (copper-oxide or germanium 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 value as an indicator of overmodulation. As explained earlier, the d.c. plate current stays constant if the amplifier is linear. 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 nonlinearity. However, since it is possible that under some operating conditions the plate current will remain constant even though the amplifier is considerably overmodulated, an indicator of this type is not wholly reliable unless it has been checked against an oscilloscope.

Suppressed-Carrier and Single-Sideband Techniques

A fully modulated a.m. signal has two-thirds of its power in the carrier and only one-third in the sidebands. The sidebands carry the intelligence to be transmitted; the carrier "goes along for the ride" and serves only to demodulate the signal at the receiver. By eliminating the carrier and transmitting only the sidebands or just one sideband, the available transmitter power is used to greater advantage. The carrier must be reinserted at the receiver, but this is no great problem, as explained later under "Receiving Suppressed-Carrier Signals."

Assuming that the same final-amplifier tube or tubes are used either for normal a.m. or for single sideband, carrier suppressed, it can be shown that the use of s.s.b. can give an effective gain of up to 9 db. over a.m. — equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that so often spoils communication in congested phone bands.

DOUBLE-SIDEBAND GENERATORS

The carrier can be suppressed or nearly eliminated by an extremely sharp filter or by using a **balanced modulator.** The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the r.f. drive in parallel, and connecting the output in push-pull. Balanced modulators can also be connected with the r.f. drive and audio inputs in push-pull and the output in parallel with equal effectiveness. The choice of a balanced modulator circuit is generally determined by constructional considerations and the method of modulation preferred by the builder. Vacuum-tube balanced modulators can be operated at high power levels and the double-sideband output can be used directly into the antenna. A d.s.b. signal can be copied by the same methods that are used for single-sideband signals, provided the receiver has sufficient selectivity to reject one of the sidebands.

In any balanced-modulator circuit there will be no output with no audio signal. When push-pull audio is applied, the balance is upset, and one branch will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not balanced for the sidebands, and they will appear in the output.

In the rectifier-type balanced modulators shown in Fig. 11-1, the diode rectifiers are connected in such a manner that, if they have equal forward resistances, no r.f. can pass from the carrier source to the output circuit via either of

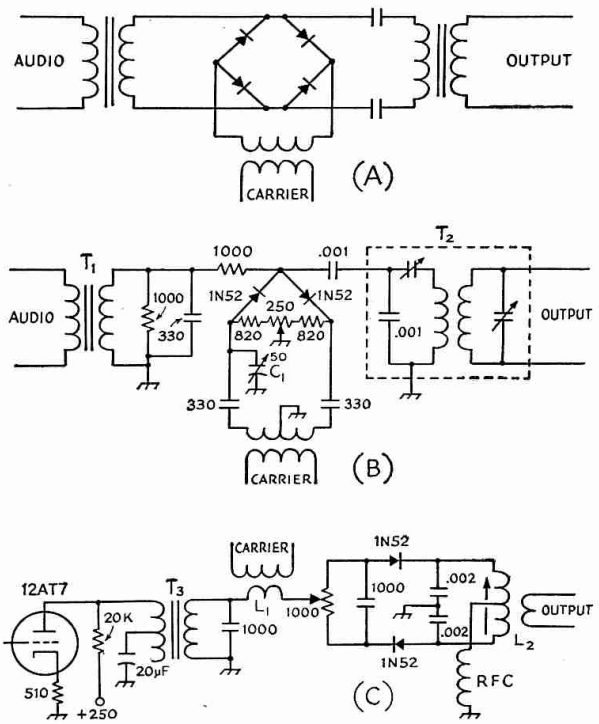


Fig. 11-1—Typical rectifier-type balanced modulators.

The circuit at A is called a "bridge" balanced modulator and has been widely used in commercial work.

The balanced modulator at B is shown with constants suitable for operation at 450 kc. It is useful for working into a crystal bandpass filter. T_1 is a transformer designed to work from the audio source into a 600-ohm load, and T_2 is an ordinary i.f. transformer with the trimmer reconnected in series with a 0.001-μf. capacitor, for impedance-matching purposes from the modulator. The capacitor C_1 is for carrier balance and may be found unnecessary in some instances—it should be tried connected on either side of the carrier input circuit and used where it is more effective. The 250-ohm potentiometer is normally all that is required for carrier balance. The carrier input should be sufficient to develop several volts across the resistor string.

The balanced modulator circuit at C is shown with constants suitable for operation at 3.9 Mc. T_3 is a small step-down output transformer (UTC R-38A), shunt-fed to eliminate d.c. from the windings. L_1 can be a small coupling coil wound on the "cold" end of the carrier-oscillator tank coil, with sufficient coupling to give two or three volts of r.f. across its output. L_2 is a slug-tuned coil that resonates to the carrier frequency with the effective 0.001 μf. across it. The 1000-ohm potentiometer is for carrier balance.

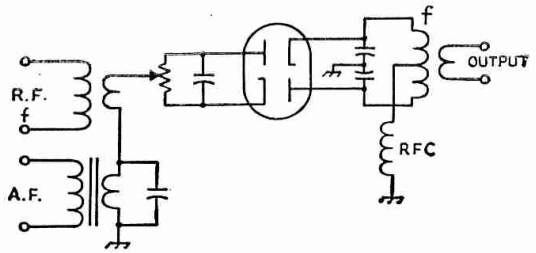


Fig. 11-2—A twin-diode balanced-modulator circuit. This is essentially the same as the circuit in Fig. 11-1C, and differs only in that a twin diode is used instead of dry rectifiers. The heater circuit for the twin diode can be connected in the usual way (one side grounded or center tap grounded).

the two possible paths. The net effect is that no r.f. energy appears in the output. When audio is applied, it unbalances the circuit by biasing the diode (or diodes) in one path, depending upon the instantaneous polarity of the audio, and hence some r.f. will appear in the output. The r.f. in the output will appear as a double-sideband suppressed-carrier signal. (For a more complete description of diode-modulator operation, see "Diode Modulators," *QST*, April, 1953, p. 39.)

In any diode modulator, the r.f. voltage should be at least 6 or 8 times the peak audio voltage, for minimum distortion. The usual operation involves a fraction of a volt of audio and several volts of r.f. The diodes should be matched as closely as possible — ohmmeter measurements of their forward resistances is the usual test.

(The circuit of Fig. 11-1B is described more fully in Weaver and Brown, "Crystal Lattice Filters for Transmitting and Receiving," *QST*, August, 1951. The circuit of Fig. 11-1C is suitable for use in a double-balanced-modulator circuit and is so described in "SSB, Jr.," *General Electric Ham News*, September, 1950.)

Vacuum-tube diodes can also be used in the two- and four-diode balanced-modulator circuits, and many operators consider them superior to the dry rectifier circuits. A typical balanced modulator circuit using a twin diode (6AL5, 6H6, etc.) is shown in Fig. 11-2. In phasing-type s.s.b. generators (described later) two of these modulators are required, and they are usually worked into a common output circuit. (For a description of a complete s.s.b. exciter using 6AL5 balanced modulators, see Vitale, "Cheap and Easy S.S.B.," *QST*, March, 1956, and May, 1958.)

Another form of balanced modulator uses a "beam-deflection" tube, and it is capable of high degrees of carrier suppression (60 db.) with good output (4 volts peak-to-peak) and low distortion (45 db.). A typical circuit, useful in the frequency range 250 to 5000 kc., is shown in Fig 11-3. A carrier signal of 10 volts peak-to-peak is applied to the No. 1 grid, and a maximum audio signal of 2.8 volts peak-to-peak is introduced at one of the deflector electrodes; the other deflector is bypassed. With no audio signal, the output can be minimized by adjustment of the balance controls R_1 and R_2. When the balance is upset by an audio signal, the beam is deflected back and forth between the two plates, and a double-sideband suppressed-carrier signal appears in the output.

Since stray magnetic fields may upset the balance, the 7360 should be mounted as far as possible from components with magnetic fields. Plate and deflection-electrode circuits should be symmetrical to minimize capacitive unbalance.

SINGLE-SIDEBAND GENERATORS

Two basic systems for generating s.s.b. signals are shown in Fig. 11-4. 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, and most filters used by amateurs are designed to work somewhere around 500 kc. Good sideband filtering can be done at frequencies as high as 5 Mc. by using multiple-crystal filters. The low-frequency oscillator output is combined with the audio output of a speech amplifier in a balanced modulator, and only the upper and lower sidebands appear in the output. One of the sidebands is passed by the filter and the other rejected, so that an s.s.b. signal is fed to the mixer. The 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. When the s.s.b. signal is generated around 500 kc. it may be necessary to convert twice to reach the operating frequency, since this sim-

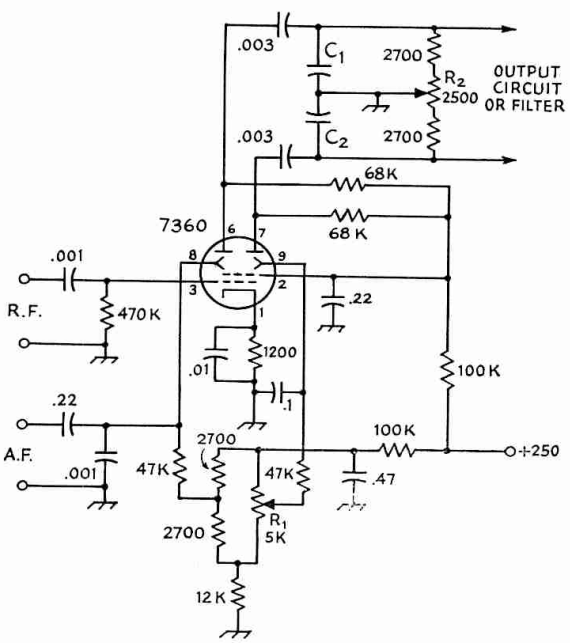


Fig. 11-3—A beam-deflection balanced modulator works well to 5 Mc., giving excellent carrier suppression with low distortion. Capacitances are in μf.

C_1, C_2—To resonate output circuit or filter.
R_1—Carrier balance control.
R_2—Quadrature balance contro'.

plifies the problem of rejecting the "image" frequencies resulting from the heterodyne process. The problem of image frequencies in the frequency conversions of s.s.b. signals differs from the problem in receivers because the beating-oscillator frequency becomes important. Either balanced modulators or sufficient selectivity must be used to attenuate these frequencies in the output and hence minimize the possibility of unwanted radiations. (Examples of filter-type exciters can be found in *QST* for June, 1958, and January, 1956.)

The second system is based on the phase 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 operating 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 balanced modulators. The carrier is suppressed in the modulators, and the relative phases of the sidebands are such that one sideband is balanced out and the other is augmented in the combined output. If the output from the balanced modulators is high enough, such an s.s.b. exciter can work directly into the antenna, or the power level can be increased in a following amplifier.

Properly adjusted, either system is capable of good results. Arguments in favor of the filter system are that it is somewhat easier to adjust without an oscilloscope, since it requires only a receiver and a v.t.v.m. for alignment, and it is more likely to remain in adjustment over a long period of time. The chief argument against it, from the amateur viewpoint, is that it requires quite a few stages and at least one frequency conversion after modulation. The phasing system requires fewer stages and can be designed to require no frequency conversion, but its alignment and adjustment are often considered to be a little "trickier" than that of the filter system. This probably stems from lack of familiarity with the system rather than any actual difficulty, and now that commercial preadjusted audio-phasing networks are available, most of the alignment difficulty has been eliminated. In most cases the phasing system will cost less to apply to an existing transmitter.

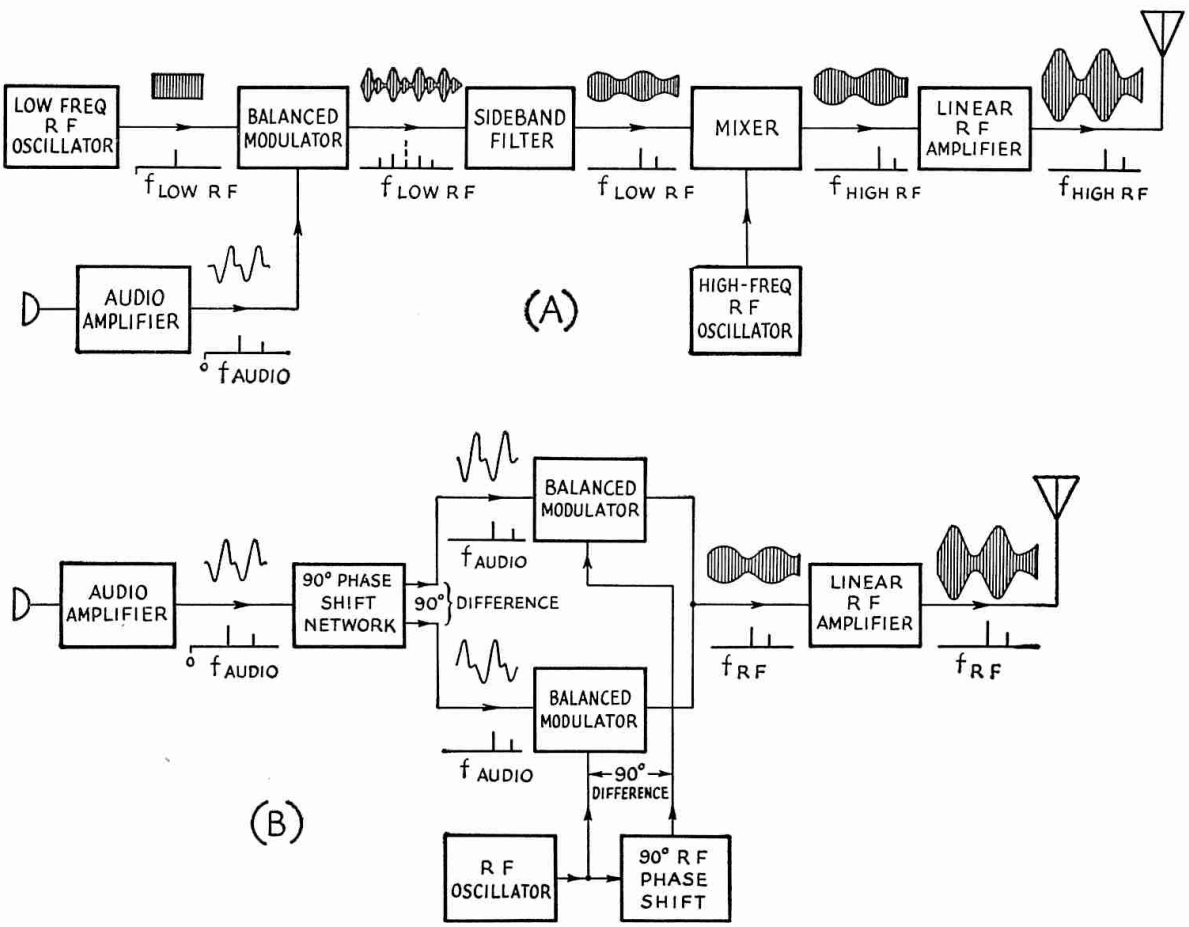


Fig. 11-4—Two basic systems for generating single-sideband suppressed-carrier signals. Representations of a typical envelope picture (as seen on an oscilloscope) and spectrum picture (as seen on a very selective panoramic receiver) are shown above and below the connecting links.

Regardless of the method used to generate a s.s.b. signal of 5 or 10 watts, the minimum cost will be found to be higher than for an a.m. transmitter of the same low power. However, as the power level is increased, the s.s.b. transmitter becomes more economical than the a.m. rig, both initially and from an operating standpoint.

Phasing-Type S.S.B. Exciters

It should be obvious that a phasing-type s.s.b. exciter can take many forms, but in general it will consist of a speech amplifier, audio phase-shift network, audio amplifier, balanced modulators, r.f. source, r.f. phase-shift network, and r.f. amplifier. If operation on a band other than that of the r.f. source, a mixer stage will also be required, for heterodyning the signal to the desired frequency. Since there are several balanced-modulator, audio- and r.f. phasing circuits, it is apparent that many different combinations are available. One of the simplest of all combinations is that shown in Fig. 11-5.

Referring to Fig. 11-5, the speech amplifier builds up the signal from a crystal microphone to a useful level. The audio signal is then fed to an audio phase-shift network, *PSN*, which applies equal-amplitude audio signals 90 degrees out of phase to the grids of the 12AT7 audio amplifier. The two audio signals, 90 degrees out of phase, are applied to two balanced modulators that have their outputs in parallel (L_3). The r.f. excitation to the balanced modulators is also 90 degrees out of phase, obtained by coupling from the two tuned circuits at L_1 and L_2. A 6AG7 linear amplifier, operating Class AB_1, follows the balanced-modulator stage and provides about 5 watts peak envelope output.

The gain control in the speech amplifier sets the gain to the proper level, depending upon the

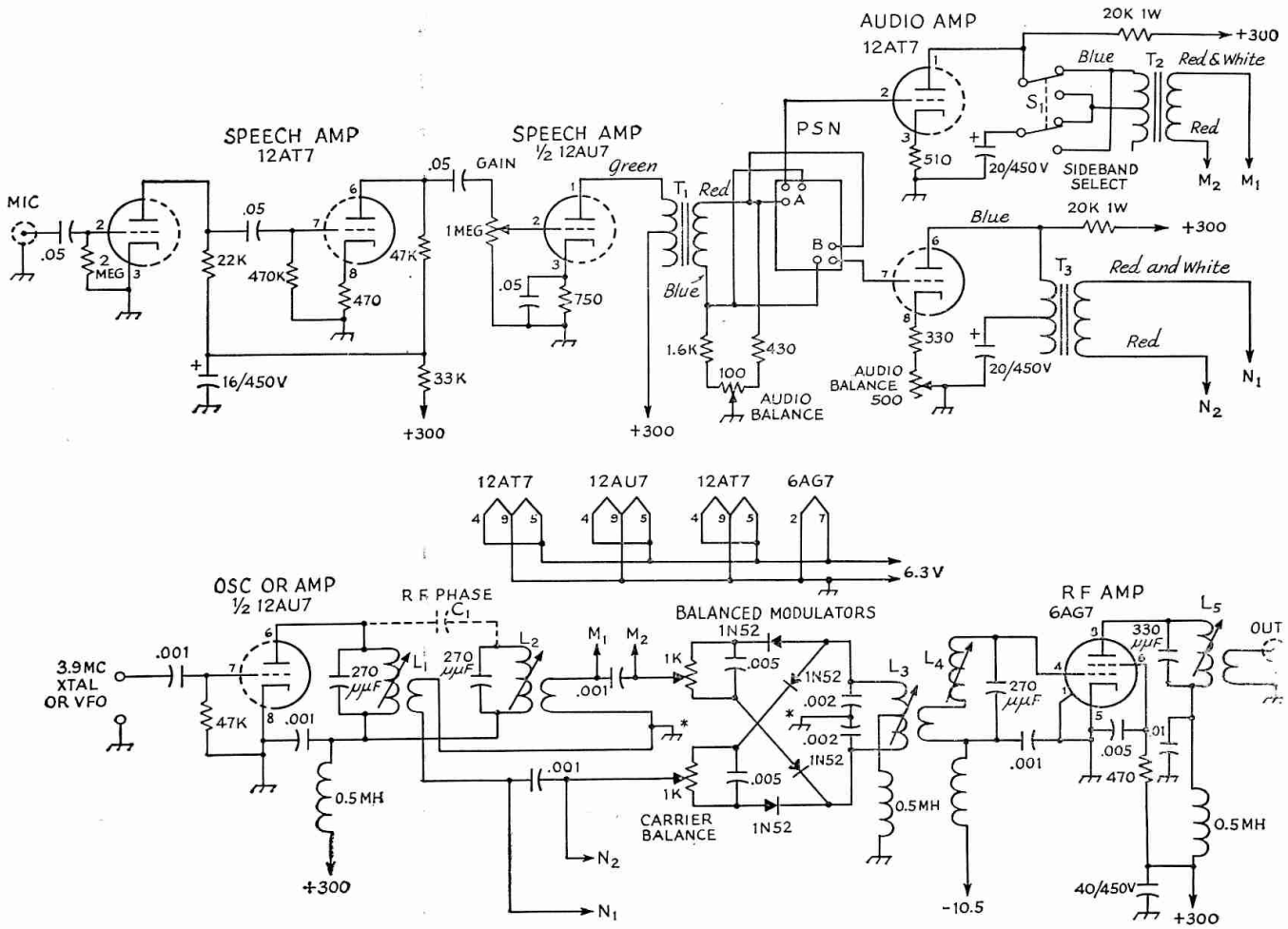


Fig. 11-5—Schematic of a phasing-type s.s.b. exciter. Capacitance in μf. unless otherwise noted—resistors are ½-watt unless otherwise noted. Chassis grounds marked * should be the same.

C_1—5 or 10 μμf. if inductive coupling between L_1 and L_2 not sufficient.

T_1—Single plate to push-pull grid, 1:3 ratio (Stancor A53C).

T_2, T_3—6-watt universal output transformer, 30 ohms output (UTC R-38A).

L_1, L_2—32 turns No. 22 enam. closewound on ½-inch diameter iron-core tuned form (Millen 69046). Link turn is 6 turns hook-up wire wound adjacent to cold end.

L_3—16 turns No. 22 enam., spaced to occupy 1-inch length on ½-inch diameter iron-core-tuned form (Millen 69046), tapped at center. One-turn link wound at center.

L_4—Same as L_1; no link.

L_5—25 turns No. 22 enam. closewound on ½-inch iron-core-tuned form (Millen 69046). Link of 4 turns at cold end.

S_1—D.p.d.t. toggle or rotary.

PSN—Audio phase-shift network (Millen 75012). See Fig. 11-6.

microphone and how the operator uses it. Since the audio phase-shift network, *PSN*, has unequal gains through its two channels, unequal-amplitude audio is required at the input to obtain equal signals in the output. This is obtained through proper adjustment of the 100-ohm input audio balance control. To compensate for lack of uniformity in audio-amplifier gains, a 500-ohm audio balance control is provided in the cathode of a 12AT7 section. R.f. carrier balance is obtained by proper setting of the 1000-ohm carrier balance controls. The sideband in use (upper or lower) is selected by S_1, which reverses the audio signal in one of the channels. The r.f. phasing adjustment is obtained by the tuning of L_1 and L_2.

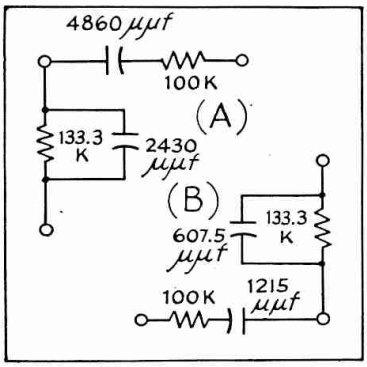


Fig. 11-6—Schematic of the phase-shift network marked *PSN* in Fig. 11-5. Resistors and capacitors should be within 1 per cent of values shown.

Construction

There are a few constructional precautions that should be observed in a unit of this type. Transformers T_2 and T_3 should preferably be mounted at right angles to each other, to minimize stray coupling. The 1N52 germanium diodes used in the balanced modulator should be checked for forward and back resistance with an ohmmeter, and the forward resistances (the lower readings) should agree within 10 per cent. The leads from the coupling loops at L_1 and L_2 should return to the balanced modulator stage in twisted pairs, and the grounding precaution mentioned in Fig. 11-5 should be observed. Coils L_1 and L_2 should be mounted parallel to each other and with a separation of about 1½ diameters — L_3 and L_4 should be mounted to minimize coupling between them and L_5 and the oscillator coils. This can be accomplished by providing shielding or using the chassis deck to separate them.

Although slug-tuned coils are shown in the schematic, capacitance-tuned circuits can of course be used. Approximately the same L/C ratios should be retained, however. If operation on another amateur band is desired, the tuned circuits can be modified accordingly, retaining the same L/C ratios, or the output of this unit can be heterodyned to the different band.

Adjustment

If v.f.o. operation is to be used, the v.f.o. signal should furnish at least 10 volts r.m.s. at the terminals. With crystal control, plug in a crystal and tune L_1 until the circuit oscillates, as indicated by a signal in a receiver tuned to the proper frequency, and then tune the circuit to a slightly higher frequency. With v.f.o. operation, the circuit is resonated in the usual manner, as indicated by a plate-current minimum.

The output from the 6AG7 stage can be checked on an oscilloscope or on a receiver. The method of coupling an oscilloscope or receiver to the exciter is shown in Fig. 11-7. When connecting to an oscilloscope, a tuned circuit is required, and the r.f. voltage developed across the tuned circuit is applied directly to the vertical deflection plates. The receiver is connected by coupling loosely through a loop and length of shielded cable; when further attenuation is required it is obtained through the use of resistors at the receiver input terminals.

With the oscillator running, tune the balanced modulator and 6AG7 circuits for maximum output — this resonates these circuits. Next adjust the carrier balance potentiometers for minimum output. Then introduce a single audio tone of around 1000 cycles at the microphone terminal. Here again it may be necessary to use a resistance voltage divider to hold the signal down and prevent overload. Advance the gain control and check the voltage at Pins 2 and 7 of the 12AT7 audio amplifier with a v.t.v.m. If they are not

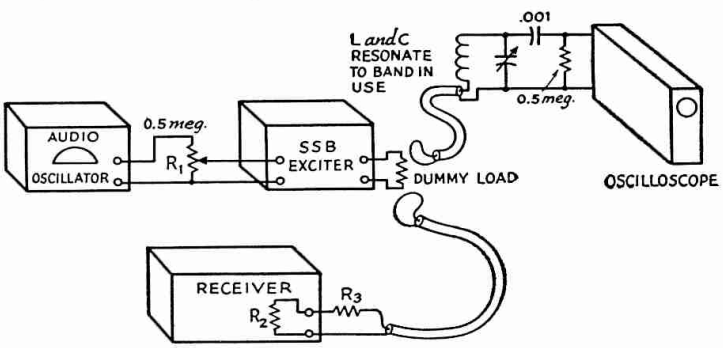


Fig. 11-7—Fundamental arrangement for using an oscilloscope and/or receiver when testing an s.s.b. exciter or transmitter. An audio oscillator is required to furnish the audio signal, and its output is best controlled by the external control R_1. The audio volume control in the s.s.b. exciter should not be turned on too far, or it should be set at the normal position if you know that position, and all volume controlling should then be done with R_1 and the output attenuator of the audio oscillator. This will reduce the chances of overloading the audio and other amplifier stages in the exciter, a common cause of distortion.

The oscilloscope is coupled to the dummy load through a loop, length of coaxial line, and an L-C circuit tuned to the operating frequency. It is necessary to go directly to the vertical deflection plates of the oscilloscope rather than through the vertical amplifier.

The receiver is coupled to the dummy load through a loop and a length of shielded line. If too much signal is obtained this way, an attenuator, R_2R_3, can be added to the input terminals of the receiver. Small values of R_2 and large values of R_3 give the most attenuation; in some cases R_2 might be merely a few inches of solid wire.

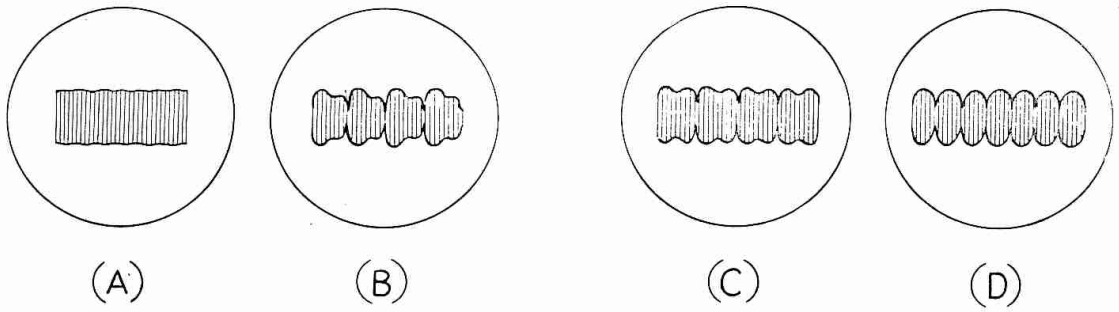


Fig. 11-8—Sketches of the oscilloscope face showing different conditions of adjustment of the exciter unit. (A) shows the substantially clean carrier obtained when all adjustments are at optimum and a sine-wave signal is fed to the audio input. (B) shows improper r.f. phase and unbalance between the outputs of the two balanced modulators. (C) shows improper r.f. phasing but outputs of the two balanced modulators equal. (D) shows proper r.f. phasing but unbalance between outputs of two balanced modulators.

equal, adjust the 100-ohm audio balance control until they are. Listening to the signal, from the 6AG7, or looking at it on the scope, should give a modulated signal. Try various settings of L_2 until the modulation is minimized, as well as touching up the 500-ohm audio balance control. With the v.t.v.m. check the r.f. voltages at the arms of the 1000-ohm carrier balance potentiometers — they should be about the same. If not, they can be brought into this condition by readjustment of the tuning conditions which, however, must be kept consistent with minimum modulation on the output signal.

The s.s.b. signal with single-tone audio input is a steady unmodulated signal. While it may not be possible to eliminate the modulation entirely, it will be possible to get it down to a satisfactorily low level. Conditions that will prevent this are improper r.f. phasing, lack of carrier balance (suppression), distortion in the audio signal (at the source or through overload in the speech amplifier), and lack of audio balance at the 12AT7 audio amplifier. Of these, the r.f. phasing is perhaps the most critical.

A final check on the signal can be made with the receiver in its most selective condition. The spectrum testing described below cannot be done with a broad receiver. Examining the spectrum near the signal, the side signals other than the main one (carrier, unwanted sidebands, and sidebands from audio harmonics) should be at least 30 db. down from the desired signal. This checking can be done with the S-meter and the a.v.c. on — in the earlier tests the a.v.c. should be off but the r.f. gain reduced low enough to avoid receiver overload.

Examples of the proper and improper scope patterns are shown in Fig. 11-8.

(For an extensive treatment of the alignment of commercial phasing-type s.s.b. exciters, see Ehrlich, "How to Adjust Phasing-Type S.S.B. Exciters," *QST*, November, 1956.)

Filter-Type S.S.B. Exciters

The basic configuration of a filter-type s.s.b. exciter was shown earlier in this chapter (Fig. 11-4). Suitable filters, sharp enough to reject the unwanted sideband above a few hundred cycles, can be built in the range 20 kc. to 5 Mc. The low-frequency filters generally use iron-cored inductors, and the new toroid forms find considerable favor at frequencies up to 50 or 60 kc. These filters are of normal band-pass constant-k and m-derived configuration. In the range 450 to 500 kc., either crystal-lattice or electromechanical filters are used. Low-frequency filters are manufactured by Barker & Williamson and by Burnell & Co., and electromechanical filters are made by the Collins Radio Co. Crystal-lattice filters are available from Hermes Electronics in the megacycle range; homemade filters generally utilize crystals from military surplus.

The frequency of the filter determines how many conversions must be made before the operating frequency is reached. For example, if the filter frequency is 30 kc. or so, it is wise to convert first to 500 or 600 kc. and then convert to the 3.9-Mc. band, to avoid the image that would almost surely result if the conversion from 30 to 3900 kc. were made without the intermediate step. When a filter at 500 kc. is used, only one conversion is necessary to operate in the 3.9-Mc. band, but 14-Mc. and higher-frequency operation would require at least two conversions to hold down the images (and local-oscillator signals if balanced mixers aren't used) and make them easy to eliminate.

The choice of converter circuit depends largely on the frequencies involved and the impedance level. At low frequencies (up to 500 kc.) and low impedances, rectifier-type balanced modulators are often used for mixers, because the balanced modulator does not show the local-oscillator frequency in its output and one source of spurious signal is minimized. At frequencies at high impedance levels, and at the higher frequencies, vacuum tubes are generally used, in straight converter or balanced-modulator circuits, de-

pending upon the need for minimizing the local-oscillator frequency in the output.

Low-frequency sideband filters in the 30- to 50-kc. range are usually low-impedance devices, and rectifier-type balanced modulators are common practice. Sideband filters in the i.f. range this can be nothing more elaborate than a shielded b.f.o. unit. The signal should be introduced at the balanced modulator, and an output indicator connected to the plate circuit of the vacuum tube following the filter. With the crystals out of the circuit, the transformers can be

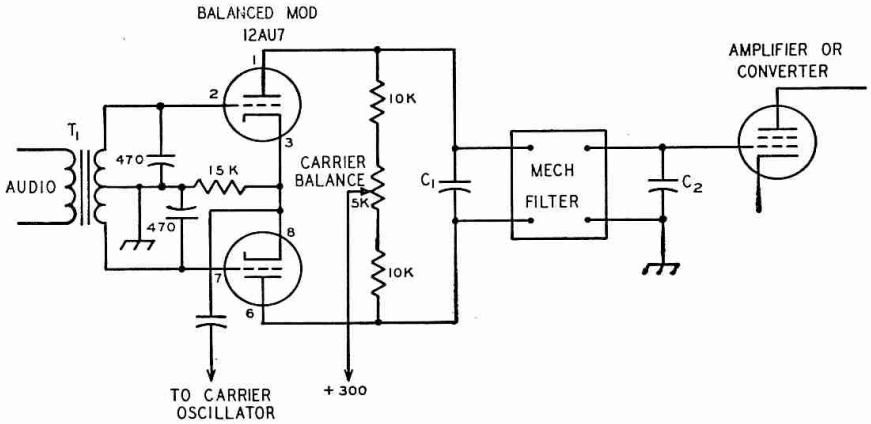


Fig. 11-9—One type of balanced-modulator circuit that can be used with a mechanical filter (Collins F455-31 or F500-31 series) in the i.f. range. The filters are furnished in various types of mountings, and the values of C_1 and C_2 will depend upon the type of filter selected.
T_1—Plate-to-push-pull grids audio transformer.

are higher-impedance circuits and vacuum-tube balanced modulators are the rule in this case. An example of one that can be used with the high-impedance (15,000 ohms) mechanical filter is shown in Fig. 11-10. The filter can be followed by a converter or amplifier tube, depending upon the signal level. Some models of the mechanical filters have a 23-db. insertion loss, while others have only 10.

Crystal-lattice filters are also used to reject the unwanted sideband. These filters can be brought close to frequency by plugging in small capacitors (10 to 25 $\mu\mu$f.) in one crystal socket in each stage and then tuning the transformers for peak output at one of the two crystal frequencies. The small capacitors can then be removed and the crystals replaced in their sockets.

Tuning the signal source slowly across the pass band of the filter and watching the output indicator will show the selectivity characteristic of the filter. The objective is a fairly flat response for about two kc. and a rapid drop-off outside

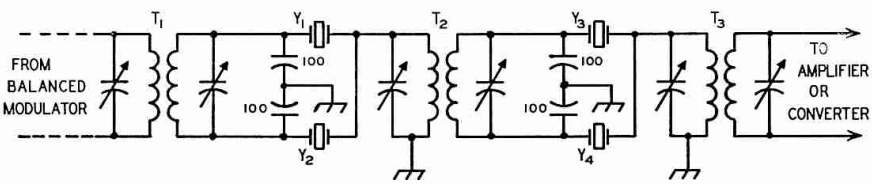


Fig. 11-10—A cascaded half-lattice crystal filter that can be used for sideband selection. The crystals are surplus type of FT-243A holders. Y_1 and Y_3 should be the same frequency and Y_2 and Y_4 should be 1.8 kc. higher. T_1, T_2, T_3—450-kc. i.f. transformers.

made from crystals in the i.f. range — many of these are still available from stores selling military surplus. A popular configuration is the "cascaded half lattice" shown in Fig. 11-10. The crystals used in this filter can be obtained at frequencies in the i.f. range, and ones that are within the ranges of the modified i.f. transformers will be satisfactory. Two 100-$\mu\mu$f. capacitors are connected across the secondary winding of two of the transformers to give push-pull output. The crystals should be obtained in pairs 1.8 kc. apart. The i.f. transformers can be either capacitor-tuned as shown, or they can be slug-tuned.

A variable-frequency signal generator of some kind is required for alignment of the filter, but this range. It will be found that small changes in the tuning of the transformers will change the shape of the selectivity characteristic, so it is wise to make a small adjustment of one trimmer, swing the frequency across the band, and observe the characteristic. After a little experimenting it will be found which way the trimmers must be moved to compensate for the peaks that will rise when the filter is out of adjustment.

The (suppressed) carrier frequency must be adjusted so that it falls properly on the slope of the filter characteristic. If it is too close to the filter mid-frequency the sideband rejection will be poor; if it is too far away there will be a lack of "lows" in the signal.

Amplification of S.S.B. Signals

AMPLIFICATION OF S.S.B. SIGNALS

When an s.s.b. signal is generated at some frequency other than the operating frequency, it is necessary to change frequency by heterodyne methods. These are exactly the same as those used in receivers, and any of the normal mixer or converter circuits can be used. One exception to this is the case where the heterodyning oscillator frequency is close to the desired output frequency. In this case, a balanced mixer should be used, to eliminate the heterodyning oscillator frequency in the output.

To increase the power level of an s.s.b. signal, a **linear amplifier** must be used. A linear amplifier is one that operates with low distortion, and the low distortion is obtained by the proper choice of tube and operating conditions. Physically there is little or no difference between a linear amplifier and any other type of r.f. amplifier stage. The circuit diagram of a tetrode r.f. amplifier is shown in Fig. 11-11; it is no different basically than the similar ones in Chapter Six. The practical differences can be found in the supply voltages for the tube and their special requirements. The proper voltages for a number of suitable tubes can be found in Table 11-I; filament-type tubes will require the addition of the filament bypass capacitors C_9 and C_{10} and the completion of the filament circuit by grounding the filament-transformer center tap. The grid bias, E_1, is furnished through an r.f. choke, although a resistor can be used if the tube is operated in Class AB_1 (no grid current). The screen voltage, E_2, must be supplied from a "stiff" source (little or no voltage change with current change) which eliminates the use of a dropping resistor from the plate supply unless a voltage-regulator tube is used to stabilize the screen voltage.

Any r.f. amplifier circuit can be adapted to linear operation through the proper choice of operating conditions. For example, the circuit in Fig. 11-11 can be modified by the use of different input and/or output coupling circuits, or by the use of another neutralizing scheme, and the resultant amplifier will still be linear if the proper operating conditions are observed. A triode or pentode amplifier circuit will differ in detail; typical circuits can be found in Chapter Six.

The simplest form of linear amplifier is the Class A amplifier, which is used almost without exception throughout receivers and low-level speech equipment. (See Chapter Three for an explanation of the classes of amplifier operation.) While its linearity can be made relatively good, it is inefficient. The theoretical limit of efficiency is 50 per cent, and most practical amplifiers run 25–35 per cent efficient at full output. At low levels this is not worth worrying about, but when the 2- to 10-watt level is exceeded something else must be done to improve this efficiency and reduce tube, power-supply and operating costs.

Class AB_1 amplifiers make excellent linear amplifiers if suitable tubes are selected. Primary advantages of Class AB_1 amplifiers are that they give much greater output than straight Class A amplifiers using the same tubes, and they do not require any grid driving power (no grid current drawn at any time). Although triodes can be used for Class AB_1 operation, tetrodes or pentodes are usually to be preferred, since Class AB_1 operation requires high peak plate current without grid current, and this is easier to obtain in tetrodes and pentodes than in most triodes.

To obtain maximum output from tetrodes, pentodes and most triodes, it is necessary to operate them in Class AB_2. Although this produces maximum peak output, it increases the driving-power requirements and, what is more important, requires that the **driver regulation** (ability to maintain wave form under varying load) be good or excellent. The usual method to improve the driver regulation is to connect a fixed resistor, R_1, across the grid circuit of the driven stage, to offer a load to the driver that is modified only slightly by the additional load of the tube when

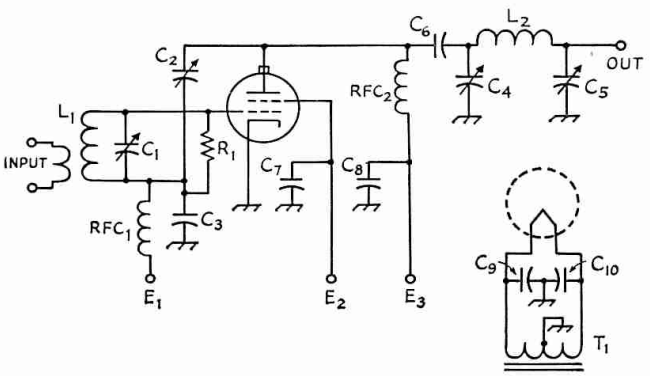


Fig. 11-11—Circuit diagram of a tetrode linear amplifier using link-coupled input tuning and pi network output coupling. The grid, screen and plate voltages (E_1, E_2 and E_3) are given in Table 11-1 for a number of tubes. Although the circuit is shown for an indirectly-heated cathode tube, the only change required when a filament type tube is used is the addition of the filament bypass capacitors C_9 and C_{10}.

Minimum voltage ratings for the capacitors are given in terms of the power supply voltages.

C_1—Grid tuning capacitor, $3E_1$
C_2—Neutralizing capacitor, $2E_3$.
C_3—Grid-circuit bypass capacitor, part of neutralizing circuit, $3E_1$.
C_4—Plate tuning capacitor, $1.5E_3$.
C_5—Output loading capacitor. 0.015 spacing for kilowatt peak.
C_6—Plate coupling capacitor, $2E_3$.
C_7—Screen bypass capacitor, $2E_2$.
C_8—H.v. bypass capacitor, $2E_3$.
C_9, C_{10}—Filament bypass capacitor.
L_1—Grid inductor.
L_2—Plate inductor.
R_1—Grid circuit swamping resistor, required for AB_2. See text.
RFC_1—Grid-circuit r.f. choke.
RFC_2—Plate r.f. choke.
T_1—Filament transformer.

TABLE 11-I — LINEAR-AMPLIFIER TUBE-OPERATION DATA FOR SINGLE SIDEBAND

Except where otherwise noted, ratings are manufacturers' for audio operation. Values given are for one tube. Driving powers represent tube losses only—circuit losses will increase the figures.

Tube	Class	Plate Voltage	Screen Voltage	D.C. Grid Voltage	Zero-Sig. D.C. Plate Current	Max.-Sig. D.C. Plate Current	Zero-Sig. D.C. Screen Current	Max.-Sig. D.C. Screen Current	Peak R.F. Grid Voltage	Max.-Sig. D.C. Grid Current	Max.-Sig. Driving Power	Max.-Rated Screen Dissipation	Max.-Rated Grid Dissipation	Avg. Plate Dissipation	Max.-Sig. Useful Power Output
2E26	AB_1	500	200	− 25	9	45	— —	10	25	0	0	2.5	—	—	15
6146	AB_1	600	200	− 50	14	115	.5	14	50	0	0	3	—	25	47
6883		750	200	− 50	12	110	.5	13	50	0	0	3	—	25	60
807	AB_2	600	300	− 30	30	100	.4	6	39	—	.1	3.5	—	25	40
1625		750	300	− 32	26	120	.3	8	46	—	.1	3.5	—	30	60
811-A	B	1000	—	0	22	175	—	—	93	—	3.8	—	—	65	124
		1250	—	0	27	175	—	—	88	13	3.0	—	—	65	155
		1500	—	− 4.5	16	157	—	—	85	—	2.2	—	—	65	170
4-65A	AB_2	1500	300	− 55[1]	35	200[2]	—	45[3]	150	15	2.3[3]	10	5	60	150
		2000	400	− 80[1]	25	270[2]	—	65[3]	190	20	3.8[3]	10	5	65	300
		2500	500	−105[1]	20	230[2]	—	45[3]	165	8	1.3[3]	10	5	65	325
7094	AB_1	2000	400	− 50	30	200	—	35	44	0	4[8]	20	—	—	250
813	AB_1	2500	750	− 95	25	145	—	27	90	0	0	—	—	—	245
	AB_2	2250	750	− 90	23	158	.8	29	115	—	.1	22	—	100	258
		2500	750	− 95	18	180	.6	28	118	—	.2	22	—	125	325
4-125A	AB_1	2000	615	−105[1]	40	135 (100)[4]	—	14 (4.0)[4]	105	0	0	20	—	—	150
		2500	555	−100[1]	35	120 (85)[4]	—	10 (3.0)[4]	100	0	0	20	—	—	180
		3000	510	− 95[1]	30	105 (75)[4]	—	6.0 (1.5)[4]	95	0	0	20	—	—	200
	AB_2	1500	350	− 41[1]	44	200	0	17	141	9	1.25	20	5	125	175
		2000	350	− 45[1]	36	150	0	3	105	7	.7	20	5	125	175
		2500	350	− 43[1]	47	130	0	3	89	6	.5	20	5	122	200
7034/ 4X150A	AB_1	1000	300	− 50	50	225	0	11	50	0	0	12	—	—	115
		1500	300	− 50	50	225	0	11	50	0	0	12	—	—	200
		1800	300	− 50	50	225	0	11	50	0	0	12	—	—	250
4-250A	AB_1	2500	600	−115	65	230 (170)[4]	—	15 (3.5)[4]	115	0	0	35	—	—	335
		3000	600	−110	55	210 (150)[4]	—	12 (2.5)[4]	110	0	0	35	—	—	400
		3500	555	−105	45	185 (130)[4]	—	9.5 (2.0)[4]	105	0	0	35	—	—	425
		4000	510	−100	40	165 (115)[4]	—	7.5 (1.5)[4]	100	0	0	35	—	—	450
	AB_2	1500	300	− 48[1]	50	243	0	17	96	11	1.1	35	10	150	214
		2000	300	− 48[1]	60	255	0	13	99	12	1.2	35	10	185	325
		2500	300	− 51[1]	60	250	0	12	100	11	1.1	35	10	205	420
		3000	300	− 53[1]	63	237	0	17	99	10	1	35	10	190	520
304TL	AB_1	1500	—	−118[1]	135	286	—	—	118	0	0	—	—	—	128
		2000	—	−170[1]	100	273	—	—	170	0	0	—	—	—	245
		2500	—	−230[1]	80	242	—	—	230	0	0	—	—	—	305
		3000	—	−290[1]	65	222	—	—	290	0	0	—	—	—	365
PL-6569	B[5]	2500	—	− 60[1]	40	300	—	—	180	80	70[6]	—	—	—	550
		3500	—	− 90[1]	30	270	—	—	220	68	75[6]	—	—	—	760
		4000	—	−105[1]	24	250	—	—	205	42	60[6]	—	—	—	800
PL-6580	B[5]	2500	—	− 50	60	350	—	—	195	95	75[6]	—	—	—	610
		3500	—	− 85	45	300	—	—	210	65	68[6]	—	—	—	765
		4000	—	−100	40	300	—	—	230	65	72[6]	—	—	—	910
PL-172	AB_1	2000	500[7]	−110	200	800	9	48	110	0	0	35	—	—	1020
		2500	500[7]	−110	220	800	9	43	110	0	0	35	—	—	1280
		3000	500[7]	−115	220	780	9	41	110	0	0	35	—	—	1540

[1] Adjust to give stated zero-signal plate current.
[2] Single-sideband suppressed-carrier ratings, voice signal.
[3] Approximate value.
[4] Values in parentheses are with two-tone test signal.
[5] Grounded-grid circuit.
[6] Includes bias loss, grid dissipation, and feed-through power.
[7] +75 v. suppressor grid.
[8] 60 Mc.

it is driven into the grid-current region. This increases the driver's output-power requirements. Further, it is desirable to make the grid circuit of the Class AB_2 stage a high-C circuit, to improve regulation and simplify coupling to the driver. A "stiff" bias source is also required, since it is important that the bias remain constant, whether or not grid current is drawn.

Class B amplifiers are theoretically capable of 78.5 per cent efficiency at full output, and practical amplifiers run at 60–70 per cent efficiency at full output. Triodes normally designed for Class B audio work can be used in r.f. linear amplifiers and will operate at the same power rating and efficiency provided, of course, that the tube is capable of operation at the radio frequency. The operating conditions for r.f. are substantially the same as for audio work — the only difference is that the input and output transformers are replaced by suitable r.f. tank circuits. Further, in r.f. circuits it is readily possible to operate only one tube if only half the power is wanted — push-pull is not a necessity in Class B r.f. work. However, the r.f. harmonics may be higher in the case of the single-ended amplifier, and this should be taken into consideration if TVI is a problem.

For proper operation of Class AB_2 and B amplifiers, and to reduce harmonics and facilitate coupling, the input and output circuits should not have a low C-to-L ratio. A good guide to the proper size of tuning capacitor will be found in Chapter Six; in case of any doubt, it is well to be on the high-capacitance side. When zero-bias tubes are used, it may not be necessary to add much "swamping" resistance across the grid circuit, because the grids of the tubes load the circuit at all times. However, in AB_2 operation, the swamping resistor should be such that it dissipates from five to ten times the power required by the grids of the tubes, insuring an almost constant load on the driver stage and good regulation of the r.f. grid voltage. In turn this means that at least five to ten times more driving power will be required than is indicated in Table 11-I. Where an excess of driving power is available, it is generally better to increase the loading (decrease the resistance of the swamping resistor) to the point where the maximum available driver power is utilized on peaks.

Before going into detail on the adjustment and loading of the linear amplifier, a few general considerations should be kept in mind. If proper operation is expected, it is essential that the amplifier be so constructed, wired and neutralized that no trace of regeneration or parasitic instability remains. Needless to say, this also applies to the stages ahead of it.

The bias supply to the Class AB_2 or B linear amplifier should be quite stiff, such as batteries or some form of voltage regulator. If nonlinearity is noticed when testing the unit, the bias supply may be checked by means of a large electrolytic capacitor. Simply shunt the supply with 100 μf. or so of capacity and see if the linearity improves. If so, rebuild the bias supply for better regulation. *Do not rely on a large capacitor alone.*

Where tetrodes or pentodes are used, the screen supply should have good regulation and its voltage should remain constant under the varying current demands. If the maximum screen current does not exceed 30 or 35 ma., a string of VR tubes in series can be used to regulate the screen voltage. If the current demand is higher, it may be necessary to use an electronically regulated power supply or a heavily bled power supply with a current capacity of several times the current demand of the screen circuit.

Where VR tubes are used to regulate the screen supply, they should be selected to give a regulated voltage as close as possible to the tube's rated voltage, but it does not have to be exact. Minor differences in idling plate current can be made up by readjusting the grid bias.

The plate voltage applied to the linear amplifier should be held as constant as possible under the varying current-demand conditions. This condition can be met by using low-resistance transformers and inductors and by using a large value of output capacitor in the power-supply filter. An output capacitor value three or four times the minimum required for normal filtering (Chapter Seven) is reasonable. Although some slight improvement can be obtained by using still higher values of output capacitance, the problem of turning on the supply without blowing fuses (from the initial surge) starts to become significant.

One should bear in mind that the same amplifier can be operated in several classes of operation by merely changing the operating conditions (bias, loading, drive, screen voltage, etc.). However, when the power sensitivity of an amplifier is increased, as by changing the operation from Class AB_2 to Class AB_1, the stability requirements for the amplifier become stringent.

From the standpoint of ease of adjustment and availability of proper operating voltages, a linear amplifier with Class AB_1 tetrodes or pentodes or one with zero-bias Class B triodes would be first choice. The Class B amplifier would require more driving power. (For examples of Class AB_1 tetrode amplifiers, see Russ, "The 'Little Firecracker' Linear Amplifier," *QST*, Sept., 1953, Eckhardt, "The Single Side-Saddle Linear," *QST*, Nov., 1953, Wolfe and Romander, "A 4X-250B Linear," *QST*, Nov., 1956, Muir, "Grounded-Grid Tetrode Kilowatt," *QST*, April, 1957, and Rinaudo, "Compact AB_1 Kilowatt," *QST*, Nov., 1957.)

Table 11-I lists a few of the more popular tubes commonly used for s.s.b. linear-amplifier operation. Except where otherwise noted, these ratings are those given by the manufacturer for audio work and as such are based on a sine-wave signal. These ratings are adequate ones for use in s.s.b. amplifier design, but they are conservative for such work and hence do not necessarily represent the maximum powers that can be obtained from the tubes in voice-signal s.s.b. service. In no case should the *average* plate dissipation be exceeded for any considerable length of time, but

the nature of a s.s.b. signal is such that the average plate dissipation of the tube will run well below the peak plate dissipation.

Getting the most out of a linear amplifier is done by increasing the peak power without exceeding the average plate dissipation over any appreciable length of time. This can be done by raising the plate voltage or the peak current (or both), provided the tube can withstand the increase. However, the manufacturers have not released any data on such operation, and any extrapolation of the audio ratings is at the risk of the amateur. A 35- to 50-per cent increase above plate-voltage ratings should be perfectly safe in most cases. In a tetrode or pentode, the peak plate current can be boosted some by raising the screen voltage.

When running a linear amplifier at considerably higher than the audio ratings, the "two-tone test signal" (described later) should never be applied at full amplitude for more than a few seconds at any one time. The above statements about working tubes above ratings apply only when a voice signal is used — a prolonged whistle or two-tone test signal may damage the tube. (For a method of adjusting amplifiers safely at high input, see Goodman, "Linear Amplifiers and Power Ratings," *QST*, August, 1957.)

Grounded-Grid Amplifiers With Filament-Type Tubes

It is not necessary to use indirectly heated cathode type tubes in grounded-grid circuits, and filament-type tubes can be used just as effectively. However, it is necessary to raise the filament above r.f. ground, and one way is shown in Fig. 11-12. Here filament chokes are used between the filament transformers and the tube socket. The inductance of the r.f. chokes does not have to be very high, and 5 to 10 μh. will usually suffice from 80 meters on down. The current-carrying capacities of the r.f. chokes must be adequate for the tube or tubes in use, and if the resistance of the chokes is too high the filament voltage *at the tube socket* may be too low and the tube life will be endangered. In such a case, a higher-voltage filament transformer can be used, with its primary voltage cut down until the voltage at the tube socket is within the proper limits.

Filament chokes can be wound on ceramic or wooden forms, using a wire size large enough to carry the filament current without undue heating. Large cylindrical ceramic antenna insulators can be used for the forms. If enameled wire is used, it should be spaced from half the diameter to the diameter of the wire; heavy string can be used for this purpose. The separate chokes indicated in Fig. 11-12 are not essential; the two windings can be wound in parallel. In this case it is not necessary to space all windings; the two parallel wires can be treated as one wire, winding them together with a single piece of string to space the turns. Enameled wire can be used because the enamel is sufficient insulation to handle the filament voltage.

When considerable power is available for driving the grounded-grid stage, the matching between driver stage and the amplifier is not too important. However, when the driving power is

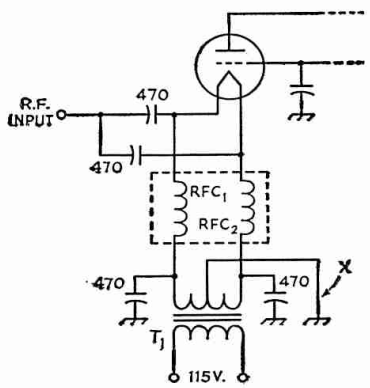


Fig. 11-12—When filament-type tubes are used in a grounded-grid circuit, it is necessary to use filament chokes to keep the filament above r.f. ground. In the portion of a typical circuit shown here, the filament chokes. RFC_1 and RFC_2, can be a manufactured unit (e.g., B&W FC15 or FC30) or homemade as described in the text. Total plate and grid current can be read on a milliammeter inserted at x.

marginal or when the driver and amplifier are to be connected by a long length of coaxial cable, a pi network matching circuit can be used in the input of the grounded-grid amplifier. The input impedance of a grounded-grid amplifier is in the range of 100 to 400 ohms, depending upon the tube or tubes and their operating conditions. When data for grounded-grid operation is available (as for two tubes in Table 11-I), the input impedance can be computed from

$$Z_{in} = \frac{(\textit{peak r.f. driving voltage})^2}{2 \times \textit{driving power}}$$

From this and the equations for a pi network, a suitable network can be devised.

Adjustment of Amplifiers

One of the more important features of the linear amplifier is that the ordinary plate and grid meters are at best only a poor indicator of what is going on. As the meters bounce back and forth, even a person who is thoroughly familiar with this kind of amplifier would be hard put to sense whether the input power registered is attributable to (a) overdrive and underload, which yield distortion, splatter, TVI, etc., or (b) underdrive and too-heavy loading, resulting in inefficiency and loss of output.

The simplest and best way to get the whole

story is to make a linearity test; that is, to send through the amplifier a signal whose amplitude varies from zero up to the peak level in a certain known manner and then observe, by means of an oscilloscope, whether this same waveform comes out of the amplifier at maximum ratings.

Test Equipment

Even the simplest type of cathode-ray oscilloscope can be used for linearity tests, so long as it has the regular internal sweep circuit. If this instrument is not already part of the regular station equipment, it might be well to purchase one of the several inexpensive kits now on the market, so that it will be on hand not only to make initial tests but also as a permanent monitor during all operation. Barring a purchase, it is recommended at least that a scope be borrowed to make the line-up checks, whereupon the regular plate and grid meters can serve thereafter to indicate roughly changes in operating conditions.

All linearity tests require that the vertical plates of the scope be supplied with r.f. from the amplifier output. To avoid interaction within the instrument, it is usually best to connect directly to the cathode-ray tube terminals at the back of the cabinet. A pick-up device and its connections to the oscilloscope are shown in Fig. 11-7. Normally, the pick-up loop should be coupled to the dummy load, antenna tuner, or transmission line; i.e., to a point in the system beyond where any tuning adjustments are to be made.

The only other piece of test equipment will be an audio oscillator. Since only one frequency is needed, the simple circuit of Fig. 11-13 works quite well. Some equipment has a circuit similar to this one built right into the exciter audio system.

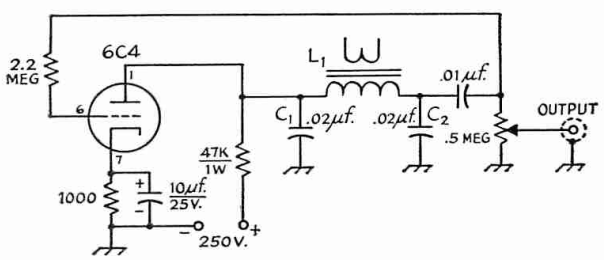


Fig. 11-13—Fixed-frequency audio oscillator having good output waveform. The frequency can be varied by changing the values of C_1 and C_2.

L_1—Small speaker output transformer, secondary not used.

Two-Tone Test

The two-tone test involves sending through the amplifier or the system a pair of r.f. signals of equal amplitude and a thousand cycles or so apart in frequency. The combined envelope of two such signals looks like two sine waves folded on one another. If this waveform comes out of the final, well and good; if not, there is work to do.

There are two commonly used ways to generate the two-tone signal, and the choice of which to use depends on the particular exciter.

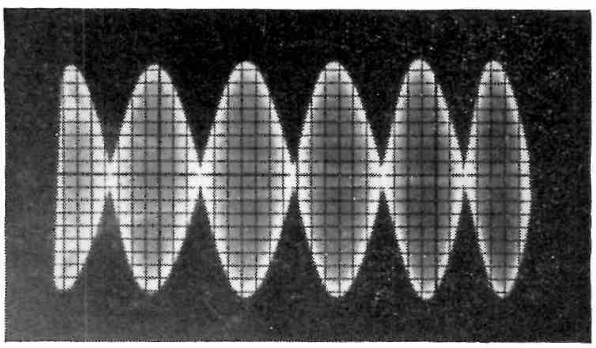
Method A — for Filter or Phasing Exciters:

1) Turn up the carrier insertion until a carrier is obtained at about half the expected output amplitude.

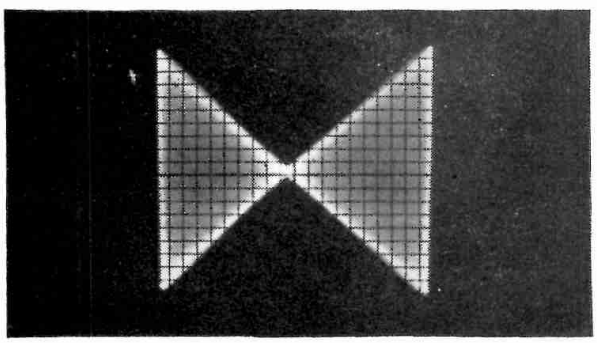
2) Connect an audio oscillator to the microphone input and advance audio gain until (when the carrier and the one sideband are equal) the scope pattern takes on the appearance of full modulation; i.e., the cusps just meet at the center line. See Fig. 11-14, photo No. 1.

3) To change the drive through the system, increase or decrease the carrier and audio settings together, maintaining equality of the two signals.

Method B — for Phasing Exciters:



(1)



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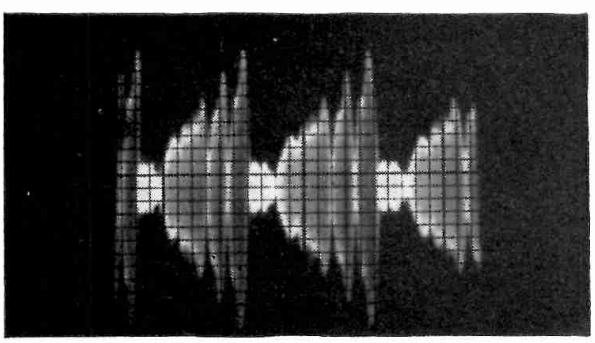


(3)

Fig. 11-14—*Correct Patterns.* 1—Desired two-tone test pattern. 2—Desired double-trapezoid test pattern. 3—Typical voice pattern in a correctly adjusted amplifier, scope set for 30-cycle sweep. Note that peaks are clean and sharp.

1) Disable the audio input to *one* balanced modulator, by removing a tube or by temporarily short-circuiting an audio transformer.

2) Connect the audio oscillator and advance audio gain to get the desired drive. Note that with one balanced modulator cut out, the resultant signal will be double-sideband with no carrier, hence two equal r.f. signals.

Double-Trapezoid Test

When Method B can be used with phasing exciters, it is possible to derive a somewhat more informative pattern by making a connection from the exciter audio system to the horizontal signal input of the oscilloscope and using this audio signal, instead of the regular internal sweep, to cause the horizontal deflection. Those who are familiar with the regular trapezoid test for a.m. transmitters will recognize this set-up as being the same, except that instead of one trapezoid, this test produces two triangles pointing toward each other.

Each individual triangle is subject to the same analysis as the regular trapezoid pattern; i.e., the sloping sides of the pattern should be straight lines for proper operation. Since it is much easier to tell whether a line is straight or not than to judge the correctness of a sine curve, the double trapezoid has the advantage of being somewhat more positive and sensitive to slight departures from linearity than is the regular two-tone pattern.

If the audio can be picked off at the plate of the audio modulator tube that is still working, the input signal need not be a pure sine wave; merely whistling or talking into the microphone should produce the appropriate pattern. If, because of the exciter layout, it is necessary to pick up the audio signal ahead of the phase-shift network, it will then be necessary to use a good sine-wave audio oscillator as before. Also, with the latter set-up, the pattern will probably have a loopy appearance at first, and phase correction

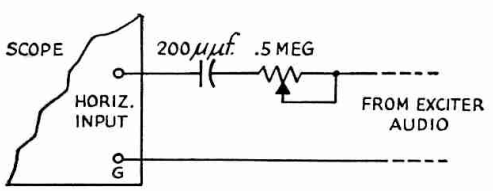


Fig. 11-15—"Phaser" circuit for the oscilloscope.

will be needed to make the figure close up. This can be done either by varying the audio frequency or by putting a phaser in series with the horizontal input to the scope, as shown in Fig. 11-15.

Ratings

Before proceeding with linearity tests, it is well to have in mind the current and power levels to expect. A suppressed-carrier signal is

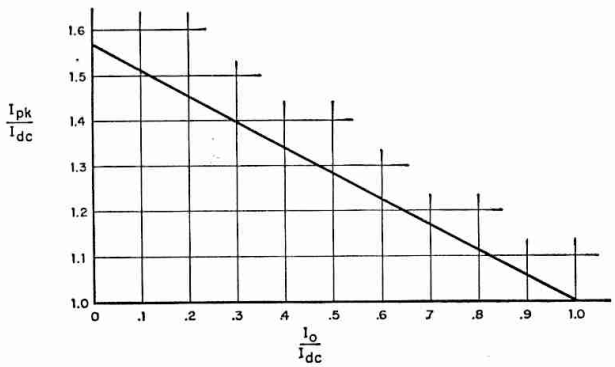


Fig. 11-16—When the two-tone test signal is used for checking the linearity of an amplifier, the peak current is higher than the current indicated by the plate meter. The ratio of these values depends upon the ratio of the idling (no-signal) current to the indicated current. The graph shows the relationship.
I_o = no-signal (idling) current,
I_{dc} = meter reading with two-tone test signal,
I_{pk} = actual peak current.

exactly like an audio signal, except for its frequency, so the audio ratings for any tube are perfectly applicable for linear r.f. service where no carrier is involved. On the other hand, the ratings sometimes shown for Class B r.f. telephony are *not* what is wanted, because they are for conventional a.m. transmission with carrier.

If audio ratings are not given for the desired tube type, it will be safe to assume that the maximum-signal input for Class B or AB_2 service is about 10 per cent less than the key-down Class C c.w. conditions. The input will have to be held somewhat lower in Class AB_1 operation because the average efficiency is lower and, also, the tube can draw only a limited amount of current at zero grid voltage.

The maximum-signal conditions determined from tube data correspond in s.s.b. work to the very peak of the r.f. envelope; when a two-tone test signal (or voice) is used, the plate milliammeter does not indicate the peak plate current. The relationship between peak current and indicated current is variable with voice signals, but with the two-tone test signal applied there is a definite relationship between indicated (d.c.) current and peak current. This relationship is plotted in Fig. 11-16. Knowing the ratio of the idling current to the plate current with the two-tone test signal, I_O/I_{DC}, one can find the factor that can be applied to give the peak current. For example, an amplifier draws 50 ma. with no signal and 250 ma. (before flattening) with the two-tone test signal. $I_O/I_{DC} = 0.2$, and $I_{PK}/I_{DC} = 1.45$, from Fig. 11-16. Thus $I_{PK} = 1.45 \times 250 = 363$ ma.

Should the resulting peak input (0.363 × plate voltage) be different than the design value for the particular amplifier tube, the drive and loading adjustments can be changed in the proper directions (always adjusting the loading so that the peaks of the envelope are on the verge of flattening) and the proper value reached.

Using the Linearity Tests

The photos (Figs. 11-14, 11-17 and 11-18) have been taken to show many of the typical patterns that may be encountered with either of the test arrangements described previously. They are classified separately as to those representing correct conditions (Fig. 11-14), faulty operation of the r.f. amplifier (Fig. 11-17), and various other patterns that look irregular but which really represent a peculiarity in the test set-up or the exciter but not in the final (Fig. 11-18).

Aside from the problem of parasitics, which may or may not be a difficult one, it should be possible without much difficulty to achieve the correct linearity pattern by taking action as indicated by the captions accompanying the photos. It can then be assumed that the amplifier is not contributing any distortion to the signal so long as the peak power level indicated by the test is not exceeded. It is entirely possible, however, that good linearity will be obtained only by holding the power down to a level considerably below what is expected, or conversely that there will be signs of excessive plate dissipation at a level that the tubes should handle quite easily. In such cases, some attention should be given to the plate loading, as discussed below.

The several patterns of Fig. 11-19 show how loading affects the output and efficiency of a linear amplifier. In the first two, loading is relatively light and limiting takes place in the final plate circuit. Reserve power is still available in the driver, evidenced by the fact that heavier loading on the final allows the peak output to increase up to the optimum level of the third pattern. With still heavier loading the output ceases to increase but in fact drops somewhat; even though the input power goes up all the time, the efficiency goes down rapidly. In the last two patterns, the driver is the limiting element in the system, and the extra power-handling *capability* of the final, due to heavier loading, is wasted by inability of the driver to do it justice.

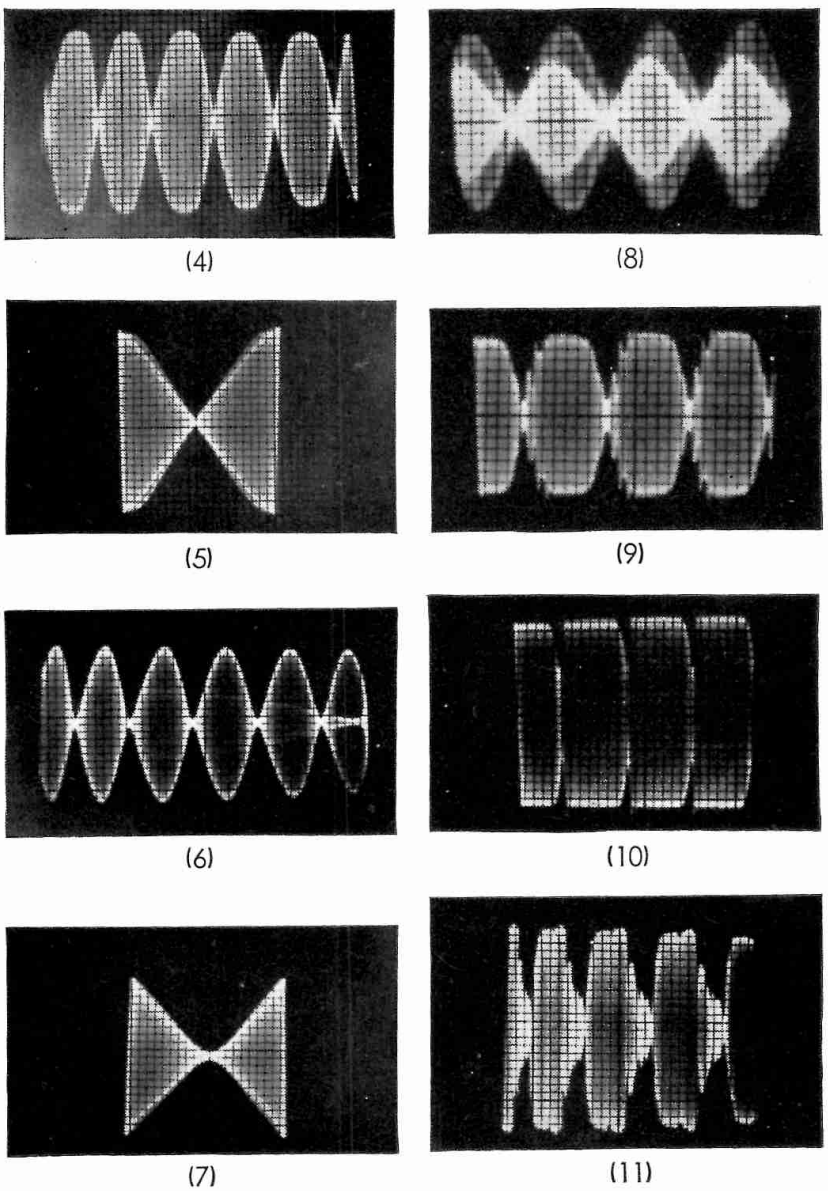


Fig. 11-17—*Improper Amplifier Operation.* 4—Overdrive, indicated by flattening of peaks. 5—Same as 4, double-trapezoid test. 6—Too much bias, causing crossover to become pinched together rather than cutting straight across center line. 7—Same as 6, double-trapezoid test. 8—Two-tone test with v.h.f. parasitics. Note fuzzy halo or fringe. In milder cases the fuzziness will appear just at the peaks. 9—Two-tone test with fundamental frequency parasitics, accompanied by overdrive. 10—Severe overdrive and parasitics. 11—Voice pattern showing flattening of peaks due to overdrive. When flattening is apparent on the voice pattern, the case is a severe one.

1) For good efficiency, the final itself must be the limiting element in the power-handling capability of the system.

2) If the final is not being driven to its limit, it should be loaded less heavily until such is the case.

3) If the power level obtained above is less than should be expected, more driving power is needed.

There are several ways to tell whether or not the final is being driven to its limit. One way is to advance the drive until peak limiting is apparent in the output, then move the oscilloscope coupling link over to the driver plate tank and see whether or not the same limiting appears there. Another way is to decrease or increase the final loading slightly and note whether the limiting output level increases or decreases correspondingly. If it does not, the final is not controlling the system. Still another but similar method is to detune the final slightly while limiting is apparent, and if proper drive conditions prevail the pattern will improve when the amplifier plate is detuned.

The intermediate and driver stages will follow the same laws, except that what is called "loading" on a final is often referred to as "impedance matching" when going between tubes. More

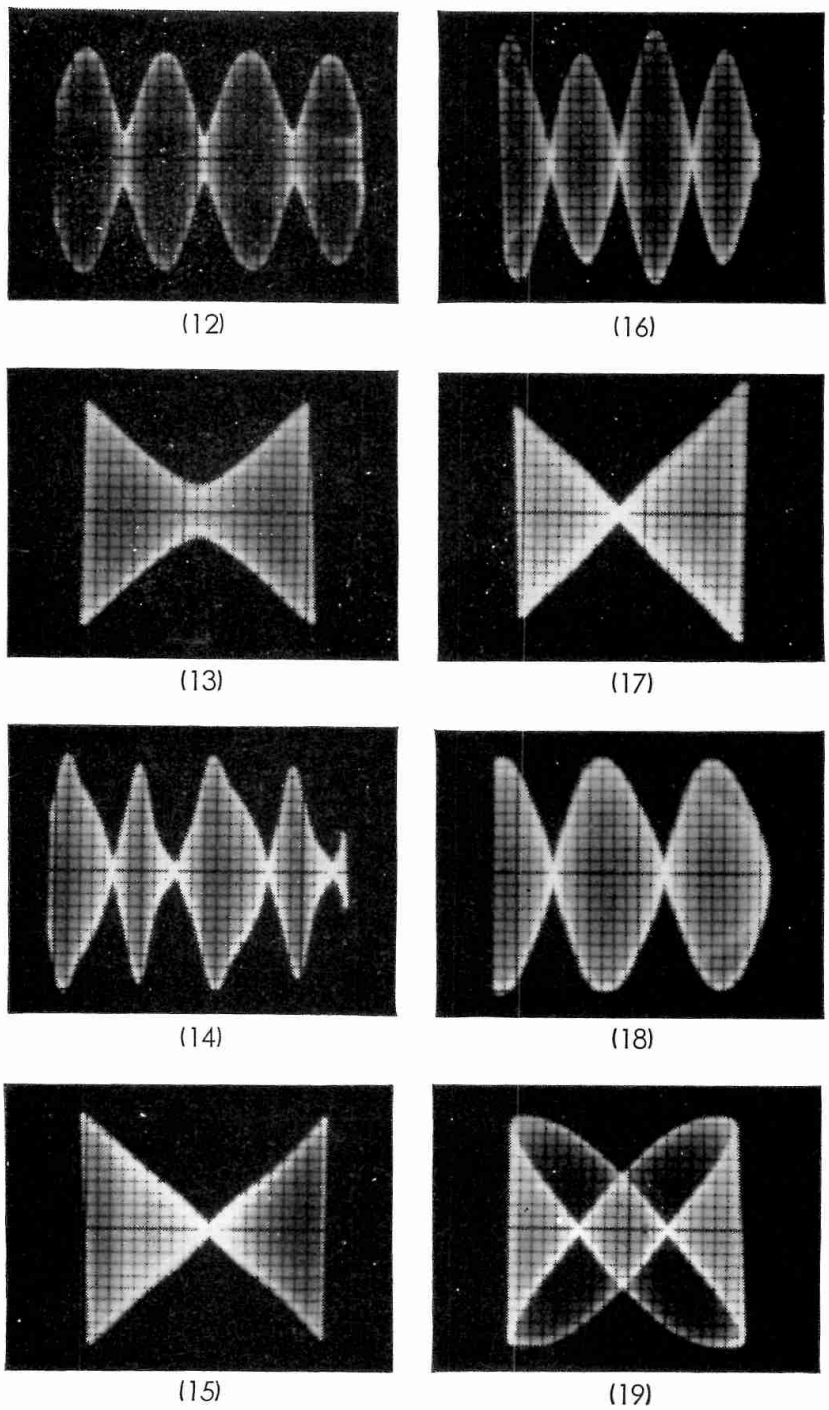


Fig. 11-18—*Improper Test Set-up* 12—Two r.f. signals unequal. in Method A, caused by improper settings of either carrier or audio control. Method B, either carrier leakage through disabled modulator or unequal sidebands due to selective action of some high-Q circuit off resonance. 13—Same as 12, double-trapezoid test (Method B). 14—Distorted audio. A clue to this defect is that successive waves are not identical. 15—Same distortion as 14, but switched to double trapezoid test pattern. Note that correct pattern prevails regardless of poor audio signal. 16—Carrier leakage through working modulator (Method B only). 17—Same as 16, double trapezoid. 18—(Note tilt to left.) Caused by incomplete suppression of unwanted sideband (Method A) or by r.f. leakage into horizontal circuits of scope. 19—Double trapezoid with audio phase shift in test set-up.

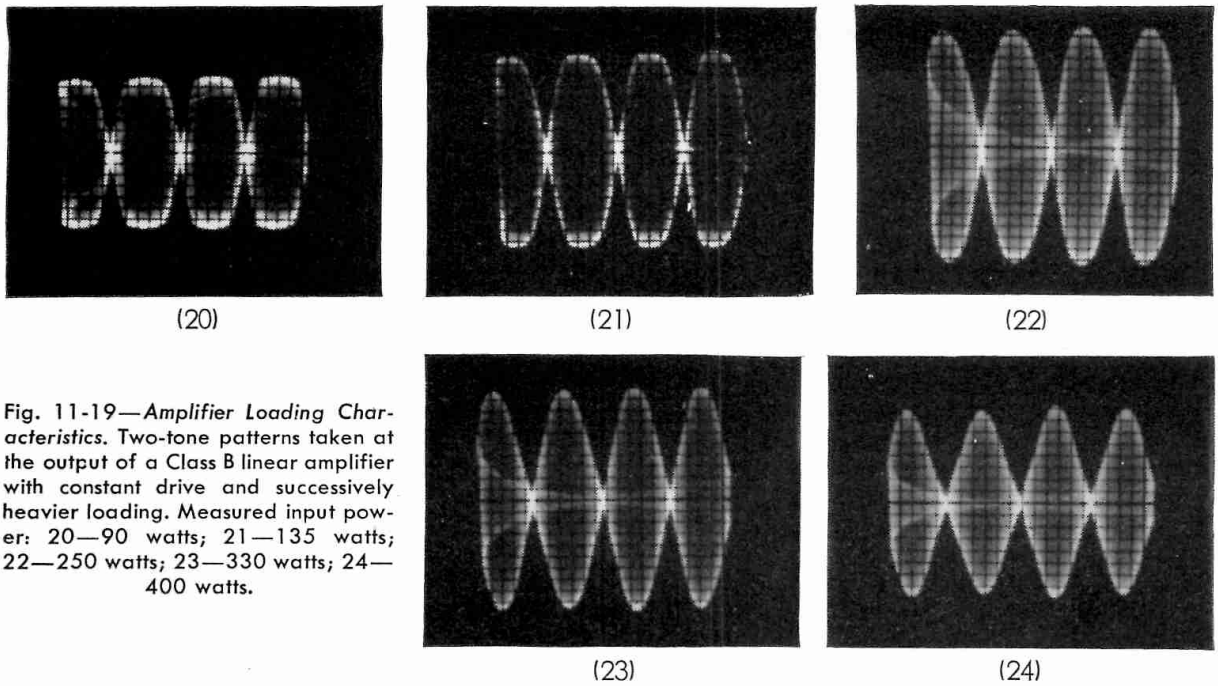


Fig. 11-19—*Amplifier Loading Characteristics.* Two-tone patterns taken at the output of a Class B linear amplifier with constant drive and successively heavier loading. Measured input power: 20—90 watts; 21—135 watts; 22—250 watts; 23—330 watts; 24—400 watts.

often than not, an apparent lack of power transfer from a driver to its succeeding stage is due to a poor match. In Class AB_2 or B service, a step-down type of coupling is required between power stages, and a person accustomed to the conventional plate-to-grid coupling capacitor technique will be surprised to find how effective it is to tap the driven stage down on its tank — or otherwise to decouple the system. For example, an 807 driving a pair of 811s requires a voltage step-down of about 3 or 4 to 1 from plate to each grid.

Dummy Load

For the sake of everyone concerned, linearity tests should be kept off the air as much as possible. They make quite a racket and spurious signals are plentiful in earlier stages of misadjustment. Ordinary lamp bulbs make a fine dummy load so long as it is recognized that their impedance is not exactly the same as the antenna and that this impedance changes somewhat as the bulbs light up. These factors can be taken into account by making careful note of plate and grid currents after the transmitter has been adjusted and is operating with a linearity test signal at maximum linear output into the lamp load. Then, having reconnected the regular antenna, the same loading conditions for the final will be reproduced by adjusting its tuning and loading until the identical combination of plate and grid currents can be obtained. This process will require only a few moments of on-the-air operation.

When the final on-the-air checks are made, it will be convenient to make a few reference marks on the oscilloscope screen to indicate the peak height of the pattern. The scope will then serve as a permanent output monitor for all operations. For best results the sweep should be set for about 30 cycles, in which case the voice patterns will stand out clearly and can easily be kept just within the reference lines. Incidentally, the pattern is really fascinating to watch.

Don't be a "meter bender." Input power isn't everything. If you have to cut your input in half to avoid overload, the fellow at the other end will hardly notice the difference in level. At the same time, your neighbors, both those on the ham band and those next door trying to watch TV, will appreciate the difference right away.

Frequency Conversion

The preferred s.s.b. transmitter is probably one that generates the s.s.b. signal at some suitable frequency and then heterodynes the signal into the desired amateur bands, although a few designs exist that generate the s.s.b. signal at the operating frequency and consequently eliminate the need for heterodyning. When the heterodyning is done at low level (involving an s.s.b. signal of not more than a few volts), standard receiving techniques are satisfactory. The converter tubes operated at manufacturer's ratings leave little to be desired.

When high-level heterodyning is required, as when an exciter delivering from 5 to 20 watts on a single band is available and multiband operation is desired, a high-level converter is used. Since the efficiency of a converter is only about one-fourth that of the same tube or tubes used in Class AB_2, using a converter stage as the output stage is not very economical, and the high-level converter is generally used to drive the output stage.

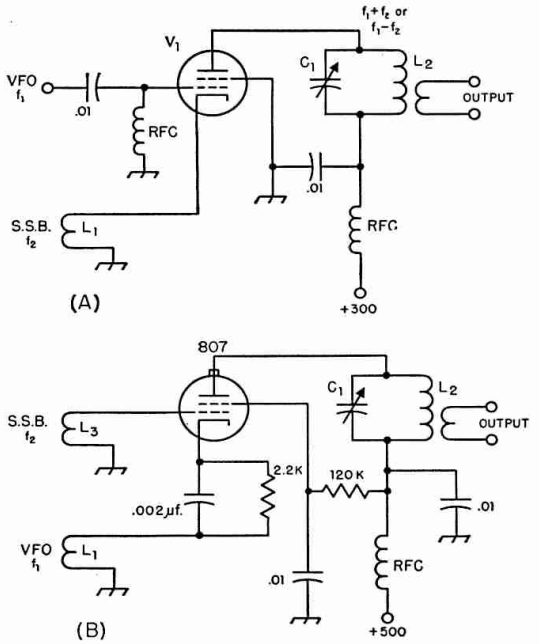


Fig. 11-20—Two examples of "high-level" mixer circuits. The circuit at A has been used with 6V6, 6L6, 6AQ5 and 6Y6 type tubes. With 300 volts on the plate the idling current is about 15 ma., kicking as high as 30 ma. with the s.s.b. signal.

The circuit in B operates with a positive screen voltage and some cathode bias, and is capable of somewhat more output than the circuit shown in A.

In either case the output circuit, C_1L_2, is tuned to the sum or difference frequency of the oscillator and s.s.b. signal. Coupling coils L_1 and L_3 will usually be three or four turns coupled to their respective driving sources.

Reference to tube manuals will disclose no information of the operation of small transmitting tubes as mixers. However, it has been found that most of the tetrodes in the 15- to 35-watt plate-dissipation class make acceptable mixers, and tubes like the 6V6, 6L6, 807 and 6146 have been used successfully. The usual procedure is to feed one of the signals (oscillator or s.s.b.) to the control grid and the other to the cathode or screen grid. Typical circuits are shown in Fig. 11–20.

(Suggestions for converting to and operating in the 50- and 144-Mc. bands can be found in Tilton, "Single-Sideband Ideas for the V.H.F. Man," *QST*, May, 1957.)

VOICE-CONTROLLED BREAK-IN

Although it is possible for two s.s.b. stations operating on widely different frequencies to work "duplex" if the carrier suppression is great enough (inadequate carrier suppression would be a violation of the FCC rules), most s.s.b. operators prefer to use voice-controlled break-in and operate on the same frequency. This overcomes any possibility of violating the FCC rules and permits "round table" operation.

Many various sytems of voice-controlled break-in are in use, but they are all basically the same. Some of the audio from the speech amplifier is amplified and rectified, and the resultant d.c. signal is used to key an oscillator and one or more stages in the s.s.b. transmitter and "blank" the receiver at the time that the transmitter is on. Thus the transmitter is on at any and all times that the operator is speaking but is off during the intervals between sentences. The voice-control circuit must have a small amount of "hold" built into it, so that it will hold in between words, but it should be made to turn on rapidly at the slightest voice signal coming through the speech amplifier. Both tube and relay keyers have been used with good success. Some voice-control systems require the use of headphones by the operator, but a loudspeaker can be used with the proper circuit. (See Nowak, "Voice-Controlled Break-In . . . and a Loudspeaker," *QST*, May, 1951, and Hunter, "Simplified Voice Control with a Loudspeaker," *QST*, October, 1953.)

If an antenna relay is used to switch the antenna from the receiver to the transmitter and back again, it is often possible to operate the output linear amplifier stage with some idling current and experience no difficulty with the "diode noise" generated by the amplifier plate current. However, when the receiver, transmitter and antenna are always connected together, as when an electronic transmit-receive switch is used (see Chapter Eight), weak signals will not be heard through the diode noise of the transmitter. To overcome this difficulty, the idling current of the amplifier must be reduced to zero during listening periods. This can be accomplished through the use of the circuit in Fig. 11-21. Here

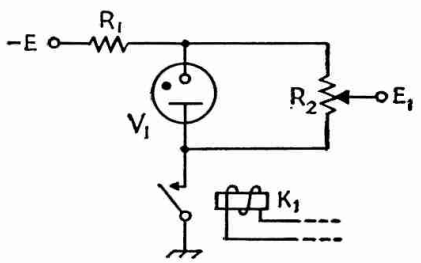


Fig. 11-21—Bias-switching circuit for use with a Class AB_1 linear amplifier and an electronic t.r. switch.

R_1—4700 ohms, 1 watt.
R_2—100,000 ohms, 2 watts.
K_1—VOX relay or relay controlled by VOX circuit.
V_1—0A2 or 0B2, depending upon amplifier requirements.

K_1 is a relay controlled by the voice-controlled break-in circuit. When the relay is closed, the operating bias E_1 for the linear amplifier is determined by the setting of the arm on R_2. When the relay is open, the grid bias jumps to the value E, which should be high enough to cut off the amplifier stage. The voltage regulator tube should be one with a nominal voltage drop in excess of the normal bias for the amplifier tube, and the negative supply voltage E should be at least 25 per cent higher than the ignition potential of the VR tube. The circuit in Fig. 11-21 is applicable to Class AB_1 amplifiers; it cannot be used when grid current is drawn during operation.

Restriction of Audio Range

In either type of s.s.b. generator, it is good practice to restrict the frequency range of the audio amplifier. In the filter-type exciter, reducing the response below 300 or 400 cycles makes it easier for the filter to eliminate the unwanted side frequencies below this range. In the phasing-type exciter, restricting the range of the audio amplifier to the frequencies at which the network gives its best performance (usually about 300 to 3000 cycles) reduces the possibility of generating unwanted side frequencies outside this range. High-frequency audio cut-off is not as important in the filter-type exciter because the filter takes care of the higher frequencies.

When a restricted audio range is used, it is a good idea to make a number of checks on the system, in an effort to obtain the best compromise between naturalness and intelligibility. Voice characteristics differ from operator to operator, and it is sometimes preferable to accentuate the "highs" slightly to give better intelligibility. No standards can be given here — it is a subject for experimentation and checking under varied conditions.

The simplest means for reducing the low-frequency response in the audio amplifier is to reduce the values of the coupling capacitors. High-frequency response can be reduced by adding capacitance across grid resistors. More elaborate means require the use of filters using inductance and capacitance combinations.

Receiving Suppressed-Carrier Signals

The reception of suppressed-carrier signals requires that the carrier be accurately reinserted at the receiver. In addition, the reception of a double-sideband suppressed-carrier signal requires that one sideband be filterd off in the receiver before demodulation or that a special type of converter be used. Because little or no carrier is transmitted, the usual a.v.c. in the receiver has nothing that indicates the average signal level, and this fact requires either manual variation of the r.f. gain control or the use of a special a.v.c. system. (As, for example, Luick, "Improved A.V.C. for Sideband and C.W.," *QST*, October, 1957.)

A suppressed-carrier 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. pass band.) 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 beat oscillator with this weak carrier. It will be noticed that with incorrect tuning of an s.s.b. signal, 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. It may require a readjustment of your tuning habits to tune the receiver slowly enough during the first few trials.

Once the proper setting of the b.f.o. has been established by the procedure above, all further tuning should be done with the main tuning control. However, it is not unlikely that s.s.b. stations will be encountered that are transmitting the other sideband, and to receive them will require shifting the b.f.o. setting to the other side of the receiver i.f. passband. The initial tuning procedure is exactly the same as outlined above, except that you will end up with a considerably different b.f.o. setting. The two b.f.o. settings should be noted for further reference, and all tuning of s.s.b. signals can then be done with the main tuning dial. With experience, it becomes a simple matter to determine which way to tune to make the signal sound lower- or higher-pitched if the receiver (or transmitter) drifts off.

When a double sideband suppressed-carrier signal is received, sufficient selectivity will be required in the receiver to eliminate one sideband and convert the signal into a single-sideband signal before detection, where it can be received by the method outlined above. Receiver bandwidths of 3 kc. or less will be required for this purpose, or the use of a "Signal Slicer," a selectivity device that uses the phasing principle. (See *GE Ham News*, Vol. 6, No. 4, July, 1951.)

Newcomers to single sideband often wonder if there is any device that can be added to a receiver that will make the tuning of sideband signals less critical. At the present time there is no device that will "lock in" automatically. However, if the receiver is lacking in selectivity, an apparent improvement can be obtained by using an adapter that adds selectivity to the receiving system. No improvement in ease of tuning will be noticed on good sideband signals (good suppression of unwanted sideband), but fair or mediocre signals will be easier to tune. The reason is that the adapter makes a better sideband signal out of the incoming signal by removing the vestiges of the unwanted sideband, and a good sideband signal will tune easier than a fair one. The sideband adapters also usually have detectors designed for best detection of sideband signals, a point that was overlooked in some of the older receivers. Good detectors for sideband signals include diodes with *sufficient* b.f.o. injection (5 to 10 times peak signal) and "product detectors" (see Chapter Five). Either detector is capable of low distortion output if the input is held down.

WHICH SIDEBAND?

It is sometimes confusing to remember how to identify, from the way your receiver tunes, which sideband the other station is using. This is especially awkward with those receivers in which the high-frequency oscillator is on the high-frequency side of the signal in some ranges and on the low side in others. With any receiver having sufficient selectivity to give a stronger signal on one side than on the other of zero beat, when the b.f.o. is offset to one side of the i.f., the chart below will help you identify which sideband is being transmitted. Set the b.f.o. to the side of the i.f. that you find is right for receiving single-sideband signals on a particular band. Then with the main dial tune through a steady carrier (or the signal from your 100-kc. standard). Note which side of zero beat gives little or no signal.

If tuning through a steady carrier gives little or no signal on the	
High Frequency	*Low Frequency*
side of zero beat, and then if tuning the receiver to a *lower* frequency makes the voice of a single-sideband signal sound *lower* pitched, he is using the	
Lower	*Upper*
sideband.	

Specialized Communication Systems

Frequency and Phase Modulation

It is possible to convey intelligence by modulating any property of a carrier, including its frequency and phase. When the frequency of the carrier is varied in accordance with the variations in a modulating signal, the result is **frequency modulation (f.m.).** Similarly, varying the phase of the carrier current is called **phase modulation (p.m.).**

Frequency and phase modulation are not independent, since the frequency cannot be varied without also varying the phase, and vice versa. The difference is largely a matter of definition.

The effectiveness of f.m. and p.m. for communication purposes depends almost entirely on the receiving methods. If the receiver will respond to frequency and phase changes but is insensitive to amplitude changes, it will discriminate against most forms of noise, particularly impulse noise such as is set up by ignition systems and other sparking devices. Special methods of detection are required to accomplish this result.

Modulation methods for f.m. and p.m. are simple and require practically no audio power. There is also the advantage that, since there is no amplitude variation in the signal, interference to broadcast reception resulting from rectification of the transmitted signal in the audio circuits of the BC receiver is substantially eliminated. These two points represent the principal reasons for the use of f.m. and p.m. in amateur work.

Frequency Modulation

Fig. 12-1 is a representation of frequency modulation. 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 **(frequency deviation)** is proportional to the instantaneous amplitude of the modulating signal, so the 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.

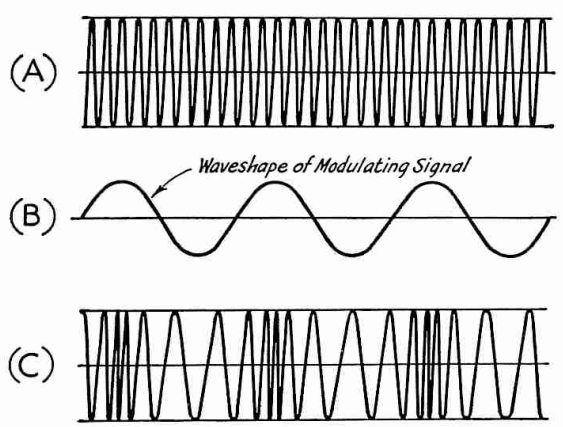


Fig. 12-1—Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

As shown by the drawing, the amplitude of the signal does not change during modulation.

Phase Modulation

If the phase of the current in a circuit is changed there is an instantaneous frequency change during the time that the phase is being shifted. The amount of frequency change, or deviation, depends on how rapidly the phase shift is accomplished. It is also dependent upon the total amount of the phase shift. In a properly operating p.m. system the amount of phase shift is proportional to the instantaneous amplitude of the modulating signal. The rapidity of the phase shift is directly proportional to the frequency of the modulating signal. Consequently, the frequency deviation in p.m. is proportional to both the amplitude and frequency of the modulating signal. The latter represents the outstanding difference between f.m. and p.m., since in f.m. the frequency deviation is proportional only to the amplitude of the modulating signal.

Modulation Depth

Percentage of modulation in f.m. and p.m. has to be defined differently than for a.m. 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 frequency deviation is greater than the receiver can accept, the receiver distorts the signal. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25 per cent modulation.

In amateur work "narrow-band" f.m. or p.m. (frequently abbreviated n.f.m.) is defined as having the same channel width as a properly modulated a.m. signal. That is, the effective channel width does not exceed twice the highest

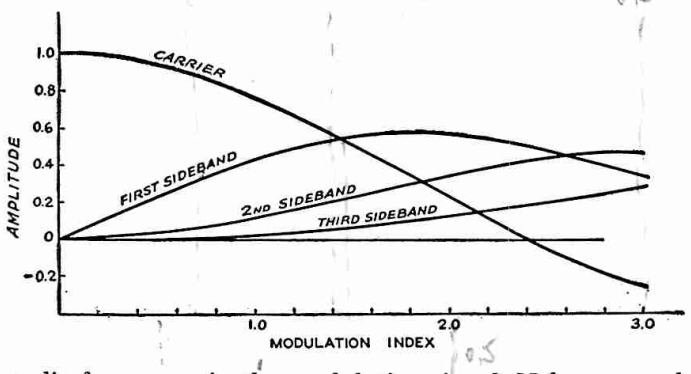


Fig. 12-2—How the amplitude of the pairs of sidebands varies with the modulation index in an f.m. or p.m. signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

audio frequency in the modulating signal. N.f.m. transmissions based on an upper audio limit of 3000 cycles therefore should occupy a channel not significantly wider than 6 kc.

F.M. and P.M. Sidebands

The sidebands set up by f.m. and p.m. differ from those resulting from a.m. in that they occur at integral multiples of the modulating frequency on either side of the carrier rather than, as in a.m., consisting of a single set of side frequencies for each modulating frequency. An f.m. or p.m. signal therefore inherently occupies a wider channel than a.m.

The number of "extra" sidebands that occur in f.m. and p.m. depends on the relationship between the modulating frequency and the 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,

$$Modulation\ index = \frac{Carrier\ frequency\ deviation}{Modulating\ frequency}$$

Example: The maximum frequency deviation in an f.m. 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.

In p.m. the modulation index is constant regardless of the modulating frequency; in f.m. it varies with the modulating frequency, as shown in the above example. In an f.m. system the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used is called the **deviation ratio.**

Fig. 12-2 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 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.

Note that, as shown by Fig. 12-2, 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. It then becomes "negative" at a higher index, meaning that its phase is reversed as compared to the phase without modulation. In f.m. and p.m. the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index.

Frequency Multiplication

Since there is no change in amplitude with modulation, an f.m. or p.m. signal can be amplified without distortion by an ordinary Class C amplifier. 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.

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, if modulation is applied on 3.5 Mc. and the final output is on 28 Mc. the total frequency multiplication is 8 times, so 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.

Narrow-Band F.M. and P.M.

"Narrow-band" f.m. or p.m., the only type that is authorized by FCC for use on the lower frequencies where the phone bands are crowded, is defined as f.m. or p.m. that *does not* occupy a wider channel than an a.m. signal having the same audio modulating frequencies.

If the modulation index (with single-tone modulation) does not exceed 0.6 or 0.7, 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 a.m. 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 com-

ponent 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 f.m. or p.m. 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 f.m. or p.m. is not as effective as a.m. with the methods of reception used by most amateurs. As shown by Fig. 12-2, 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 a.m. transmitter. So far as effectiveness is concerned, a narrow-band f.m. or p.m. transmitter is about equivalent to a 100 per cent modulated a.m. transmitter operating at one-fourth the carrier power.

Comparison of F.M. and P.M.

Frequency modulation cannot be applied to an amplifier stage, but phase modulation can; p.m. is therefore readily adaptable to transmitters employing oscillators of high stability such as the crystal-controlled type. The amount of phase shift that can be obtained with good linearity is such that the maximum practicable modulation index is about 0.5. Because the phase shift is proportional to the modulating frequency, this index can be used only at the highest frequency present in the modulating signal, assuming that all frequencies will at one time or another have equal amplitudes. Taking 3000 cycles as a suitable upper limit for voice work, and setting the modulation index at 0.5 for 3000 cycles, the frequency response of the speech-amplifier system above 3000 cycles must be sharply attenuated, to prevent sideband splatter. Also, if the "tinny" quality of p.m. as received on an f.m. receiver is to be avoided, the p.m. must be changed to f.m., in which the modulation index decreases in inverse proportion to the modulating frequency. This requires shaping the speech-amplifier frequency-response curve in such a way that the output voltage is inversely proportional to frequency over most of the voice range. When this is done the maximum modulation index can only be used at some relatively low audio frequency, perhaps 300 to 400 cycles in voice transmission, and must decrease in proportion to the increase in frequency. The result is that the maximum linear frequency deviation is only one or two hundred cycles, when p.m. is changed to f.m. To increase the deviation for n.f.m. requires a frequency multiplication of 8 times or more.

It is relatively easy to secure a fairly large frequency deviation when a self-controlled oscillator is frequency-modulated directly. (True frequency modulation of a crystal-controlled oscillator results in only very small deviations and so requires a great deal of frequency multiplication.) The chief problem is to maintain a satisfactory degree of carrier stability, since the greater the inherent stability of the oscillator the more difficult it is to secure a wide frequency swing with linearity.

Methods of Frequency and Phase Modulation

A simple and satisfactory device for producing f.m. in the amateur transmitter 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. 12-3 is a representative circuit. The control grid of the modulator tube, V_2, is connected across the oscillator tank circuit, C_1L_1, through resistor R_1 and blocking capacitor C_2. C_8 represents the input capacitance of the modulator tube. The resistance of R_1 is made large compared to the reactance of C_8, so the r.f. current through R_1C_8 will be practically in phase with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_8 will lag the current by 90 degrees. The r.f. current in the plate circuit of the modulator will be in phase with the grid voltage, and consequently is 90 degrees behind the current through C_8, 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 audio voltage, introduced through a radio-frequency choke, RFC_1, varies the transconductance of the tube and thereby varies the r.f. plate current.

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.

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 R_1 is made smaller in comparison with C_8. It also increases with an increase in L/C ratio in the oscillator tank circuit. However, for highest carrier stability it is desirable to use the largest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation.

A change in *any* of the voltages on the modu-

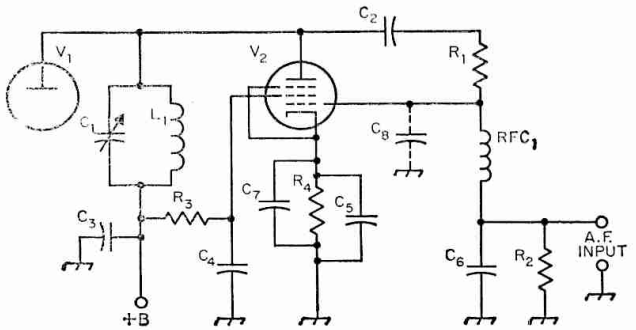


Fig. 12-3—Reactance modulator using a high-transconductance pentode (6BA6, 6CL6, etc.).
C_1—R.f. tank capacitance (see text).
C_2, C_3—0.001-μf. mica.
C_4, C_5, C_6—0.0047-μf. mica.
C_7—10-μf. electrolytic.
C_8—Tube input capacitance.
R_1—47,000 ohms.
R_2—0.47 megohm.
R_3—Screen dropping resistor; select 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_1—2.5-mh. r.f. choke.

lator 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 voltage used (250 volts or less) 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 taken from it and the a.f. voltage required 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 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 f.m. can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for p.m. Hence the modulator circuit of Fig. 12-3 can be used for p.m. if the reactance tube works on an amplifier tank instead of directly on a self-controlled oscillator.

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 phase-shift 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 stage where receiving-type tubes are used.

Reactance modulation of an amplifier stage usually also results in simultaneous amplitude modulation because the modulated stage is detuned from resonance as the phase is shifted. This must be eliminated by feeding the modulated signal through an amplitude limiter or one or more "saturating" stages — that is, amplifiers that are operated Class C and driven hard enough so that variations in the amplitude of the grid excitation produce no appreciable variations in the final output amplitude.

For the same type of reactance modulator, the speech-amplifier gain required is the same for p.m. as for f.m. However, as pointed out earlier, the fact that the actual frequency deviation increases with the modulating audio frequency in p.m. makes it necessary to cut off the frequencies above about 3000 cycles before modulation takes place. If this is not done, unnecessary sidebands will be generated at frequencies considerably away from the carrier.

Checking F.M. and P.M. Transmitters

Accurate checking of the operation of an f.m. or p.m. transmitter requires different methods than the corresponding checks on an a.m. 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 measuring instrument that indicates frequency deviation directly.

However, there is one favorable feature in f.m. or p.m. 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 unneces-

sary 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 f.m. or p.m. transmitter, particularly for narrow-band f.m. or p.m.

The quantities to be checked in an f.m. or p.m. transmitter are the linearity and frequency deviation. Because of the essential difference between f.m. and p.m. the methods of checking differ in detail.

Reactance-Tube F.M.

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. 12-4. The battery should have a

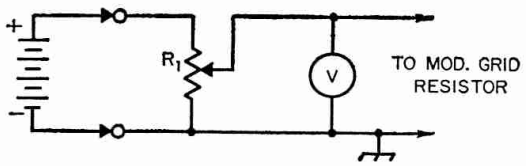


Fig. 12-4—D.c. method of checking frequency deviation of a reactance-tube-modulated oscillator. A 500- or 1000-ohm potentiometer may be used at R_1.

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 (disconnecting 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. 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 several readings have been taken a curve may be plotted to demonstrate the relationship between grid voltage and frequency deviation.

A sample curve is shown in Fig. 12-5. 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

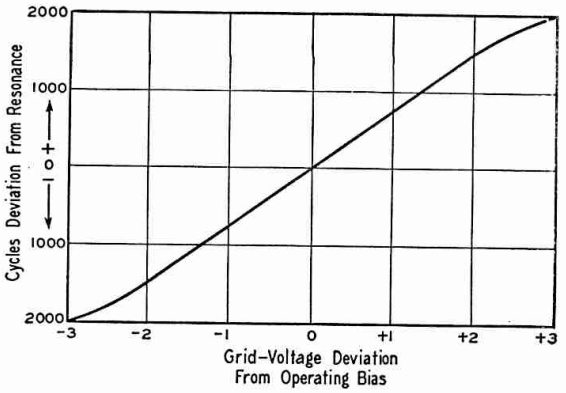


Fig. 12-5—A typical curve of frequency deviation vs. modulator grid voltage.

either side of the center or carrier 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. 12-6. Note its deflection (using the d.c. voltage method as in Fig. 12-4) at the maximum deviation to be used. For narrow-band f.m. the proper deviation is approximately 2000 cycles (this maximum deviation is based on an upper a.f. limit of 3000 cycles and a deviation ratio of 0.7) at the *output* frequency. 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

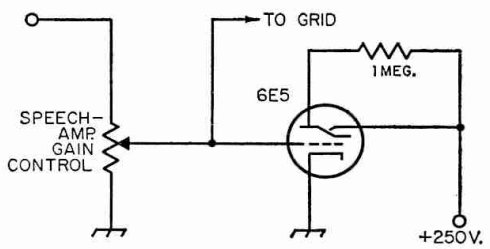


Fig. 12-6—6E5 modulation indicator for f.m. or p.m. modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.

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 example, 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 p.m. the d.c. method of checking just described cannot be used, because the frequency deviation at zero frequency (d.c.) also is zero. For narrow-band p.m. it is necessary to check the actual width of the channel occupied by the transmission. (The same method also can be used to check f.m.) For this purpose it is necessary to have a crystal-filter receiver and

an a.f. oscillator that generates a 3000-cycle sine wave.

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 (through an attenuator, if necessary, to avoid overloading; see chapter on audio amplifiers) and increase the audio gain until there is a small amount of modulation. Tuning the receiver near 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 f.m. (not p.m.) 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 capable of shifting 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 f.m. and p.m., 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 a.m. phone operation.

The output power of the transmitter should be checked for amplitude modulation. It should not change from the unmodulated-carrier value when the transmitter is modulated. If no output indicator is available, 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 f.m. or p.m. is just as much to be avoided as frequency or phase modulation that accompanies a.m. A mixture of a.m. with either of the other two systems results in the generation of spurious sidebands and consequent widening of the channel. If the presence of a.m. is indicated by variation of antenna current with modulation, the cause is almost certain to be nonlinearity in the modulator.

Reception of F.M. and P.M. Signals

Receivers for f.m. and p.m. signals differ from those for a.m. and s.s.b. principally in two features — there is no need for linearity in the amplifier stages preceding detection (in fact, it is advantageous if the amplitude variations in the signal and background noise can be "washed out"), and the detector must be capable of converting the frequency variations in the incoming signal into amplitude variations. These amplitude variations, combined with rectification, produce an audio voltage corresponding to the frequency or phase modulation on the signal.

Frequency- or phase-modulated signals can be received after a fashion on any ordinary receiver that has a selectivity curve with sloping sides. As shown in Fig. 12-7A, the receiver is tuned so that the carrier frequency is placed part-way down on one side of the selectivity curve so that the amplitude is less than the maximum that would be possible with normal tuning. When the frequency of the signal varies with modulation it swings between some such limits as are indicated in Fig. 12-7A, resulting in an amplitude-modulated output varying between *X* and *Y*. After this f.m.-to-a.m. conversion the signal goes to a conventional detector (usually a diode) and is rectified in the same way as an a.m. signal.

With most receivers, particularly those having steep-sided selectivity curves, the method is not very satisfactory because the distortion is quite severe unless the frequency deviation is small, because the relationship between frequency deviation and output amplitude is linear over only a small part of the selectivity curve.

A detector designed expressly for f.m. or p.m. will have a characteristic similar to that shown in Fig. 12-7B. The output is zero when the unmodulated carrier is tuned to the center, *0*, of

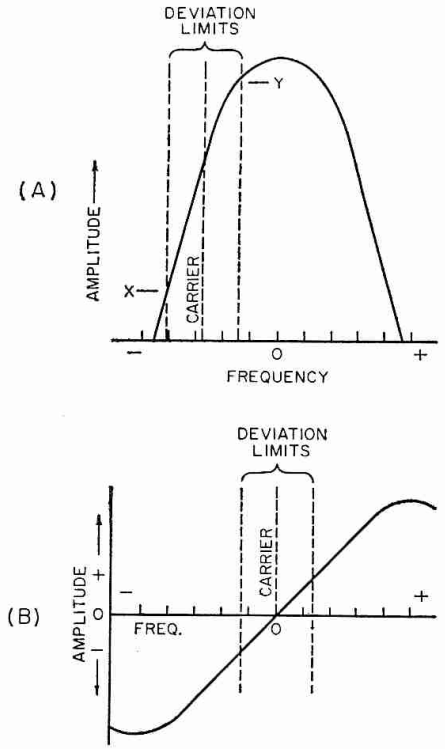


Fig. 12-7—F.m. or p.m. detection characteristics. A—"Slope detection," using the sloping side of the receiver's selectivity curve to convert f.m. or p.m. to a.m. for subsequent rectification. B—Typical discriminator characteristic. The straight portion of this curve between the two peaks is the useful region. The peaks should always lie outside the pass band of the receiver's selectivity curve.

the characteristic. When the frequency swings higher, the rectified output amplitude increases in the positive direction (as chosen in this example), and when the frequency swings lower the output amplitude increases in the negative direction. Over the range in which the characteristic is a straight line the conversion from f.m. to a.m. is linear and there is no distortion. One type of detector that operates in this way is the **frequency discriminator,** which combines the f.m.-to-a.m. conversion with rectification to give an audio-frequency output from the frequency-modulated r.f. signal.

Limiter and Discriminator

A practical discriminator circuit is shown in Fig. 12-8. The f.m.-to-a.m. conversion takes place in transformer T_1, which operates at the intermediate frequency of a superheterodyne receiver. The voltage induced in the transformer secondary, S, is 90 degrees out of phase with the primary current. The primary voltage is introduced at the center tap on the secondary through C_1 and combines with the secondary voltages on each side of the center tap in such a way that the resultant voltage on one side of the secondary leads the primary voltage and the voltage on the other side lags by the same phase angle, when the circuits are resonated to the unmodulated carrier frequency. When rectified, these two voltages are equal and of opposite polarity. If the frequency changes, there is a shift in the relative phase of the voltage components that results in an increase in output amplitude on one side of the secondary and a corresponding decrease in amplitude on the other side. Thus the voltage applied to one diode of V_2 increases while the voltage applied to the other diode decreases. The difference between these two voltages, after rectification, is the audio-frequency output of the detector.

The output amplitude of a simple discriminator depends on the amplitude of the input r.f. signal, which is undesirable because the noise-reducing benefits of f.m. are not secured if the receiving system is sensitive to amplitude variations. A discriminator is always preceded by some form of amplitude limiting, therefore. The conventional type of limiter also is shown in Fig. 12-8. It is simply a pentode i.f. amplifier, V_1, with its operating conditions chosen so that it "saturates" on a relatively small signal voltage. The limiting action is aided by grid rectification, with grid-leak

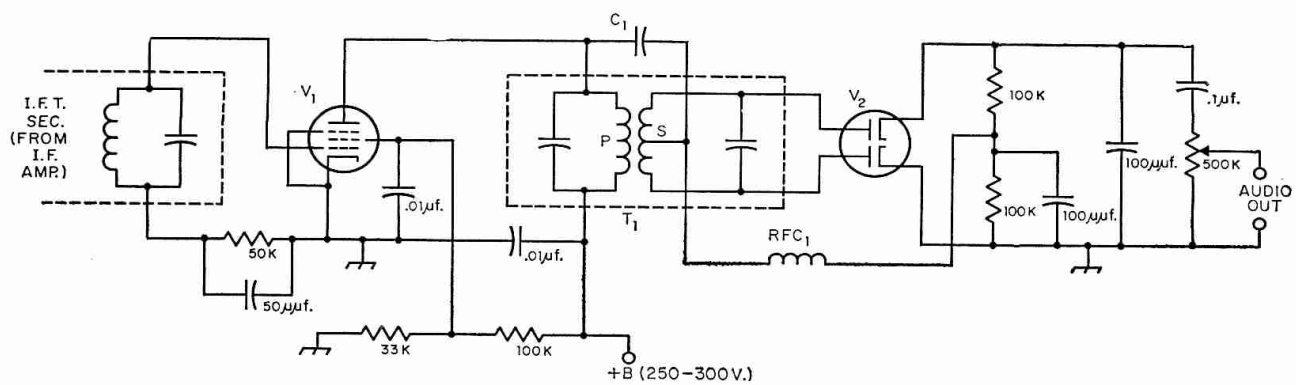


Fig. 12-8—Limiter-discriminator circuit. This type of circuit is frequently used at 455 kc. in the form of an "adapter" for communications receivers, for reception of narrow-band f.m. signals.

C_1—App. 100 μμf. for 455-kc. i.f.; 50 μμf. for higher frequencies.

T_1—Discriminator transformer for intermediate frequency used. Push-pull diode transformer may be substituted.

RFC_1—10 mh. r.f. choke for 455-kc. i.f.; 2.5 mh. satisfactory for frequencies above 3 Mc.

V_1—6AU6 or equivalent.

V_2—6AL5 or equivalent.

bias developed in the 50,000-ohm resistor in the grid circuit. Another contributing factor is low screen voltage, the screen voltage-divider constants being chosen to result in about 50 volts on the screen.

Receiver Tuning with an F.M. Detector

In tuning a signal with a receiver having a discriminator or other type of f.m. detector the tuning controls should be adjusted to center the carrier on the detector characteristic. At this point the noise suppression is most marked, so the proper setting is easily recognized. An amplitude-modulated signal tuned at the same point will have its modulation "washed off" if the signal is completely limited in amplitude and the discriminator alignment is symmetrical. With either f.m. or a.m. signals, there will be a distorted audio-frequency output if the receiver is tuned "off center."

Radioteletype

Radioteletype (abbreviated **RTTY**) is a form of telegraphic communication employing typewriter-like machines for 1) generating a coded set of electrical impulses when a typewriter key corresponding to the desired letter or symbol is pressed, and 2) converting a received set of such impulses into the corresponding printed character. The message to be sent is typed out in much the same way that it would be written on a typewriter, but the printing is done at the distant receiving point. The teletypewriter at the sending point also prints the same material, for checking and reference.

The machines used for RTTY are far too complex mechanically for home construction, and if purchased new would be highly expensive. However, used teletypewriters in good mechanical condition are available at quite reasonable prices. These are machines retired from commercial service but capable of entirely satisfactory operation in amateur work. They may be obtained from a number of sources (latest information on this may be obtained from ARRL, West Hartford, Conn.) on condition that they will be used purely for amateur purposes and will not be resold for commercial use.

Types of Machines

There are two general types of machines, the **page printer** and the **tape printer.** The former prints on a paper roll about the same width as a business letterhead. The latter prints on paper tape, usually gummed on the reverse side so it may be cut to letter-size width and pasted on a sheet of paper in a series of lines. The page printer is the more common type in the equipment available to amateurs.

The operating speed of most machines is such that characters are sent at the rate of about 60 words per minute. Ordinary teletypewriters are of the **start-stop** variety, in which the pulse-forming mechanism (motor driven) is at rest until a typewriter key is depressed. At this time it begins operating, forms the proper pulse sequence, and then comes to rest again before the next key is depressed to form the following character. The receiving mechanism operates in similar fashion, being set into operation by the first pulse of the sequence from the transmitter. Thus, although the actual transmission speed cannot exceed about 60 w.p.m. it can be considerably slower, depending on the typing speed of the operator.

It is also possible to transmit by using perforated tape. This has the advantage that the complete message may be typed out in advance of actual transmission, at any convenient speed; when transmitted, however, it is sent at the machine's normal maximum speed. A special transmitting head and tape perforator are required for this process. A **reperforator** is a device that may be connected to the conventional teletypewriter for punching tape when the machine is operated in the regular way. It may thus be used either for an original message or for "taping" an incoming message for retransmission.

Teletype Code

In the special code used for teletype every character has five "elements" sent in sequence. Each element has two possible states, either "mark" or "space," which are indicated by different types of electrical impulses (i.e., mark might be indicated by a negative voltage and space by a positive voltage). In customary practice each element occupies a time of 22 milliseconds. In addition, there is an initial "start" element (space), also 22 milliseconds long, to set the transmitting and receiving mechanisms in operation, and a terminal "stop" element (mark) 31 milliseconds long, to shut down the operation and ready the machine for the next character.

This sequence is illustrated in Fig. 12-9, which

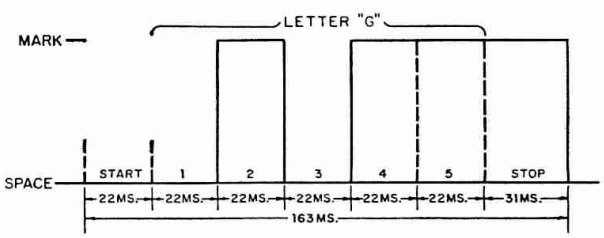


Fig. 12-9—Pulse sequence in the teletype code. Each character begins with a start pulse, always a "space," and ends with a "stop" pulse, always a "mark." The distribution of marks and spaces in the five elements between start and stop determines the particular character transmitted.

shows the letter G with its start and stop elements. The letter code as it would appear on perforated tape is shown in Fig. 12-10, where the black dots indicate marking pulses. Figures and arbitrary signs — punctuation, etc. — use the

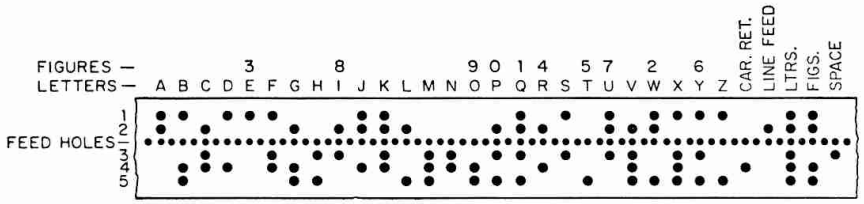


Fig. 12-10—Teletype letter code as it appears on perforated tape. Start and stop elements do not appear on tape. Elements are numbered from top to bottom, and dots indicate marking pulses. Numerals, punctuation signs, and other arbitrary symbols are secured by carriage shift.

There are no lower-case letters on a teletypewriter. Where blanks appear in the above chart in the "FIGS" line, characters may differ on different machines.

same set of code impulses as the alphabet, and are selected by shifting the carriage as in the case of an ordinary typewriter. The carriage shift is accomplished by transmitting either the "LTRS" or "FIGS" code symbol as required. There is also a "carriage return" code character to bring the carriage back to the starting position after the end of the line is reached on a page printer, and a "line feed" character to advance the page to the next line after a line is completed.

Additional System Requirements

To be used in radio communication, the pulses (d.c.) generated by the teletypewriter must be utilized in some way to key a radio transmitter so they may be sent in proper sequence and usable form to a distant point. At the receiving end the incoming signal must be converted into d.c. pulses suitable for operating the printer. These functions, shown in block form in Fig. 12-11, are

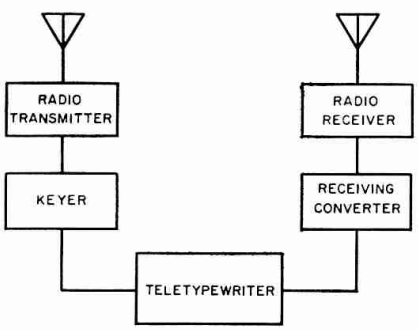


Fig. 12-11—Radioteletype system in block form.

performed by electronic units known respectively as the **keyer** and **receiving converter.**

The radio transmitter and receiver are quite conventional in design. Practically all the special features needed can be incorporated in the keyer and converter, so that any ordinary amateur equipment is suitable for RTTY with little modification.

Transmission Methods

It is quite possible to transmit teletype signals by ordinary "on-off" or "make-break" keying such as is used in regular hand-keyed c.w. transmission. In practice, however, **frequency-shift keying** is preferred because it gives definite pulses on both mark and space, which is an advantage in printer operation. Also, since f.s.k. can be received by methods similar to those used for f.m. reception, there is considerable discrimination against noise, both natural and man-made, distributed uniformly across the receiver's pass band, when the received signal is not too weak. Both factors make for increased reliability in printer operation.

Frequency-Shift Keying

General practice with f.s.k. is to use a frequency shift of 850 cycles per second, although FCC regulations permit the use of any value of frequency shift up to 900 cycles. The smaller values of shift have been shown to have a signal-to-noise-ratio advantage in commercial circuits, and are currently being experimented with by amateurs. At present, however, the major part of amateur RTTY work is done with the 850-cycle shift. This figure also is used in much commercial work. The nominal transmitter frequency is the mark condition and the frequency is shifted 850 cycles (or whatever shift may be chosen) lower for space.

On the v.h.f. bands where A2 transmission is permitted **audio frequency-shift keying (a.f.s.k.)** is generally used. In this case the r.f. carrier is transmitted continuously, the pulses being transmitted by frequency-shifted tone modulation. The audio frequencies used have been more-or-less standardized at 2125 and 2975 cycles per second, the shift being 850 cycles as in the case of straight f.s.k. (These frequencies are the 5th and 7th harmonics, respectively, of 425 cycles, which is half the shift frequency, and thus are convenient for calibration and alignment purposes.) With a.f.s.k. the lower audio frequency is customarily used for mark and the higher for space.

The Receiving Converter

In receiving an f.s.k. teletype signal, the receiver's beat-frequency oscillator is turned on as for ordinary c.w. reception and the receiver tuning is then adjusted so that the mark and space signals produce audio beat tones of 2125 and 2975 cycles. Either frequency can be used for either mark or space, but no matter which may be used at the transmitter, the mark and space frequencies can be reversed at the receiver simply by tuning to the "other side of zero beat." (This cannot be done with a.f.s.k., of course, but the reversal can be accomplished quite simply, if

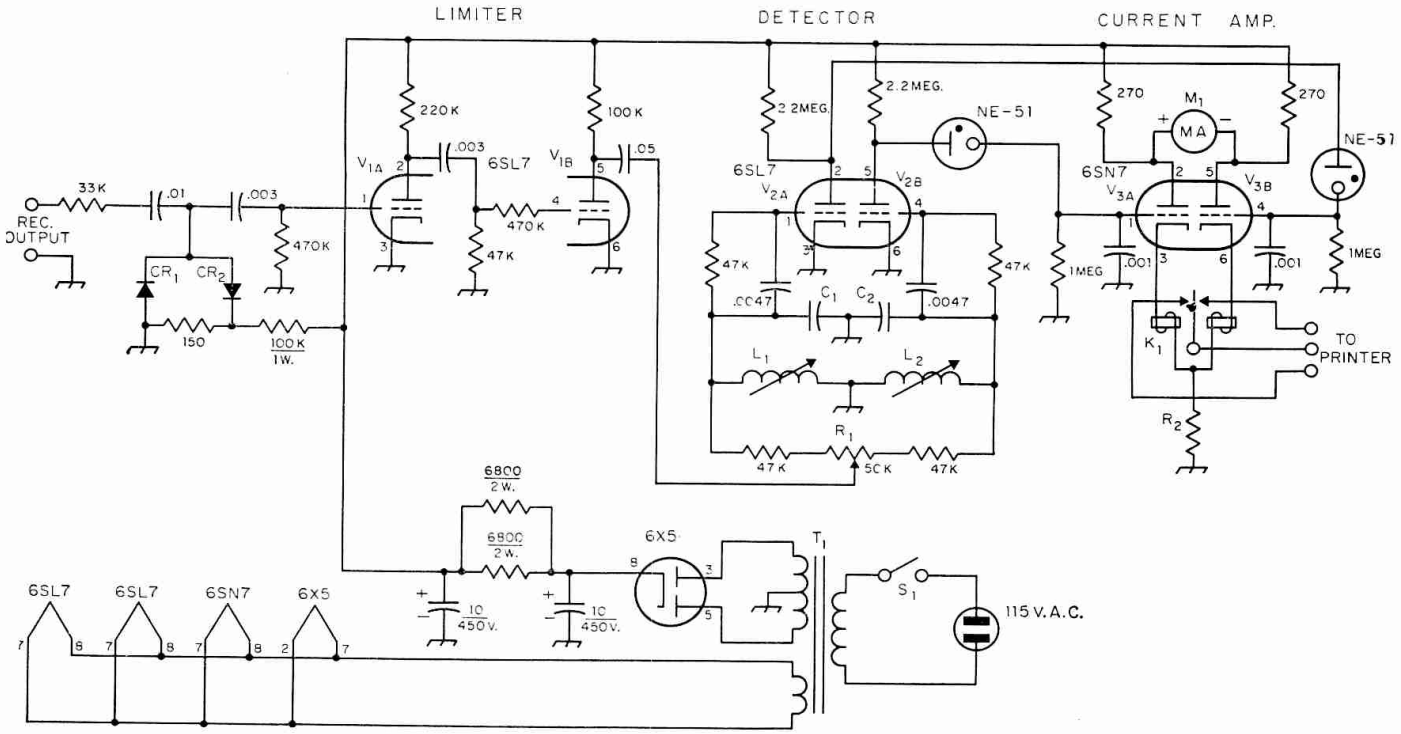


Fig. 12-12—Receiving converter for f.s.k. teletype signals (W2PAT). Unless otherwise indicated, capacitances are in μf. resistances are in ohms, resistors are ½ watt. Capacitors of 0.01 μf. or less may be mica or ceramic; larger values may be paper. Capacitors with polarities indicated are electrolytic.

C_1—0.15-μf. paper.
C_2—0.1-μf. paper.
CR_1, CR_2—1N34 or equivalent.
K_1—Polar relay, to operate on 20 ma.
L_1—36 mh. (TV width control, GE type RLD-019).
L_2—29 mh. (TV width control, GE type RLD-014).
M_1—Zero-center d.c. milliammeter, 20 ma. or more full scale (may be a 100-0-100 microammeter appropriately shunted).
R_1—50,000-ohm volume control, linear taper.
R_3—1000 ohms, 1 watt.
S_1—S.p.s.t. toggle.
T_1—Power transformer, 500 volts c.t., 30 ma; 6.3 volts 3 amp.
V_1, V_2—6SL7 (or 12AX7).
V_3—6SN7GT (or 12AU7).

necessary, by interchanging the outputs from the two frequencies as applied to the printer.) The audio-frequency tones are applied to separate rectifiers to convert them into d.c. impulses, which may then be further amplified to the power level required to operate the printer.

The receiving converter which performs these functions generally will include means for clipping or limiting the signals so they are held at constant amplitude, and may also include provision for some shaping of the pulses to overcome distortion that occurs in transmission. There are many ways by which these results can be accomplished, and the higher the order of performance the more complicated the circuits become. However, satisfactory results under reasonably good receiving conditions can be secured with relatively simple equipment, and the "basic" circuit shown in Fig. 12-12 has proved to be quite successful in practice. It operates as follows:

When audio output from the receiver is applied, the two diodes, CR_1 and CR_2, which are biased with approximately 0.3 volt, limit the peak voltage at the grid of the limiter tube, V_{1A}, to 0.6 volt or less for signal voltages up to 30 volts or more. Additional limiting in V_{1A} further stabilizes the voltage level. V_{1B} is primarily an amplifier, and delivers approximately 15 volts output, constant to within 1 db. for receiver output voltages varying between about 0.5 volt and more than 30 volts.

The two tones, thus limited in amplitude, are applied to two simple filter circuits, L_1C_1 and L_2C_2, tuned to 2125 and 2975 cycles, respectively. The two tones are thus separated, one being applied to the grid of V_{2A} and the other to the grid of V_{2B}. V_{2A} and V_{2B} operate as grid-leak detectors, and when a signal is applied to, say, V_{2A}, the flow of grid current causes the grid to be driven practically to plate-current cutoff. As a result the plate voltage on V_{2A}, normally 15 volts with no signal, rises to 50 volts. This is sufficient to ignite the neon lamp connected between the plate of V_{2A} and the grid of V_{3B}, and a positive bias of about 25 volts is applied to the grid of V_{3B}. V_{3B} then takes a plate current of about 20 ma. and a bias of 20 volts is developed across the common cathode resistor, R_2. This is sufficient to cut off the plate current of V_{3A}, hence the left-hand magnet of the polarized relay, K_1, is inoperative while the right-hand magnet closes the contacts on its side. A similar action takes place when a signal is applied to the grid of V_{2B} but not to V_{2A}; in this

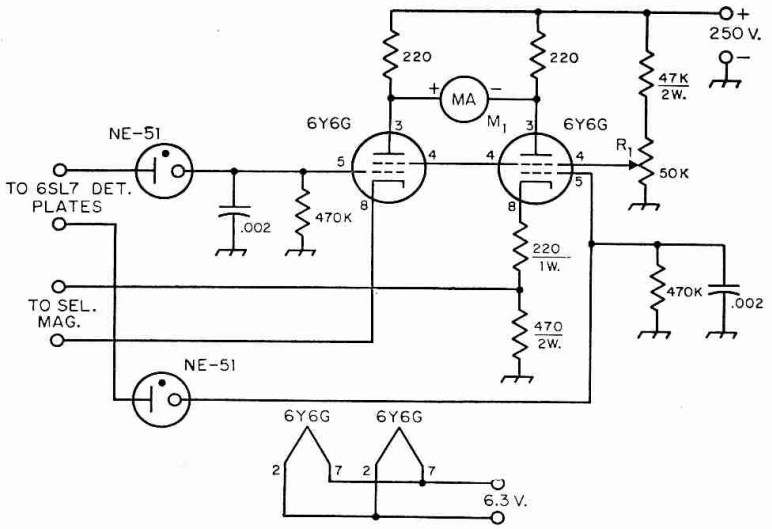


Fig. 12-13—Modification of converter circuit for use with single-magnet printers. Unless otherwise indicated, capacitances are in μf., resistances in ohms, resistors are ½ watt.

M_1—Zero-center d.c. milliammeter, 100 ma. full scale (may be microammeter with appropriate shunt).

R_1—50,000-ohm volume control.

case the relay contacts are pulled to the left. The relay thus keys the mark and space voltages applied to the printer.

Potentiometer R_1 is adjusted so that incoming noise (which will affect both channels equally) is balanced out and does not cause K_1 to operate. The neon lamps improve the operation of the circuit by acting as switches, thus making a sharp demarcation between mark and space pulses.

The zero-center meter, M_1, is not a necessity but is a convenience in making adjustments. R_1 should be adjusted on receiver noise for zero reading. With a 2125-cycle tone the pointer will swing to the left and L_1 should be adjusted for maximum deflection. With a 2975-cycle tone the pointer will swing to the right and L_2 should be adjusted for maximum deflection. Equal deflections should be obtained from both channels.

The keying circuit shown in Fig. 12-12 is for use with the Model 12 machine which requires an external power supply. For machines having a single selector magnet the modification shown in Fig. 12-13 may be used so the printer may be operated directly. These machines usually require a current of 60 ma., which will be furnished by this circuit and may be adjusted to the correct value by means of R_1.

Frequency-Shift Keyers

The keyboard contacts of the teletypewriter actuate a direct-current circuit that operates the printer magnets, and a pair of terminals is provided at which a keyed d.c. signal of the order of 100 volts is available. (Some machines, such as the Model 12, require an external d.c. power supply for this purpose; others have self-contained power supplies.) In the "resting" or non-operating condition the contacts are closed (mark) and the voltage at the terminals, which are in parallel with the contacts, is zero. In operation, the contacts open for "space" and the full voltage appears across the terminals. As normally connected, the spacing signal is of positive polarity.

This keyed d.c. voltage may be used to operate a keyer circuit for the radio transmitter, provided it is not "loaded" to such an extent that it affects the operation of the printer. Alternatively, the keyed current, rather than the voltage, may be used for external keying. This can be done by using an auxiliary keying relay with its coil connected in series with the printer magnet or relay circuit. A fast-acting relay must be used, and the coil must be one that will operate satisfactorily on the current available in the printer circuit. This will usually be either 20 or 60 milliamperes, depending on the type of machine.

F.S.K. with Variable-Frequency Oscillators

Perhaps the simplest satisfactory circuit for frequency-shift keying a v.f.o. is the one shown in Fig. 12-14A. This operates from the voltage available at the keyboard contact terminals and uses a reactance tube to obtain the required frequency shift.

The frequency shift is obtained by changing the plate resistance of the reactance tube, V_2, so that in effect the variable capacitor C_2 is alternately disconnected or connected in parallel with the tuning capacitor in the v.f.o. tank circuit. With no voltage applied to the grid, V_2 is biased so that the plate current is low and the effect of C_2 on the oscillator frequency is small. When a positive voltage from the keyboard contacts is applied to the grid the plate resistance is low and the oscillator frequency becomes lower because of the greater effect of C_2. The amount of frequency shift depends on the capacitance of C_2 and the amplitude of the positive voltage applied to the grid of V_2. The latter can be controlled by R_1.

C_1, the associated 20,000-ohm resistor, and the neon bulb, V_1, constitute a filter for removing clicks generated at the keyboard contacts. The value of C_1 depends somewhat on the machine,

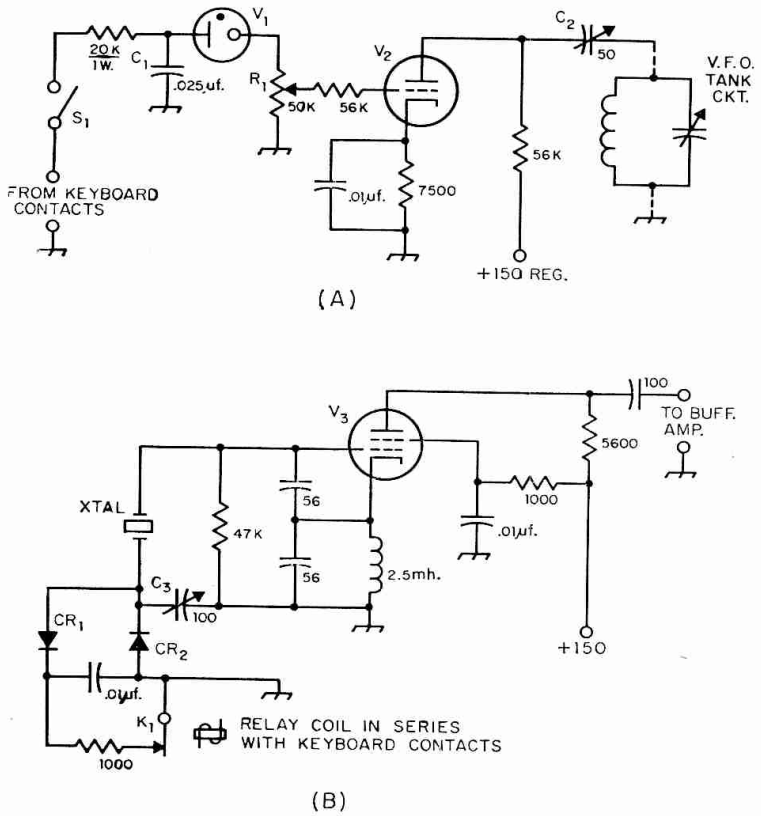


Fig. 12-14—Frequency-shift keyer circuits. Unless otherwise indicated, capacitances are in μμf., resistances are in ohms, resistors are ½ watt. A—Reactance-tube keyer for use with variable-frequency oscillator (W6OWP). B—Crystal oscillator circuit (W2PAT). It is essential that all leads associated with the crystal portion of the circuit be held to a small fraction of an inch in length if maximum shift is desired.

C_1—Paper (see text).
C_2—50-μμf. midget variable.
C_3—100-μμf. midget variable.
CR_1, CR_2—1N34 or equivalent.
K_1—Normally closed relay, fast operating, coil current according to printer magnet or relay current
R_1—Volume control.
S_1—S.p.s.t. toggle.
V_1—1-watt neon bulb without base resistor.
V_2—6C4 or equivalent.
V_3—6AK5 or equivalent.

and values up to 0.25 μf. can be used, if necessary, without objectionable distortion of the keying pulses. The capacitance should be adjusted for clickless keying.

The frequency-shift circuit should be initially adjusted at the lowest radio frequency to be used, since the shift will be smallest in this case. If C_2 is set so a shift of 850 cycles is obtained at this frequency, further adjustment of the shift may be made by means of R_1. If the transmitter output is on a higher-frequency band than that on which the v.f.o. operates, the shift at the v.f.o. fundamental frequency must be reduced accordingly.

F.S.K. With Crystal Oscillators

Fig. 12-14B is a circuit which has been found to give a frequency shift of 850 cycles or more with crystals of the type ordinarily used for frequencies of the order of 3.5 Mc. and higher. This is an oscillator of the "grid-plate" type discussed in Chapter 6 on transmitters, with the addition of a variable capacitor, C_3, in series with the crystal. C_3 reduces the total capacitance across the crystal and thus raises the oscillation frequency. When it is shorted out the capacitance across the crystal is higher and the resulting frequency is lower.

Although relay contacts could be used for shorting the capacitor, the diode arrangement shown in Fig. 12-14B is more reliable in practice. With the contacts of K_1 open there is no d.c. path through CR_2 and it acts simply as a small capacitance (about 1 μμf.) in parallel with C_3. When the contacts of K_1 are closed there is a d.c. circuit through CR_1, CR_2 and the 1000-ohm resistor. Thus there is a path for direct current flow as a result of rectification of the r.f. voltage across CR_2. Because of the d.c. bias the resistance of CR_2 drops to a low value and C_3 is effectively shorted out.

Adjustment of the circuit consists simply of determining the setting of C_3 at which the operating frequency is 850 cycles (or the desired shift) higher with the contacts of K_1 open than the frequency when the relay contacts are closed. A normally closed relay is used in order to make the mark frequency lower than the space frequency, in accordance with usual practice.

Frequency Adjustment

The frequency shift, whatever the type of circuit, should be made as nearly exact as available equipment will permit, since the shift must match the frequency difference between the filters in the receiving converter if the signals are to be usable at the receiving end. An accurately calibrated audio oscillator is useful for this purpose. To check, the mark frequency should be tuned in on the station receiver, with the b.f.o. on, and the receiver set to exact zero beat (see Chapter 21 on measurements for identification of exact zero beat). The space frequency should then be adjusted to exactly the desired shift. This may be done by adjusting for an auditory zero beat between the beat tone from the receiver and the tone from the audio oscillator. If an oscilloscope is available, the frequency adjustment may be accomplished by feeding the receiver tone to the vertical plates and the audio-oscillator tone to the horizontal plates, and then adjusting the space frequency for the elliptical pattern that indicates the two frequencies are the same.

Transmission Lines

The place where r.f. power is generated is very frequently not the place where it is to be utilized. A transmitter and its antenna are a good example: The antenna, to radiate well, should be high above the ground and should be kept clear of trees, buildings and other objects that might absorb energy, but the transmitter itself is most conveniently installed indoors where it is readily accessible.

The means by which power is transported from point to point is the r.f. transmission line. At radio frequencies a transmission line exhibits entirely different characteristics than it does at commercial power frequencies. This is because the speed at which electrical energy travels, while tremendously high as compared with mechanical motion, is not infinite. The peculiarities of r.f. transmission lines result from the fact that a time interval comparable with an r.f. cycle must elapse before energy leaving one point in the circuit can reach another just a short distance away.

Operating Principles

If a source of e.m.f. — a battery, for example — is connected to the ends of a pair of insulated parallel wires that extend outward for an infinite distance, electric currents will immediately become detectable in the wires near the battery terminals. The electric field of the battery will cause free electrons in the wire connected to the positive terminal to be attracted to the battery, and an equal number of free electrons in the wire connected to the negative terminal will be repelled from the battery. These currents do not flow instantaneously throughout the length of the wires; the electric field that causes the electron movement cannot travel faster than the speed of light, so a measurable interval of time elapses before the currents become evident even a relatively short distance away.

For example, the currents would not become detectable 300 meters (nearly 1000 feet) from the battery until at least a microsecond (one millionth of a second) after the connection was made. By ordinary standards this is a very short length of time, but in terms of radio frequency it represents the time of one complete cycle of a 1000-kilocycle current — a frequency considerably lower than those with which amateurs communicate.

The current flows to charge the capacitance between the two wires. However, the conductors of this "linear" capacitor also have appreciable inductance. The line may be thought of as being

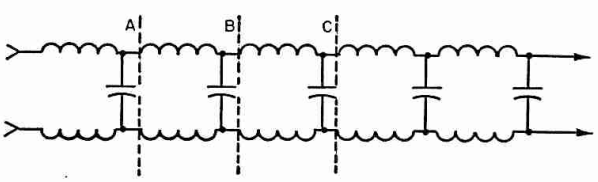


Fig. 13-1—Equivalent of a transmission line in lumped circuit constants.

composed of a whole series of small inductances and capacitances connected as shown in Fig. 13-1, where each coil is the inductance of a very short section of one wire and each capacitor is the capacitance between two such short sections.

Characteristic Impedance

An infinitely long chain of coils and capacitors connected as in Fig. 13-1, where the small inductances and capacitances all have the same values, respectively, has an important property. To an electrical impulse applied at one end, the combination appears to have an impedance — called the **characteristic impedance** or **surge impedance** — approximately equal to $\sqrt{L/C}$, where L and C are the inductance and capacitance per unit length. This impedance is purely resistive.

In defining the characteristic impedance as $\sqrt{L/C}$, it is assumed that the conductors have no inherent resistance — that is, there is no I^2R loss in them — and that there is no power loss in the dielectric surrounding the conductors. There is thus no power loss in or from the line no matter how great its length. This may not seem consistent with calling the characteristic impedance a pure resistance, which implies that the power supplied is all dissipated in the line. But in an infinitely long line the effect, so far as the source of power is concerned, is exactly the same as though the power were dissipated in a resistance, because the power leaves the source and travels outward forever along the line.

The characteristic impedance determines the amount of current that can flow when a given voltage is applied to an infinitely long line, in exactly the same way that a definite value of actual resistance limits current flow when a voltage is applied.

The inductance and capacitance per unit length of line depend upon the size of the conductors and the spacing between them. The closer the two conductors 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.

"Matched" Lines

Actual transmission lines do not extend to infinity but have a definite length and are connected to, or terminate 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 line is said to be **matched.** To current traveling along the line such a load just looks like still more transmission line of the same characteristic impedance.

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. In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

R.F. on Lines

The principles discussed above, although based on direct-current flow from a battery, also hold when an r.f. voltage is applied to the line. The difference is that the alternating voltage causes the amplitude of the current at the input terminals of the line to vary with the voltage, and the direction of current flow also periodically reverses when the polarity of the applied voltage reverses. The current at a given instant at any point along the line is the result of a voltage that was applied at some *earlier* instant at the input terminals. Since the distance traveled by the electromagnetic fields in the time of one cycle is equal to one wavelength (Chapter 2), the instantaneous amplitude of the current is different at all points in a one-wavelength section of line. In fact, the current flows in opposite directions in the same wire in successive half-wavelength sections. However, at any given point along the line the current goes through similar variations with time that the current at the input terminals did.

Thus the current (and voltage) travels along the wire as a series of waves having a length equal to the speed of travel divided by the frequency of the a.c. voltage. On an infinitely long line, or one properly matched by its load, an ammeter inserted anywhere in the line will show the same current, because the ammeter averages out the variations in current during a cycle. It is only when the line is not properly matched that the wave motion becomes apparent through observations made with ordinary instruments.

● STANDING WAVES

In the infinitely long line (or its matched counterpart) the impedance is the same at any point on the line because the ratio of voltage to current is always the same. However, the impedance at the end of the line in Fig. 13-2 is zero — or at least extremely small — because the line is short-circuited at the end. The outgoing power, on meeting the short-circuit, reverses its direction of flow and goes back along the transmission line toward the input end. There is a 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 **(incident power)** toward the short-circuit, and a second voltage and current representing the **reflected power** traveling back toward the source.

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 phase of the incident and reflected currents will be such that the currents cancel each other while at others the amplitude will be doubled. At in-between points the amplitude is between these two extremes. 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.

In the short-circuit at the end of the line the two current components are in phase and the total current is large. 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 resultant current will again have its maximum value. This is also

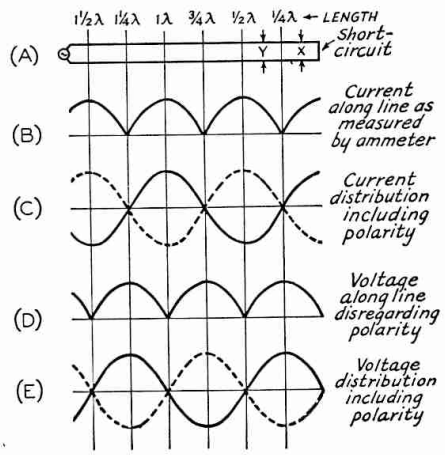


Fig. 13-2—Standing waves of voltage and current along short-circuited transmission line.

true at any point that is a multiple of a half wavelength from the short-circuited end of the line.

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. 13-2B. 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. 13-2C. Furthermore, the current in the second wire is flowing in the opposite direction to the current

in the adjacent section of the first wire. This is indicated by the broken curve in Fig. 13-2C. 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** or **current antinode** and the point of minimum line current is called 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 a reflected voltage of equal amplitude and opposite polarity. In other words, 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. 13-2, 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 is called a **voltage loop** or **antinode** and a voltage minimum is called a **voltage node.**

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 incident power is reflected back toward the source. The incident and reflected components of current must be equal and opposite in phase at the open circuit in order for the total current at the end of the line to be zero. The incident and reflected components of voltage are in phase and add together. The result is again that there are standing waves, but the conditions are reversed as compared with a short-circuited line. Fig. 13-3 shows the open-circuited line case.

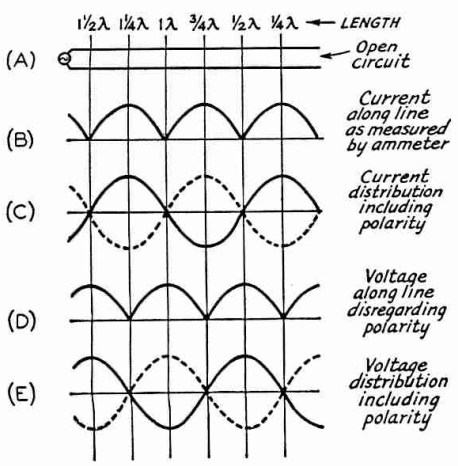


Fig. 13-3—Standing waves of current and voltage along an open-circuited transmission line.

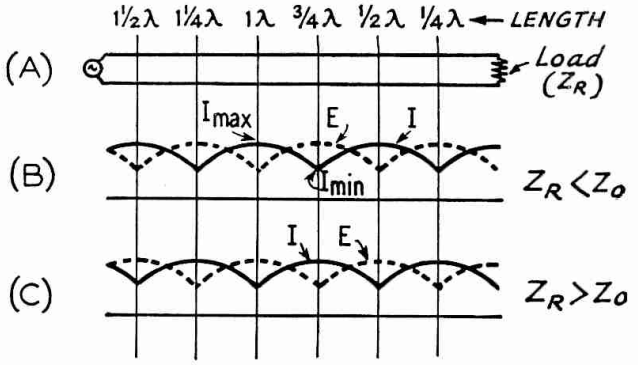


Fig. 13-4—Standing waves on a transmission line terminated in a resistive load.

Lines Terminated in Resistive Load

Fig. 13-4 shows a line terminated in a resistive load. In this case at least part of the incident 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 incident components. Therefore neither voltage nor current cancel completely at any point along the line. However, the *speed* at which the incident 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, Z_R, is equal to the characteristic impedance, Z_0, 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 change-over point between "short-circuited" and "open-circuited" lines. If Z_R is less than Z_0, the current is largest at the load, while if Z_R is greater than Z_0 the voltage is largest at the load. The two conditions are shown at B and C, respectively, in Fig. 13-4.

The resistive termination is an important practical case. The termination is seldom an actual resistor, the most common terminations being resonant circuits or resonant antenna systems, both of which have essentially resistive impedances. If the load is reactive as well as resistive, the operation of the line resembles that shown in Fig. 13-4, but the presence of reactance in the load causes two modifications: The loops and nulls are shifted toward or away from the load; and the amount of power reflected back toward the source is increased, as compared with the amount reflected by a purely resistive load of the same total impedance. Both effects become more pronounced as the ratio of reactance to resistance in the load is made larger.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, Fig. 13-5, is called the **standing-wave ratio.** The same ratio holds for maximum voltage and minimum voltage. It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is per-

fectly matched. (In that case the "maximum" and "minimum" are the same, since the current and voltage do not vary along the line.) When the line is terminated in a purely resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_R}{Z_0} \text{ or } \frac{Z_0}{Z_R} \qquad \textbf{(13-A)}$$

Where $S.W.R.$ = Standing-wave ratio

Z_R = Impedance of load (must be pure resistance)

Z_0 = 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

$$S.W.R. = \frac{Z_0}{Z_R} = \frac{300}{25} = 12 \text{ to } 1$$

It is customary to put the larger of the two quantities, Z_R or Z_0, 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. In practical lines, the power loss in the line itself increases with the s.w.r., as shown later.

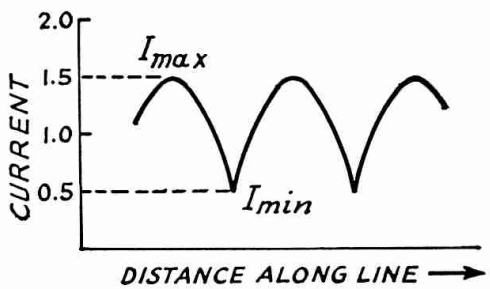


Fig. 13-5—Measurement of standing-wave ratio. In this drawing, I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. = I_{max}/I_{min} = 1.5/0.5 = 3 to 1.

INPUT IMPEDANCE

The input impedance of a transmission line is the impedance seen looking into the sending-end or input terminals; it is the impedance into which the source of power must work when the line is connected. If the load is perfectly matched to the line the line appears to be infinitely long, as stated earlier, and the input impedance is simply the characteristic impedance of the line itself. However, if there are standing waves this is no longer true; the input impedance may have a wide range of values.

This can be understood by referring to Figs. 13-2, 13-3, or 13-4. If the line length is such that standing waves cause the voltage at the input terminals to be high and the current low, then the input impedance is higher than the Z_0 of the line, since impedance is simply the ratio of voltage to current. Conversely, low voltage and high current at the input terminals mean that the input impedance is lower than the line Z_0. Comparison of the three drawings also shows that the range of input impedance values that may be encountered is greater when the far end of the line is open- or short-circuited than it is when the line has a resistive load. In other words, the higher the s.w.r. the greater the range of input impedance values when the line length is varied.

In addition to the variation in the absolute value of the input impedance with line length, the presence of standing waves also causes the input impedance to contain both reactance and resistance, even though the load itself may be a pure resistance. The only exceptions to this occur at the exact current loops or nodes, at which points the input impedance is a pure resistance. These are the only points at which the outgoing and reflected voltages and currents are exactly in phase: At all other distances along the line the current either leads or lags the voltage and the effect is exactly the same as though a capacitance or inductance were part of the input impedance.

The input impedance can be represented either by a resistance and a capacitance or by a resistance and an inductance, as shown in Fig. 13-6. Whether the impedance is inductive or capacitive depends on the characteristics of the load and the length of the line. It is possible to represent the input impedance by an equivalent circuit having resistance and reactance either in series or parallel, so long as the total impedance and phase angle are the same in either case. For a given impedance and phase angle, different values of resistance and reactance are required in the series circuit as compared with the parallel equivalent circuit.

The magnitude and character of the input impedance is quite important, since it determines the method by which the power source must be coupled to the line. The calculation of input impedance is rather complicated and its measurement is not feasible without special equipment. Fortunately, in amateur work it is unnecessary either to calculate or measure it. The proper coupling can be achieved by relatively simple methods described later in this chapter.

Lines Without Load

The input impedance of a short-circuited or open-circuited line not an exact multiple of one-quarter wavelength long is practically a pure reactance. This is because there is very little power lost in the line. Such lines are frequently used as "linear" inductances and capacitances.

If a shorted line is less than a quarter-wave long, as at X in Fig. 13-2, it will have inductive reactance. The reactance increases with the line length up to the quarter-wave point. Beyond that, as at Y, the reactance is capacitive, high near the quarter-wave point and becoming lower as the half-wave point is approached. It then alternates between inductive and capacitive in successive

quarter-wave sections. Just the reverse is true of the open-circuited line.

At exact multiples of a quarter wavelength the impedance is purely resistive. It is apparent, from examination of B and D in Fig. 13-2, 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

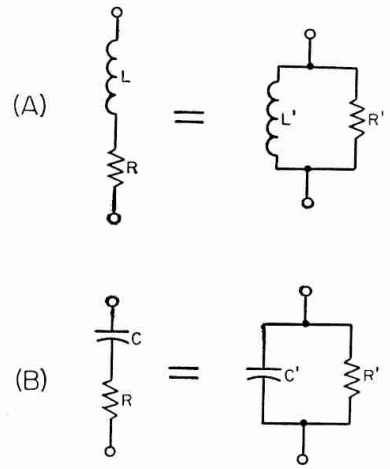


Fig. 13-6—Series and parallel equivalents of a line whose input impedance has both reactive and resistive components. The series and parallel equivalents do not have the same values; e.g., in A, *L* does not equal *L'* and *R* does not equal *R'*.

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.)

Impedance Transformation

The fact that the input impedance of a line depends on the s.w.r. and line length can be used to advantage when it is necessary to transform a given impedance into another value.

Study of Fig. 13-4 will show that, just as in the open- and short-circuited cases, if the line is one-half wavelength long the voltage and current are exactly the same at the input terminals as they are at the load. This is also true of lengths that are integral multiples of a half wavelength. It is also true for all values of s.w.r. Hence the input impedance of any line, no matter what its Z_0, that is a multiple of a half wavelength long is exactly the same as the load impedance. Such a line can be used to transfer the impedance to a new location without changing its value.

When the line is a quarter wavelength long, or an odd multiple of a quarter wavelength, the load impedance is "inverted." That is, if the current is low and the voltage is high at the load, the input impedance will be such as to require high current and low voltage. The relationship between the load impedance and input impedance is given by:

$$Z_S = \frac{Z_0^2}{Z_R} \qquad \textbf{(13-B)}$$

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_0 = 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

$$Z_S = \frac{Z_0^2}{Z_R} = \frac{(500)^2}{75} = \frac{250{,}000}{75} = 3333 \text{ ohms}$$

If the formula above is rearranged, we have

$$Z_0 = \sqrt{Z_S Z_R} \qquad \textbf{(13-C)}$$

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.

Resonant and Nonresonant Lines

The input impedance of a line operating with a high s.w.r. is critically dependent on the line length, and resistive only when the length is some integral multiple of one-quarter wavelength. Lines cut to such a length and operated with a high s.w.r. are called "tuned" or "resonant" lines. On the other hand, if the s.w.r. is low the input impedance is close to the Z_0 of the line and does not vary a great deal with the line length. Such lines are called "flat," or "untuned," or "nonresonant."

There is no sharp line of demarcation between tuned and untuned lines. If the s.w.r. is below 1.5 to 1 the line is essentially flat, and the same input coupling method will work with all line lengths. If the s.w.r. is above 3 or 4 to 1 the type of coupling system, and its adjustment, will depend on the line length and such lines fall into the "tuned" category.

It is usually advantageous to make the s.w.r. as low as possible. A resonant line becomes necessary only when a considerable mismatch between the load and the line has to be tolerated. The most important practical example of this is when a single antenna is operated on several harmonically related frequencies, in which case the antenna impedance will have widely different values on different harmonics.

● 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 (for a given current) will be small. However, a transmission line used to feed power to an antenna is not short; 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. 13-2C and 13-3C. It 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.

Practically, the fields do not quite cancel out because for them to do so the two conductors would have to occupy the same space, whereas they are actually slightly separated. However, the cancelation is substantially complete if the distance between the conductors is very small compared to the wavelength. Transmission line radiation will be negligible if the distance between the conductors is 0.01 wavelength or less, provided the currents in the two wires are balanced.

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 — and is just as serious when the line is flat as when the s.w.r. is high.

Practical Line Characteristics

The foregoing discussion of transmission lines has been based on a line consisting of two parallel conductors. The **parallel-conductor** line is but one of two general types, the other being the **coaxial** or **concentric** line. The coaxial line consists of a conductor placed in the center of a 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 the construction and use of parallel and coaxial lines.

PARALLEL-CONDUCTOR LINES

A type of parallel-conductor line sometimes used in amateur installations is one 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. 13-7. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 13-8. The characteristic impedance of such "open-wire" lines is between 400 and 600 ohms, depending on the wire size and spacing.

Parallel-conductor lines also are occasionally constructed of metal tubing of a diameter of ¼ to ½ inch. This reduces the characteristic impedance

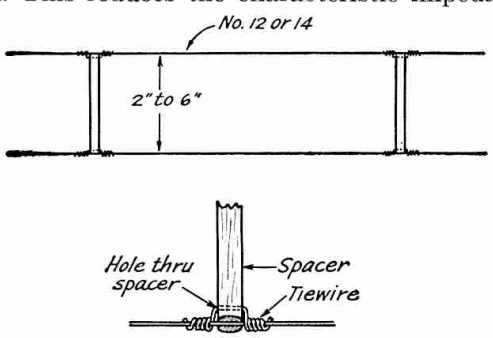


Fig. 13-7—Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

of the line. Such lines are mostly used as quarter-wave transformers, when different values of impedance are to be matched.

Prefabricated parallel-conductor line with air insulation, developed for television reception, can be used in transmitting applications. This line consists of two conductors separated one-half to one inch by molded-on spacers. The characteristic impedance is 300 to 450 ohms, depending on the wire size and spacing.

A convenient type of manufactured line is one in which the parallel conductors are imbedded in low-loss insulating material (polyethylene). It is commonly used as a TV lead-in and has a charac-

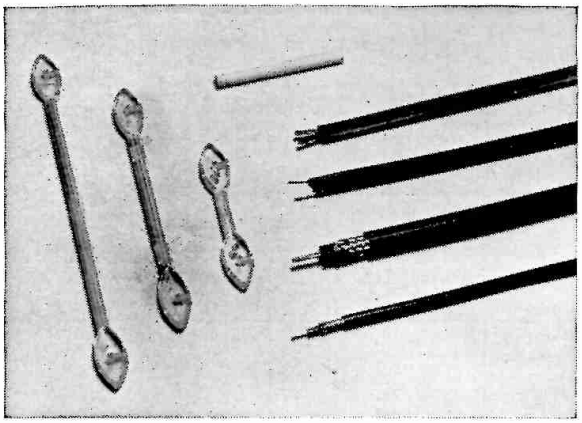


Fig. 13-8—Typical manufactured transmission lines and spacers.

teristic impedance of about 300 ohms. It is sold under various names, the most common of which is "Twin-Lead." This type of line has the advantages of light weight, close and uniform conductor spacing, flexibility and neat appearance. However, the losses in the solid dielectric are higher than in air, and dirt or moisture on the line tends to change the characteristic impedance. Moisture effects can be reduced by coating the line with silicone grease. A special form of 300-ohm Twin-Lead for transmitting uses a polyethylene tube with the conductors molded diametrically opposite; the longer dielectric path in such line reduces moisture troubles.

In addition to 300-ohm line, Twin-Lead is obtainable with a characteristic impedance of 75 ohms for transmitting purposes. Light-weight 75- and 150-ohm Twin-Lead also is available.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_0 = 276 \log \frac{b}{a} \qquad \textbf{(13-D)}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

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. 13-9 for a number of common conductor sizes.

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 dielectric.

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. However, 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 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

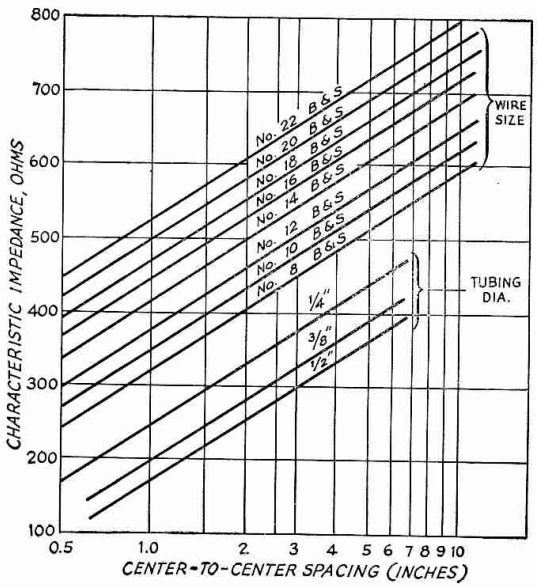


Fig. 13-9—Chart showing the characteristic impedance of spaced-conductor parallel transmission lines with air dielectric. Tubing sizes given are for outside diameters.

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.

● COAXIAL LINES

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. 13-8. 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 rarely 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" placed at regular intervals.

Characteristic Impedance

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_0 = 138 \log \frac{b}{a} \qquad \textbf{(13-E)}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Curves for typical conductor sizes are given in Fig. 13-10.

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.

● 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 material dielectrics 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.

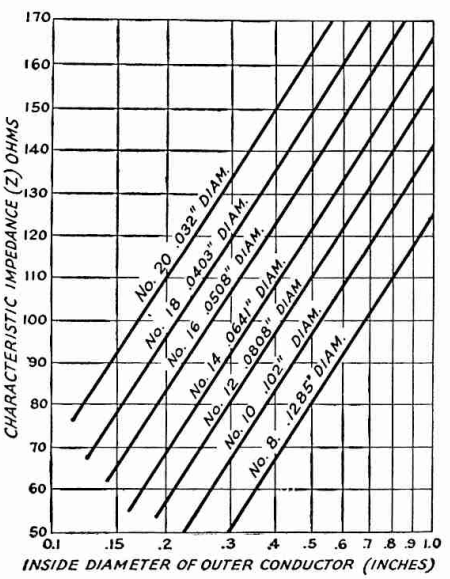


Fig. 13-10—*Chart showing characteristic impedance of various air-insulated concentric lines.*

TABLE 13-I
Transmission-Line Data

Type	Description or Type Number	Characteristic Impedance	Velocity Factor	Capacitance per foot; $\mu\mu$f.
Coaxial	Air-insulated	50–100	0.85[1]	
	RG-8/U	53	0.66	29.5
	RG-58/U	53	0.66	28.5
	RG-11/U	75	0.66	20.5
	RG-59/U	73	0.66	21.0
Parallel-Conductor	Air-insulated	200–600	0.975[2]	
	214-080[3]	75	0.68	19.0
	214-023[3]	75	0.71	20.0
	214-079[3]	150	0.77	10.0
	214-056[3]	300	0.82	5.8
	214-076[3]	300	0.84	3.9
	214-022[3]	300	0.85	3.0

[1] Average figure for small-diameter lines with ceramic beads.
[2] Average figure for lines insulated with ceramic spacers at intervals of a few feet.
[3] Amphenol type numbers and data. Line similar to 214-056 is made by several manufacturers, but rated loss may differ from that given in Fig. 13-11. Types 214-023, 214-076, and 214-022 are made for transmitting applications.

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

$$\textit{Length in feet} = \frac{984V}{f} \qquad \textbf{(13-F)}$$

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 13-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 13-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is

$$\textit{Length (feet)} = \frac{984V}{f} = \frac{984}{7.15} \times 0.82$$

$$= 137.6 \times 0.82 = 112.8 \text{ ft.}$$

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 con-

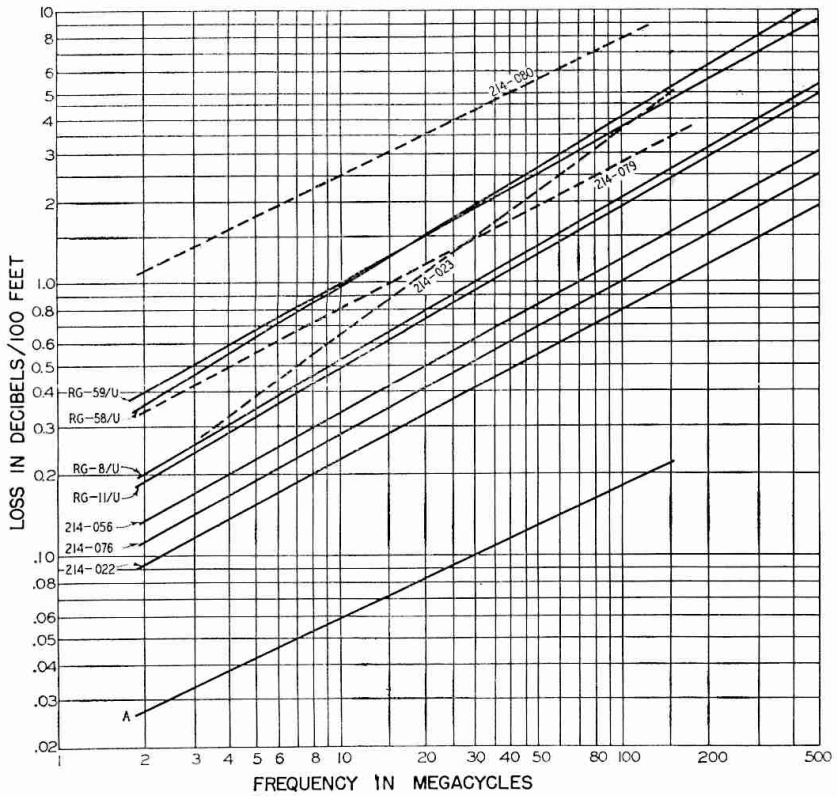


Fig. 13-11—Attenuation data for common types of transmission lines. Curve A is the nominal attenuation of 600-ohm open-wire line with No. 12 conductors, not including dielectric loss in spacers nor possible radiation losses. Additional line data is given in Table 13-I.

venient to calculate the length of a quarter-wave line directly. The formula is

$$Length\ (\text{feet}) = \frac{246V}{f} \qquad \textbf{(13-G)}$$

where the symbols have the same meaning as above.

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. Radiation losses are in general the result of "antenna currents" on the line, resulting from undesired coupling to the radiating antenna. They cannot readily be estimated or measured, so the following discusssion is based only on conductor and dielectric losses.

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.

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 graphical form in Fig. 13-11. In these curves the radiation loss is assumed to be negligible.

When there are standing waves on the line the power loss increases as shown in Fig. 13-12. 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.,

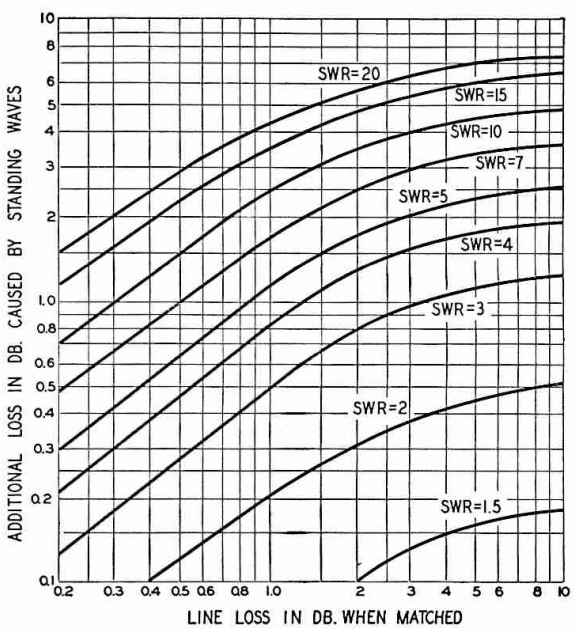


Fig. 13-12—Effect of standing-wave ratio on line loss. The ordinates give the *additional* loss in decibels for the loss, under perfectly matched conditions, shown on the horizontal scale.

the ratio of the power delivered to the load to the power put into the line.

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 Fig. 13-11 would be 1.5 × 0.4 = 0.6 db. From Fig. 13-12 the additional loss because of the s.w.r. is 0.73 db. The total loss is therefore 0.6 + 0.73 = 1.33 db.

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. 13-12, 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.

Loads and Balancing Devices

The most important practical load for a transmission line is an antenna which, in most cases, will be "balanced" — that is, symmetrically constructed with respect to the feed point. Aside from considerations of matching the actual impedance of the antenna at the feed point to the characteristic impedance of the line (if such matching is attempted) a balanced antenna should be fed through a balanced transmission line in order to preserve symmetry with respect to ground and thus avoid difficulties with unbalanced currents on the line. Such currents, as pointed out earlier in this chapter, will result in undesirable radiation from the transmission line itself.

If, as is often the case, the antenna is to be fed through coaxial line (which is inherently unbalanced) some method should be used for connecting the line to the antenna without upsetting the symmetry of the antenna itself. This requires a circuit that will isolate the balanced load from the unbalanced line while providing efficient power transfer. Devices for doing this are called **baluns**. The types used between the antenna and transmission line are generally "linear," consisting of transmission-line sections as described in Chapter 14.

The need for baluns also arises in coupling a transmitter to a balanced transmission line, since the output circuits of most transmitters have one side grounded. (This type of output circuit is desirable for a number of reasons, including TVI reduction.) The most flexible type of balun for this purpose is the inductively coupled matching network described in a subsequent section in this chapter. This combines impedance matching with balanced-to-unbalanced operation, but has the disadvantage that it uses resonant circuits and thus can work over only a limited band of frequencies without readjustment. However, if a fixed impedance ratio in the balun can be tolerated, the coil balun described below can be used without adjustment over a frequency range of about 10 to 1 — 3 to 30 Mc., for example. Alternatively, a similarly wide band can be covered by a properly designed transformer (with the same impedance limitation) but the design principles and materials used in such transformers are quite specialized. Their construction is beyond the scope of this *Handbook*.

Coil Baluns

The type of balun known as the "coil balun" is based on the principles of a linear transmission-line balun as shown in the upper drawing of Fig. 13-13. Two transmission lines of equal length having a characteristic impedance Z_0 are connected in series at one end and in parallel at the other. At the series-connected end the lines are balanced to ground and will match an impedance equal to $2Z_0$. At the parallel-connected end the lines will be matched by an impedance equal to $Z_0/2$. One side may be connected to ground at the parallel-connected end, provided the two lines have a length such that, considering each line as a single wire, the balanced end is effectively decoupled from the parallel-connected end. This requires a length that is an odd multiple of ¼ wavelength. The impedance transformation from the series-connected end to the parallel-connected end is 4 to 1.

A definite line length is required only for decoupling purposes, and so long as there is adequate decoupling the system will act as a 4-to-1 impedance transformer regardless of line length. If each line is wound into a coil, as in the lower drawing, the inductances so formed will act as choke coils and will tend to isolate the series-connected end from any ground connection that may be placed on the parallel-connected end. Balun coils made in this way will operate over a wide frequency range, since the choke inductance is not critical. The lower frequency limit is where the coils are no longer effective in isolating one end from the other; the length of line in each coil should be about equal to a quarter wavelength at the lowest frequency to be used.

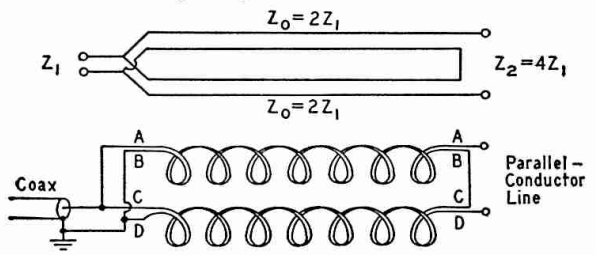


Fig. 13-13—Baluns for matching between push-pull and single-ended circuits. The impedance ratio is 4 to 1 from the push-pull side to the unbalanced side. Coiling the lines as shown in the lower drawing increases the frequency range over which satisfactory operation is obtained.

The principal application of such coils is in going from a 300-ohm balanced line to a 75-ohm coaxial line. This requires that the Z_0 of the lines forming the coils be 150 ohms. Design data for winding the coils is not available; however, Equation 13-D can be used for determining the approximate wire spacing. Allowance should be made for the fact that the effective dielectric constant will be somewhat greater than 1 if the coil is wound on a form. The proximity effect between turns can be reduced by making the turn spacing somewhat larger than the conductor spacing. For operation at 3.5 Mc. and higher frequencies the length of each conductor should be about 60 feet. The conductor spacing can be adjusted to the proper value by terminating each line in a noninductive 150-ohm resistor and adjusting the spacing until an impedance bridge at the input end shows the line to be matched to 150 ohms.

A balun of this type is simply a fixed-ratio transformer, when matched. It cannot compensate for inaccurate matching elsewhere in the system. With a "300-ohm" line on the balanced end, for example, a 75-ohm coax cable will not be matched unless the 300-ohm line actually is terminated in a 300-ohm load.

● NONRADIATING LOADS

Typical examples of nonradiating loads for a transmission line are the grid circuit of a power amplifier (considered in the chapter on transmitters), the input circuit of a receiver, and another transmission line. This last case includes the "antenna tuner" — a misnomer because it is actually a device for coupling a transmission line to the transmitter. Because of its importance in amateur installations, the antenna coupler is considered separately in a later part of this chapter.

Coupling to a Receiver

A good match between an antenna and its transmission line does not guarantee a low standing-wave ratio on the line when the antenna system is used for receiving. The s.w.r. is determined wholly by what the line "sees" at the receiver's antenna-input terminals. For minimum s.w.r. the receiver input circuit must be matched to the line. The rated input impedance of a receiver is a nominal value that varies over a considerable range with frequency. Methods for bringing about a proper match are discussed in the chapter on receivers.

It should be noted that *if* the receiver is matched to the line, then it is desirable that the antenna and line also be matched, since this results in maximum signal transfer from the antenna to the line. If the receiver is *not* matched to the line, the input impedance of the line (at the terminals of the antenna itself) in turn cannot match the antenna impedance. In such a case the signal input to the receiver depends on the coupling system used between the line and the receiver. For greatest signal strength the coupling system has to be adjusted to the best compromise between receiver input impedance and load appearing at the input (antenna) end of the line. The proper adjustments must be determined by experiment.

A similar situation exists when the receiver input impedance inherently matches the line Z_0, but the line and antenna are mismatched. Under these conditions perfect matching at the receiver does not result in greatest signal strength; a deliberate mismatch has to be introduced so that the maximum power will be taken from the antenna.

The most desirable condition is that in which the receiver is matched to the line Z_0 and the line in turn is matched to the antenna. This transfers maximum power from the antenna to the receiver with the least loss in the transmission line.

Coupling the Transmitter to the Line

The type of coupling system that will be needed to transfer power adequately from the final r.f. amplifier to the transmission line depends almost entirely on the input impedance of the line. As shown earlier in this chapter, the input impedance is determined by the standing-wave ratio and the line length. The simplest case is that where the line is terminated in its characteristic impedance so that the s.w.r. is 1 to 1 and the input impedance is equal to the Z_0 of the line, regardless of line length.

Coupling systems that will deliver power into a flat line are readily designed. For all practical purposes the line can be considered to be flat if the s.w.r. is no greater than about 1.5 to 1. That is, a coupling system designed to work into a pure resistance equal to the line Z_0 will have enough leeway to take care of the small variations in input impedance that will occur when the line length is changed, if the s.w.r. is higher than 1 to 1 but no greater than 1.5 to 1.

Current practice in transmitter design is to provide an output circuit that will work into such a line, usually a coaxial line of 50 to 75 ohms characteristic impedance. The design of such output circuits is discussed in the chapter on high-frequency transmitters. If the input impedance of the transmission line that is to be connected to the transmitter differs appreciably from the value of impedance into which the transmitter output circuit is designed to operate, an impedance-matching network must be inserted between the transmitter and the line input terminals.

● IMPEDANCE-MATCHING CIRCUITS FOR PARALLEL CONDUCTOR LINES

As shown earlier in this chapter, the input impedance of a line that is operating with a high standing-wave ratio can vary over quite wide

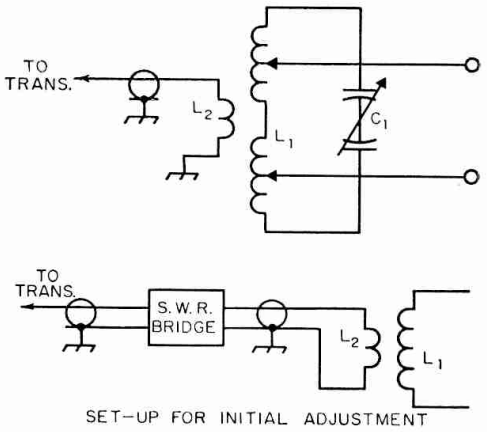


Fig. 13-14—Matching circuits using a coaxial link, for use with parallel-conductor transmission lines. Adjustment setup using an s.w.r. bridge is shown in the lower drawing. Design considerations and method of adjustment are discussed in the text.

limits. The simplest type of circuit that will match such a range of impedances to 50 to 75 ohms is a parallel-tuned circuit approximately resonant at the operating frequency. In its ordinary form, such a circuit will be connected to a short length of coaxial line or "link" by inductive coupling as shown in Fig. 13-14, the other end of the cable being attached to the output terminals of the transmitter. The cable may be any convenient length if the impedance that it "sees" at the matching circuit is equal to its own characteristic impedance. This method has the further advantage that the coaxial link offers an ideal spot for the insertion of a low-pass filter for preventing harmonic interference to television and f.m. reception.

The constants of the tuned circuit C_1L_1 are not particularly critical; the principal requirement is that the circuit must be capable of being tuned to the operating frequency. Constants similar to those used in the plate tank circuit will be satisfactory. The construction of L_1 must be such that it can be tapped at least every turn. L_2 must be tightly coupled to L_1, and the inductance of L_2 should be approximately the value that gives a reactance equal to the Z_0 of the connecting line at the frequency in use. An average reactance of about 60 ohms will suffice for either 52- or 75-ohm coaxial line.

The most satisfactory way to set up the system initially is to connect a coaxial s.w.r. bridge in the link as shown in Fig. 13-14. The "Monimatch" type of bridge, which can handle the full transmitter power and may be left in the line for continuous monitoring, is excellent for this purpose. However, a simple resistance bridge such as is described in the chapter on measurements is perfectly adequate, requiring only that the transmitter output be reduced to a very low value so that the bridge will not be overloaded. To adjust the circuit, take a trial position of the line taps on L_1, keeping them equidistant from the center of the coil, and adjust C_1 for minimum s.w.r. as indicated by the bridge. If the s.w.r. is not close to 1 to 1, try new tap positions and adjust C_1 again, continuing this procedure until the s.w.r. is practically 1 to 1. The setting of C_1 and the tap positions may then be logged for future reference. At this point, check the link s.w.r. over the frequency range normally used in that band, without changing the setting of C_1. No readjustment will be required if the s.w.r. does not exceed 1.5 to 1 over the range, but if it goes higher it is advisable to note as many settings of C_1 as may be necessary to keep the s.w.r. below 1.5 to 1 at any part of the band. Changes in the link s.w.r. are caused chiefly by changes in the s.w.r. on the main transmission line with frequency, and relatively little by the coupling circuit itself. A single setting of C_1 at mid-frequency will suffice if the antenna itself is broad-tuning.

If it is impossible to get a 1-to-1 s.w.r. at any settings of the taps or C_1, the s.w.r. on the main transmission line is high and the line length is probably unfavorable. Ordinarily there should be no difficulty if the transmission-line s.w.r. is not more than about 3 to 1, but if the line s.w.r. is higher it may not be possible to bring the link s.w.r. down except by using the methods for reactance compensation described in a subsequent section in this chapter.

The matching adjustment can be considerably facilitated by using a variable capacitor in series with the matching-circuit coupling coil as shown in Fig. 13-15. The additional adjustment thus provided makes the tap settings on L_1 much less critical since varying C_2 has the effect of varying the coupling between the two circuits. For optimum control of coupling, L_2 should be somewhat larger than when C_2 is not used — perhaps twice the reactance recommended above — and the reactance of C_2 at maximum capacitance should be the same as that of L_2 at the operating frequency. L_1 and C_1 are the same as before. The method of adjustment is the same, except that for each trial tap position C_1 and C_2 are alternately adjusted, a little at a time, until the s.w.r. is brought to its lowest possible value. In general, the adjustment sought should be the one that keeps C_2 at the largest possible capacitance, since this broadens the frequency response. Also, the taps on L_1 should be kept as far apart as possible, while still permitting a match, since this also broadens the frequency response of the circuit.

Once the matching circuit is properly adjusted, the s.w.r. bridge may be removed, if necessary, and full power applied to the transmitter. The power input should be adjusted by the coupling or loading control built into the transmitter, *not*

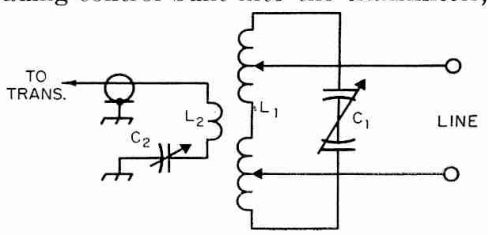


Fig. 13-15—Using a series capacitor for control of coupling between the link and line circuits with the coax-coupled matching circuit.

by making any changes in the matching-circuit adjustments. If an amplifier having a parallel-tuned tank circuit will not load properly, tuned coupling should be used into the coax link.

It is possible to use a circuit of this type without initially setting it up with the s.w.r. bridge. In such a case it is a matter of cut-and-try until adequate power transfer between the amplifier and main transmission line is secured. However, this method frequently results in a high s.w.r. in the link, with consequent power loss, "hot spots" in the coaxial cable, and tuning that is critical with frequency. The bridge method is simple and gives the optimum operating conditions quickly and with certainty.

Untuned Coupling

A simple coil can be used for coupling to a line having a high standing-wave ratio providing the line length is adjusted so there is a current loop near the point where it connects to the pick-up coil. The coupling will be maximum, for a given degree of separation between the pick-up coil and the amplifier tank coil, if the line is pruned to a length such that the input impedance is just sufficiently capacitive to cancel the inductive reactance of the pick-up coil. This can be done by cut-and-try. The higher the s.w.r. on the line the easier it becomes to load the amplifier with loose coupling between the two coils. The sharper the antenna and the higher the line s.w.r. the more difficult it becomes to operate with this system over a band without progressively changing the line length.

Series and Parallel Tuning

Lines classified as "tuned" or "resonant" — i.e., cut to lengths approximately equal to integral multiples of one-quarter wavelength, and operating with a high standing-wave ratio — are characterized by having either very high or very low input impedances. Also, the input impedances of such lines are essentially resistive.

Under these conditions the circuit arrangements shown in Fig. 13-16 will work satisfactorily.

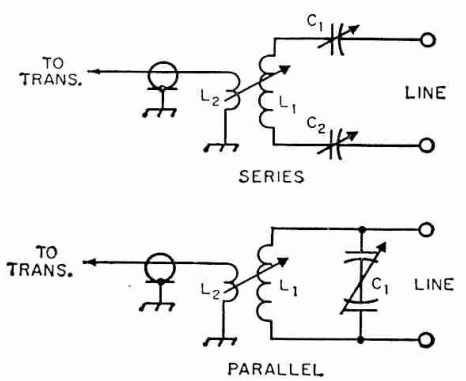


Fig. 13-16—Link-coupled series and parallel tuning.

Their advantage over the circuit of Fig. 13-14 is that it is not necessary to provide for taps on the matching-circuit coil, L_1. "Series" tuning is used when a current loop occurs at or near the input end of the line; i.e., when the input impedance is low. "Parallel" tuning is used when there is a voltage loop at or near the input end; i.e., when the input impedance is high.

In the series case, the circuit formed by L_1, C_1 and C_2 with the line terminals short-circuited should tune to the operating frequency. C_1 and C_2 should be maintained at equal capacitance. In the parallel case, the circuit formed by L_1 and C_1 should tune to resonance with the line disconnected.

The L/C ratio in either circuit depends on the transmission line Z_0 and the standing-wave ratio. With series tuning, a high L/C ratio must be used if the s.w.r. is relatively low and the line Z_0 is high. With parallel tuning, a low L/C ratio must be used if the s.w.r. is relatively low and the transmission-line Z_0 also is low. With either series or parallel tuning the L/C ratio becomes less critical when the s.w.r. is high. As a first approximation, coil and capacitor values of the same order as those used in the plate tank circuit may be tried. The coupling coil, L_2, should have a reactance about equal to the Z_0 of the coaxial line, just as in the case of the circuit of Fig. 13-14. The coupling between L_1 and L_2 should be continuously adjustable.

Two capacitors are used in the series-tuned circuit in order to keep the line balanced to ground. This is because two identical capacitors, both connected with either their stators or rotors to the line, will have the same capacitance to ground. A single capacitor would be perfectly usable so far as the operation of the coupling circuit is concerned, but will slightly unbalance the circuit because the frame has more capacitance to ground than the stator. The unbalance is not especially serious unless the capacitor is mounted near a large mass of metal, such as a chassis or shield assembly.

A balanced capacitor is used in the parallel circuit, in preference to a single unit, for the same reason. An alternative scheme to maintain balance is to use two single-ended capacitors 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 capacitors may be switched in series when series tuning is to be used.

As an alternative to adjustable coupling between L_1 and L_2, fixed coupling may be used and a variable capacitor connected in series with L_2 as shown in Fig. 13-15.

These circuits should be set up and adjusted in the same way as the tapped matching circuit, Fig. 13-14. That is, an s.w.r. bridge should be used to indicate the impedance match, which is brought about by alternately adjusting C_1 and the coupling between L_1 and L_2 until the bridge shows a null.

In the event that there is difficulty in bringing the s.w.r. down to 1 to 1 in the coaxial link, the probable cause is that the input impedance of the transmission line is neither very high nor very low. In such a case, if series tuning does not

work it may pay to try parallel tuning, and vice versa. If a match cannot be secured with either, the circuit should be changed to that of Fig. 13-14.

Adjustment Without the S.W.R. Bridge

Use of the s.w.r. bridge with the circuits described above is the only certain way of arriving at optimum adjustments. However, if a bridge is not available, the transmitter usually can be made to take the proper load by a cut-and-try method of adjustment. In the case of Fig. 13-14, take a trial position of the taps fairly close to the center of L_1. With loose coupling between L_1 and L_2 (this may be controlled either by adjustment of the mutual inductance or by means of the series capacitor C_2) and with the amplifier plate tank circuit tuned to resonance as indicated by the plate-current dip, vary C_1 until a setting is found that causes the plate current to rise to a peak. This peak should be less than the expected normal loaded plate current. Then increase the coupling between L_1 and L_2, readjust C_1 for maximum plate current, and readjust the amplifier tank for the plate-current dip. Continue until the amplifier is fully loaded at the plate-current dip, increasing the coupling between the transmitter tank and the coax line if necessary to obtain full loading. Then spread the taps on L_1 a little farther apart and go through the same procedure. The object is to use the widest spread between taps that will permit proper loading of the transmitter.

The procedure with series or parallel tuning is similar except that there are no taps to adjust. If full loading cannot be secured with either, the circuit should be changed to Fig. 13-14.

Although this cut-and-try method generally will lead to adequate transmitter loading, the adjustments seldom are optimum from the standpoint of low s.w.r. in the coax link. This may lead to excessive power dissipation in the link, with overheating the result. Also, the loading may change more rapidly with small frequency changes than would be the case with a matching circuit adjusted for optimum performance with the aid of the s.w.r. bridge.

Lines of Random Length

Series or parallel tuning will always work satisfactorily with lines having a high standing-wave ratio so long as either a current loop or node occurs at the input end of the transmission line. This will be the case if the antenna is resonant and the line length is a multiple of one-quarter wavelength. However, it is not always possible to couple satisfactorily when intermediate line lengths are used. This is because at some lengths the input impedance of the line has a considerable reactive component, and because the resistive component is too large to be connected in series with a tuned circuit and too small to be connected in parallel.

The coupling system shown in Fig. 13-14 is capable of handling the resistive component of the input impedance of the transmission lines used in most amateur installations, regardless of the standing-wave ratio on the line. Consequently, it can generally be used wherever either series or parallel tuning would normally be called for, simply by setting the taps properly on the coil. (A possible exception is where the s.w.r. is considerably higher than 10 to 1 and the line length is such as to bring a current loop at the input end. In such a case the resistance may be only a few ohms, which is difficult to match by means of taps on a coil.)

Within limits, the same circuit is capable of being adjusted to compensate for the reactive component of the input impedance; this merely means that a 1-to-1 s.w.r. in the link will be obtained at a different setting of C_1 than would be the case if the line "looked like" a pure resistance. Sometimes, however, C_1 does not have enough range available to give complete compensation, particularly when (as is the case with some line lengths when the s.w.r. is high) the input impedance is principally reactive.

Under such conditions it is necessary, if the line length cannot be changed to a more satisfactory value, to provide additional means for compensating for or "canceling out" the reactive component of the input impedance. As described earlier in this chapter (Fig. 13-6) the input impedance can be considered to be equivalent to a circuit consisting either of resistance and inductance or resistance and capacitance. It is generally more convenient to consider these elements as a parallel combination, so if the line "looks like" $L'R'$ at A in Fig. 13-6, it is apparent that if we connect a capacitance of the right value across L' the circuit will become resonant and will appear to be a pure resistance of the value R'. Similarly, connecting an inductance of the right value across C' in Fig. 13-6B will resonate the circuit and the impedance will be equal to R'. The resistive impedance that remains can easily be matched to the coax link by means of the circuit of Fig. 13-14.

The practical application of this principle is shown in Fig. 13-17, where L and C are the reactances required to cancel out the line reactance, L for cases where the line is capacitive, C for lines having inductive reactance. The amount of either

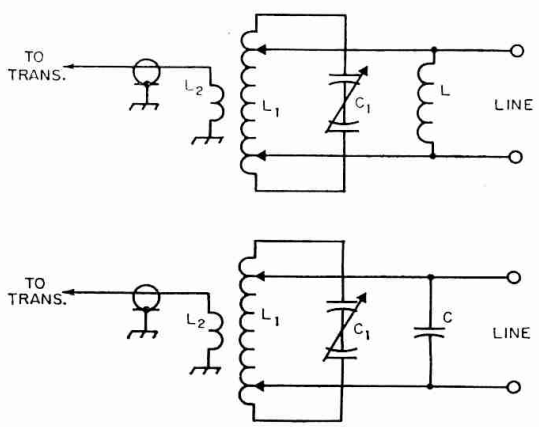


Fig. 13-17—Reactance cancellation on random-length lines having a high standing-wave ratio.

inductance or capacitance required is easily determined by trial, using the s.w.r. bridge in the coax link. First disconnect the main transmission line from L_1 and connect a noninductive resistor in its place. A 1-watt carbon resistor of about the same resistance as the line Z_0 will do, if a low-power bridge of the resistance type is used. With the "Monimatch" bridge, a suitable load may be made by connecting carbon resistors in parallel; for example, five 1500-ohm 2-watt resistors in parallel will make a 300-ohm load capable of handling 10 watts of r.f. Adjust the coil taps and C_1 for a 1-to-1 standing-wave ratio in the link, as described earlier. This determines the proper setting of C_1 for a purely resistive load. Then take off the resistor and connect the line, again adjusting the taps and C_1 to make the s.w.r. as low as possible, and compare the new setting of C_1 with the original setting. If the capacitance has increased, the line reactance is inductive and a capacitor must be connected at C in Fig. 13-17. The amount of capacitance needed to bring the proper setting of C_1 near the original setting can be determined by trial. On the other hand, if the capacitance of C_1 is less than the original, an inductance must be connected at L. Trial values will show when the proper tuning conditions have been reached.

It is not necessary that C_1 be at exactly the original setting after the compensating reactance has been adjusted; it is sufficient that it be in the same vicinity.

Using this procedure practically any length of line can be coupled properly to the transmitter, even when the line s.w.r. is quite high. Unfortunately, no specific values can be suggested for L and C, since they vary widely with Z_0, line length and s.w.r. Their values usually are comparable with the values used in the regular coupling circuits at the same frequency.

MATCHING TO COAXIAL LINES

Coaxial transmission lines usually are (or at least should be) operated at a low-enough standing-wave ratio so that no special matching circuits are needed; the line simply may be connected to the transmitter output terminals. A properly designed transmitter output circuit (see chapter on high-frequency transmitters) will be capable of handling variations in s.w.r. that are acceptable from the standpoint of line losses.

However, there are cases where it becomes necessary to provide some frequency selectivity between the transmitter and antenna system in order to prevent undesirable radiation of harmonics. A matching circuit of the same general type as those discussed above can provide a considerable degree of selectivity in addition to matching the input impedance of the transmission line to the Z_0 of the coaxial link. The difference in the circuit arrangement is simply that the secondary or output side need not be balanced with respect to ground.

Fig. 13-18 shows a typical circuit. Except for

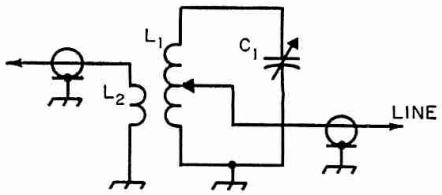


Fig. 13-18—Inductively coupled matching circuit for coupling between coaxial lines. The principles are the same as in Fig. 13-14; the secondary circuit is simply made single-ended for use with a coaxial transmission line.

the fact that there is only one coil tap, the design considerations and adjustment procedure are the same as described for Fig. 13-14. Also, the series capacitor, C_2, shown in Fig. 13-15 may be used with this circuit for fine variation of the effective coupling between L_1 and L_2. Constants for the circuit L_1C_1 are not critical; any convenient values that will tune to the operating frequency may be used. The Q of this circuit, and hence the selectivity, is controlled principally by the position of the line tap. As the tap is moved farther up the coil the Q and selectivity decrease.

The practical matching circuits described in the following section may be used with coaxial line simply by connecting the outer conductor of the line to the center of the coil and tapping the inner conductor along one side. The balanced circuit may still be used, although if the coupler is to be used only with coaxial line the circuit may be made single-ended as shown in Fig. 13-18.

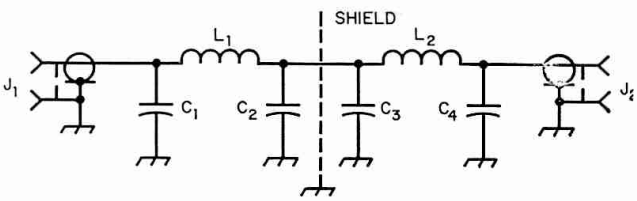


Fig. 13-19—Half-wave filter for harmonic suppression. The two sections of the filter should be shielded from each other as indicated by the dashed line, and the whole filter should be constructed in a shield enclosure to insure effective operation. A separate filter is required for each amateur band. All capacitors have the same value, as do all inductors, for a given band. Suggested constants are as follows:

Band	Capacitance	Inductance
3.5 Mc.	820 $\mu\mu$f.	2.2 μh.
7 Mc.	390 $\mu\mu$f.	1.3 μh.
14 Mc.	220 $\mu\mu$f.	0.57 μh.
21 Mc.	150 $\mu\mu$f.	0.375 μh.
28 Mc.	100 $\mu\mu$f.	0.3 μh.

Design is based on standard values of fixed mica capacitors. Larger capacitances may be made up by using smaller-capacitance units in parallel, if necessary. See text for voltage ratings. Inductances may be adjusted to proper value by resonating to center of band with the capacitance value given in the above table.

"Half-wave" Filters for Harmonic Suppression

If impedance matching is not a consideration — i.e., the transmission line to the antenna is operating at a low s.w.r. — but harmonic sup-

pression is desirable, the circuit of Fig. 13-19 may be used as an alternative to Fig. 13-18. This is a "half-wave" filter circuit, so called because it has similar properties to a half-wave transmission line. When inserted in a line, the impedance at the input terminals of the filter is the same impedance that the filter "sees" at its output terminals. Thus if the line input impedance is a pure resistance of 50 ohms, the impedance at the filter input terminals also will be 50 ohms.

Just as in the half-wave line case, the characteristic impedance of the filter can be any value without altering its performance with respect to input and output impedance. However, it is desirable in the interests of broad-band operation to make the filter characteristic impedance approximately the same as the Z_0 of the line. The constants given in Fig. 13-19 will serve for either 50- or 75-ohm line. The filter can be used without adjustment at any frequency within the amateur band for which it is designed.

The capacitance values required are fairly large, but under the assumed conditions (low s.w.r. on the line, filter Z_0 approximately equal to line Z_0) the voltages across the capacitors are low. Mica capacitors having a voltage rating suitable for the power level are satisfactory. The peak rating required is equal to $\sqrt{2PZ_0}$, where P is the r.f. power and Z_0 is the characteristic impedance of the line. This value should be doubled for 100 per cent amplitude modulation, and it is advisable to allow a safety factor in addition. A rating of 1500 volts d.c. will be sufficient for a kilowatt a.m. transmitter if the line is well matched by the antenna.

The attenuation of a filter of this type is about 30 db. at the second harmonic and greater at higher harmonics, until limited by self-resonances at high frequencies that occur in the inductors. These usually are not important at harmonics below the fourth.

Coupler or Matching-Circuit Construction

The design of matching or "antenna coupler" circuits has been covered in the preceding section, and the adjustment procedure also has been outlined. Since circuits of this type are most frequently used for transferring power from the transmitter to a parallel-conductor transmission line, a principal point requiring attention is that of maintaining good balance to ground. If the coupler circuit is appreciably unbalanced the currents in the two wires of the transmission line will also be unbalanced, resulting in radiation from the line.

In most cases the matching circuit will be built on a metal chassis, following common practice in the construction of transmitting units. The chassis, because of its relatively large area, will tend to establish a "ground" — even though not actually grounded — particularly if it is assembled with other units of the transmitter in a rack or cabinet. The components used in the coupler, therefore, should be placed so that they are electrically symmetrical with respect to the chassis and to each other.

In general, the construction of a coupler circuit should physically resemble the tank layouts used with push-pull amplifiers. In parallel-tuned circuits a split-stator capacitor should be used. The capacitor frame should be insulated from the chassis because, depending on line length and other factors, harmonic reduction and line balance may be improved in some cases by grounding and in others by not grounding. It is therefore advisable to adopt construction that permits either. Provision also should be made for grounding the center of the coil, for the same reason. The coil in a parallel-tuned circuit should be mounted so that its hot ends are symmetrically placed with respect to the chassis and other components. This equalizes stray capacitances and helps maintain good balance.

When the coupler is of the type that can be shifted to series or parallel tuning as required, two separate single-ended capacitors will be satisfactory. As described earlier, they should be connected so that both frames go to corresponding parts of the circuit — i.e., either to the coil or to the line — for series tuning, and when used in parallel for parallel tuning should be connected frame-to-stator.

A coupler designed and adjusted so that the connecting link acts as a matched transmission line may be placed in any convenient location. Some amateurs prefer to install the coupler at the point where the main transmission line enters the station. This helps maintain a tidy station lay-

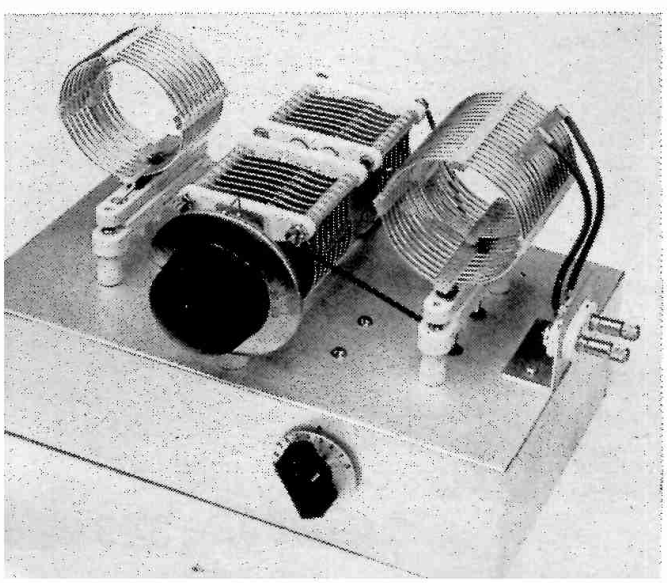


Fig. 13-20—Matching circuit for coupling balanced line to a coaxial link. It may also be used between two coaxial lines as described in the text. The coil at the left is simply "stored" on the chassis as a convenience for changing between two favorite bands. A "Monimatch" bridge is mounted under the 7 × 11 × 3 inch chassis.

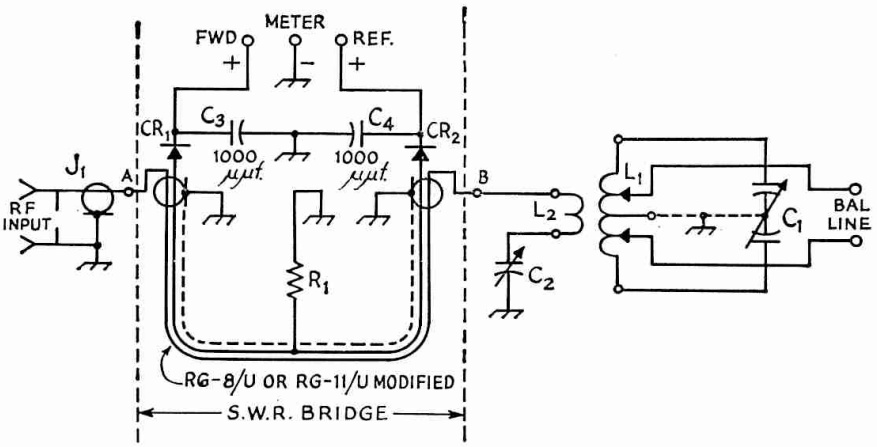


Fig. 13-21—Circuit of the coax-coupled matching circuit of Fig. 13-20. The s.w.r. bridge, a highly useful aid in adjustment, may be omitted if desired, in which case points *A* and *B* are simply connected together. See text for data on modified line.

C_1—100 $\mu\mu$f. per section variable, 0.075-inch spacing (Johnson 154-505).
C_2—700 to 800 $\mu\mu$f.; dual-section 365- to 400-$\mu\mu$f. broadcast-receiver type capacitor with sections in parallel.
C_3, C_4—0.001-μf. disk ceramic.
CR_1, CR_2—1N34A or equivalent.
J_1—Coax receptacle, chassis-mounting type.
L_1, L_2—See coil table.
R_1—See text.

out when an air-insulated parallel-conductor transmission line is used. With solid-dielectric lines, which lend themselves well to neat installation indoors, it is probably more desirable to install the coupler where it can be reached easily for adjustment and band-changing.

COAX-COUPLED MATCHING CIRCUIT

The matching unit shown in Fig. 13-20 is constructed according to the design principles outlined earlier in this chapter. It uses a parallel-tuned circuit with taps for matching a parallel-conductor line through a link coil to a coaxial line to the transmitter. It will handle about 500 watts of r.f. power and will work, without modification, into lines of any length if the s.w.r. is below 3 or 4 to 1. If the s.w.r. is high, it may be necessary to compensate for the reactive part of the input impedance of the line, at certain line lengths, by using an additional coil or capacitor as discussed earlier. The necessity for such compensation can be avoided, on lines having a high s.w.r., by making the electrical length of the line a multiple of a quarter wavelength.

As shown by the circuit diagram, Fig. 13-21, the link circuit is adjusted by means of a variable capacitor, C_2, to facilitate matching between the main transmission line from the antenna and the coax line to the transmitter. The coils are constructed from commercially available coil material, and the link (L_2) inductances are chosen to provide adequate coupling for flat lines. The link coil, of smaller diameter than the tank coil L_1, is mounted inside the latter at the center. Duco cement is used to hold the coils together at their bottom tie strips. The coils are mounted on Millen type 40305 plugs and require no other support than the stiffness of the short lengths of wire going into the end prongs of the plug from the tank coil. Short lengths of spaghetti tubing are slipped over the leads to the link coil where they go between the tank coil turns to reach the plug.

Taps on the tank coil for connection to a parallel-conductor transmission line are made by means of Johnson type 235-860 clips. If coils are changed frequently it will be convenient, after finding the proper tap points for each band, to bend ordinary soldering lugs around the wire and solder them in place so they project radially from the coil. The clips can then be adjusted to fit snugly over the lugs when pushed on sidewise. Used this way, the clips provide an easy and rapid method of connecting and disconnecting the line.

Monimatch

The circuit as shown in Fig. 13-21 includes a bridge or directional coupler of the Monimatch type to assist in adjusting the circuit to match the coax line. It is constructed from a 24-inch length of either RG-8/U or RG-11/U (depending on the Z_0 of the coax line between the transmitter and the matching circuit) as described in the section on measurements. The pickup line, to

Coil Data for Fig. 13-21

Band, Mc.	L_1				L_2			
	Turns	*Wire Size*	*Dia., In.*	*Turns/ In.*	*Turns*	*Wire Size*	*Dia., In.*	*Turns/ In.*
3.5	44	16	2½	10	10	16	2	10
7	18	12	2½	6	6	16	2	10
14	10	12	2½	6	3	16	2	10
21–28	6	12	2½	6	2	16	2	10

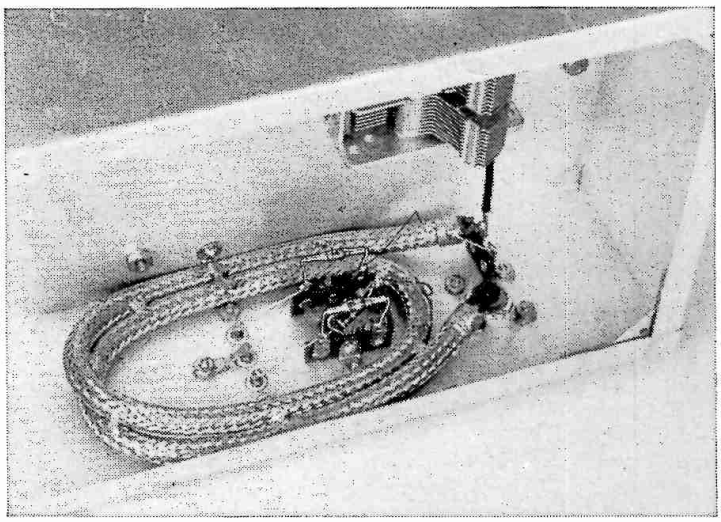


Fig. 13-22—Below-chassis view of the matching circuit, showing Monimatch made from a section of coax cable. The crystal rectifiers are mounted on dual tie-point strips, with R_1 between them.

which R_1 and the crystal rectifiers are connected, is a length of No. 30 enameled wire inserted between the insulation and the shield-braid outer conductor of the coax cable. In constructing this line section be careful not to scrape the enamel from the wire, and after the braid has been smoothed out to its original length check between it and the pickup wire with an ohmmeter to make sure the two are not short-circuited. The cable is formed into a double turn so that the center, where R_1 connects to the pickup wire, is close to the ends. This keeps the ground paths to minimum length and helps in obtaining proper balance in the bridge. The braided outsides of the turns are spot soldered together at several points to reduce the effect of unwanted currents on the surface, and also to improve the assembly mechanically.

Bridge Adjustment

Adjusting the bridge is simply a matter of finding the value of R_1 that gives a good null reading with the indicating meter connected to the "reflected" position when the output end is terminated in a resistive load of either 52 or 75 ohms, depending on whether RG-8/U or RG-11/U is used. If a suitable dummy load is available (see chapter on measurements) the wiring to L_2 should be disconnected at B in Fig. 13-21 and the dummy load connected between B and ground (that is, to the output terminals of the Monimatch). R_1 may be set to the proper value by trying several values of half-watt carbon resistors, or combinations in parallel, to find the resistance that gives the deepest null. A value of about 35 ohms proved to be optimum with RG-8/U in the bridge shown in the photograph.

Alternatively, a dummy load may be connected to the balanced line terminals, and the Monimatch disconnected at B. If a suitable bridge can be borrowed, it can be connected at B and r.f. power fed through it to the matching circuit, which should then be adjusted to match the coax line. This establishes a load of known value which may then be used for adjustment of the built-in Monimatch as described above, after the connection at B has been restored.

A suitable indicator unit, including meter, variable resistor, and forward-reflected switch, is described in the chapter on measurements.

Matching-Circuit Adjustment

The method of adjusting a matching circuit of this type has been described earlier in this chapter in connection with Figs. 13-14 and 13-15. The construction is such that either the center tap of L_1 or the rotor of C_1 may be grounded to the chassis, since C_1 is mounted on small stand-off insulators. Insofar as normal balanced-line operation is concerned, it makes no difference which is grounded (or neither). Grounding will, however, affect any parallel or "antenna" currents on the line. In general, the effect of such currents will be minimized if the ground connection showing the least r.f. current is chosen. This test should also be tried with and without an actual earth connection to the matching-circuit chassis.

The coupler may be used between coaxial lines by grounding the center tap of L_1 and connecting the outer braid of the coax line to the chassis and the inner conductor to a single tap on the coil. The method of adjustment is otherwise the same as for balanced lines.

The matching circuit should be adjusted with the aid of an s.w.r. bridge, as described earlier in this chapter. In general, the tuning will be less critical, and the circuit will work over a wider frequency range without readjustment, if the taps are kept as far toward the ends of the coil as possible and C_2 is set at the largest capacitance that will permit bringing the s.w.r. in the coax link down to 1 to 1.

● ANTENNA MATCHING CIRCUIT FOR HIGH OR LOW IMPEDANCE

The unit shown in Figs. 13-23 and 13-25 can be used to match the coaxial-line output of a transmitter to either a high- or low-impedance load. To facilitate tuning it includes an s.w.r. indicator that can be set for a wide range of power levels. The power-handling ability of a circuit of this type will depend to some extent upon the imped-

Coupler Construction

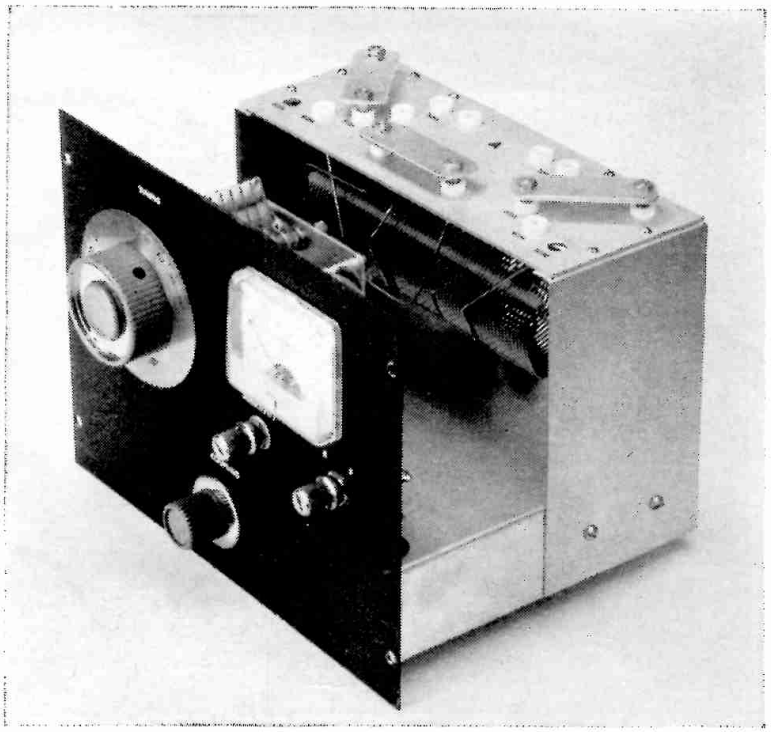


Fig. 13-23—Antenna coupler out of its case. The large dial controls a 100-$\mu\mu$f. tuning capacitor, and the smaller dial (bottom center) turns a 320-$\mu\mu$f. coupling capacitor. Two knobs control the sensitivity and direction of the s.w.r. bridge. Simple band switches on top of the aluminum arch are made from banana plugs and insulated jacks

ance of the load, but as shown the matching circuit will handle up to 300 or 400 watts under practically any condition. If higher power is involved, the circuit can be "scaled upward" with heavier inductances and greater capacitor spacings.

Referring to the circuit in Fig. 13-24, a series-tuned circuit, L_3C_1, is coupled to a balanced circuit, $C_2L_2L_4$. This latter circuit is series-tuned if the load is connected to terminals A-A and parallel-tuned if a jumper is used between A-A and the load is connected at B-B. Low-impedance loads (high-current) call for series tuning, and high-impedance loads (high voltage) couple better with parallel tuning.

A simple version of the "Monimatch" s.w.r. indicator is included by wrapping the necessary length of RG-58/U around the indicating meter (see Fig. 13-25).

The unit shown here was built on a 7 × 9 × 2-inch aluminum chassis, but dimensions are not critical so long as the inductance is not crowded against the metal parts of the chassis or housing. Capacitor C_2 is insulated from the chassis and panel by using small stand-off insulators for its support and a ceramic insulating shaft coupling.

The switches S_1 and S_2 are made from nylon-insulated banana jacks (Johnson 108-901) mounted on an arch of $\frac{3}{32}$-inch sheet aluminum. In each switch one jack serves as the rotor

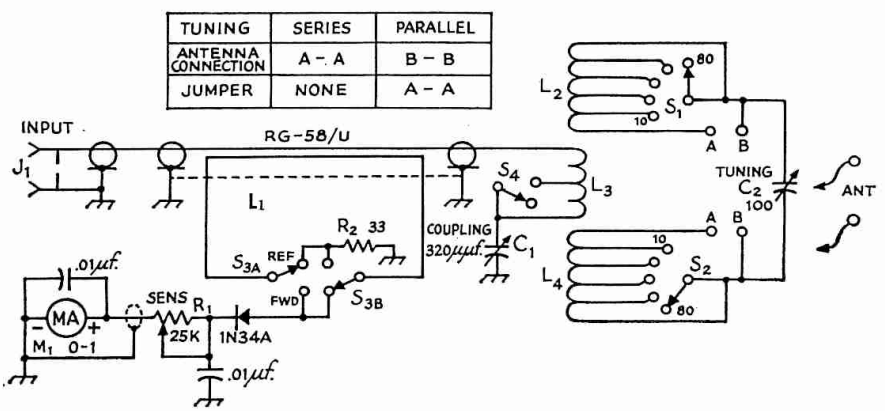


Fig. 13-24—Circuit diagram of the antenna coupler.

C_1—320-$\mu\mu$f. midget variable (Hammarlund MC-325-M).
C_2—100-$\mu\mu$f. tuning, .077-inch spacing (National TMC-100).
J_1—Coaxial receptacle, type SO-239.
L_1—Wire inside coaxial line. See text.
L_2, L_3, L_4—See Fig. 13-26.
M_1—0-1 milliammeter (Triplett 227-PL).
R_1—25,000-ohm volume control (Mallory U-28).
R_2—33 ohms, ½ watt. Must be composition, not wirewound.
S_1, S_2—See text.
S_3—D.p.d.t. rotary switch (Centralab 1462).
S_4—Tap on L_3, shorted to end of coil by copper test clip (Mueller 45C).

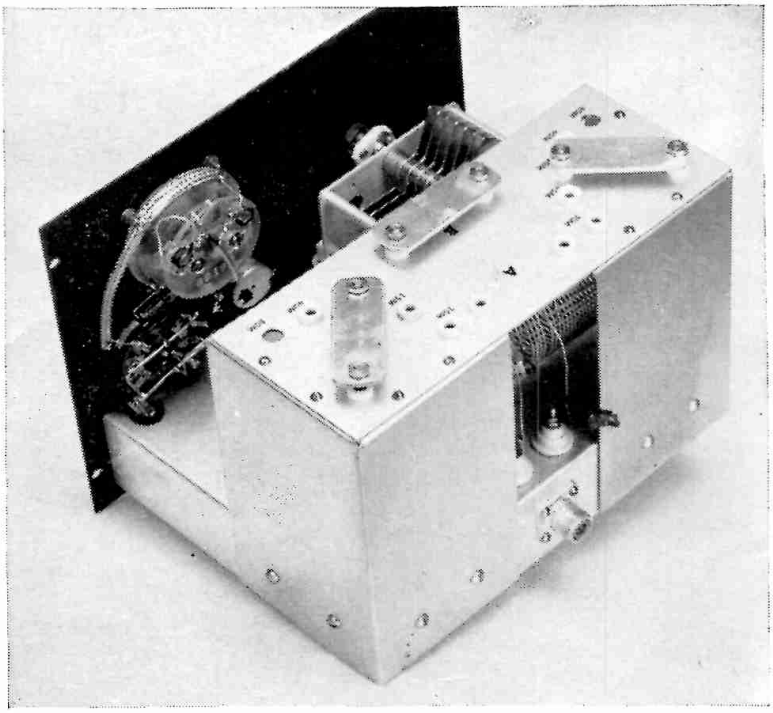


Fig. 13-25—Rear view of the coupler shows the coaxial line of the s.w.r. indicator wrapped around the meter. The test clip of S_4 is parked on one of the feedthrough insulators for L_3. Shorting bar in B-B (center) only for photograph; it is used only in A-A.

terminal and the others serve as the contacts. A shorting bar of aluminum with two banana plugs (Johnson 108-750) mounted on it at the proper distance is used as the switch arm. The shorting bar for the A-A connection is made similarly. Two feedthrough insulators at the rear of the cabinet (Bud C-1746) are used as antenna terminals; flexible leads connected to them have banana plugs at the other end to connect to A-A or B-B as required.

The inductors L_2, L_3 and L_4 are made from a length of 2-inch diameter transmitting coil stock, as indicated in Fig. 13-26. While the over-all sizes of the coils will suffice for practically any installation, it is suggested that the taps be made temporarily until the unit can be tested with the antenna to be used. The taps as indicated will be correct for most cases, but variations in antenna systems will account for some discrepancies. The inductors are supported by their leads from the banana jacks. Switch S_4 is merely two solder lugs on the proper wires; they can be shorted together by clipping them with a copper test clip. (It is recommended that screws and hardware be tested with a magnet before using near the coils; iron will get hot in the fields surrounding the coils.)

The s.w.r. bridge is made by first peeling the vinyl outer coating from a 3½-foot length of RG-58/U. Measure 6½ inches out either side of the center and open the shield braid slightly with a pointed tool. Thread a length of insulated wire (No. 22 or 24) in one hole and out the other, being careful not to scratch off the insulation of the wire; test with an ohmeter to make sure. Smooth out the shield braid on the RG-58/U and wrap the coaxial line for two turns around the meter housing. The coaxial line can then be threaded through a rubber grommet in the chassis and led to J_1 and the feedthrough from L_3, both at the rear of the chassis. The length of insulated wire, L_1, will have its ends conveniently situated for soldering to S_3.

In operation, the antenna feed line can be connected for series tuning if coaxial line is used and for parallel tuning if open-wire line is used. This is not an iron-clad rule, however, particularly when a high s.w.r. exists on the line to the antenna. Capacitors C_1 and C_2 are then adjusted for minimum reflected reading and maximum forward reading of M_1. If the maximum reading tends to send the meter off scale, increase the resistance at R_1. If the reflected reading cannot be brought down to a very low value, it may be necessary to try the opposite series/parallel connection or, as mentioned earlier, to change the location of the taps on L_2 and L_4.

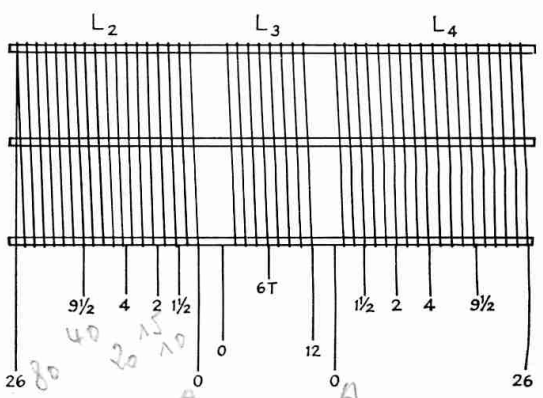


Fig. 13-26—Details of coil tapping. Material is No. 16 wound 10 t.p.i. on 2-inch diameter (B&W 3907-1). Half turns peeled off between L_2-L_3 and L_3-L_4 to give one-turn separation. Tap placement may vary somewhat with antenna system.

CHAPTER 14

Antennas

An *antenna system* can be considered to include the antenna proper (the portion that radiates the r.f. energy), the feed line, and any coupling devices used for transferring power from the transmitter to the line and from the line to the antenna. Some simple systems may omit the transmission line or one or both of the coupling devices. This chapter will describe the antenna proper, and in many cases will show popular types of lines, as well as line-to-antenna couplings where they are required. However, it should be kept in mind that *any* antenna proper can be used with *any* type of feedline if a suitable coupling is used between the antenna and the line. Changing the line does not change the type of antenna.

Selecting an Antenna

In selecting the type of antenna to use, the majority of amateurs are somewhat limited through space and structural limitations to simple antenna systems, except for v.h.f. operation where the small space requirements make the use of multielement beams readily possible. This section will consider antennas for frequencies as high as 30 Mc. — a later chapter will describe the popular types of v.h.f. antennas. However, even though the available space may be limited, it is well to consider the propagation characteristics of the frequency band or bands to be used, to insure that best possible use is made of the available facilities. The propagation characteristics of the amateur-band frequencies are described in Chapter Fifteen. 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 vertical angle of radiation and the plane of polarization may be of relatively little importance; at 28 Mc. they may be all-important.

Definitions

The **polarization** of a straight-wire antenna is determined by its position with respect to the earth. Thus a vertical antenna radiates vertically polarized waves, while a horizontal antenna radiates horizontally polarized waves in a direction broadside to the wire and vertically polarized waves at high vertical angles off the ends of the wire. The wave from an antenna in a slanting position, or from the horizontal antenna in directions other than mentioned above, contains components of both horizontal and vertical polarization.

The **vertical angle of maximum radiation** of an antenna is determined by the free-space pattern of the antenna, its height above ground, and the nature of the ground. The angle is measured in a vertical plane with respect to a tangent to the earth at that point, and it will usually vary with the horizontal angle, except in the case of a simple vertical antenna. The **horizontal angle of maximum radiation** of an antenna is determined by the free-space pattern of the antenna.

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 to the line offered by the antenna. It can be either resistive or complex, depending upon whether or not the antenna is resonant.

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.

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. The comparison antenna is generally a half-wave antenna at the same height and having the same polarization as the antenna under consideration. Gain usually is expressed in decibels.

In unidirectional beams (antennas with most of the 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.

The **bandwidth** of an antenna refers to the frequency range over which a property falls within acceptable limits. The **gain bandwidth,** the **front-to-back-ratio bandwidth** and the **standing-wave-ratio bandwidth** are of prime interest in amateur work.

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 height of the antenna above ground is a major factor in determining the 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

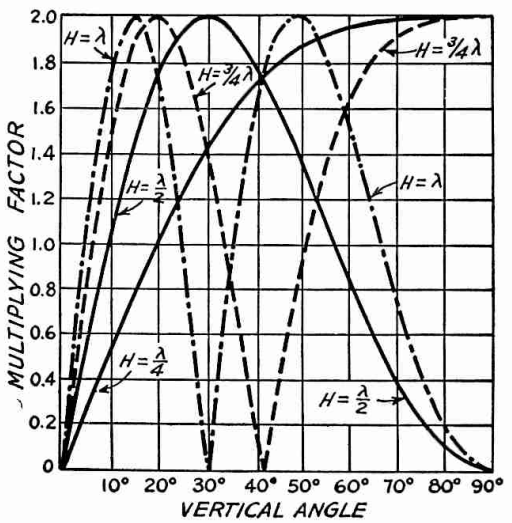


Fig. 14-1—Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly conducting ground.

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. 14-1 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 of maximum 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 angles usually are most effective, this generally means that the antenna should be high — at least one-half wavelength at 14 Mc., and preferably three-quarters or one wavelength, and at least one wavelength, and preferably higher, at 28 Mc. 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 useful 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. 14-1 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 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 an-

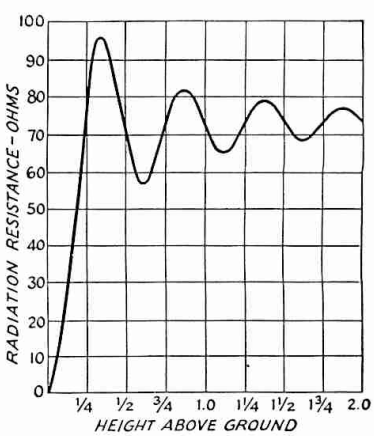


Fig. 14-2—Theoretical curve of variation of radiation resistance for a very thin half-wave horizontal antenna as a function of height in wavelength above perfectly reflecting ground.

tenna 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 a very thin half-wave antenna above perfectly reflecting ground is shown in Fig. 14-2. 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 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

A 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 known as a **dipole antenna.**

The length of a half-wave in space is:

$$Length\ (\text{feet}) = \frac{492}{Freq.\ (\text{Mc.})} \qquad \textbf{(14-A)}$$

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. 14-3, 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.:

$$Length\ of\ half\text{-}wave\ antenna\ (\text{feet}) = \frac{492 \times 0.95}{Freq.\ (\text{Mc.})} = \frac{468}{Freq.\ (\text{Mc.})} \qquad \textbf{(14-B)}$$

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. 14-3.

$$Length\ of\ half\text{-}wave\ antenna\ (\text{feet}) = \frac{492 \times K}{Freq.\ (\text{Mc.})} \qquad \textbf{(14-C)}$$

$$or\ length\ (\text{inches}) = \frac{5905 \times K}{Freq.\ (\text{Mc.})} \qquad \textbf{(14-D)}$$

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. 14-A. Ratio of half wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2} = 101.8$. From Fig. 14-3, $K = 0.963$ for this ratio. The length of the antenna, from Eq. 14-C, 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. 14-D: $\frac{5905 \times 0.963}{29} = 196$ inches.

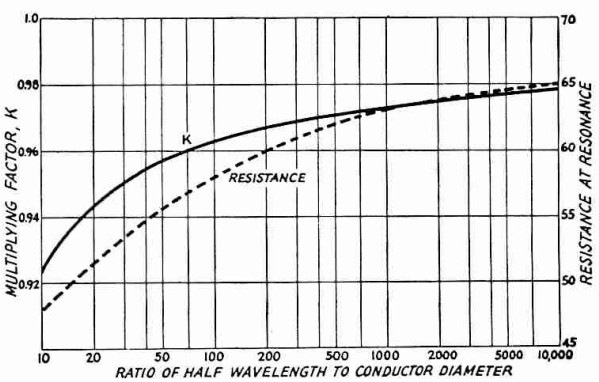


Fig. 14-3—Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half wavelength (Equation 14-A). The effect of conductor diameter on the center impedance also is shown.

Current and Voltage Distribution

When power is fed to an antenna, the current and voltage vary along its length. The current is maximum **(loop)** at the center and nearly zero **(node)** 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

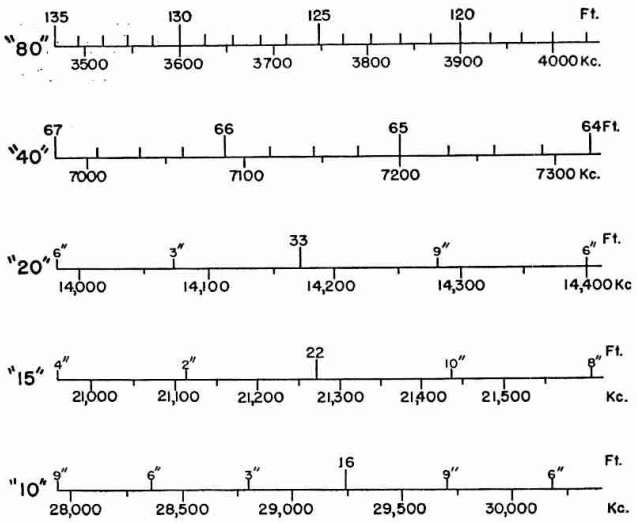


Fig. 14-4—The above scales, based on Eq. 14-B, can be used to determine the length of a half-wave antenna of wire.

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, compared 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 is about 73 ohms. The value under practical conditions is commonly taken to be in the neighborhood of 60 to 70 ohms, although it varies with height in the manner of Fig. 14-2. It 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 indicated in Fig. 14-3. If the diameter of the conductor is increased 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 dipole 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. 14-5, 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, 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. 14-16 and 14-7.

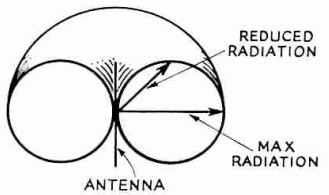


Fig. 14-5—The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position, and the actual "doughnut" pattern is cut in half to show how the line from the center of the antenna to the surface of the pattern varies. In practice this pattern is modified by the height above ground and if the antenna is vertical or horizontal. Fig. 14-1 shows some of the effects of height on the vertical angle of radiation.

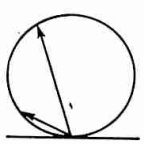


Fig. 14-6—Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Off the end, the radiation is greater at higher angles. Ground reflection is neglected in this drawing of the free-space pattern of a horizontal antenna.

● FEEDING A DIPOLE 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. 14-8. In the method shown at A, C_1 and C_2 should be about 150 $\mu\mu$f. each for the 3.5-Mc. band, 75 $\mu\mu$f. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil connected between them should resonate to 3.5 Mc. with about 60 or 70 $\mu\mu$f., for the 80-meter band, for 40 meters it should resonate with 30 or 35 $\mu\mu$f., 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 indi-

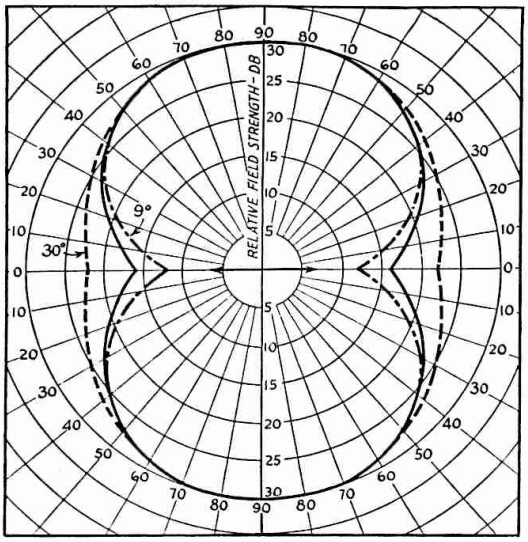


Fig. 14-7—Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will 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.

cated 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 re-resonate the transmitter tank circuit as the coupling is increased, but the change should be small.

The circuits in Fig. 14-8B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the

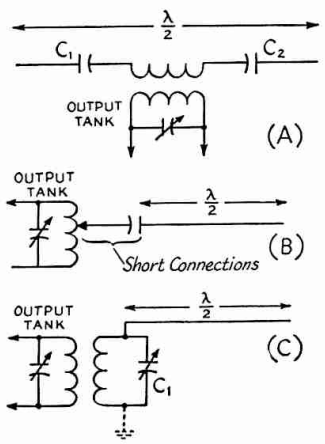


Fig. 14-8—Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitive coupling; C, voltage feed, with inductively coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper. Link coupling can be used in A and C.

tap toward the "hot" or plate end of the tank coil — the series capacitor 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.

Transmission-Line Feed for Dipoles

Since the impedance at the center of a dipole is in the vicinity of 70 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

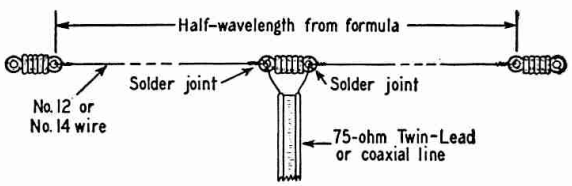


Fig. 14-9—Construction of a dipole fed with 75-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

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 14-B, 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. 14-9.

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 dipole**, a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 14-10. The open-wire line shown in Fig. 14-10 is made of No. 12 or No. 14 enameled wire, separated by

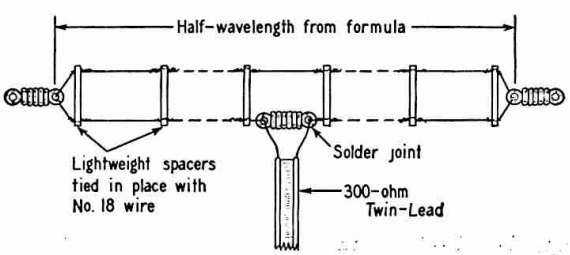


Fig. 14-10—The construction of an open-wire folded dipole fed with 300-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

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 spacing can be used at 3.5 Mc.

The half wavelength antenna can also be made from the proper length of 300-ohm line, opened on one side in the center and connected to the feedline. After the wires have been soldered together, the joint can be strengthened by molding some of the excess insulating material (polyethylene) around the joint with a hot iron, or a suitable lightweight clamp of two pieces of Lucite can be devised.

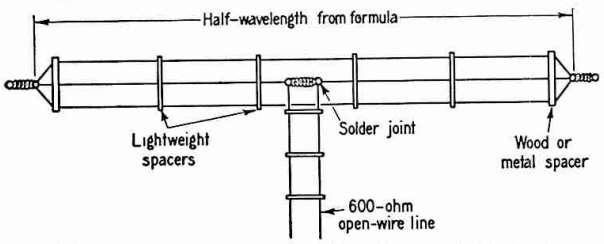


Fig. 14-11—The construction of a 3-wire folded dipole is similar to that of the 2-wire folded dipole. The end spacers may have to be slightly stronger than the others because of the greater compression force on them. The length of the antenna is obtained from Equation 14-B or Fig. 14-4. A suitable line can be made from No. 14 wire spaced 5 inches, or from No. 12 wire spaced 6 inches.

Similar in some respects to the two-wire folded dipole, the three-wire folded dipole of Fig. 14-11 offers a good match for a 600-ohm line. It is favored by amateurs who prefer to use an open-wire line instead of the 300-ohm insulated line. The three wires of the antenna proper should all be of the same diameter.

Another method for offering a match to a 600-ohm open-wire line with a half wavelength antenna is shown in Fig. 14-12. 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 **14-B** or Fig. 14-4. The length of section C is computed from:

$$C \text{ (feet)} = \frac{118}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-E)}$$

The feeder clearance, E, is found from

$$E \text{ (feet)} = \frac{148}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-F)}$$

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.

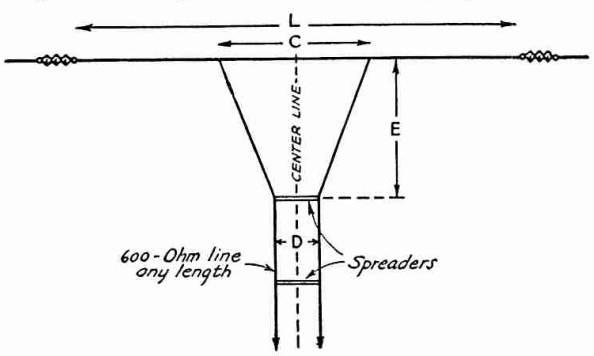


Fig. 14-12—Delta-matched antenna system. The dimensions C, D, and E are found by formulas given in the text. It is important that the matching section, E, come straight away from the antenna without any bends.

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires 5-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or 3¾-inch spaced No. 16 wire.

If a half wavelength antenna is fed at the center with other than 75-ohm line, or if a two-wire dipole 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 in Chapter 13. 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 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 boiled in paraffin are satisfactory. Mechanical details of half wavelength antennas fed with open-wire lines are given in Fig. 14-13. Regardless of the power level, solid-dielectric Twin-Lead is not recommended for this use.

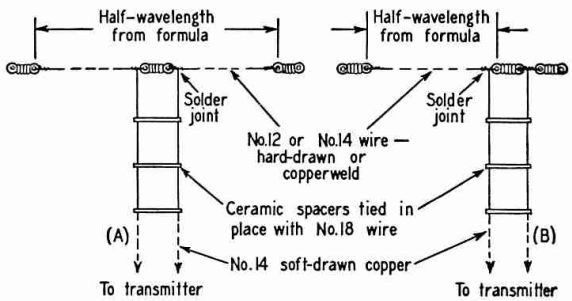


Fig. 14-13—The half-wave antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 14-B or Fig. 14-4.

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. 14-14 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is

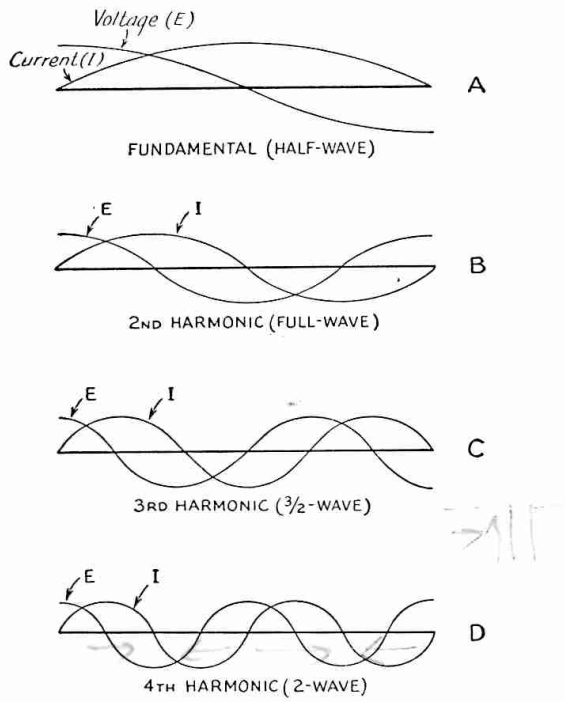


Fig. 14-14—Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.

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 adjacent 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.

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

$$Length\ (\text{feet}) = \frac{492\ (N-0.05)}{Freq.\ (\text{Mc.})} \qquad \textbf{14-G}$$

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 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

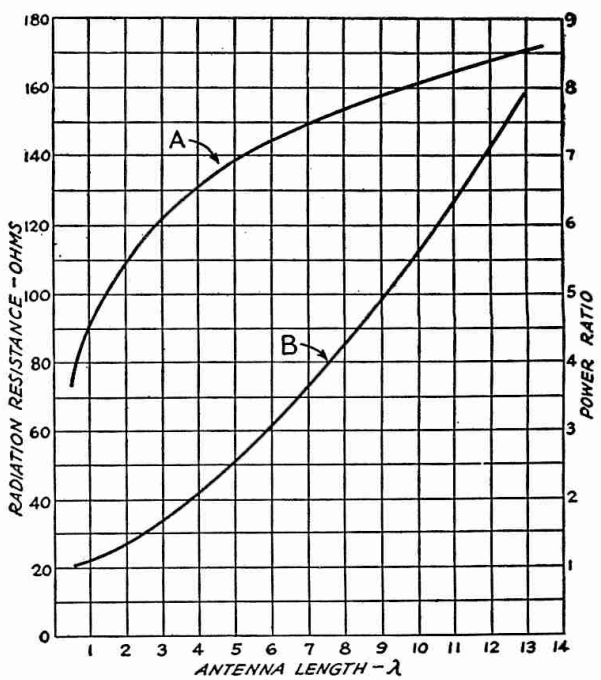


Fig. 14-15—Curve A shows variation in radiation resistance with antenna length. Curve B shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

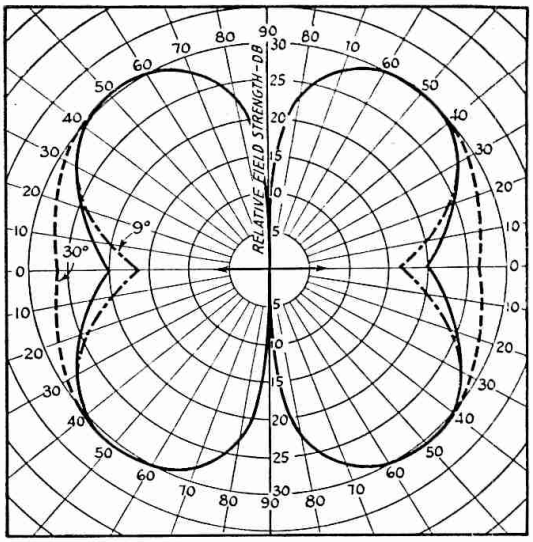


Fig. 14-16—Horizontal patterns of radiation from a full-wave antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale; actual amplitudes will depend upon the height of the antenna.

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 on all but one frequency in each harmonic range.

Impedance and Power Gain

The radiation resistance as measured at a current loop becomes higher as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable direction than does a half-wave antenna in its most favorable direction. This power gain is secured at the expense of radiation in other directions. Fig. 14-15 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

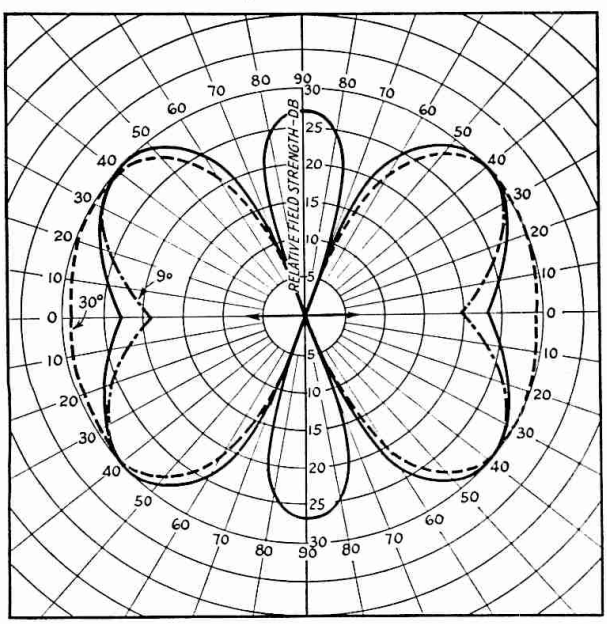


Fig. 14-17—Horizontal patterns of radiation from an antenna three half-waves long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

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. 14-16, 14-17 and 14-18, 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.

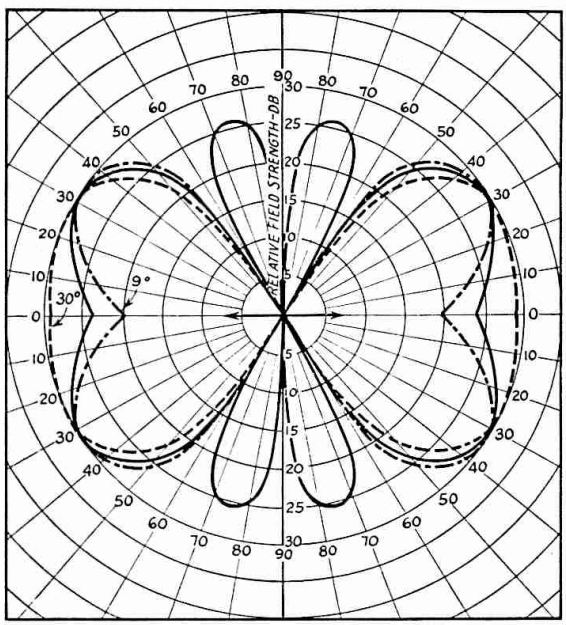


Fig. 14-18—Horizontal patterns of radiation from an antenna two wavelengths long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

Methods of Feeding

In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 14-14. The feeder system must not upset this phase relationship. This is satisfied 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. A long wire antenna is usually made a half wavelength at the lowest frequency and fed at the end.

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 tuned 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.

A dipole antenna that is center-fed by a solid-dielectric line is useless for even harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch causes a large standing-wave ratio and consequently high losses arise in the solid dielectric. It is wise not to attempt to use on its even harmonics a half-wave antenna center-fed with coaxial cable. On odd harmonics, as between 7 and 21 Mc.. a current loop will appear in the center of the antenna and a fair match can be obtained. High-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used in an emergency, provided the power does not exceed a few hundred watts, but it is an inefficient feed method.

When the same antenna is used for work in several bands, the directional characteristics 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 length of the antenna is a half wavelength. 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. 14-16, but if it is fed in the center the pattern will be somewhat similar to Fig. 14-7, 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 should be paid to the length of the feedline if convenient transmitter-coupling arrangements are to be obtained. Table 14-I gives some suggested antenna and feeder lengths for multiband operation. In general, the length of the feedline can be other than that indicated, but the type of coupling circuit may change.

Open-wire line feed is recommended for an antenna of this type, since the losses will run too high in solid-dielectric line. For low-power applications up to a few hundred watts, open-wire TV line is convenient and satisfactory to use. However, for high-power installations up to the kilowatt limit, an open-wire line with No. 14 or No. 12 conductors should be used. This can be built from soft-drawn wire and ceramic or other suitable spacers, or it can be bought ready-made.

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 will 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.

Tuned feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance.

TABLE 14-I
Multiband Tuned-Line-Fed Antennas

Antenna Length (Ft.)	*Feeder Length (Ft.)*	*Band*	*Type of Coupling Circuit*
With end feed:			
135	45	3.5 – 21 28	Series Parallel
67	45	7 – 21 28	Series Parallel
With center feed:			
135	42	3.5 – 21 28	Parallel Series
135	77½	3.5 – 28	Parallel
67	42½	3.5 7 – 28	Series Parallel
67	65½	3.5, 14, 28 7, 21	Parallel Series

Antenna lengths for end-fed antennas are approximate and should be cut to formula length at favorite operating frequency.

Where parallel tuning is specified, it will be necessary in some cases to tap in from the ends of the coil for proper loading — see Chapter 13 for examples of antenna couplers.

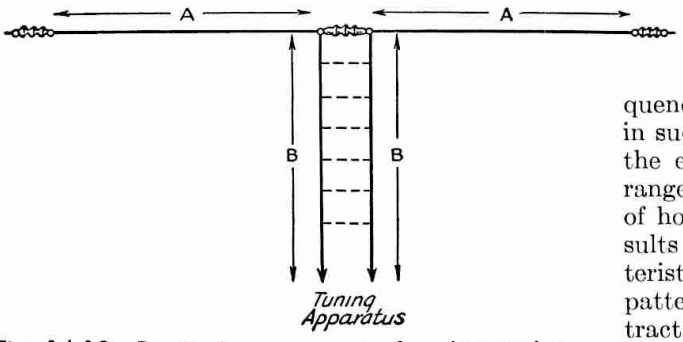


Fig. 14-19—Practical arrangement of a shortened antenna. When the total length, $A + B + B + A$, is the same as the antenna length plus twice the feeder length of the center-fed antennas of Table 14-I, the same type of coupling circuit will be used. When the feeder length or antenna length, or both, makes the sum different, the type of coupling circuit may be different but the effectiveness of the antenna is not changed, unless $A + A$ is less than a quarter wavelength.

With end feed the feeder currents become badly unbalanced.

With center feed, practically any convenient length of antenna can be used. If the total length of antenna plus twice feedline is the same as in Table 14-I, the type of tuning will be the same as stated. This is illustrated in Fig. 14-19. If the total length is not the same, different tuning conditions can be expected on some bands. This should not be interpreted as a fault in the antenna, and any tuning system (series or parallel) that works well without any trace of heating is quite satisfactory. Heating may result when the taps with parallel tuning are made too close to the center of the coil — it can often be corrected by using less total inductance and more capacitance.

Bent Antennas

Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a dipole 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 building 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. 14-20. 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. For one-band operation, end-loading with coils (5 feet or so in from each end) is practical and efficient.

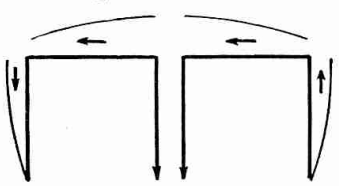


Fig. 14-20—Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

"Windom" or Off-Center-Fed Antenna

A multiband antenna that enjoyed considerable popularity in the 1930s is the "off-center feed" or "Windom," named after the amateur who wrote a comprehensive article about it. Shown in Fig. 14-21A, it consists of a half wavelength antenna on the lowest-frequency band to be used, with a *single-wire* feeder connected 14% off center. The antenna will operate satisfactorily

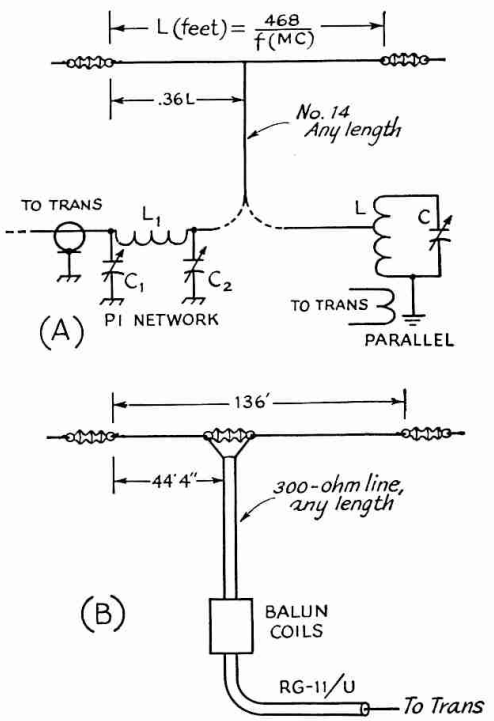


Fig. 14-21—Two versions of the off-center-fed antenna.

(A) Single-wire feed shows approximately 600 ohms impedance to ground and is most conveniently coupled to the transmitter as shown. The pi-network coupling will require more capacity at C_1 than at C_2. L_1 is best found by experiment—an inductance of about the same size as that used in the output stage is a good starting point. The parallel-tuned circuit will be a tuned circuit that resonates at the operating frequency with L and C close to those used in the output stage. The tap is found by experiment, and it should be as near the top of L as it can and still give good loading of the transmitter.

(B) Two-wire off-center feed uses 300-ohm TV line. Although the 300-ohm line can be coupled directly to some transmitters, it is common practice to step down the impedance level to 75 ohms through a pair of "balun" coils.

on the even-harmonic frequencies, and thus a single antenna can be made to serve on the 80-, 40-, 20-, and 10-meter bands. The single-wire feeder shows an impedance of approximately 600 ohms to ground, and consequently the antenna coupling system must be capable of matching this value to the transmitter. A tapped parallel-tuned circuit or a properly-proportioned pi-network coupler is generally used. Where TVI is a problem, the antenna coupler is required, so that a low-pass filter can be used in the connecting link of coaxial line.

Although theoretically the feed line can be of any length, some lengths will tend to give trouble with "too much r.f. in the shack," with the consequence that r.f. sparks can be drawn from the transmitter's metal cabinet and/or v.f.o. notes will develop serious modulation. If such is found to be the case, the feeder length should be changed.

A newer version of the off-center-feed antenna uses 300-ohm TV Twin-Lead to feed the antenna, as shown in Fig. 14-21B. It is claimed that the antenna offers a good match for the 300-ohm line on four bands and, although this is more wishful thinking than actual truth, the system is widely used and does work satisfactorily. It is subject to the same feed line length and "r.f.-in-the-shack" troubles that the single-wire version enjoys. However, in this case a pair of "balun" coils can be used to step down the impedance level to 75 ohms and at the same time alleviate some of the feed line troubles. This antenna system is popular among amateurs using multiband transmitters with pi-network-tuned output stages.

With either of the off-center-fed antenna systems, the feed line should run away from the antenna at right angles for as great a distance as possible before bending. No sharp bends should be allowed anywhere in the line.

Multiband Operation with Coaxial Line Feed

The proper use of coaxial line requires that the standing-wave ratio be held to a low value, preferably below 2:1. Since the impedance of an ordinary antenna changes widely from band to band, it is not possible to feed a simple antenna with coaxial line and use it on a number of bands without tricks of some kind. The single exception to this is the use of 75-ohm coaxial line to feed a 7-Mc. half-wave antenna, as in Fig. 14-19; this antenna can also be used on 21 Mc. and the s.w.r. in the line will not run too high.

One multiband antenna system that can be used by anyone without much trouble is shown in Fig. 14-22. Here separate dipoles are connected to one feedline. The 7-Mc. dipole also serves on 21 Mc. A low s.w.r. will appear on the feedline in each band if the dipoles are of the proper length. The antenna system can be built by suspending one set of elements from the one above, using insulator-terminated wood spreaders about one foot long. An alternative is to let one antenna droop several feet under the other, bring ropes attached to the insulators back to a common support point. It has been found that a separation of only an inch or two between dipoles is satisfactory. By using a length of the Twin-Lead used for folded dipoles (one Copperweld conductor and one soft-drawn), the strong wire can be used for the low-frequency dipole. The soft-drawn wire is then used on a higher band, supported by the solid dielectric.

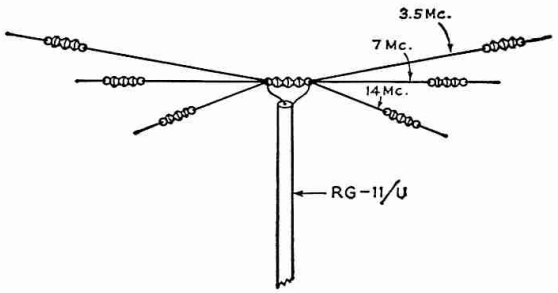


Fig. 14-22—An effective "all-band" antenna fed with a single length of coaxial line can be constructed by joining several half wavelength antennas at their centers and feeding them at the common point. In the example above, a low s.w.r. will be obtained on 80, 40, 20 and 15 meters. (The 7-Mc. antenna also works at 21 Mc.) If a 28-Mc. antenna were added, 10-meter operation could also be included.

The antenna lengths can be computed from formula 14-B. The shorter antennas can be suspended a foot or two below the longest one.

A vertical antenna can be operated on several bands and fed with a single length of coaxial line provided the antenna is no longer than 0.6 wavelength at the highest frequency and that a suitable matching network for each band is used at the base. A good radial or ground system is required. The matching sections can be housed in a weatherproof box and changed manually or by stepping relays; their form will vary from parallel-tuned circuits to L sections. (See McCoy, *QST*, December, 1955, for description of L-section coupler.)

Multiband "Trap" Antennas

Another approach to the problem of multiband operation with a single untuned feed line is the use of parallel-tuned circuits installed in the antenna at the right points to "divorce" the remainder of the antenna from the center section (part fed by coaxial line) as the transmitter is changed to a higher-frequency band. This principle of the divorcing circuits is utilized in a commercial "all-band" vertical antenna, and a 5-band kit for horizontal antennas is also available commercially. The divorcing circuits are also used in several commercial multiband beams for the 14-, 21- and 28-Mc. bands.

The multiband antenna system shown in Fig. 14-23 may be of interest to the ham who wishes to work on several bands but doesn't have sufficient space for an 80-meter antenna and consequently is limited to 40 meters and below. (A five-band antenna requires more than a 100-foot span; see Greenberg, *QST*, October, 1956.)

On 40 meters the traps serve as inductors to load the system to 7 Mc. On 20, the traps (resonant to 14.1 Mc.) divorce the B sections from the

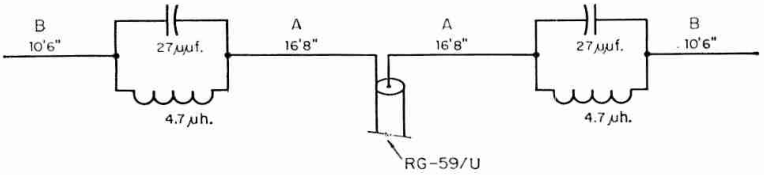


Fig. 14-23—Sketch showing dimensions of a trap dipole covering the 40-, 20- and 10-meter bands. The total span is less than 60 feet.

antenna proper. On 28 Mc. the entire antenna becomes approximately a 5/2-radiator.

As shown in Fig. 14-23, each trap is literally built around an "egg" or "strain" insulator. In this type of insulator, the hole at one end is at right angles to the hole at the other end, and the wires are fastened as in Fig. 14-25. These insulators have greater compressive strength than tensile strength and will not permit the antenna to fall should the insulator break, since the two interlooped wires prevent it. There is ample space within the inductor for both the insulator and capacitor. The plastic covers are not essential but are considered desirable because they provide mechanical protection and prevent the accumulation of ice or soot and tars which may not wash off the traps when it rains.

Electrically, each trap consists of a 25-$\mu\mu$f. capacitor shunted by 4.7 μh. of inductance. A Centralab ceramic transmitting capacitor 857-25Z, rated at 15,000 volts d.c., is shown and will safely handle a kilowatt. Other ceramic capacitors rated at approximately 6000 volts would be satisfactory, as well as cheaper. The inductors are made of No. 12 wire, 2½ inches in diameter, 6 turns per inch (B & W 3905-1 coil stock).

One may wish to choose a different frequency in the 20-meter band for which optimum results are desired; for example, 14.05 Mc. for c.w. operation, 14.25 Mc. for phone operation, or perhaps 14.175 Mc. for general coverage. In any case, the number of inductor turns is adjusted accordingly.

Trap Adjustment

As a preliminary step, loops of No. 12 wire are fitted to one of the egg insulators in the normal manner (see Fig. 14-25), except that after the wraps are made, the end leads are snipped off close to the wraps. A capacitor is then placed in position and bridged with short leads across the insulator and soldered sufficiently to provide temporary support. The combination is then slipped inside about 10 turns of the inductor, one end of which should be soldered to an insulator-capacitor lead. Adjustment to the resonant frequency can now proceed, using a grid-dip meter.

Coupling between the g.d.o. and the trap should be very loose. To insure accuracy, the station receiver should be used to check the g.d.o. frequency. The inductance should be reduced ¼ turn at a time. If one is careful, the resonant frequency can easily be set to within a few kilocycles of the chosen figure.

The reason for snipping the end leads close to the wraps and the inclusion of the loops through the egg insulator soon becomes apparent. The resonant frequency of the capacitor and inductor alone is reduced about 20 kc. per inch of end lead length and about 350 kc. by the insulator loops. The latter add approximately 2 $\mu\mu$f. to the fixed capacitor value and account for the total of 27 $\mu\mu$f. shown in Fig. 14-23.

Assembly

Having determined the exact number of inductor turns, the trap is taken apart and reassembled with leads of any convenient length. One may, of course, connect the entire lengths of sections *A* and *B* to the trap at this time, if desired. But, if more convenient, a foot or two of wire can be fastened and the remaining lengths soldered on just before the antenna is raised.

The protective covers are most readily formed by wrapping two turns (plus an overlap of ½ inch) of 0.020-inch polystyrene or lucite sheeting around a 3-inch plastic disk held at the center of the cylinder so formed. The length of the cover should be about 4 inches. A very small amount of plastic solvent (a cohesive cement that actually softens the plastic surfaces) should then be applied under the edge of the overlap and the joint held firmly for about

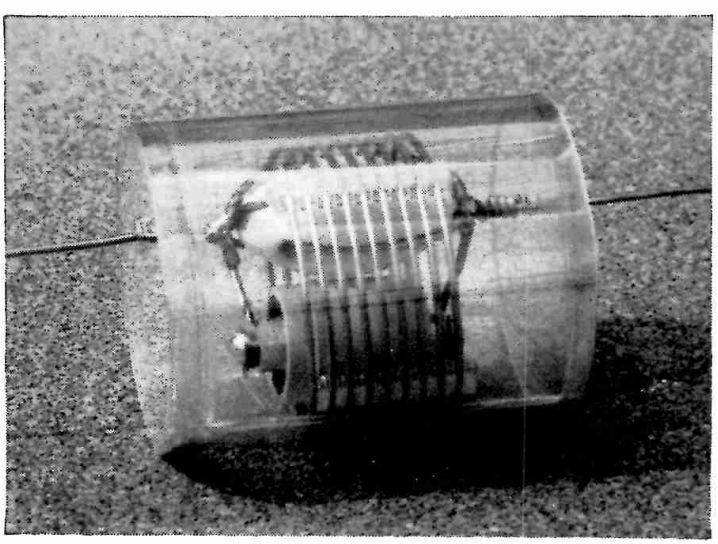


Fig. 14-24—The 14-Mc. trap is enclosed in a weatherproof cover made of plastic sheet. The ceramic capacitor and strain insulator are inside the coil.

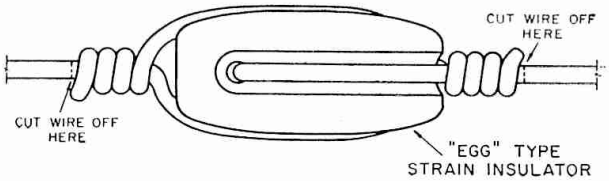


Fig. 14-25—Method of connecting the antenna wire to the strain insulator. The antenna wire is cut off close to the wrap before checking the resonant frequency of the trap.

two minutes to insure a strong, tight seal. The disk is pushed out and the inner seam of the sheeting sealed.

The trap is then placed in the plastic cylinder and the end disks marked where the antenna wires are to pass through. After drilling these holes, the disks are slipped over the leads, pressed into the ends of the cylinder and a small amount of solvent applied to the periphery to obtain a good seal. Some air can flow in and out of the trap through the antenna-wire holes, and this will prevent the accumulation of condensation.

Length Adjustment

Standing-wave ratios are not uniform throughout the band or bands for which an antenna is designed. In a trap antenna, the choice of frequencies for best performance is a compromise. After making the traps resonant at 14.1 Mc., sections *A* are adjusted for resonance. Sections *B* are then adjusted for resonance at approximately 7.2 Mc. For the dimensions shown, with the antenna about 250 ft. above street level and 35 ft. above electrical ground, an s.w.r. of virtually 1 to 1 was obtained at 7.2 Mc., with maximums of 1.3 and 1.1 at 7.0 and 7.3 Mc., respectively. In the 20-meter band, the s.w.r. was also 1 to 1 at 14.1 Mc., 1.1 at 14.0 Mc. and 1.3 at 14.3 Mc. In the 10-meter band, the s.w.r. was 1.3 to 1 at 28.0 Mc., 1.1 at 28.4 Mc., 1.5 at 29 Mc., and only 2.4 at the upper extreme of the band. The s.w.r. on 21 Mc. will be high because the antenna is not resonant in that band.

RG-59/U 73-ohm coaxial cable forms the transmission line and is connected to the antenna through a Continental Electronics & Sound Co. "Dipole Dri-Fit Connector." After connecting the cable and antenna wires, the connector should be coated with several layers of insulating varnish to make certain that the junction is watertight.

Vertical Antennas

A vertical quarter-wavelength antenna is often used in the low-frequency amateur bands to obtain low-angle radiation. It is also used when there isn't enough room for the supports for a horizontal antenna. For maximum effectiveness it should be located free of nearby objects and it should be operated in conjunction with a good ground system, but it is still worth trying where these ideal conditions cannot be obtained.

Four typical examples and suggested methods for feeding a vertical antenna are shown in Fig. 14-26. The antenna may be wire or tubing supported by wood or insulated guy wires. When tubing is used for the antenna, or when guy wires (broken up by insulators) are used to reinforce the structure, the length given by the formula is likely to be long by a few per cent. A check of the standing-wave ratio on the line will indicate the frequency at which the s.w.r. is minimum, and the antenna length can be adjusted accordingly.

A good ground connection is necessary for the most effective operation of a vertical antenna (other than the ground-plane type). In some cases a short connection to the cold-water system of the house will be adequate. But maximum performance usually demands a separate ground system. A single 4- to 6-foot ground rod driven into the earth at the base of the antenna is usually not sufficient, unless the soil has exceptional conductivity. A minimum ground system that can be depended upon is 6 to 12 quarter wavelength radials laid out as the spokes of a wheel from the base of the antenna. These radials can be made of heavy aluminum wire, of the type used for grounding TV antennas, buried at least

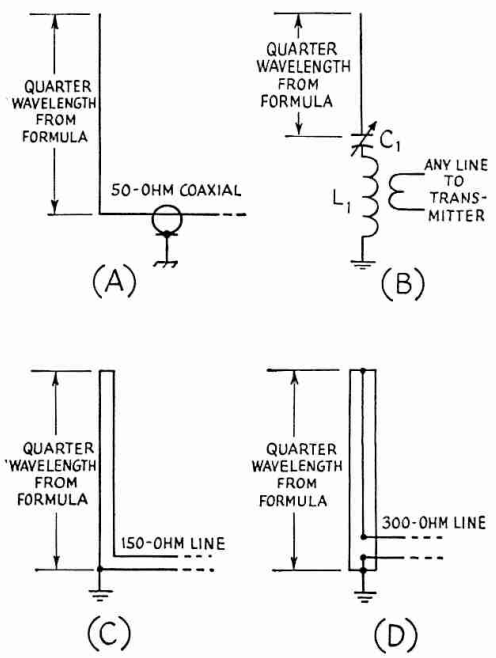


Fig. 14-26—A quarter-wavelength antenna can be fed directly with 50-ohm coaxial line (A) with a low standing-wave ratio, or a coupling network can be used (B) that will permit a line of any impedance to be used. In (B), L_1 and C_1 should resonate to the operating frequency, and L_1 should be larger than is normally used in a plate tank circuit at the same frequency. By using multiwire antennas, the quarter-wave vertical can be fed with (C) 150- or (D) 300-ohm line.

6 inches in the ground. This is normally done by slitting the earth with a spade and pushing the wire into the slot, after which the earth can be tamped down.

The examples shown in Fig. 14-26 all require an antenna insulated from the ground, to provide for the feed point. A *grounded* tower or pipe can be used as a radiator by employing "shunt feed," which consists of tapping the inner conductor of the coaxial-line feed up on the tower until the best match is obtained, in much the same manner as the "gamma match" (described later) is used on a horizontal element. If the antenna is not an electrical quarter wavelength long, it is necessary to tune out the reactance by adding capacity or inductance between the coaxial line and the shunting conductor. A metal tower supporting a TV antenna or rotary beam can be shunt-fed only if all of the wires and leads from the supported antenna run down the center of the tower and underground away from the tower.

● THE GROUND-PLANE ANTENNA

A ground-plane antenna is a vertical quarter-wavelength antenna using an artificial metallic ground, usually consisting of four rods or wires perpendicular to the antenna and extending radially from its base. Unlike the quarter-wavelength vertical antennas without an artificial ground, the ground-plane antenna will give low-angle radiation regardless of the height above actual ground. However, to be a true ground-plane antenna, the plane of the radials should be at least a quarter wavelength above ground. Despite this one limitation, the antenna is useful for DX work in any band below 30 Mc.

The vertical portion of the ground-plane antenna can be made of self-supported aluminum tubing or a top-supported wire depending upon the necessary length and the available supports. The radials are also made of tubing or heavy wire depending upon the available supports and necessary lengths. They need not be exactly symmetrical about the base of the vertical portion.

The radiation resistance of a ground-plane antenna varies with the diameter of the vertical element. Since the radiation resistance is usually in the vicinity of 30 to 32 ohms the antenna can be fed with 75-ohm coaxial line if a quarter wavelength matching section of 50-ohm coaxial line is used between the line and the antenna. (See "Quarter-Wave Transformers" in this chapter.)

For multiband operation, a ground-plane antenna can be fed with tuned open-wire line.

Three-Band Ground-Plane Antenna

A three-band ground-plane antenna using wire elements and fed with coaxial line is shown in Fig. 14-27. The builder (K5AYJ) elected to mount it on top of a 34-foot length of galvanized iron pipe, since a ground-plane antenna close to the ground is not a ground-plane antenna at all. Four 17-foot "drooping radials" form the ground plane and double as guy wires. These four wires are fastened to a pipe flange at the top of the mast. At one point on the mast the pipe sections are joined by a T fitting, which provides a convenient point for bringing out the RG-8/U feed line. If it is more convenient to bring out the coax at the base of the mast, one can eliminate the T fitting and use an ordinary coupling.

A cane fishing pole supports the three separate vertical elements. These elements, made of No. 12 wire, are taped to the pole every three inches with Scotch electrical tape. The bottom end of the pole is jammed tight into the upper end of the support pipe and the coaxial line is brought out of the pipe through a small hole just below the bottom of the flange. The inner conductor of the coaxial line is soldered to the junction of the three vertical elements and the braid of the coaxial line is connected to the pipe flange. Anyone worrying about the insulating ability of a cane pole can forget it; it is being used at a low-impedance point.

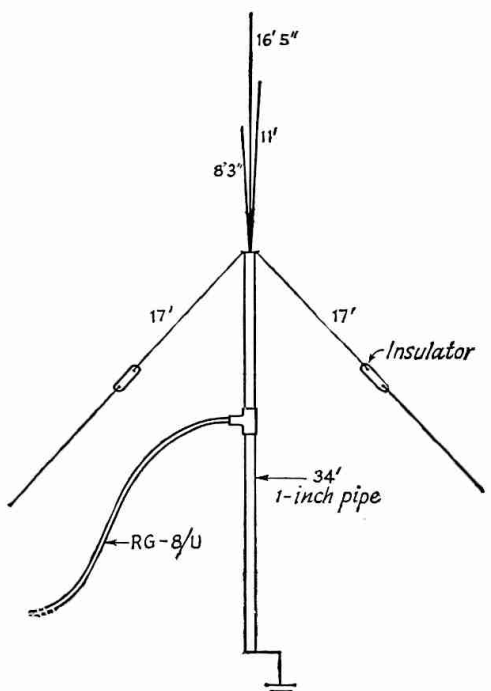


Fig. 14-27—The three-band ground-plane antenna uses wire elements. Vertical elements are taped to a cane pole; the four radials also serve as guy wires. The radials "droop" a little, making a 40-degree angle with the supporting 1-inch pipe.

Antennas for 160 Meters

Results on 1.8 Mc. will depend to a large extent on the antenna system and the time of day or night. Almost any random long wire that can be tuned to resonance will work during the night but it will generally be found very ineffective during the day. A vertical antenna — or rather an an-

tenna from which the radiation is predominantly vertically polarized — is probably the best for 1.8-Mc. operation. A horizontal antenna (horizontally-polarized radiation) will give better results during the night than the day. The vertically-polarized radiator gives a strong ground wave that is effective day or night, and it is to be preferred on 1.8 Mc.

The low-angle radiation from a horizontal antenna ⅛ or ¼ wavelength above ground is almost insignificant. Any reasonable height is small in terms of wavelength, so that a horizontal antenna on 160 meters is a poor radiator at angles useful for long distances ("long," that is, for this band). Its chief usefulness is over relatively short distances at night.

Bent Antennas

Since ideal vertical antennas are generally out of the question for practical amateur work, the best compromise is to bend the antenna in such a way that the high-current portions of the antenna run vertically. It is advisable to place the antenna so that the highest currents in the antenna occur at the highest points above actual ground. Two antenna systems designed along these lines are shown in Fig. 14-28. The antenna of Fig. 14-28B uses a full half wavelength of wire but is bent so that the high-current portion runs vertically. The horizontal portion running to L_1C_1 should run 8 or 10 feet above ground.

Grounds

A good ground connection is generally important on 160 meters. The ideal system is a number of wire radials buried a foot or two underground and extending 50 to 100 feet from the central connection point. The use of any less than six or eight radials is inadvisable.

If the soil is good (not rocky or sandy) and generally moist, a low-resistance connection to the cold-water pipe system in the house will often serve as an adequate ground system. The connection should be made close to where the pipe enters the ground, and the surface of the pipe should be scraped clean before tightening the ground clamp around the pipe.

A 6- or 8-foot length of 1-inch water pipe, driven into the soil at a point where there is considerable natural moisture, can be used for the ground connection. Three or four pipes driven into the ground 8 or 10 feet apart and all joined

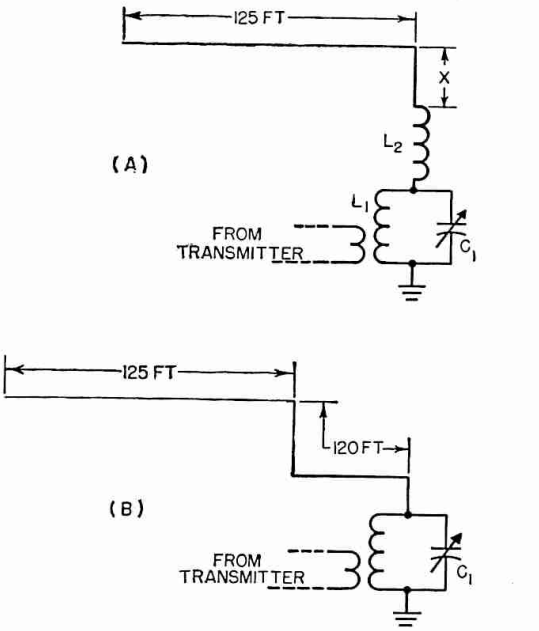


Fig. 14-28—Bent antenna for the 160-meter band. In the system at A, the vertical portion (length *X*) should be made as long as possible. In either antenna system, L_1C_1 should resonate at 1900 kc., roughly. To adjust L_2 in antenna A, resonate L_1C_1 alone to the operating frequency, then connect it to the antenna system and adjust L_2 for maximum loading. Further loading can be obtained by increasing the coupling between L_1 and the link.

together at the top with heavy wire are more effective than the single pipe.

The use of a counterpoise is recommended where a buried system is not practicable or where a pipe ground cannot be made to have low resistance because of poor soil conditions. A counterpoise consists of a number of wires supported from 6 to 10 feet above the surface of the ground. Generally the wires are spaced 10 to 15 feet apart and located to form a square or polygonal configuration under the vertical portion of the antenna.

Long-Wire Directive Arrays

As the length (in wavelengths) of an antenna is increased, the lobes of maximum radiation make a more acute angle with the wire. Two long wires can be combined in the form of a horizontal "V", in the form of a horizontal rhombus, or in parallel, to provide a long-wire directive array. In the "V" and rhombic antennas the main lobes reinforce along a line bisecting the acute angle between the wires; in the parallel antenna the reinforcement is along the line of the lobe. This reinforcement provides both gain and directivity along the line, since the lobes in other directions tend to cancel out. In general, the power gain depends upon the length in wavelengths of the wires, assuming that the proper configuration for a given length and height above ground is used.

Rhombic and "V" antennas are normally bidirectional along the bisector line mentioned above. They can be made unidirectional by terminating the ends of the wires away from the feed point in the proper value of resistance. When properly terminated, "V" and rhombic antennas of sufficient length work well over a three-to-one or four-to-one frequency range and hence are useful for multiband operation.

Antenna gains of the order of 10 to 15 db. can

be obtained with properly-constructed long-wire arrays. However, the pattern is rather sharp with gains of this order, and rhombic and "V" beams are not used by amateurs as commonly as they were, having been displaced by the rotatable multi-element Yagi beam. Further information on these antennas can be found in *The ARRL Antenna Book*.

Beams 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 radiation from the elements add up along a single direction and form a beam. In other directions the radiation tends to cancel, so a power gain is obtained in one direction at the expense of radiation in other directions. 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 **broad-side** when the phase of the current is the same in all, and **end-fire** when the currents are not in phase.

Collinear Arrays

Simple forms of collinear arrays, with the current distribution, are shown in Fig. 14-29.

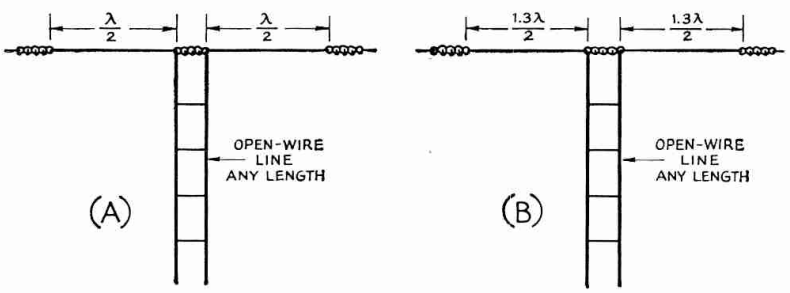


Fig. 14-29—Collinear antennas in phase. The system at A is known as "two half waves in phase" and has a gain of 1.8 db. over a half-wave antenna. By lengthening the antenna slightly, as in B, the gain can be increased to 3 db. Maximum radiation is at right angles to the antenna. The antenna at A is sometimes called a "double Zepp" antenna, and that at B is known as an "extended double Zepp."

The two-element array at A is popularly known as "two half-waves in phase" or a **double Zepp** antenna. It will be recognized as simply a center-fed dipole operated at its second harmonic.

By extending the antenna, as at B, the additional gain of an **extended double Zepp** antenna can be obtained. Carrying the length beyond that shown will result in an "X"-shaped pattern that no longer has the maximum radiation at right angles to the wire.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mounting gives increased horizontal directivity, while the vertical directivity remains the same as for a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles.

Broadside Arrays

Parallel antenna elements with currents in phase may be combined as shown in Fig. 14-30 to form a **broadside** array, so named because the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the spacing of the elements.

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 tuned open-wire lines or through quarter-wave matching sections and flat lines. In Fig. 14-30B, note the "crossing over" of the phasing section, which is necessary to bring the elements into proper

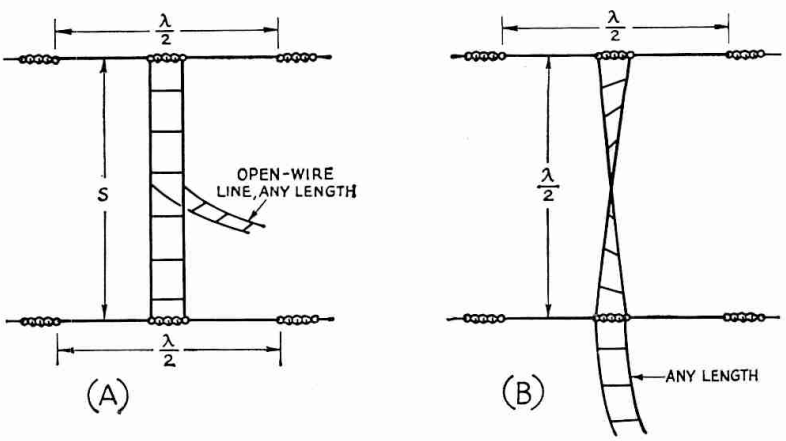


Fig. 14-30—Simple broadside array using horizontal elements. By making the spacing *S* equal to ⅜ wavelength, the antenna at A can be used at the corresponding frequency and up to twice that frequency. Thus when designed for 14 Mc. it can also be used on 21 and 28 Mc. The antenna at B can be used on only the design band. This array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page). Gain varies with the spacing *S*, running from 2½ to almost 5 db. (See Fig. 14-32).

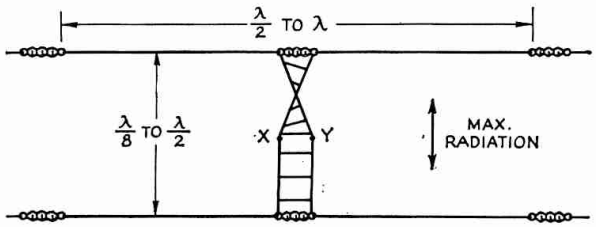


Fig. 14-31—Top view of a horizontal end-fire array. The system is fed with an open-wire line at *x* and *y*; the line can be of any length. Feed points *x* and *y* are equidistant from the two insulators, and the feed line should drop down vertically from the antenna. The gain of the system will vary with the spacing, as shown in Fig. 14-32, and is a maximum at ⅛ wavelength. By using a length of 33 feet and a spacing of 8 feet, the antenna will work on 20, 15 and 10 meters.

phase relationship.

End-Fire Arrays

Fig. 14-31 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 plane of the antennas, as shown.

The end-fire array may be used either vertically or horizontally (elements at the same height), and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 14-32 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.

Either tuned or untuned lines may be used with this type of array. Untuned lines preferably are matched to the antenna through a quarter-wave matching section or phasing stub.

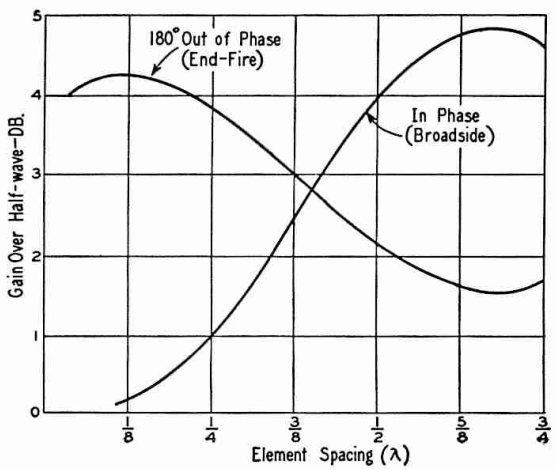


Fig. 14-32—Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

Combined Arrays

Broadside, collinear and end-fire arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. 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.

Although arrays can be fed at one end as in Fig. 14-30B, it 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.

A four-element array, known as the **"lazy-H"** antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 14-33. (Compare with Fig. 14-30B). For best results, the bottom section should be at least a half wavelength above ground.

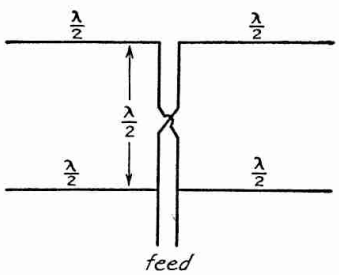


Fig. 14-33—A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into an untuned transmission line, or tuned feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

It will usually suffice to make the length of each element that given by Equations **14-B** or **14-C.** The phasing line between the parallel elements should be of open-wire construction, and its length can be calculated from:

Length of half-wave line (feet) =

$$\frac{480}{Freq.\ (\text{Mc.})} \qquad \textbf{(14-H)}$$

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.

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

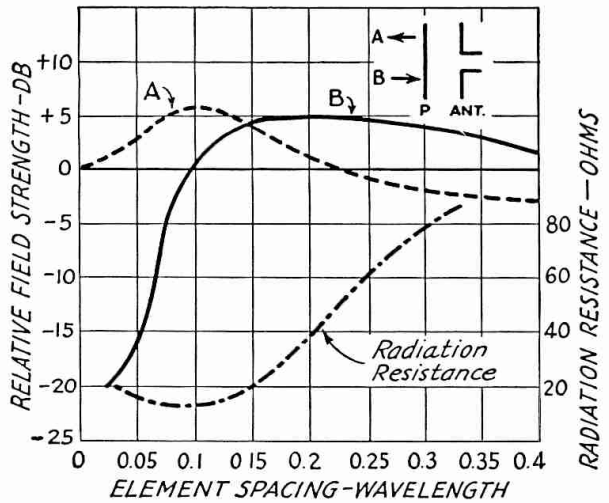


Fig. 14-34—Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back 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.

in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line.

The parasitic element is called a **director** when it reinforces radiation on a line pointing to it from the antenna, and a **reflector** when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic-element tuning, which usually is adjusted by changing its length.

Gain vs. Spacing

The gain of an antenna with parasitic elements varies with the spacing and tuning of the elements and thus for any given spacing there is a tuning condition that will give maximum gain at this spacing. The maximum front-to-back ratio seldom if ever, occurs at the same condition that gives maximum forward gain. The impedance of the driven element also varies with the tuning and spacing, and thus the antenna system must be tuned to its final condition before the match between the line and the antenna can be completed. However, the tuning and matching may interlock to some extent, and it is usually necessary to run through the adjustments several times to insure that the best possible tuning has been obtained.

Two-Element Beams

A 2-element beam is useful where space or other considerations prevent the use of the larger structure required for a 3-element beam. The general practice is to tune the parasitic element as a reflector and space it about 0.15 wavelength from the driven element, although some successful antennas have been built with 0.1-wavelength spacing and director tuning. Gain *vs.* element spacing for a 2-element antenna is given in Fig. 14-34, for the special case where the parasitic element is resonant. It is indicative of the performance to be expected under maximum-gain tuning conditions.

Three-Element Beams

Where room is available for an over-all length greater than 0.2 wavelength, a 3-element beam is preferable to one with only 2 elements. Once the over-all length has been decided upon, the curves of Fig. 14-35 can be used to determine the proper spacing of director and reflector. If, for example, the distance between director and reflector can be made 0.4 wavelength, Fig. 14-35 shows that a spacing of 0.15D-0.25R gives a gain of 7.8 db., and a spacing of 0.25D-0.15R gives a gain of 8.2 db. Obviously the latter is the better choice, although the practical difference might be difficult to measure, and practical (mechanical) considerations might call for using the more balanced 0.2D-0.2R construction and a gain of 8.1 db.

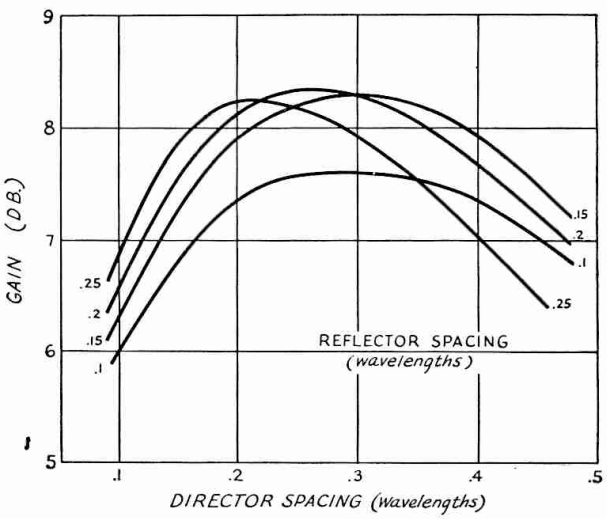


Fig. 14-35—Gain *vs.* element spacing for 3-element beams using a driven element and a director and a reflector. The 0-db. reference level is the field strength from a half-wavelength antenna alone. These curves are for the system tuned for maximum forward gain.

The element spacing shown is the fraction of wavelength determined by $\frac{984}{f\,(\text{Mc.})}$. Thus a wavelength at 14.2 Mc. = 984/14.2 = 69.3 feet. A spacing of 0.15 wavelength at 14.2 Mc. would be 0.15 × 69.3 = 10.4 feet = 10 feet 5 inches.

When the over-all length has been decided upon, and the element spacing has been determined, the element lengths can be found by referring to Fig. 14-36. It must be remembered that the lengths determined by these charts will vary slightly in actual practice with the element diameter and the method of supporting the elements, and the tuning of a beam should always be checked after installation. However, the lengths obtained by the use of the charts will be

close to correct in practically all cases, and they can be used without checking if the beam is difficult of access.

The preferable method for checking the beam is by means of a field-strength meter or the S-meter of a communications receiver, used in conjunction with a dipole antenna located at least 10 wavelengths away and as high as or higher than the beam that is being checked. A few watts of power fed into the antenna will give a useful signal at the observation point, and the power input to the transmitter (and hence the antenna) should be held constant for all of the readings. Beams tuned on the ground and then lifted into place are subject to tuning errors and cannot be depended upon. The impedance of the driven element will vary with the height above ground, and good practice dictates that all final matching between antenna and line be done with the antenna in place at its normal height above ground.

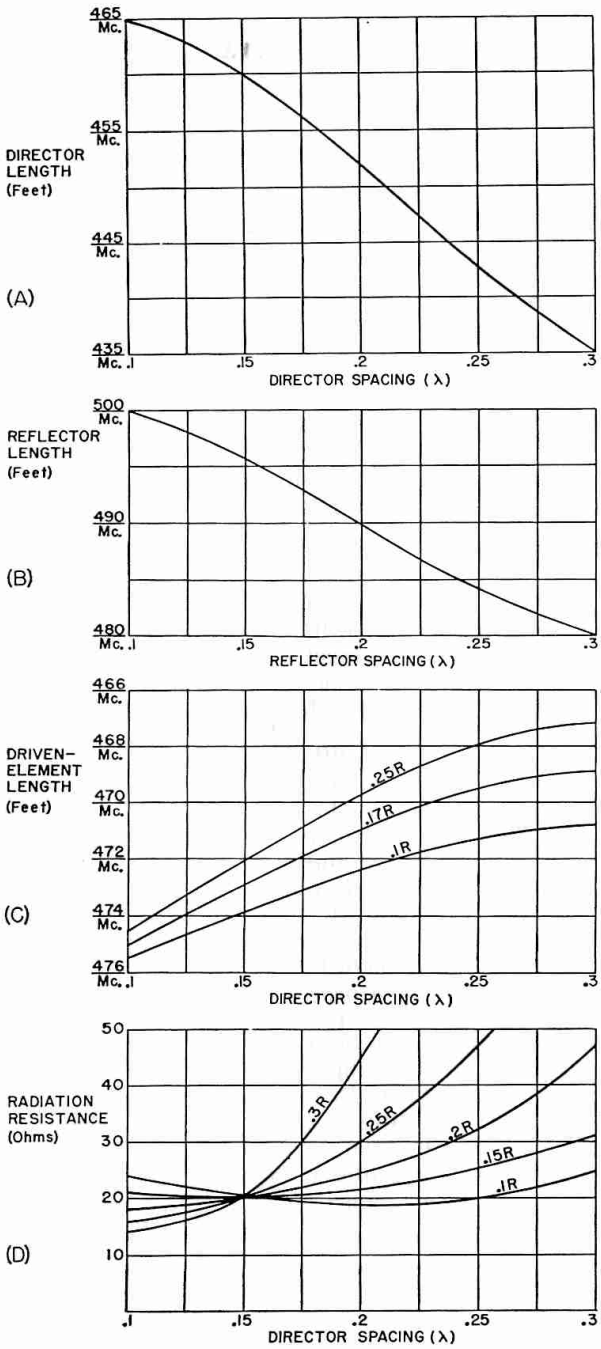


Fig. 14-36—Element lengths for a 3-element beam. These lengths will hold closely for tubing elements supported at or near the center. The radiation resistance (D) is useful information in planning for a matching system, but it is subject to variation with height above ground and must be considered an approximation.

The driven-element length (C) may require modification for tuning out reactance if a T- or gamma-match feed system is used, as mentioned in the text.

A 0.2D-0.2R beam cut for 28.6 Mc. would have a director length of 452/28.6 = 15.8 = 15 feet 10 inches, a reflector length of 490/28.6 = 17.1 = 17 feet 1 inch, and a driven-element length of 470.5/28.6 = 16.45 = 16 feet 5 inches.

Simple Systems: the Rotary Beam

Two- and 3-element systems 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.

A 4-element beam will give still more gain than a 3-element one, provided the support is sufficient for about 0.2 wavelength spacing between elements. The tuning for maximum gain involves many variables, and complete gain and tuning data are not available.

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 simple dipole antenna. With 3- and 4-element close-spaced arrays the radiation resistance of the driven element may be so low that ohmic losses in the conductor can consume an appreciable fraction of the power.

Feeding the Rotary Beam

Any of the usual methods of feed (described later under "Matching the Antenna to the Line") can be applied to the driven element of a rotary beam. Tuned feeders are not recommended for lengths greater than a half wavelength unless open lines of copper-tubing conductors are used. The popular choices for feeding a beam are the gamma match with series capacitor and the T match with series capacitors and a half-wavelength phasing section, as shown in Fig. 14-37. These methods are preferred over any others because they permit adjustment of the matching and the use of coaxial line feed. The variable capacitors can be housed in small plastic cups for weatherproofing; receiving types with close spacing can be used at powers up to a few hundred watts. Maximum capacity required is usually 140 $\mu\mu$f. at 14 Mc. and proportionately less at the higher frequencies.

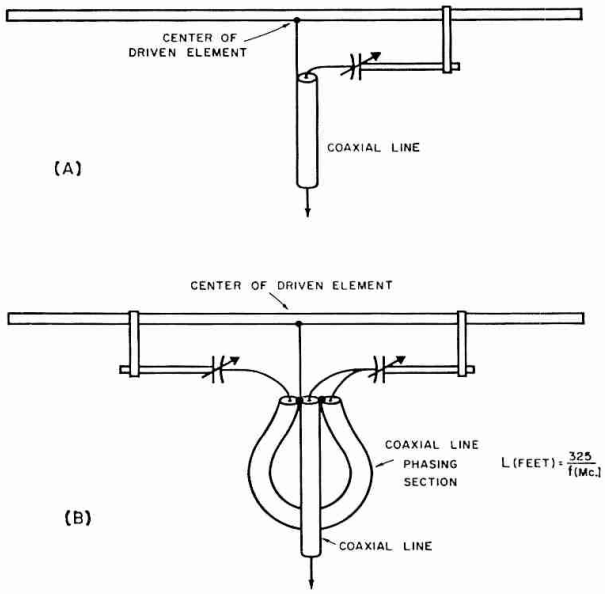


Fig. 14-37—The most popular methods of feeding the driven element of a beam antenna are (A) the gamma match and (B) the T match. The aluminum tubing or rod used for the matching section is usually of smaller diameter than the antenna element; its length will vary somewhat with the spacing and number of elements in the beam. The coaxial line in the phasing section can be coiled in a 2- or 3-foot diameter coil instead of hanging as shown.

If physically possible, it is better to adjust the matching device after the antenna has been installed at its ultimate height, since a match made with the antenna near the ground may not hold for the same antenna in the air.

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 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.

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 can be treated in the same fashion.

The Quad Antenna

The "cubical quad" antenna shown in Fig. 14-38 uses a square loop driven element and a square loop parasitic reflector. The spacing is usually between .15 and .20 wavelength and it is not critical, since the reflector element is tuned for maximum gain after installation.

Quad antennas are popular because they are lightweight and have low wind resistance. Their gain has been measured at 8 db. or more. A 14-Mc. quad will have approximately a 17-foot "wing span" and a boom length of about 12 feet, and it can be built light enough to be turned by a good TV rotator. Suggestions for the construction of a quad antenna are given in Fig. 14-39. The design was intended to be as light as possible and while the antenna will whip some in the wind, this should not cause any noticeable change in

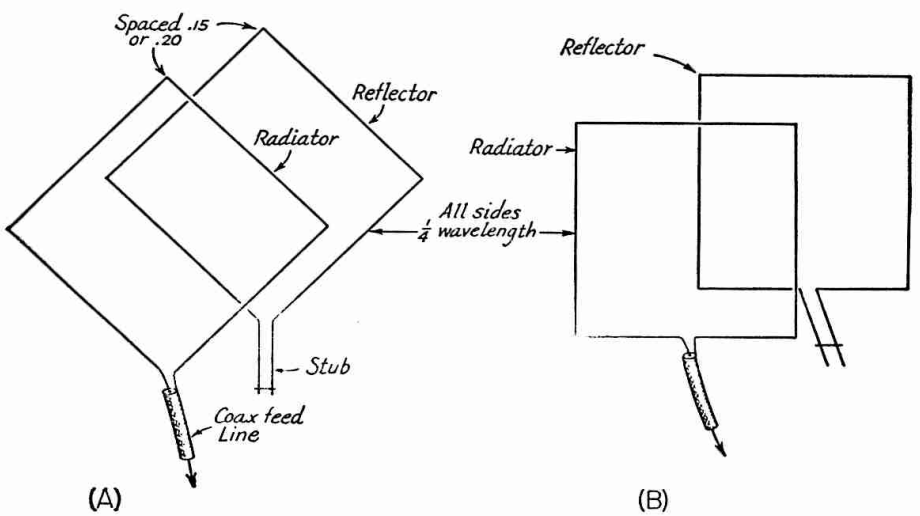


Fig. 14-38—Two different arrangements of cubical quad antennas.

Quad Antenna

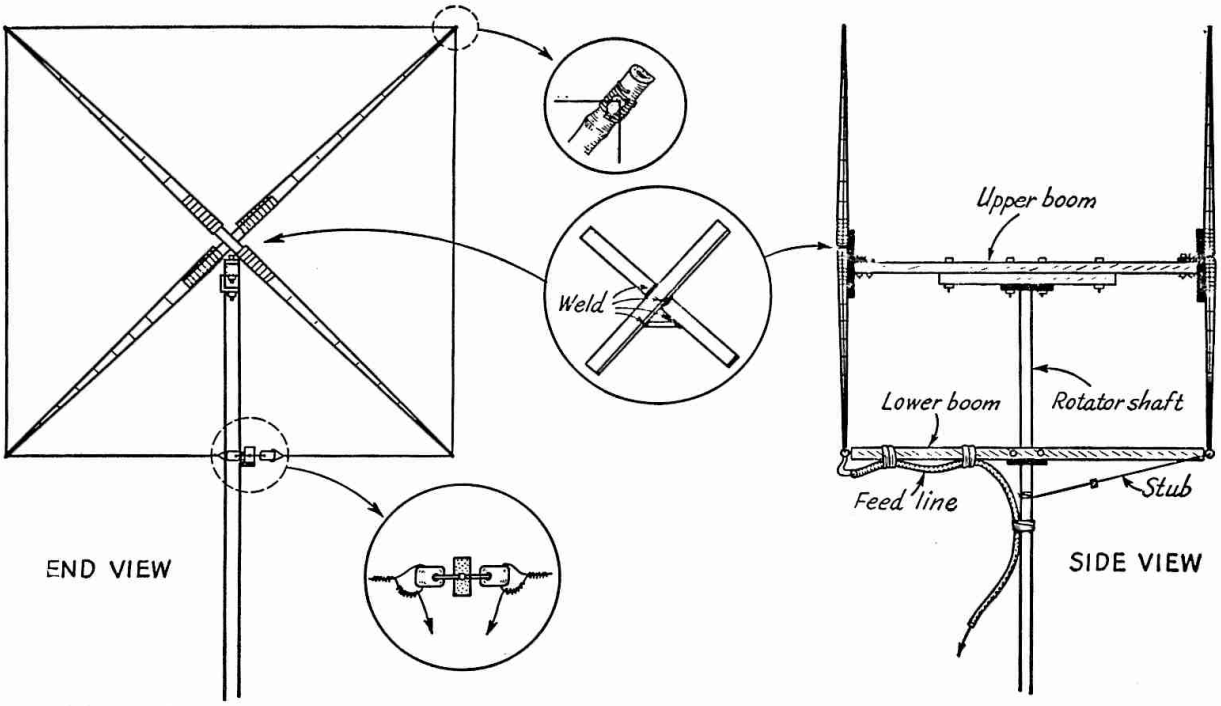


Fig. 14-39—End and side views of a quad. Upper insert shows method of fastening antenna wire to support arms. Center insert shows construction of support-arm mounting bracket. Lower insert shows method of attaching feed line and stub to the center insulators. Two small egg insulators are used, fastened to end of lower boom as shown with a small nail.

$$\text{The length of one side is found from } L \text{ (feet)} = \frac{251}{f \text{(Mc.)}}$$

loading or on received signals. There is nothing critical in the construction except the length of the wire elements. One quad used 1 × 2-inch pine for the support arms but this beam was much too heavy and blew down in the first light wind. The support arms shown in the drawing are ordinary bamboo fishing poles about 16 feet long, with the butt ends wrapped with friction tape to prevent the metal mounting bracket and wire from biting into the bamboo. These arms are fastened to the mounting brackets with several turns of No. 14 galvanized wire, and the far ends are not trimmed until the antenna wire has been fastened in place. Two mounting brackets and eight bamboo support arms are required. The mounting brackets serve to hold the arms in place and to fasten them to the end of the boom. These brackets are made by welding two 24-inch lengths of 1-inch angle iron together back to back to form a large "X" 90 degrees between legs, and welding a 5-inch length of 1½-inch strap iron between two of the legs to fasten the "X" to the boom end. The arms are assembled and the antenna wire is fastened in place before attaching the brackets to the boom.

If the fishing poles are well treated with a weatherproofing compound they will last several years. Weatherproofing compounds are available at all lumber dealers. Get straight poles with no splits in them. No insulators are necessary, the poles themselves acting as long insulators. The easiest way to mount the antenna wire on the arms is to lay a long length of wire on the ground and mark it at the approximate quarter-wave intervals, and use these marks to indicate where the wire fastens to the pole.

Dual and triple quads can be built for the bands 20 through 10 meters. One such antenna is shown in Fig. 14-40, a dual quad for 15 and 10 meters. The same supporting structure is used for the two antennas, making the boom length equal to 0.15 to 0.2 wavelengths at the lower-frequency band. Separate coaxial cable feed lines are brought down from the two driven elements. In a two-band quad (20/15 or 15/10) the length of one side is obtained from

$$L \text{ (feet)} = 250 \div \text{(Mc.)}$$

In the case of any quad or combination of

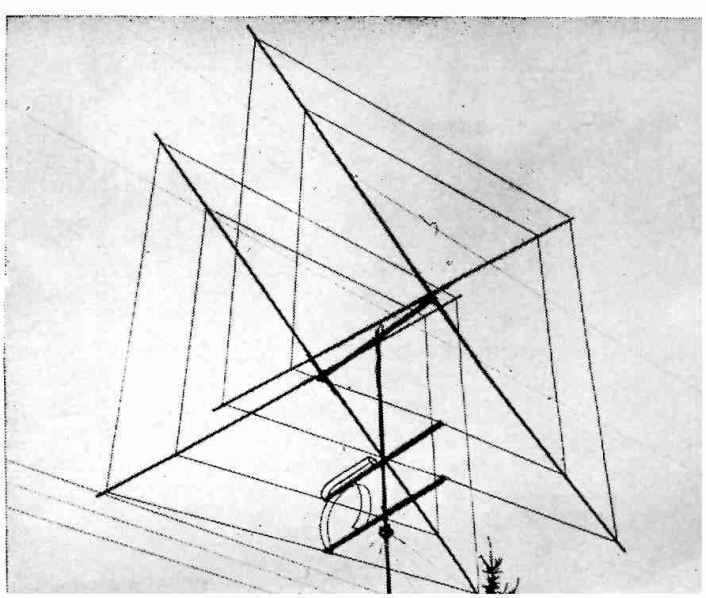


Fig. 14-40—A 15/10-meter quad. Tuning stubs for the reflectors are looped back along the tie bars. Total weight of this assembly, not including the mast, is 13 pounds.

quads, each quad should be tuned up separately for maximum forward gain by adjusting the stub length on the reflector element and checking the field strength with a nearby ham. If accessible, the reflector element can be resonated with a grid-dip meter to a frequency just below the lowest to be used; this is a good starting place for further adjustment. The resonance of the antenna system can be checked by finding the frequency that gives the lowest s.w.r. on the feed line; this lowest s.w.r. is not necessarily 1.0. If the resonant frequency is higher than the desired frequency, lengthen the driven element; shorten the element if the resonant frequency is too low. In the dual antennas that have been constructed, there has been little or no evidence of interaction of tuning.

Matching the Antenna to the Line

The load for a transmission line may be any device capable of dissipating r.f. power. When lines are used for transmitting applications the most common type of load is an antenna. When a transmission line is connected between an antenna and a receiver, the receiver input circuit (not the antenna) is the load, because the power taken from a passing wave is delivered to the receiver.

Whatever the application, the conditions existing at the load, and *only* the load, determine the standing-wave ratio on the line. If the load is purely resistive and equal in value to the characteristic impedance of the line, there will be no standing waves. If the load is not purely resistive, and/or is not equal to the line Z_0, there will be standing waves. No adjustments that can be made at the input end of the line can change the s.w.r., nor is it affected by changing the line length.

Only in a few special cases is the load inherently of the proper value to match a practicable transmission line. In all other cases it is necessary either to operate with a mismatch and accept the s.w.r. that results, or else to take steps to bring about a proper match between the line and load by means of transformers or similar devices. Impedance-matching transformers may take a variety of physical forms, depending on the circumstances.

Note that it is essential, if the s.w.r. is to be made as low as possible, that the load at the point of connection to the transmission line be purely resistive. In general, this requires that the load be tuned to resonance. If the load itself is not resonant at the operating frequency the tuning sometimes can be accomplished in the matching system.

THE ANTENNA AS A LOAD

Every antenna system, no matter what its physical form, will have a definite value of impedance at the point where the line is to be connected. The problem is to transform this **antenna input impedance** to the proper value to match the line. In this respect there is no one "best" type of line for a particular antenna system, because it is possible to transform impedances in any desired ratio. Consequently, any type of line may be used with any type of antenna. There are frequently reasons other than impedance matching that dictate the use of one type of line in preference to another, such as ease of installation, inherent loss in the line, and so on, but these are not considered in this section.

Although the input impedance of an antenna system is seldom known very accurately, it is often possible to make a reasonably close estimate of its value. The information earlier in this chapter can be used as a guide.

Matching circuits may be constructed using ordinary coils and capacitors, but are not used very extensively because they must be supported at the antenna and must be weatherproofed. The systems to be described use **linear transformers.**

The Quarter-Wave Transformer or "Q" Section

As described earlier in this chapter, a quarter-wave transmission line may be used as an impedance transformer. Knowing the antenna impedance and the characteristic impedance of the

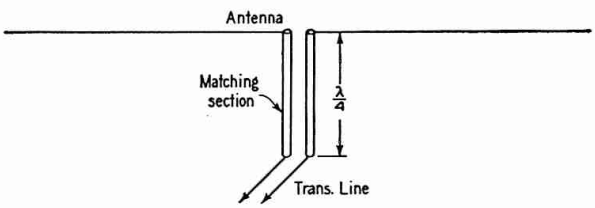


Fig. 14-41—"Q" matching section, a quarter-wave impedance transformer.

transmission line to be matched, the required characteristic impedance of a matching section such as is shown in Fig. 13-13 is

$$Z = \sqrt{Z_1 Z_0} \qquad \textbf{(14-I)}$$

where Z_1 is the antenna impedance and Z_0 is the characteristic impedance of the line to which it is to be matched.

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 the chapter on "Transmission Lines." (With ½-inch tubing, the spacing in the example above should be 1.5 inches for an impedance of 208 ohms.)

The length of the quarter-wave matching section may be calculated from

$$\textit{Length (feet)} = \frac{246V}{f} \qquad \textbf{(14-J)}$$

where V = Velocity factor
f = Frequency in Mc.

Example: A quarter-wave transformer of RG-11/U is to be used at 28.7 Mc. From the table in Chapter

Thirteen, $V = 0.66$.

$$\text{Length} = \frac{246 \times 0.66}{28.7} = 5.67 \text{ feet}$$

$$= 5 \text{ feet } 8 \text{ inches}$$

The antenna must be resonant at the operating frequency. Setting the antenna length by formula is amply accurate with single-wire antennas, but in other systems, particularly close-spaced arrays, the antenna should be adjusted to resonance before the matching section is connected.

When the antenna input impedance is not known accurately, it is advisable to construct the matching section so that the spacing between conductors can be changed. The spacing then may be adjusted to give the lowest possible s.w.r. on the transmission line.

Folded Dipoles

A half-wave antenna element can be made to match various line impedances if it is split into two or more parallel conductors with the transmission line attached at the center of only one of them. Various forms of such "folded dipoles" are shown in Fig. 14-42. Currents in all conductors are in phase in a folded dipole, and since the conductor spacing is small the folded dipole is equivalent in radiating properties to an ordinary single-conductor dipole. However, the current flowing into the input terminals of the antenna from the line is the current in one conductor only, and the entire power from the line is delivered at this value of current. This is equivalent to saying that the input impedance of the antenna has been raised by splitting it up into two or more conductors.

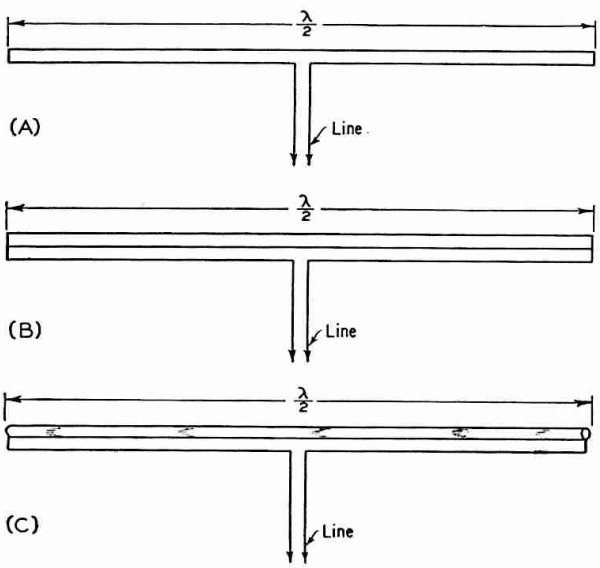


Fig. 14-42—The folded dipole, a method for using the antenna element itself to provide an impedance transformation.

The ratio by which the input impedance of the antenna is stepped up depends not only on the number of conductors in the folded dipole but also on their relative diameters, since the distribution of current between conductors is a function of their diameters. (When one conductor is larger

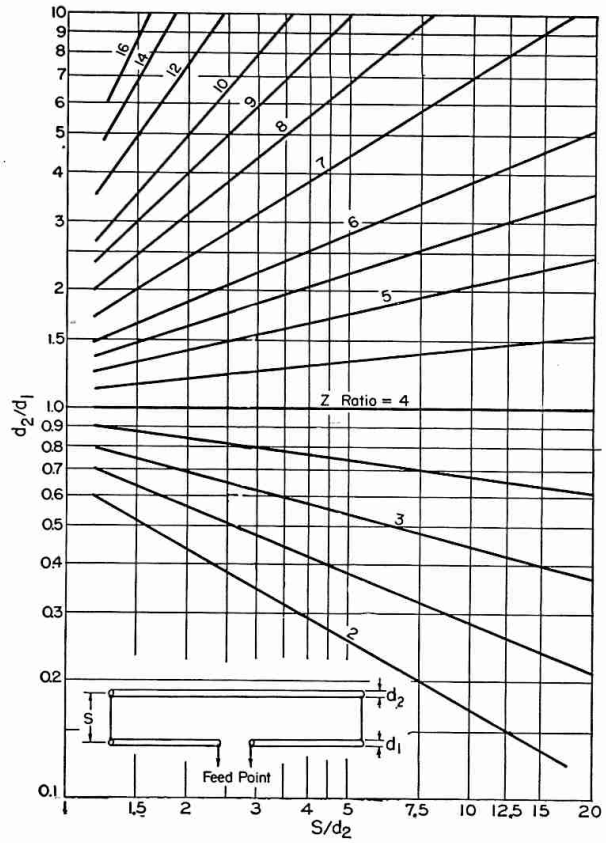


Fig. 14-43—Impedance transformation ratio, two-conductor folded dipole. The dimensions d_1, d_2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

than the other, as in Fig. 14-42C, the larger one carries the greater current.) The ratio also depends, in general, on the spacing between the conductors, as shown by the graphs of Figs. 14-43 and 14-44. An important special case is the 2-conductor dipole with conductors of equal diameter; as a simple antenna, not a part of a directive array, it has an input resistance close enough to 300 ohms to afford a good match to 300-ohm Twin-Lead.

The required ratio of conductor diameters to give a desired impedance ratio using two conductors may be obtained from Fig. 14-43. Similar information for a 3-conductor dipole is given in Fig. 14-44. This graph applies where all three conductors are in the same plane. The two conductors not connected to the transmission line must be equally spaced from the fed conductor, and must have equal diameters. The fed conductor may have a different diameter, however. The unequal-conductor method has been found particularly useful in matching to low-impedance antennas such as directive arrays using close-spaced parasitic elements.

The length of the antenna element should be such as to be approximately self-resonant at the median operating frequency. The length is usually not highly critical, because a folded dipole tends to have the characteristics of a "thick" antenna

and thus has a relatively broad frequency-response curve.

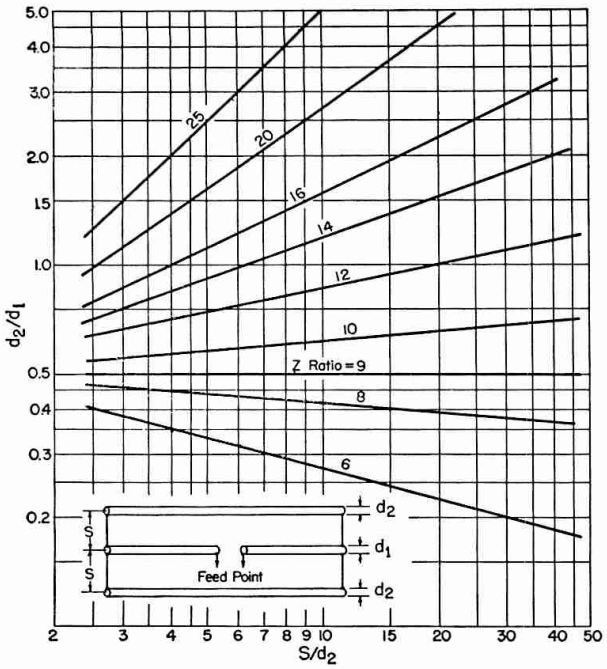


Fig. 14-44—Impedance transformation ratio, three-conductor folded dipole. The dimensions d_1, d_2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

"T" and "Gamma" Matching Sections

The method of matching shown in Fig. 14-45A is based on the fact that the impedance between any two points along a resonant antenna is resistive, and has a value which depends on the spacing between the two points. It is therefore possible to choose a pair of points between which the impedance will have the right value to match a transmission line. In practice, the line cannot be connected directly at these points because the distance between them is much greater than the conductor spacing of a practicable transmission line. The "T" arrangement in Fig. 14-45A overcomes this difficulty by using a second conductor paralleling the antenna to form a matching section

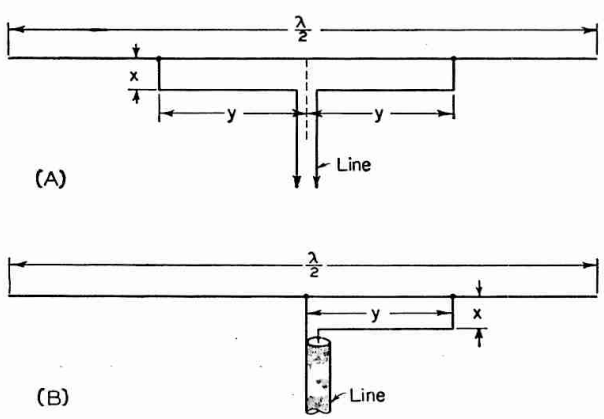


Fig. 14-45—The "T" match and "gamma" match.

to which the line may be connected.

The "T" is particularly suited to use with a parallel-conductor line, in which case the two points along the antenna should be equidistant from the center so that electrical balance is maintained.

The operation of this system is somewhat complex. Each "T" conductor (y in the drawing) forms with the antenna conductor opposite it a short section of transmission line. Each of these transmission-line sections can be considered to be terminated in the impedance that exists at the point of connection to the antenna. Thus the part of the antenna between the two points carries a transmission-line current in addition to the normal antenna current. The two transmission-line matching sections are in series, as seen by the main transmission line.

If the antenna by itself is resonant at the operating frequency its impedance will be purely resistive, and in such case the matching-section lines are terminated in a resistive load. However, since these sections are shorter than a quarter wavelength their input impedance — i.e., the impedance seen by the main transmission line looking into the matching-section terminals — will be reactive as well as resistive. This prevents a perfect match to the main transmission line, since its load must be a pure resistance for perfect matching. The reactive component of the input impedance must be tuned out before a proper match can be secured.

One way to do this is to detune the antenna just enough, by changing its length, to cause reactance of the opposite kind to be reflected to the input terminals of the matching section, thus cancelling the reactance introduced by the latter. Another method, which is considerably easier to adjust, is to insert a variable capacitor in series with the matching section where it connects to the transmission line, as shown in Fig. 14-37. The capacitor must be protected from the weather.

The method of adjustment commonly used is to cut the antenna for approximate resonance and then make the spacing x some value that is convenient constructionally. The distance y is then adjusted, while maintaining symmetry with respect to the center, until the s.w.r. on the transmission line is as low as possible. If the s.w.r. is not below 2 to 1 after this adjustment, the antenna length should be changed slightly and the matching-section taps adjusted again. This process may be continued until the s.w.r. is as close to 1 to 1 as possible.

When the series-capacitor method of reactance compensation is used (Fig. 14-37) the antenna should be the proper length to be resonant at the operating frequency. Trial positions of the matching-section taps are taken, each time adjusting the capacitor for minimum s.w.r., until the standing waves on the transmission line are brought down to the lowest possible value.

The unbalanced ("gamma") arrangement in Fig. 14-45B is similar in principle to the "T," but is adapted for use with single coax line. The method of adjustment is the same.

BALANCING DEVICES

An antenna with open ends, of which the half-wave type is an example, is inherently a balanced radiator. When opened at the center and fed with a parallel-conductor line this balance is maintained throughout the system, so long as the causes of unbalance discussed in the transmission-line chapter are avoided.

If the antenna is fed at the center through a coaxial line, as indicated in Fig. 14-46A, this balance is upset because one side of the radiator is connected to the shield while the other is connected to the inner conductor. On the side connected to the shield, a current can flow down over the *outside* of the coaxial line, and the fields thus set up cannot be canceled by the fields from the inner conductor because the fields *inside* the line cannot escape through the shielding afforded by the outer conductor. Hence these "antenna" currents flowing on the outside of the line will be responsible for radiation.

Linear Baluns

Line radiation can be prevented by a number of devices whose purpose is to detune or decouple the line for "antenna" currents and thus greatly reduce their amplitude. Such devices generally are known as **baluns** (a contraction for "balanced to unbalanced"). Fig. 14-46B shows one such arrangement, known as a **bazooka,** which uses a sleeve over the transmission line to form, with the outside of the outer line conductor, a shorted quarter-wave line section. As described earlier in this chapter, the impedance looking into the open end of such a section is very high, so that the end of the outer conductor of the coaxial line is effectively insulated from the part of the line below the sleeve. The length is an *electrical* quarter wave, and may be physically shorter if the insulation between the sleeve and the line is other than air. The bazooka has no effect on the impedance relationships between the antenna and the coaxial line.

Another method that gives an equivalent effect is shown at C. Since the voltages at the antenna terminals are equal and opposite (with reference to ground), equal and opposite currents flow on the surfaces of the line and second conductor. Beyond the shorting point, in the direction of the transmitter, these currents combine to cancel out. The balancing section "looks like" an open circuit to the antenna, since it is a quarter-wave parallel-conductor line shorted at the far end, and thus has no effect on the normal antenna operation. However, this is not essential to the line-balancing function of the device, and baluns of this type are sometimes made shorter than a quarter wavelength in order to provide the shunt inductive reactance required in certain types of matching systems.

Fig. 14-46D shows a third balun, in which equal and opposite voltages, balanced to ground, are taken from the inner conductors of the main transmission line and half-wave phasing section. Since the voltages at the balanced end are in series while the voltages at the unbalanced end are in

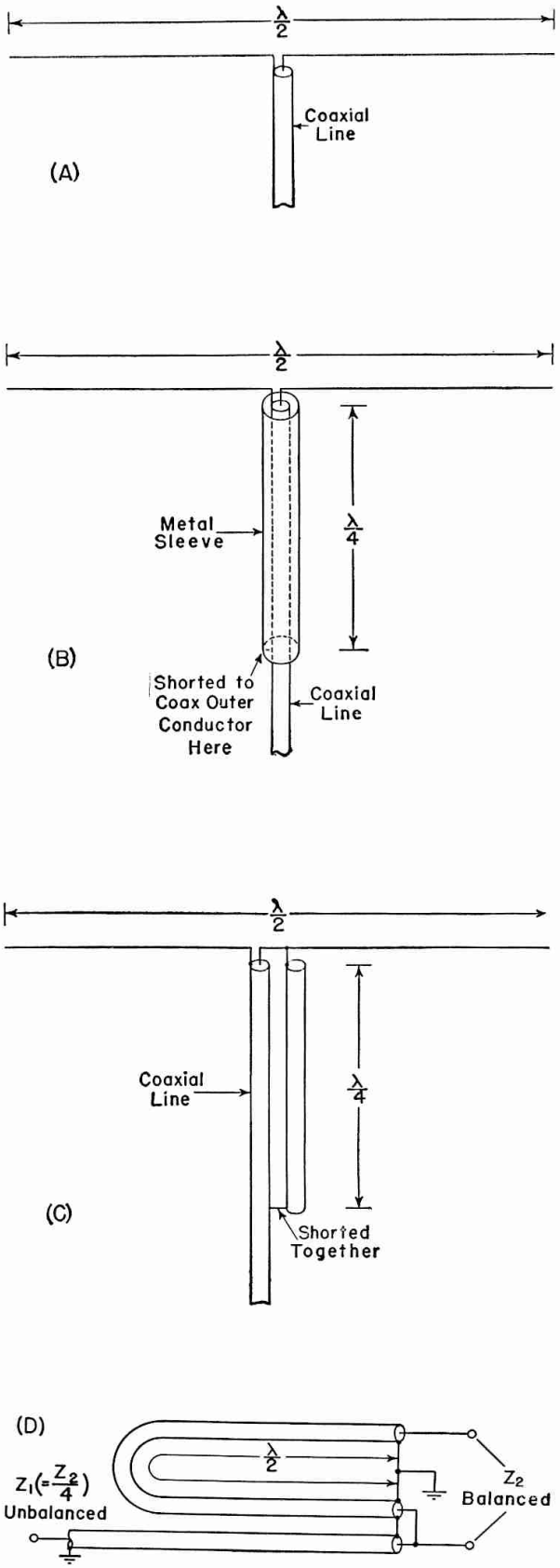


Fig. 14-46—Radiator with coaxial feed (A) and methods of preventing unbalance currents from flowing on the outside of the transmission line (B and C). The half-wave phasing section shown at D is used for coupling between an unbalanced and a balanced circuit when a 4-to-1 impedance ratio is desired or can be accepted.

parallel, there is a 4-to-1 step-down in impedance from the balanced to the unbalanced side. This arrangement is useful for coupling between a balanced 300-ohm line and a 75-ohm coaxial line, for example.

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 and higher the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, sometimes only a short length of wire strung around the room is used for a receiving antenna, but such an antenna cannot be expected to give good performance, although it is adequate for loud signals on the 3.5- and 7-Mc. bands. It will serve in emergencies, but a longer wire outdoors is always better.

The use of a tuned antenna improves the operation of the receiver, because the signal strength is greater than with a wire of random length. Where local electrical noise is a problem, as from an electrical appliance, a measure of relief can often be obtained by locating the antenna as high above and as far as possible from the noise source and power lines. The lead-in wire, from the center of the antenna, should be a coaxial line or shielded twin-conductor cable (RG-62/U). If the twin-conductor cable is used, the conductors connect to the antenna binding posts and the shield to the ground binding post of the receiver.

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.

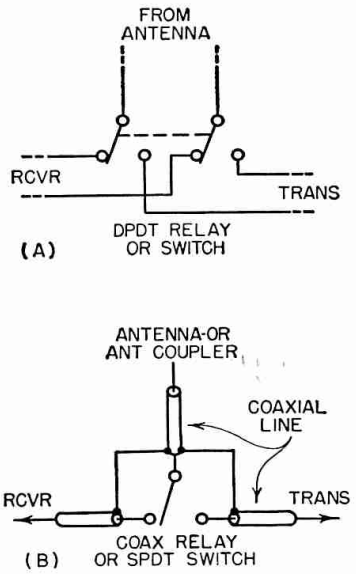


Fig. 14-47—Antenna changeover for receiving and transmitting in two-wire line (A) and coaxial line (B). The low-pass filter for TVI reduction should be connected between switch or relay and the transmitter.

Typical arrangements are shown in Fig. 14-47. If coaxial line is used, a coaxial relay is recommended, although on the lower-frequency bands a regular switch or change-over relay will work almost as well. The relay or switch contacts should be rated to handle at least the maximum power of the transmitter.

An additional refinement is the use of an electronic transit-receive switch, which permits full break-in operation even when using the transmitting antenna for receiving. For details and circuitry on t.r. switches, see Chapter Eight.

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. Open-wire TV line is excellent up to several hundred watts.

In building a two-wire open line, 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.

At points of maximum voltage, insulation is

most important, and Pyrex glass or ceramic insulators with long leakage paths are recommended for the antenna. Insulators should be cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung from one chimney to another or from a housetop 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.

Telephone poles, if they can be purchased and installed economically, make excellent supports because they do not ordinarily require guying in heights up to 40 feet or so. Many low-cost television-antenna supports are now available, and they should not be overlooked as possible antenna aids.

"A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 14-48 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.

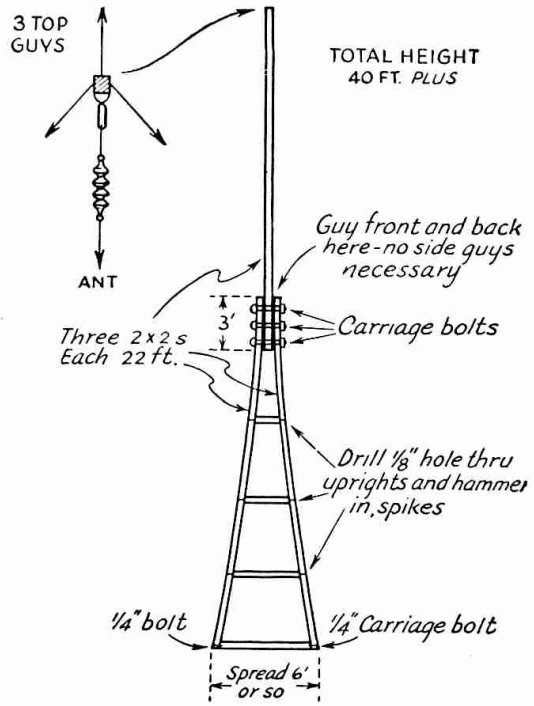


Fig. 14-48—Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

SIMPLE 40-FOOT MAST

The mast shown in Fig. 14-49 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

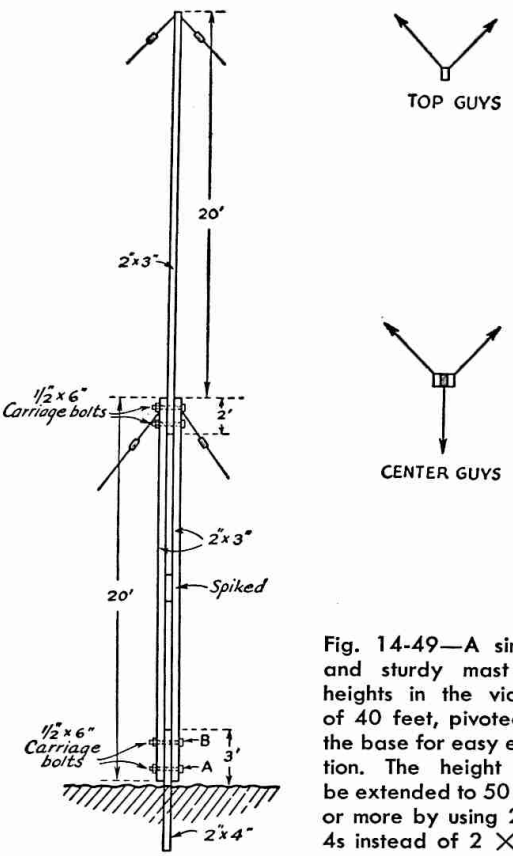


Fig. 14-49—A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.

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 measure continually supported. When the mast is vertical, bolt *B* should be slipped in place and both *A* and *B* tightened. The lower guys can then be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they should be adjusted to pull the top section into line.

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 likely to be 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. 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 available facilities, 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. A simple time- and finger-saving

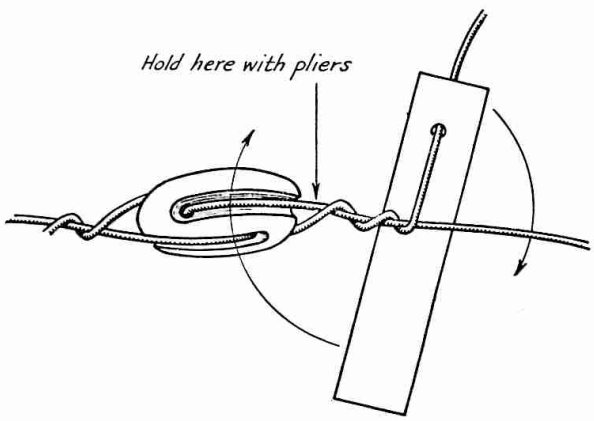


Fig. 14-50—Using a lever for twisting heavy guy wires.

device (piece of heavy iron or steel) can be made 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 Fig. 14–50. 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.

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.

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.

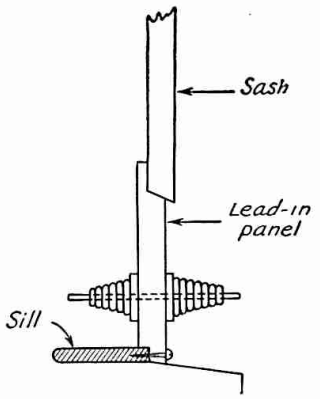


Fig. 14-51—An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Substituting a smaller height sash in half the window will simplify the weatherproofing problem where the sash overlaps.

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 FEED LINE INTO THE STATION

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 14-52, 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. Cement or rubber gaskets may be used to waterproof the exposed joints.

Where such a procedure is not permissible, the window itself usually offers the best opportunity. One satisfactory method is to drill holes in the glass near the top of the upper sash. If the glass is replaced by plate glass, a stronger job will result. Plate glass may be obtained from automobile junk yards and drilled before placing in the frame. The glass itself provides insulation and the transmission line may be fastened to bolts fitting the holes. 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. 14-52B 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. 14-51, or by using weatherstrip material where necessary.

Coaxial line can be brought through clearance holes without additional insulation.

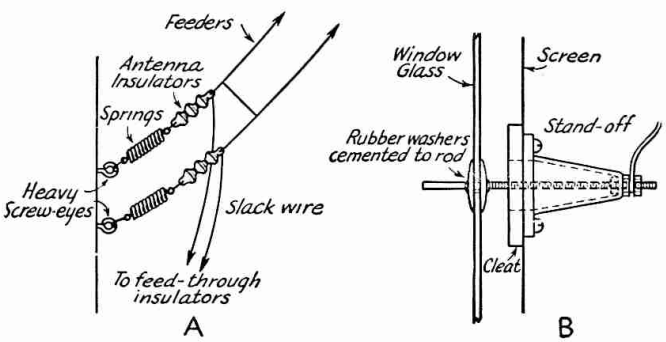


Fig. 14-52—A—Anchoring feeders takes the strain from feed-through insulators or window glass. B—Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

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. The use of such rotatable antennas is usually 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-, 21- 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 another chapter.

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.

Elements

The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the

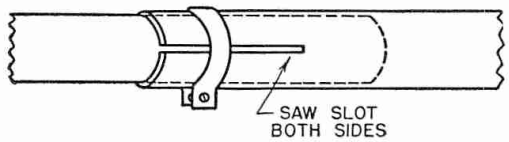


Fig. 14-53—Details of telescoping tubing for beam elements.

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.

Aluminum alloy tubes are generally used for the elements. 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. Regardless of the tubing used, the ends should be plugged up with corks sealed with glyptal varnish.

The element lengths are made adjustable by sawing a 6- to 12-inch slot in the ends of the larger-diameter tubing and clamping the smaller tubing inside. Homemade clamps of aluminum can be built, or hose clamps of suitable size can be used. An example of this construction is shown in Fig. 14-53. If steel clamps are used, they should be cadmium- or zinc-plated before installation.

Supports

Metal is commonly 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. Fittings for TV antennas can often be used on 21- and 28-Mc. beams. "Irrigation pipe" is a good source of aluminum tubing up to diameters of 6 inches and lengths of 20 feet. Muffler clamps can be used to hold beam elements to a boom.

Most of the TV antenna rotators are satisfactory for turning the smaller beams.

With all-metal construction, delta, "gamma" or "T"-match are the only practical matching methods 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.

"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. Some suggestions for the constructional details are given in Figs. 14-54, 14-55 and 14-56. These show portions of a 4-element 10-meter beam, but the same principles hold for 15- and 20-meter beams.

Boom material can be the irrigation pipe suggested earlier (available from Sears Roebuck). Muffler clamps and homemade brackets (aluminum or cadmium-plated steel) can be used to hold the parasitic elements to the boom, as shown in Fig. 14-54. The muffler clamps and all hardware should be cadmium-plated to forestall corrosion; the plating can be done at a plating shop and will not be very expensive if it is all done at the same time.

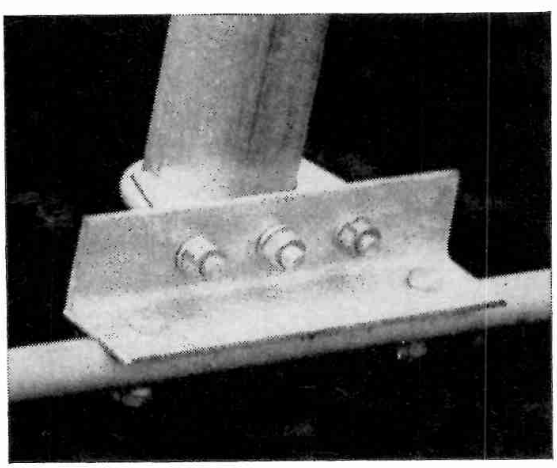


Fig. 14-54—Muffler clamps can be used to hold beam elements to the boom. The angle can be aluminum angle or angle iron; if iron is used it should be cadmium plated. This example shows a ¾-inch-diameter element held to a 2-inch diameter boom.

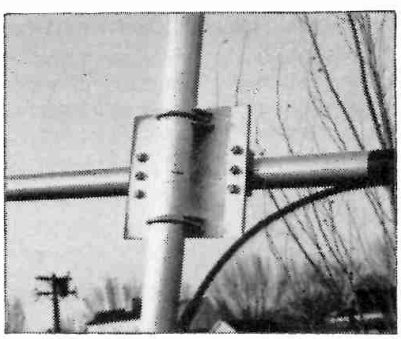
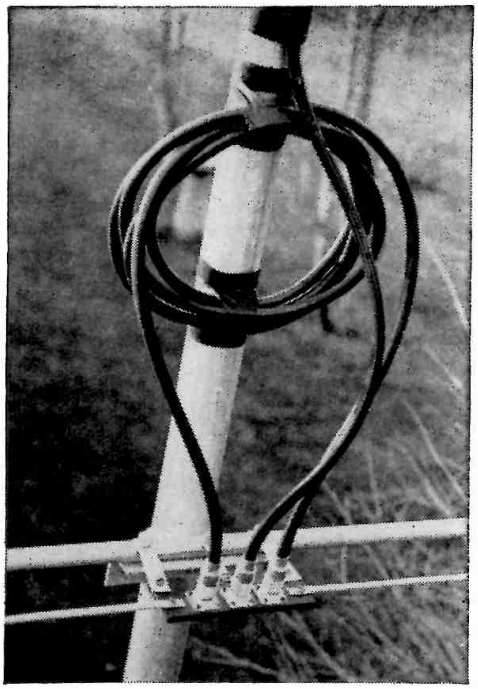


Fig. 14-55—The boom can be tied to the mast with muffler clamps and a steel plate. The coaxial line from the driven element is taped to the boom and mast.

Muffler clamps and a steel plate can be used to hold the boom to the supporting mast, as shown in Fig. 14-55. For maximum strength, the mast section should be a length of galvanized iron pipe. The plate thickness should run from $\frac{3}{16}$ inch for a 10-meter beam to ½ inch or more for a 20-meter beam. Steel plates of this thickness are best cut in a welding shop, where it can be done quickly for a nominal fee. After the plate has been cut and the muffler-clamp holes drilled, the plate, clamps and hardware should be plated.

The photograph in Fig. 14-56 shows one way

Rotary Beam Construction



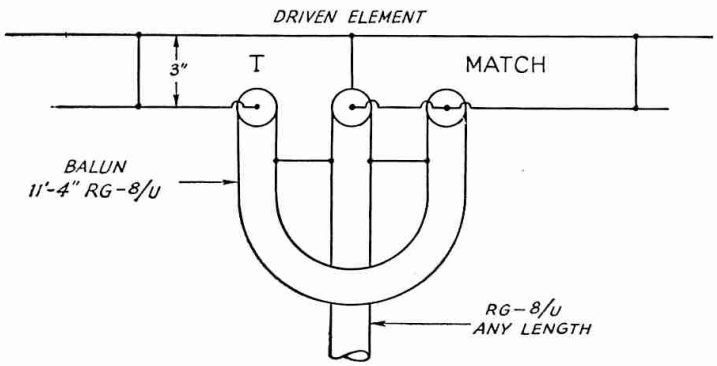


Fig. 14-56—Details of a coaxial-line termination board and T-match support for a 10-meter beam. The balun of a half-wavelength of coaxial line is coiled and then fastened to the boom with tape.

a T-matched driven element can be assembled with its half-wave balun. Three coaxial chassis receptacles are fastened to a ¼-inch thick sheet of phenolic that is supported below the driven element by three aluminum straps. The two T rods are also supported by the phenolic sheet at the inner ends and by suitable straps at the outer ends where they make up to the driven element.

Rotation

It is convenient but not essential to use a motor to rotate the beam. If a rope-and-pulley arrangement can be brought into the operating room or if the pole can be mounted near a window in the operating room, hand rotation will work.

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. Generally speaking, light-weight units are better because they reduce the tower load.

The speed of rotation should not be too great — one or 1½ 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 usual beam does not require a great deal of power for rotation at slow speed, and a ⅛-hp. motor will be ample. A reversible motor should be used. War-surplus "prop pitch" motors have found wide application for rotating 14-Mc. beams, while TV rotators can be used with many 28-Mc. lightweight beams.

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 that the surfaces are clean and free from grease before painting. Grease can be removed by brushing with kerosene and then squirting the surface with a solid stream of water. The work can then be wiped dry with a rag.

The power and control leads to the rotator should be run in electrical conduit or in lead covering, and the metal should be grounded.

A Compact 14-Mc. 3-Element Beam

A 20-meter beam no larger than the usual 10-meter beam can be made by using center-loaded elements and close spacing. Such an antenna will show good directivity and can be rotated with a TV-antenna rotator.

Constructional details of the elements are shown in Figs. 14-57 and 14-58. The loading coils are space-wound by interwinding plumb line (sometimes known as chalk line) with the No. 12 wire coils. The coil ends are secured by drilling small holes through the polystyrene bar, as shown in Fig. 14-60. The coils should be sprayed or painted with Krylon before installing the protective Lucite tubes.

The beam will require 4-foot lengths of the tubings indicated in Fig. 14-57A. For good telescoping, element wall thickness of 0.058 inch is recommended. The ends of the tubing sections should be slotted to permit adjustment, and secured with clamps, so that the joints will not work loose in the wind. Perforated ground clamps can be used for this purpose. The boom is a 12-foot length of 1½-inch o.d. 61ST aluminum tubing, with 0.125-inch wall.

The line is coupled and matched at the center of the driven element through adjustment of the link wound on the outside of the Lucite tubing. To check the adjustment of the elements, first resonate the driven element to the desired frequency in the 14-Mc. band with a grid-dip oscil-

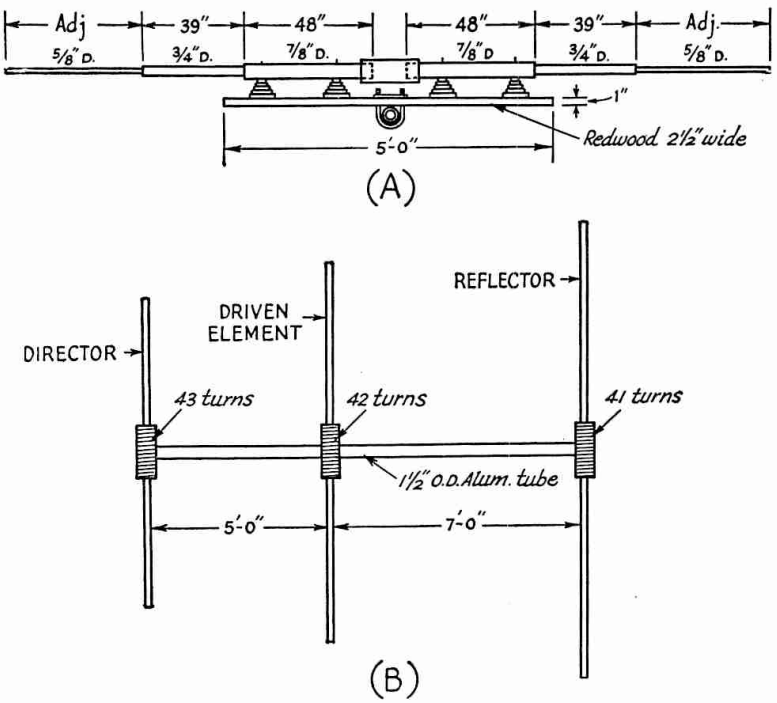


Fig. 14-57—Dimensions of a compact 14-Mc. beam. A—Side view of a typical element. TV-antenna "U" clamps hold the support arms to the boom. Birnbach 4176 insulators support the elements. B—Top plan of the beam showing element spacing and loading-coil dimensions. Elements are made of aluminum tubing. Construction of the loading coils and adjustment of the elements are discussed in the text. End-section lengths of 41 inches for the reflector, 40 inches for the driven element, and 10 inches for the director will be close to optimum.

lator. Then resonate the director to approximately 14.8 Mc., and the reflector to approximately 13.6 Mc. This is not critical and only serves as a rough point for the final tuning, which is done by use of a conventional field-strength indicator. Check the transmitter loading and readjust if necessary. Adjust the director for maximum forward gain, and then adjust the reflector for maximum forward gain. At this point, check the driven element for resonance and readjust if necessary. Turn the reflector toward the field-strength indicator and adjust for back cut-off. This must be done in small steps. Do not expect the attenuation off the sides of a short beam to be as high as that obtained with full-length elements. The s.w.r. of the line feeding the antenna can be checked with a bridge, and after the elements have been tuned, a final adjustment of the s.w.r. can be made by adjusting the coupling at the antenna loading coil turns and spacing. As in any beam, the s.w.r. will depend upon this adjustment and not on any that can be made at the transmitter. Transmitter coupling is the usual for any coaxial line. (From *QST*, May, 1954.)

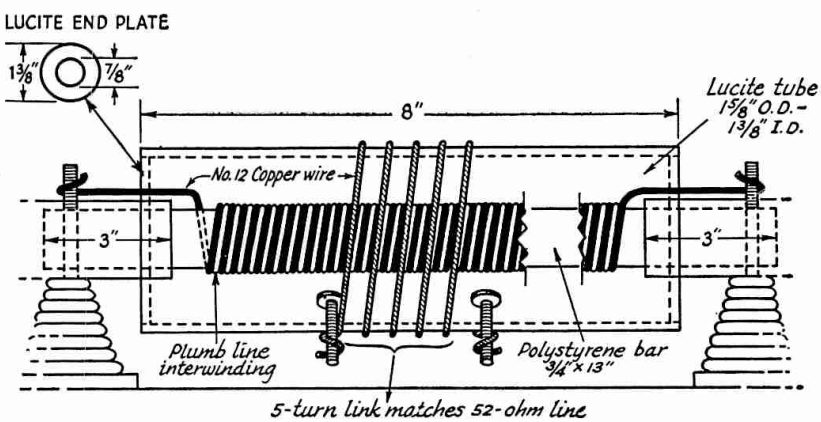


Fig. 14-58—Detailed sketch of the loading and coupling coils at the center of the driven element, and its mounting. Similar loading coils (see text) are used at the centers of the director and reflector.

A "One-Element Rotary" for 21 Mc.

The directional properties of a simple half-wavelength antenna become more apparent at higher frequencies, and it is possible to take advantage of this fact to build a "one-element rotary" for 21 or 28 Mc. To take advantage of the directional properties of the antenna, it is only necessary to rotate it 180 degrees. It can be rotated by hand, as will be described, or by a small TV antenna rotator. A 28-Mc. antenna should be made full size (14-C) and fed at the center with RG-11/U.

The 21-Mc. antenna is made from two pieces of ½-inch diameter electrical thin-wall steel tubing or conduit. This tubing is readily available at any electric supply shop. It comes in 10-foot lengths and, while 20 feet is short for a half-wave antenna

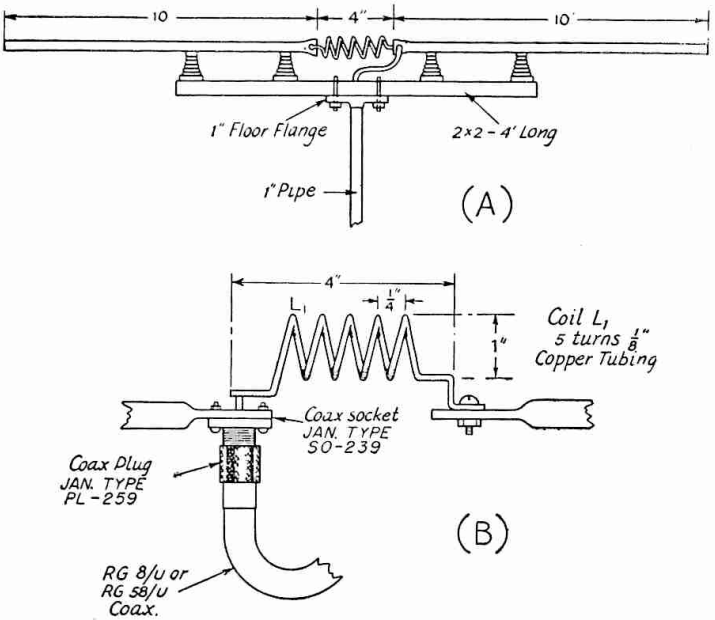


Fig. 14-59—(A) Diagram of the 21-Mc. antenna and mounting. The U-bolts that hold the 2 by 2 to the floor flange are standard 2-inch TV mast type bolts. (B) A more detailed drawing of the coil and coax-fitting mountings. The ¼-inch spacing between turns is not critical, and they can vary as much as 1/16 inch without any apparent harm to the match.

at 21-Mc., with loading the length is just about right for 52-ohm line feed. (A half-wavelength antenna would normally be fed with 72-ohm cable, since the antenna offers a good match for this impedance value. In this antenna system, the shorter elements, plus the small coil, offer a good match for 52-ohm cable.) If aluminum tubing is available, it can be used in place of the conduit, and the antenna will be lighter in weight. As shown in Figs. 14-59 and 14-60, the two pieces of tubing are supported by four stand-off insulators on a four foot long 2 by 2. The coax fitting for the feed line is mounted on the end of one of the lengths of tubing. A mounting point is made by flattening the end of the tubing for a length of about 1½ inches. The tubing can be flattened by squeezing it in a vise or by laying the end of the tubing on a hard surface and then hammering it flat. This will provide enough space to accommodate the coax fitting (Amphenol type 83-1R). A ⅝-inch hole will be needed in the flat section to clear the shell of the coax fitting.

The coil, L_1, is made from ⅛-inch diameter copper tubing. It consists of 5 turns spaced ¼ inch apart and is 1 inch inside diameter. The coil is connected in series with the inner conductor pin on the coax fitting and the other half of the antenna. To secure a good connection at the coax fitting, the coil lead should be wound around the inner-conductor pin and soldered. The other end of the coil can be connected with a screw and nut.

Mounting

The antenna can be mounted on a 1-inch floor flange and held in place by two 2-inch bolts, as shown in Fig. 14-61. The floor flange can be connected to a 12-foot length of 1-inch pipe which will serve as a mast. Television antenna wall mounts can be used to support the mast.

In the installation shown in Fig. 14-61, 19-inch wall mounts were used in order to clear the eaves of the house. A 2-inch long piece of 1¼-inch pipe was used as a sleeve, and it was clamped in the U bolt on the bottom wall mount. A ¼-inch hole

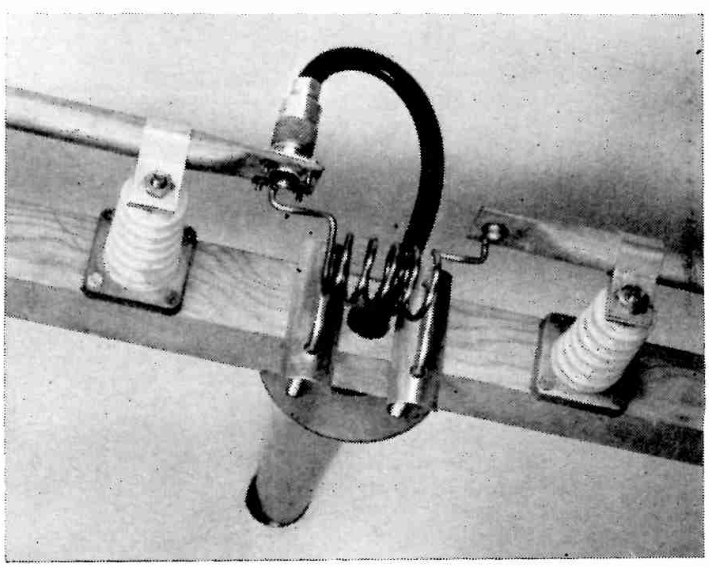


Fig. 14-60—A close-up of the coil and coax fitting mountings. Be sure that the coil doesn't short out to the outer conductor when soldering the coil end to the inner conductor pin on the coax fitting.

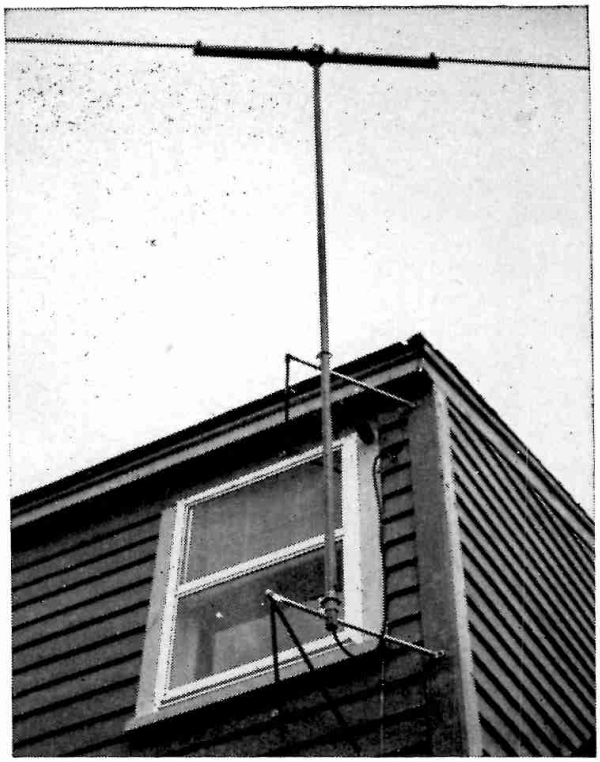


Fig. 14-61—Over-all view of the antenna and mounting. The feed line comes out of the bottom of the mast and through the wall into the shack.

was drilled through the mast pipe approximately 6 inches from the bottom. Then a 1½-inch bolt was slipped through the hole and the mast was then mounted in the sleeve on the bottom wall mount. The bolt acted as a bearing point against the top of the sleeve. Another ¼-inch hole was drilled through the mast about three feet above the bottom wall mount. A piece of ¼-inch metal rod, six inches long, was forced through the hole so that the rod projected on each side of the mast. To turn the mast, a piece of rope was attached to each end of the rod and the rope was brought into the shack, so that the antenna could be rotated by the "arm-strong" method. Obviously, one could spend more money for a "de luxe" version and use a TV antenna rotator and mast.

RG-8/U 52-ohm coax cable is recommended to feed the antenna. For power inputs up to 100 watts, the smaller and less expensive RG-58/U can be used. However, when you buy RG-58/U, be sure that the line is made by a reputable manufacturer (such as Amphenol or Belden). Some of the line made for TV installations is of inferior quality and is likely to have higher losses. The feedline was fed up through the mast pipe and through a ¾-inch hole in the 2 by 2. An Amphenol 83-1SP fitting on the end of the coax line connects to the female fitting on the antenna.

Coupling to the Transmitter

It may be found that, when the feed line is coupled to the transmitter, the antenna won't take power. Since the line is terminated at the antenna in its characteristic impedance of 52 ohms, the output of the final r.f. amplifier must be adjusted to couple into a 52-ohm load. Where the output coupling device is a variable link, all that may be needed is the correct setting of the link. If the link is fixed, one end of the link can be grounded to the transmitter chassis and the other end of the link connected in series with a small variable capacitor to the inner conductor of the feed line. The outer conductor of the coax is grounded to the transmitter chassis. The capacitor is tuned to the point where the final amplifier is properly loaded. For transmitters having a pi-network output circuit, it is merely a matter of adjusting the network to the point where the amplifier is properly loaded.

(From *QST*, January, 1955.)

Wave Propagation

Much of the appeal of amateur communication lies in the fact that the results are not always predictable. Transmission conditions on the same frequency vary with the year, season and with the time of day. Although these variations usually follow certain established patterns, 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 worthwhile contributions to the science, provided he has sufficient background to understand his results. He may discover new facts about propagation at the very-high frequencies or in the microwave region, as amateurs have in the past. In fact, it is through amateur efforts that most of the extended-range possibilities of various radio frequencies have been discovered, both by accident and by long and careful investigation.

Characteristics of Radio Waves

Radio waves, like other forms of electromagnetic radiation such as light, travel at a speed of 300,000,000 meters per second in free space, and can be reflected, refracted, and diffracted.

An electromagnetic wave is composed of moving fields of electric and magnetic force. The lines of force in the electric and magnetic fields are at right angles, and are mutually perpendicular to the direction of travel. A simple representation of a wave is shown in Fig. 15-1. In this drawing the electric lines are perpendicular to the earth and the magnetic lines are horizontal. They could, however, have any position with respect to earth so long as they remain perpendicular to each other.

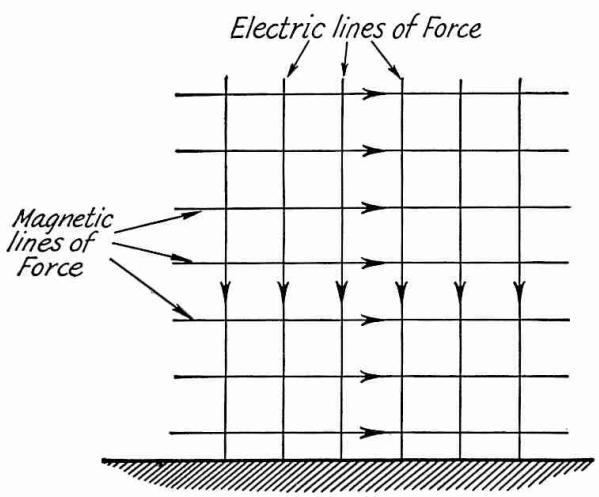


Fig. 15-1—Representation of electric and magnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

The plane containing the continuous lines of electric and magnetic force shown by the grid- or mesh-like drawing in Fig. 15-1 is called the wave **front.**

The **medium** in which electromagnetic waves travel has a marked influence on the speed with which they move. When the medium is empty space the speed, as stated above, is 300,000,000 meters per second. It is almost, but not quite, that great in air, and is much less in some other substances. In dielectrics, for example, the speed is inversely proportional to the square root of the dielectric constant of the material.

When a wave meets a good conductor it cannot penetrate it to any extent (although it will travel through a dielectric with ease) because the electric lines of force are practically short-circuited.

Polarization

The **polarization** of a radio wave is taken as the direction of the lines of force in the electric field. If the electric lines are perpendicular to the earth, the wave is said to be **vertically polarized;** if parallel with 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.

Spreading

The field intensity of a wave is inversely proportional to the distance from the source. Thus if in a uniform medium one receiving point is twice as far from the transmitter as another, the field strength at the more distant point will be just half the field strength at the nearer point. This results from the fact that the energy in the wave front must be distributed over a greater area as the wave moves away from the source. This **inverse-distance law** is based on the assumption that there is nothing in the medium to absorb energy from the wave as it travels. This is not the case in practical communication along the ground and through the atmosphere.

Types of Propagation

According to the altitudes of the paths along which they are propagated, radio waves may

be classified as **ionospheric waves, tropospheric waves** or **ground waves.**

The ionospheric wave or **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 wave length, 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 may occur at 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. 15-2.

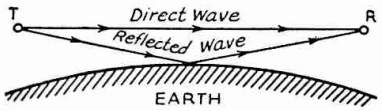


Fig. 15-2—Showing how both direct and reflected waves may be received simultaneously.

Ionospheric Propagation

PROPERTIES OF THE IONOSPHERE

Except for distances of a few miles, nearly all amateur communication on frequencies below 30 Mc. is by means of the sky 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 have an appreciable effect on wave travel.

The ionization in the upper atmosphere is believed to be caused by 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

The greater the intensity of ionization in a layer, the more the path of the wave is bent. The bending, or refraction (often also called reflection), also depends on the wavelength; the longer the wave, the more the path is bent for a given degree of ionization. Thus low-frequency waves are more readily bent than those of high frequency. For this reason the lower frequencies — 3.5 and 7 Mc. — are more "reliable" than the higher frequencies — 14 to 28 Mc.; there are times when the ionization is of such low value that waves of the latter frequency range are not bent enough to return to earth.

Absorption

In traveling through the ionosphere the wave gives up some of its energy by setting the ionized particles into motion. When the moving ionized particles collide with others this energy is lost. The **absorption** from this cause is greater at lower frequencies. It also increases with the intensity of ionization, and with the density of the atmosphere in the ionized region.

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 bending that actually takes place, as illustrated in Fig. 15-3. 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 having equal sides of a total length proportional to the time taken for the wave to travel from *T* to *R*.

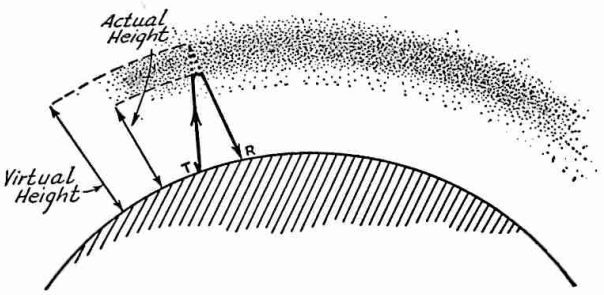


Fig. 15-3—Bending in the ionosphere, and the echo or reflection method of determining virtual height.

Normal Structure of the Ionosphere

The lowest useful ionized layer is called the ***E* layer.** The average height of the region of maximum ionization is about 70 miles. The air at this height is sufficiently dense so that the ions and electrons set free by the sun's radiation do not travel far before they meet and recombine to form neutral particles, so the layer can maintain its normal intensity of ionization only in the presence of continuing radiation from the sun. Hence the ionization is greatest around local noon and practically disappears after sundown.

In the daytime there is a still lower ionized

area, the ***D* region.** *D*-region ionization is proportional to the height of the sun and is greatest at noon. The lower amateur-band frequencies (1.8 and 3.5 Mc.) are almost completely absorbed by this layer, and only the high-angle radiation is reflected by the *E* layer. (Lower-angle radiation travels farther through the *D* region 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. 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.

SKY-WAVE PROPAGATION

Wave Angle

The smaller the angle at which a wave leaves the earth, the less the bending required in the ionosphere to bring it back. Also, the smaller the angle the greater the distance between the point where the wave leaves the earth and that at which it returns. This is shown in Fig. 15-4. The vertical angle that the wave makes with a tangent to the earth is called the **wave angle** or **angle of radiation.**

Skip Distance

More bending is required to return the wave to earth when the wave angle is high, and at times the bending will not be sufficient unless the wave angle is smaller than some critical value. This is illustrated in Fig. 15-4, where *A* and smaller angles give useful signals while waves sent at higher angles penetrate the layer and are not returned. The distance between *T* and *R*$_1$ is, therefore, the shortest possible distance, at that particular frequency, over which communication by 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,** and the distance from the transmitter to the nearest point where the sky wave returns to earth is called the **skip distance.** The extent of the skip zone depends upon the frequency and the state of the ionosphere, and also upon the height of the layer in which the refraction takes place. The higher layers give 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.

Critical and Maximum Usable Frequencies

If the frequency is low enough, a wave sent vertically to the ionosphere will be reflected back down to the transmitting point. If the frequency is then gradually increased, eventually a frequency will be reached where this vertical reflection just fails to occur. This is the **critical frequency** for the layer under consideration. When the operating frequency is below the critical value there is no skip zone.

The critical frequency is a useful index to the highest frequency that can be used to transmit over a specified distance — the **maximum usable frequency (m.u.f.).** If the wave leaving the transmitting point at angle *A* in Fig. 15-4 is, for example, at a frequency of 14 Mc., and if a higher frequency would skip over the receiving point *R*$_1$, then 14 Mc. is the m.u.f. for the distance from *T* to *R*$_1$.

The greatest possible distance is covered when the wave leaves along the tangent to the earth; that is, at zero wave angle. Under average conditions this distance is about 4000 kilometers or 2500 miles for the *F*$_2$ layer, and 2000 km. or 1250 miles for the *E* layer. The distances vary with the layer height. Frequencies above these limiting m.u.f.'s will not be returned to earth at any distance. The 4000-km. m.u.f. for the *F*$_2$ layer is approximately 3 times the critical frequency for that layer, and for the *E* layer the 2000-km. m.u.f. is about 5 times the critical frequency.

Absorption in the ionosphere is least at the maximum usable frequency, and increases very rapidly as the frequency is lowered below the m.u.f. Consequently, best results with low power always are secured when the frequency is as close to the m.u.f. as possible.

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.

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

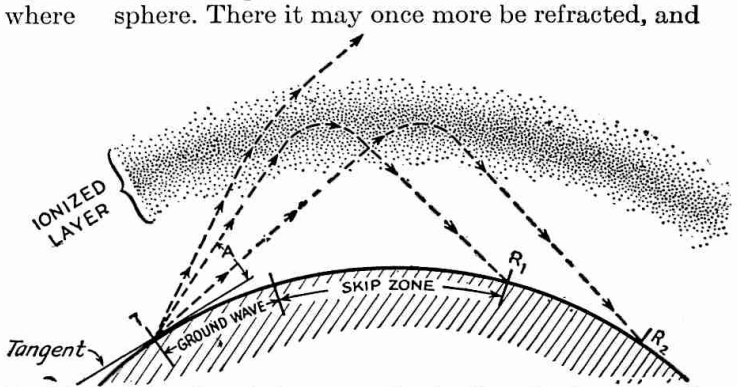


Fig. 15-4—Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than A) are not bent enough to be returned to earth. As the angle is decreased, the waves return to earth at increasingly greater distances.

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, which restrict the maximum one-hop distance to the values mentioned in the preceding section. 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), and there is also absorption in the ionosphere at each reflection. Hence the smaller the number of hops the greater the signal strength at the receiver, other things being equal.

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 total field strength will be the sum of the components and may be larger or smaller than one component alone, since the phases may be such as either to aid or oppose. 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.

Fading may be either 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 side band components may not be propagated in the same relative amplitudes and phases they had at the transmitter. This effect, known as **selective fading,** causes severe distortion of the signal.

Back Scatter

Even though the operating frequency is above the m.u.f. for a given distance, it is usually possible to hear signals from within the skip zone. This phenomenon, called **back scatter,** is caused by reflections from distances beyond the skip zone. Such reflections can occur when the transmitted energy strikes the earth at a distance and some of it is reflected back into the skip zone to the receiver. Such scatter signals are weaker than those normally propagated, and also have a rapid fade or "flutter" that makes them easily recognizable.

A certain amount of scattering of the wave also takes place in the ionosphere because the ionized region is not completely uniform. Scattering in the normal propagation direction is called **forward scatter,** and is responsible for extending the range of transmission beyond the distance of a regular hop, and for making communication possible on frequencies greater than the actual m.u.f.

● OTHER FEATURES OF IONOSPHERIC PROPAGATION

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 critical frequency is 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 daytime maximum critical frequencies for the F_2 are highest in winter (10 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. These values are representative of latitude 40 deg. North in the Western hemisphere, and are subject to considerable variation in other parts of the world.

Very marked changes in ionization also occur in step with the **11-year sunspot cycle.** Although there is no apparent direct correlation between sunspot activity and critical frequencies on a given day, there is a definite correlation between *average* sunspot activity and critical frequencies. The critical frequencies are highest during sunspot maxima and lowest during sunspot minima. During the period of minimum sunspot activity the lower frequencies — 7 and 3.5 Mc. — frequently are the only usable bands at night. 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.

Ionosphere Storms

Certain types of sunspot activity cause considerable disturbances in the ionosphere **(ionosphere storms)** and are accompanied by disturbances in the earth's magnetic field **(magnetic storms).** Ionosphere storms are characterized by a marked increase in absorption, so that radio conditions become poor. The critical frequencies also drop to relatively low values during a storm, so that only the lower frequencies are useful for communication. Ionosphere storms may last from a few hours to several days. Since the sun rotates on its axis once every 28 days, disturbances tend to recur at such intervals, if the sunspots responsible do not become inactive in the meantime. Absorption is usually low, and radio conditions therefore good, just preceding a storm.

Sporadic-E Ionization

Scattered patches or clouds of relatively dense ionization occasionally appear at heights approximately the same as that of the E layer, for rea-

sons not yet known. This sporadic-*E* ionization is most prevalent in the equatorial regions, where it is substantially continuous. In northern latitudes it is most frequent in the spring and early summer, but is present in some degree a fair percentage of the time the year 'round. It accounts for a good deal of the night-time short distance work on the lower frequencies (3.5 and 7 Mc.) and, when more intense, for similar work on 14 to 28 Mc. Exceptionally intense sporadic-*E* ionization is responsible for work over distances exceeding 400 or 500 miles on the 50-Mc. band.

There are indications of a relationship between sporadic-*E* ionization and average sunspot activity, but it does not appear to be directly related to daylight and darkness since it may occur at any time of the day. However, there is an apparent tendency for the ionization to peak at mid-morning and in the early evening.

Tropospheric Propagation

Changes in temperature and humidity of air masses in the lower atmosphere often permit work over greater than normal ground-wave 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 later.

PREDICTION CHARTS

The Central Radio Propagation Laboratory of National Bureau of Standards offers prediction charts three months in advance, by means of which 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 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. They are called "*CRPL-D Basic Radio Propagation Predictions.*" The use of the charts is explained in Circular 462, "*Ionospheric Radio Propagation,*" available for $1.00 from the same address. This publication also contains much information of value to those who wish to pursue the subject of ionospheric propagation in more detail.

PROPAGATION IN THE 3.5 TO 30-MC. BANDS

The 1.8-Mc., or "160-meter," band offers reliable working over ranges up to 25 miles or so during daylight. On winter nights, ranges up to several thousand miles are not impossible. Only small sections of the band are currently available to amateurs, because of the presence of the loran service in that part of the spectrum.

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 200 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.

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. 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 the high portion of the sunspot cycle it is open to some part of the world during practically all of the 24 hours, while during a sunspot minimum it is generally useful only during daylight hours and the dawn and dusk periods. There is practically always a skip zone on this band.

The 21-Mc., or "15-meter," band shows highly variable characteristics depending on the sunspot cycle. During sunspot maxima it is useful for long-distance work during a large part of the 24 hours, but in years of low sunspot activity it is almost wholly a daytime band, and sometimes unusable even in daytime. However, it is often possible to maintain communication over distances up to 1500 miles or more by sporadic-*E* ionization which may occur either day or night at any time in the sunspot cycle.

The 28-Mc. ("10-meter") band is generally considered to be a DX band during the daylight hours (except in summer) and good for local work during the hours of darkness, for about half the sunspot cycle. At the very peak of the sunspot cycle, it may be "open" into the late evening hours for DX communication. At the sunspot minimum the band is usually "dead" for long-distance communication, by means of the F_2 layer, in the northern latitudes. Nevertheless, sporadic-*E* propagation is likely to occur at any time, just as in the case of the 21-Mc. band.

Propagation Above 50 Mc.

The importance to the amateur of having some knowledge of wave propagation was stressed at the beginning of this chapter. An understanding of the means by which his signals reach their destination is an even greater aid to the v.h.f. worker. Each of his bands shows different characteristics, and knowledge of their peculiarities is as yet far from complete. The observant user of the amateur v.h.f. assignments has a good opportunity to contribute to that knowledge, and

his enjoyment of his work will be greatly enhanced if he knows when to expect unusual propagation conditions.

CHARACTERISTICS OF THE V.H.F. BANDS

An outstanding feature of our bands from 50 Mc. up is their ability to provide consistent and interference-free communication within a limited range. All lower frequencies are subject to varying conditions that impair their effectiveness for work over distances of 100 miles or less at least part of the time, and the heavy occupancy they support results in severe interference problems in areas of dense population. The v.h.f. bands, being much wider, can handle many times the amateur population without crowding, and their characteristics for local work are more stable. It is thus to the advantage of amateur radio as a whole to make use of 50 Mc. and higher bands for short-range communication wherever possible.

In addition to reliable local coverage, the v.h.f. bands also exhibit several forms of long-distance propagation at times, and use of 50 and 144 Mc. has been taken up in recent years by many isolated amateurs who must depend on these propagation peculiarities for all or most of their contacts. It is particularly important to these operators that they understand common propagation phenomena. The material to follow supplements information presented earlier in this chapter, but deals with wave propagation only as it affects the occupants of the world above 50 Mc. First let us consider the bands individually.

50 to 54 Mc.: This band is borderline territory between the DX frequencies and those normally employed for local work. Thus just about every form of wave propagation found throughout the radio spectrum appears, on occasion, in the 50-Mc. region. This has contributed greatly to the popularity of the 50-Mc. band.

During the peak years of a sunspot cycle it is occasionally possible to work 50-Mc. DX of world-wide proportions, by reflection of signals from the F_2 layer. Sporadic-E skip provides contacts over distances from 400 to 2500 miles or so during the early summer months, regardless of the solar cycle. Reflection from the aurora regions allows 100- to 1000-mile work during pronounced ionospheric disturbances. The ever-changing weather pattern offers extension of the normal coverage to as much as 300 to 500 miles. This develops most often during the warmer months, but may occur at any season. In the absence of any favorable propagation, the average well-equipped 50-Mc. station should be able to work regularly over a radius of 75 to 100 miles or more, depending on local terrain.

144 to 148 Mc.: Ionospheric effects are greatly reduced at 144 Mc. F_2-layer reflection is unlikely, and sporadic-E skip is rare. Aurora DX is fairly common, but signals are generally weaker than on 50 Mc. Tropospheric effects are more pronounced than on 50 Mc., and distances covered during favorable weather conditions are greater than on lower bands. Air-mass boundary bending has been responsible for communication on 144 Mc. over distances in excess of 2500 miles, and 500-mile work is fairly common in the warmer months. The reliable range under normal conditions is slightly less than on 50 Mc., with comparable equipment.

220 Mc. and Higher: Ionospheric propagation is unlikely at 220 Mc. and up, but tropospheric bending is more prevalent than on lower bands. Amateur experience on 220 and 420 Mc. is showing that they can be as useful as 144 Mc., when comparable equipment is used. Under minimum conditions the range may be slightly shorter, but when signals are good on 144 Mc., they may be better on 220 or 420. Even above 1000 Mc. there is evidence of tropospheric DX.

PROPAGATION PHENOMENA

The various known means by which v.h.f. signals may be propagated over unusual distances are discussed below.

F_2-Layer Reflection: Most contacts made on 28 Mc. and lower frequencies are the result of reflection of the 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 follows other well-defined cycles, daily, monthly, and seasonal, 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 in the early '50s, 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 instances of long-distance work on 50 Mc. by F_2-layer reflection, and as late as 1950 contacts were made in the more favorable areas of the world by this medium. The rising curve of the current solar cycle again made F_2 DX on 50 Mc. possible in the low latitudes in the winter of 1955–6. DX was worked over much of the earth in 1956–8 and may be expected through 1959. Loss of the 50-Mc. band to television in some countries will limit the scope of 50-Mc. DX in years to come.

The F_2 m.u.f. is readily determined by observation, and it may be estimated quite accurately for any path at any time. It is predictable for months in advance, enabling the v.h.f. worker to arrange test schedules with distant stations at propitious times. As there are numerous commercial signals, both harmonics and fundamental transmissions, on the air in the range between 28 and 50 Mc., it is possible to determine the approximate m.u.f. by careful listening in this range. Daily observations will show if the m.u.f. is rising or falling, and once the peak for a given month is determined it can be assumed that another will occur about 27 days later, this cycle coinciding with the turning of

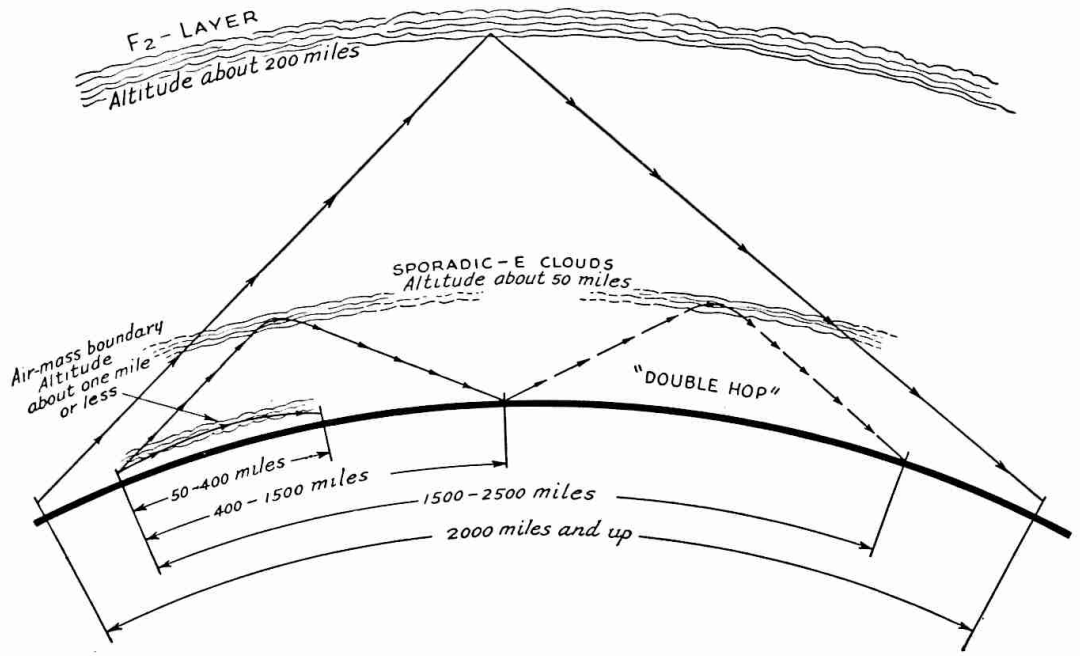


Fig. 15-5—The principal means by which v.h.f. signals may be returned to earth, showing the approximate distances over which they are effective. The F_2 layer, highest of the reflecting layers, may provide 50-Mc. DX at the peak of the 11-year sunspot cycle. Such communication may be world-wide in scope. Sporadic ionization of the *E* region produces the familiar "short skip" on 28 and 50 Mc. It is most common in early summer and in late December, but may occur at any time, regardless of the sunspot cycle. Refraction of v.h.f. waves also takes place at air-mass boundaries, making possible communication over distances of several hundred miles on all v.h.f. bands. Normally it exhibits no skip zone.

the sun on its axis. The working range, via F_2 skip, is roughly comparable to that on 28 Mc., though the *minimum* distance is somewhat longer. Two-way work on 50 Mc. by reflection from the F_2 layer has been accomplished over distances from 2200 to 12,000 miles. The maximum frequency for F_2 reflection is believed to be about 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 morning and early evening hours, but it may occur at any time or season. Multiple-hop effects may appear, when ionization develops simultaneously over large areas, making possible work over distances of more than 2500 miles.

The upper limit of frequency for sporadic-*E* skip is not positively known, but scattered instances of 144-Mc. propagation over distances in excess of 1000 miles indicate that *E*-layer reflection, possibly aided by tropospheric effects, may be responsible.

Aurora Effect: Low-frequency communication is occasionally wiped out by absorption in the ionosphere, when ionospheric storms, associated with variations in the earth's magnetic field, occur. During such disturbances, however, v.h.f. signals may be reflected back to earth, making communication possible over distances not normally workable in the v.h.f. range. Magnetic storms may be accompanied by an aurora-borealis display, if the disturbance occurs at night and visibility is good. Aiming a directional array at the auroral curtain will bring in signals strongest, regardless of the true direction to the transmitting station.

Aurora-reflected signals are characterized by a rapid flutter, which lends a "dribbling" sound to 28-Mc. carriers and may render modulation on 50- and 144-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 1000 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 is observed frequently on 50 and 144 Mc. in northeastern U. S. A., usually in the early evening hours or after midnight. The highest frequency for auroral reflection is not yet known, but pronounced disturbances have permitted work by this medium in the 220-Mc. band.

Tropospheric Bending: The most common form of v.h.f. DX is the extension of the normal operating range associated with easily observed weather phenomena. It is the result of the change in refractive index of the atmosphere at the boundary between air masses of differing temperature and humidity characteristics. Such air-mass boundaries usually lie along the western or southern edges of a stable slow-moving area of high barometric pressure (fair, calm weather) in the period prior to the arrival of a storm.

A typical upper-air sounding showing temperature and water-vapor gradients favorable to v.h.f. DX is shown in Fig. 15-6. An increase in temperature and a sharp drop in water-vapor

gradient are seen at about 4000 feet, in comparison to the U. S. Standard Atmosphere curves at the left.

Such a favorable condition develops most often in the late summer or early fall, along the junction between air masses that may have come together from such widely separated points as the Gulf of Mexico and Northern Canada. Under stable weather conditions the two air masses may retain their original character for several days at a time, usually moving slowly eastward across the country. When the path between two v.h.f. stations separated by fifty to several hundred miles lies along such a boundary, signal levels run far above the average value.

Many factors other than air-mass movement of a continental character provide increased v.h.f. operating range. The convection along 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 to rise, may be the best of the day for extended v.h.f. range, particularly in clear, calm weather, when the barometer is high and the humidity low.

The 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 can 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. The 50-Mc. band is 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 lower frequencies are relatively inactive. It is probable that this tendency continues on up through the microwave range, and there is good evidence to indicate 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.

Scatter: Forward scatter, both ionospheric and tropospheric, may be used for marginal communication in the v.h.f. bands. Both provide very weak but consistent signals over distances that were once thought impossible on frequencies higher than about 30 Mc.

Tropospheric scatter is prevalent all through the v.h.f. and microwave regions, and is usable over distances up to about 400 miles. Ionospheric scatter, augmented by meteor bursts, brings in signals over 600 to 1300 miles, on frequencies up to about 100 Mc. Either form of scatter requires high power, large antennas and c.w. technique to provide effective communication.

Back scatter, of the type heard on lower bands, is also heard occasionally on 50 Mc., when F_2 or sporadic-E skip is present.

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. Ordinarily such reflections are of little value in communication, since the increases in signal strength are of short duration, but meteor showers of considerable magnitude and duration may provide fluttery signals from distances up to 1500 miles or more on both 50 and 144 Mc.

As meteor-burst signals are relatively weak, their detection is greatly aided if high power and high-gain antennas are used. Two-way communication of sorts has been carried on by this medium on 50 and 144 Mc. over distances of 600 to 1300 miles, through the use of short c.w. transmissions and frequent repetition.

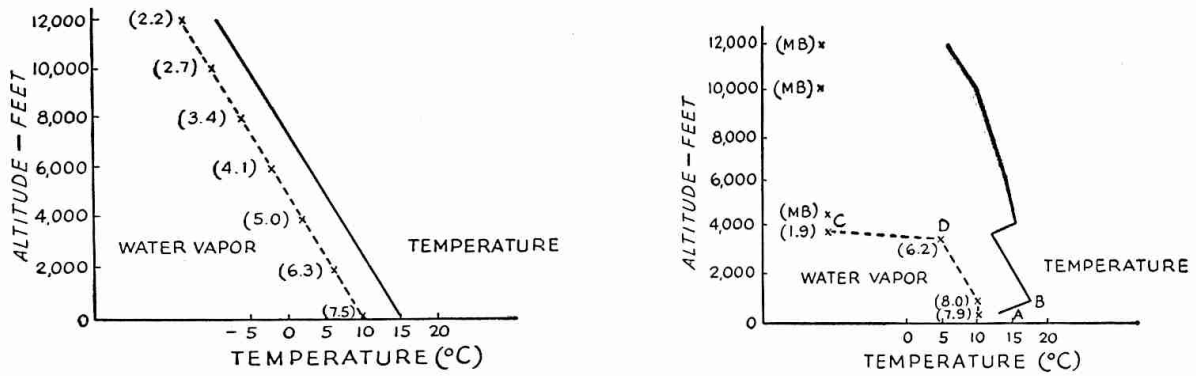


Fig. 15-5—Upper-air conditions that produce extended-range communication on the v.h.f. bands. At the left is shown the U. S. Standard Atmosphere temperature curve. The humidity curve (dotted) is that which would result if the relative humidity were 70 per cent from the ground level to 12,000 feet elevation. There is only slight refraction under this standard condition. At the right is shown a sounding that is typical of marked refraction of v.h.f. waves. Figures in parentheses are the "mixing ratio"—grams of water vapor per kilogram of dry air. Note the sharp break in both curves at about 4000 feet. (From Collier, "Upper-Air Conditions for 2-Meter DX," *QST*, September, 1955.)

V.H.F. Receivers

Good receiving facilities are all-important in v.h.f. work. High sensitivity, adequate stability and good signal-to-noise ratio, necessary attributes in a receiving system for 50 Mc. and higher frequencies, are most readily attained through the use of a converter working into a communications receiver designed for lower frequencies. Though receivers and converters for the v.h.f. bands are available on the amateur market, the amateur worker can build his own with fully as good results, usually at a considerable saving in cost.

Basically, modern v.h.f. receiving equipment is little different from that employed on lower frequencies. The same order of selectivity may be used on all amateur frequencies up to at least 450 Mc. The greatest practical selectivity should be employed in v.h.f. reception, as it not only allows more stations to operate in a given band, but is an important factor in improving the signal-to-noise ratio. The effective sensitivity of a receiver having "communication" selectivity can be made much better than is possible with broadband systems.

This rules out converted radar-type receivers and others using high intermediate frequencies. The superregenerative receiver, a simple but broadband device that was popular in the early days of v.h.f. work, is now used principally for portable operation, or for other applications where high sensitivity and selectivity are not of prime importance. It is capable of surprising performance, for a given number of tubes and components, but its lack of selectivity, its poor signal-to-noise ratio, and its tendency to radiate a strong interfering signal have eliminated the superregenerator as a fixed-station receiver in areas where there is appreciable v.h.f. activity.

R. F. AMPLIFIER DESIGN

The noise generated within the receiver itself is an important factor in the effectiveness of v.h.f. receiving gear. At lower frequencies, and to a considerable extent on 50 Mc., external noise is a limiting factor. At 144 Mc. and higher the receiver noise figure, gain and selectivity determine the ability of the system to respond to weak signals. Proper selection of r.f. amplifier tubes and appropriate circuit design aimed at low noise figure are more important in the v.h.f. receiver "front end" than mere gain.

Triode or Pentode?

Certain triode tubes have been developed with this end in view. Their superiority over pentode types is more pronounced as we go higher in frequency. Because of the limitation on sensitivity imposed by external noise at that frequency, triode or pentode r.f. amplifiers give about the same results at 50 Mc. Thus the pentode types, which offer the advantages of better selectivity and simpler circuitry, are often used for 50-Mc. work. But at 144 Mc., the newer triodes designed for r.f. amplifier service give fully as much gain as the pentodes, and with lower internal noise. With the exception of the simplest unit, the equipment described in the following pages incorporates low-noise r.f. amplifier techniques.

Neutralizing Methods

When triodes are used as r.f. amplifiers some form of neutralization of the grid-plate capacitance is required. This can be capacitive, as is commonly used in transmitting applications, or inductive. The alternative to neutralization is the use of grounded-grid technique. Circuits for v.h.f. triode r.f. amplifier stages are given in Figs. 16-1 through 16-4.

A dual triode operated as a neutralized push-pull amplifier is shown at 16-1. This arrangement is well adapted to v.h.f. preamplifier applications, or as the first stage in a converter, particularly when a balanced transmission line such as the popular 300-ohm Twin-Lead is used. It is relatively selective and may require resistive loading of the plate circuit, when used as a preamplifier. The loading effect of the following circuit may be sufficient to give the required band width, when the push-pull stage is inductively coupled to the mixer.

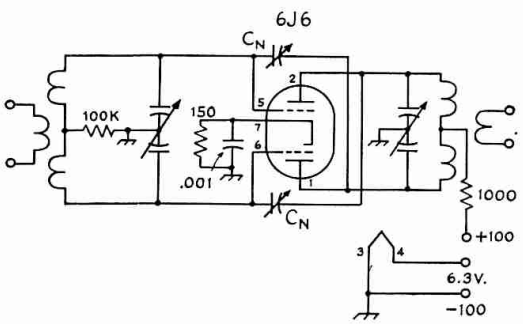


Fig. 16-1—Schematic diagram of a push-pull r.f. amplifier for v.h.f. applications. This circuit is well-suited to use with antenna systems having balanced lines. Coil and capacitor values not given depend on the frequency at which the amplifier is to be used. Neutralizing capacitance, C_N, may be built up by twisting ends of insulated leads together.

A triode amplifier having excellent noise figure and broadband characteristics is shown in Fig.

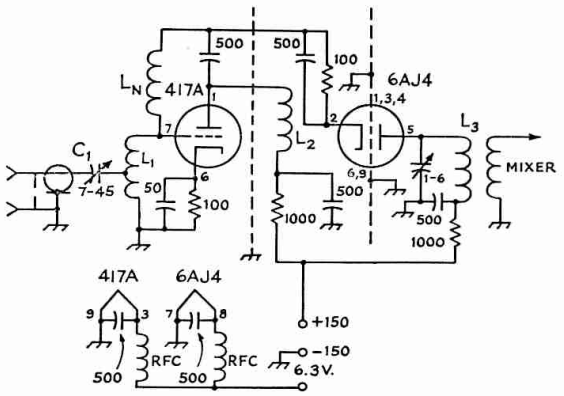


Fig. 16-2—Circuit of the cascode r.f. amplifier. Coupling capacitor, C_1, may be omitted if spurious receiver responses are not a problem. Neutralizing winding, L_N should resonate at the signal frequency with the grid-plate capacitance of the first tube. Base connections are for 417A and 6AJ4, but other small triodes may be used.

16-2. Commonly called the cascode, it uses a triode or triode-connected pentode followed by a triode grounded-grid stage. This circuit is extremely stable and uncritical in adjustment. At 50 Mc. and higher its over-all gain is at least equal to the best single-stage pentode amplifier and its noise figure is far lower.

Neutralization is accomplished by the coil L_N, whose value is such that it resonates at the signal frequency with the grid-plate capacitance of the tube. Its inductance is not critical; it may be omitted from the circuit without the stage going into oscillation, but neutralization results in a lower noise figure than is possible without it. Any of several v.h.f. tubes may be used in the cascode circuit. The example shown in Fig. 16-2 uses the 417A, followed by a 6AJ4. Two 6AJ4s would work almost equally well, as would the 6AM4, 6AN4 and 6BC4. Pin connections in Fig. 16-2 should be changed to suit the tubes selected.

A simplified version of the cascode, using a dual triode tube designed especially for this application, is shown in Fig. 16-3. By reducing stray capacitance, through direct coupling between the two triode sections, this circuit makes for improved performance at the frequencies above 100 Mc. The two sections of the tube are in series, as far as plate voltage is concerned, so

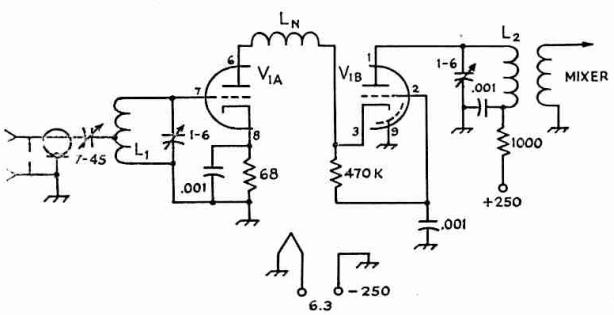


Fig. 16-3—Simplified cascode circuit for use with dual triodes having separate cathodes. Coil and capacitance values not given depend on frequency. Bifilar r.f. chokes are occasionally used in heater leads.

it requires higher voltage than the other circuits shown.

The neutralization process for the cascode and neutralized-triode amplifiers is somewhat similar. With the circuit operating normally the neutralizing adjustments (capacitance of C_N in Fig. 16-1; inductance of L_N in Figs. 16-2 and 16-3) can be set for best signal-to-noise ratio. The best results are obtained using a noise generator, adjusting for lowest noise figure, but careful adjustment on a weak signal provides a fair approximation. Noise generators and their use in v.h.f. receiver adjustment are treated in July, 1953, *QST*, p. 10, and in this *Handbook*, Chapter 21.

Grounded-grid r.f. amplifier technique is illustrated in Figs. 16-4 and 16-14. Here the input is in the cathode lead, with the grid of the tube grounded, to act as a shield between cathode and plate. The grounded-grid circuit is stable and easily adjusted, and is well adapted to broadband applications. The gain per stage is low, so that two or more stages may be required.

Tubes well-suited to grounded-grid amplifier service include the 6J4, 6AN4, 6AJ4, 6AM4, 6BC4, 417A and 416B. Disk-seal tubes such as the "lighthouse" and "pencil tube" types are often used as r.f. amplifiers above 500 Mc., and the new ceramic tubes show great possibilities for r.f. amplifier service in the u.h.f. range.

Great care should be used in adjusting the r.f. portion of a v.h.f. receiver, whatever circuit is used. If it is working properly it will control the noise figure of the entire system.

Reducing Spurious Responses

In areas where there is a high level of v.h.f. activity or extensive use of other frequencies in the v.h.f. range, the ability of the receiver to operate properly in the presence of strong signals may be an important consideration. Special tube types, otherwise similar to older numbers, have been developed for low overload and cross-modulation susceptibility. The 6BC8, which may be used as a replacement for the 6BQ7A or 6BZ7, is one of these.

Modification of the converter design can also improve performance in these respects. In general, the gain ahead of the mixer stage should be made no more than is necessary to achieve good noise figure characteristics. The plate voltage on the r.f. amplifier should be kept as high as practical, to prevent easy overloading.

Rejection of signals outside the desired frequency range can be improved by the use of high-Q tuned circuits ahead of the first r.f. amplifier stage. Television transmitters are particularly troublesome in this respect, and one or more coaxial-type circuits inserted in the lead from the antenna to the converter may be necessary to keep such signals from interfering with normal reception.

A common cause of unwanted signals appearing in the tuning range is the presence of oscillator harmonics in the energy being fed to the mixer of a crystal-controlled converter. This may be pre-

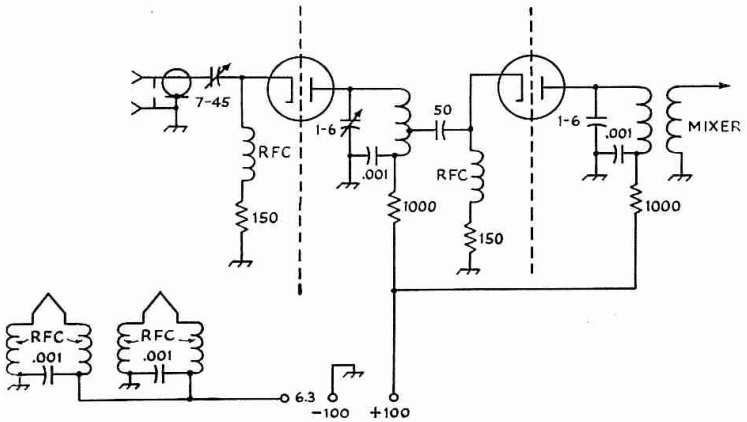


Fig. 16-4—Grounded-grid amplifier. Position of tap on plate coil should be adjusted for lowest noise figure. Low gain with this circuit makes two stages necessary for most applications. R.f. choke and coil values depend on frequency.

vented by using a high oscillator frequency, to keep down the number of multiplications, and by shielding the oscillator and multiplier stages from the rest of the converter.

Signals at the intermediate frequency may ride through a converter. This can be prevented by keeping down capacitive interstage coupling in the r.f. circuitry, and by shielding the converter and the receiver antenna terminals. The problem of receiver responses is dealt with in *QST* for April, 1955, p. 56, and February, 1958, p. 27.

MIXER CIRCUITS

The mixer in a v.h.f. converter may be either a pentode or a triode tube. Pentodes give generally higher output, and may require less injection. When used without a preceding r.f. amplifier stage, the triode mixer may provide a better noise figure. With either tube, the grid circuit is tuned to the signal frequency, and the plate circuit to the intermediate frequency.

A simple triode mixer is shown in Fig. 16-5A, with a pentode mixer at B. A dual-triode version (push-push mixer) is shown at C. The push-push mixer is well adapted to use at 420 Mc., and may, of course, be used at any lower frequency. Dual tubes may be used as both mixer and oscillator, combining the circuits of Figs. 16-5 and 16-6. A 6U8 could use its pentode as a mixer (16-5B) and the oscillator portion (16-6A) would be a triode. Dual-triode tubes (6J6, 12AT7 and many others) would combine 16-5A and 16-6A. In dual triodes having separate cathodes some external coupling may be required, but the common cathode of the 6J6 will provide sufficient injection in most cases. If the injection is more than necessary it can be reduced by dropping the oscillator plate voltage, either directly or by increasing the value of the dropping resistor.

A pentode mixer is less subject to oscillator pulling than a triode, and it will probably require less injection voltage. In a pentode mixer with no r.f. amplifier, plate current should be held to the lowest usable value, to reduce tube noise. This may be controlled by varying the mixer screen voltage. When a good r.f. amplifier is used the mixer plate current may be run higher, for better operation with strong signals.

Occasionally oscillation near the signal frequency may be encountered in v.h.f. mixers. This usually results from stray lead inductance in the mixer plate circuit, and is most common with triode mixers. It may be corrected by connecting a small capacitance from plate to cathode, *directly* at the tube socket. Ten to 25 $\mu\mu$f. will be sufficient, depending on the signal frequency.

OSCILLATOR STABILITY

When a high-selectivity i.f. system is employed in v.h.f. reception, the stability of the oscillator is extremely important. Slight variations in oscillator frequency that would not be noticed when a broadband i.f. amplifier is used become intolerable when the passband is reduced to crystal-filter proportions.

One satisfactory solution to this problem is the use of a crystal-controlled oscillator, with frequency multipliers if needed, to supply the injection voltage. Such a converter usually employs one or more broadband r.f. amplifier stages, and tuning is done by tuning the receiver with which the converter is used to cover the desired intermediate frequency range.

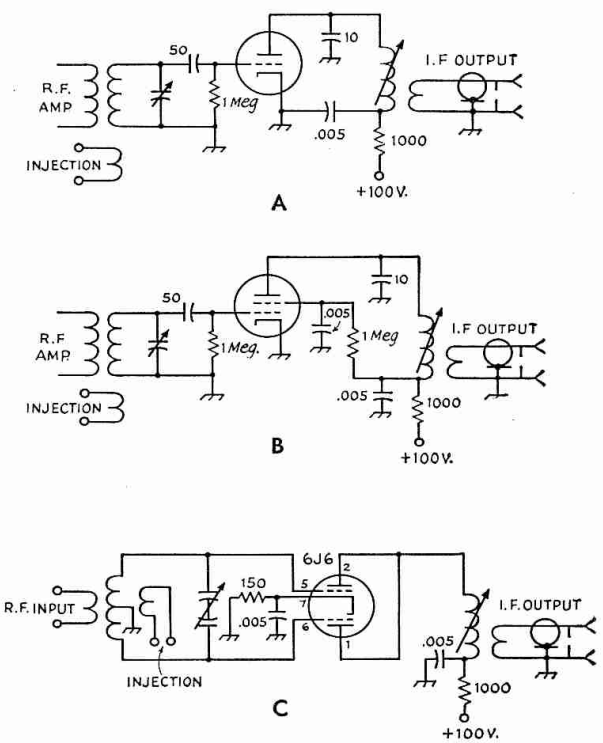


Fig. 16-5—Typical v.h.f. mixer circuits for triode (A), pentode (B) and push-push triode (C). Circuits A and B may be used with one portion of various dual-purpose tubes. Plate current of pentode (B) should be held at lowest usable value if no r.f. stage is used.

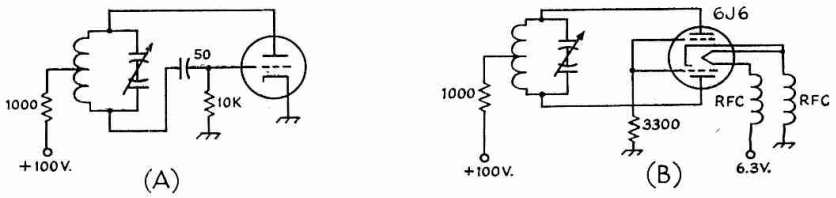


Fig. 16-6—Recommended oscillator circuits for tunable v.h.f. converters. Dual-triode-version (B) is recommended for 220 or 420 Mc. R.f. choke coil and capacitor values not given depend on frequency.

When a tunable oscillator and a fixed intermediate frequency are used, special attention must be paid to the oscillator design, to be sure that it is mechanically and electrically stable. The tuning capacitor should be solidly built, preferably of the double-bearing type. Split-stator capacitors specifically designed for v.h.f. service, usually having ball-bearing end plates and special construction to insure short leads, are well worth their extra cost. Leads should be made with stiff wire, to reduce vibration effects. Mechanical stability of air-wound coils can be improved by tying the turns together with narrow strips of household cement at several points.

Recommended oscillator circuits for v.h.f. work are shown in Fig. 16-6. The single-ended oscillator may be used for 50 or 144 Mc. with good results. The push-pull version is recommended for higher frequencies and may also be used on the two lower bands, as well. Circuit A works well with almost any small triode, or one half of a 6J6 or 12AT7. The 6J6 is well suited to push-pull applications, as shown in circuit 16-6B.

THE I.F. AMPLIFIER

Superheterodyne receivers for 50 Mc. and up should have fairly high intermediate frequencies, to reduce both oscillator pulling and image response. Approximately 10 per cent of the signal frequency is commonly used, with 10.7 Mc. being set up as the standard i.f. for commercially-built f.m. receivers. This particular frequency has a disadvantage for 50-Mc. work, in that it makes the receiver subject to image response from 28-Mc. signals, if the oscillator is on the low side of the signal frequency. A spot around 7 Mc. is favored for amateur converter service, as practically all communications receivers are capable of tuning this range.

For selectivity with a reasonable number of i.f. stages, double conversion is usually employed in complete receivers for the v.h.f. range. A 7-Mc. intermediate frequency, for instance, is changed to 455 kc., by the addition of a second mixer-oscillator. This procedure is, of course, inherent in the use of a v.h.f. converter ahead of a communications receiver.

If the receiver so used is lacking in sensitivity, the over-all gain of the converter-receiver combination may be inadequate. This can be corrected by building an i.f. amplifier stage into the converter itself. Such a stage is useful even when the gain of the system is adequate without it, as the gain control can be used to permit operation of the converter with receivers of widely different performance. If the receiver has an S-meter, its adjustment may be left in the position used for lower frequencies, and the converter gain set so as to make the meter read normally on v.h.f. signals.

Where reception of wide-band f.m. or unstable signals of modulated oscillators is desired, a converter may be used ahead of an f.m. broadcast receiver. 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 microwave region, where stabilized transmission is extremely difficult at the current state of the art.

THE SUPERREGENERATIVE RECEIVER

The simplest type of v.h.f. receiver is the superregenerator. It affords fair sensitivity with few tubes and elementary circuits, but its weaknesses, listed earlier, have relegated it to applications where small size and low power consumption are important considerations.

Its sensitivity results from the use of an alternating quenching voltage, usually in the range between 20 and 200 kc., to interrupt the normal oscillation of a regenerative detector. The regeneration can thus be increased far beyond the amount usable in a straight regenerative circuit.

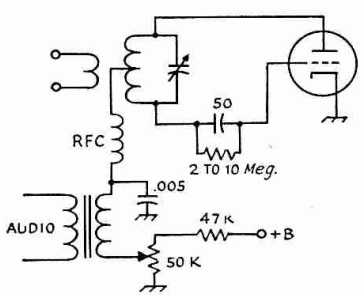


Fig. 16-7—Superregenerative detector circuit for self-quenched detector. Pentode tube may be used, varying screen voltage by means of the potentiometer to control regeneration.

The detector itself can be made to furnish the quenching voltage, or a separate oscillator tube can be used. Regeneration is usually controlled by varying the plate voltage in triode detectors, or the screen voltage in the case of pentodes. A typical circuit is shown in Fig. 16-7.

Crystal-Controlled Converters for 50, 144 and 220 Mc.

The three converters and their power supply, shown below, were designed to meet the special requirements of each of the v.h.f. bands, insofar as possible. They offer high stability and reasonably low noise figure, and special attention was paid to the reduction of spurious responses, particularly in the converters for 50 and 220 Mc. Each unit plugs into the power supply, which also includes the i.f. output circuitry. Anyone interested in one or two of the bands can thus build for his own purposes and omit the other band or bands. The i.f. tuning range is 7 to 11 Mc. for 50- and 144-Mc. coverage, and 7-12 Mc. for the 220-Mc. band.

THE 50-MC. CONVERTER

A pentode r.f. amplifier stage is used in the 50-Mc. converter, Figs. 16-9 and 16-10. With proper design and adjustment such a stage will have a noise figure sufficiently low that it will respond to the weakest signals that can be heard with other and more complex stages. The tube shown is a 6CB6, but other pentodes such as the 6AK5 may be substituted.

A gain control is included in the cathode circuit. Normally this is run all-out, for optimum noise figure and gain, but in the presence of strong local signals it can be cut in to reduce overloading. This causes some impairment of the noise figure, but may still make possible reception of distant signals through the locals.

Note the double-tuned coupling circuits in the r.f. input and between the r.f. amplifier and the mixer. The capacitors C_1 and C_2 are kept as small as possible, and the coils are not coupled together otherwise. A value of 1 to 2 $\mu\mu$f. gives sufficient coupling at the desired frequency, but the system responds only very slightly to lower frequencies. This helps to prevent interference from signals on the intermediate frequency.

The mixer is also a 6CB6. Its operating conditions are set up for resistance to overloading and cross-modulation from strong signals, rather than for optimum noise figure, as the latter is taken care of by the r.f. amplifier. Note that the plate circuit of the mixer is omitted from the converters. It is built into the power unit, and thus only one coil need be made for all the converters.

The oscillator is a 6AF4 triode. Any other small triode could be substituted. Input is held to a low level (note 47,000-ohm resistor in series with L_7) in the interest of stability. The oscillator circuitry is isolated from the rest of the converter, so that injection can be controlled readily. Energy from the oscillator is carried to the mixer grid circuit through a shielded link.

Mechanical Features

Each converter is built on a flat plate, which screws onto a standard aluminum chassis. Connection to the power unit is made through a 4-pin plug mounted on the side of the case. This carries the heater voltage, the plate voltage, the mixer plate lead and the common chassis connection. The plug on the converter is the male type. It may be fastened to the chassis conveniently by soldering 4-40 nuts to the back of the flanges used for mounting the plug. Flat-head machine screws in countersunk holes, in both the converter and the power supply unit allow the two to fit snugly together. This is important in preventing pickup of signals in the i.f. range.

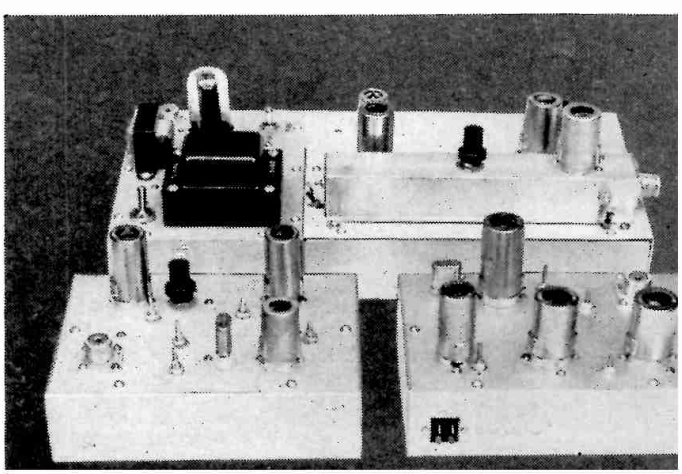
In the bottom view, Fig. 16-9, the antenna connector is seen at the lower right. Just to the left, separated by a small shield, are the two r.f. coils, L_1 and L_2. The coupling capacitor, C_1, made of two wires twisted together, is on the low side of the shield, its lead to L_2 running through a hole in the shield.

The lead from L_2 to the amplifier grid pin runs through the main lengthwise shield. This lead was made of shielded wire, with the shielding removed from the part of the lead that is in the coil compartment. The portion of the wire in the tube compartment must be shielded to prevent feedback between the plate coil, L_3, and the grid circuit. The coupling capacitor, C_2, the gain control, the plate coil and all other amplifier components are in this section, upper right.

Mixer components are at the upper left, with the oscillator section below. The coupling link between L_5 and L_6 is made of shielded wire, running through the main shield partition.

The leads from the mixer to the plug, J_2, and all power leads, are made with shielded wire. The common connection for ground and heater lead is the shielding over the other three wires. These leads should be long enough so that the converter can be lifted from the box without removing the plug. A length of vinyl sleeving slipped over the leads will help to prevent shorts. Transparent sleeving was used, so it does not show in the

Fig. 16-8—Converters for the three v.h.f. bands, with their power supply and i.f. output unit. The 220-Mc. converter is shown plugged into the power unit. At the left is the 50-Mc. converter. The one for 144 Mc. is at the right.



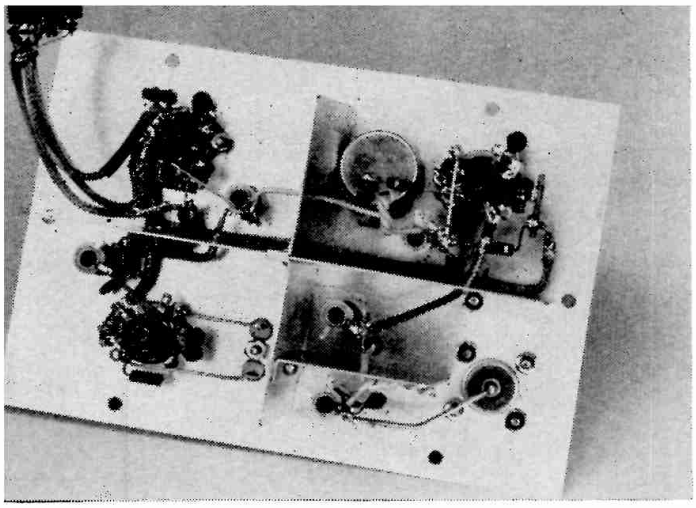


Fig. 16-9—Bottom view of the 50-Mc. converter. R.f. input circuit is at the lower right, with the amplifier itself above. Crystal oscillator components at lower left; mixer and output cable above.

photographs.

The main shield is 6 by $1\frac{15}{16}$ inches in size, with a ¼-inch lip folded over for mounting to the plate. The two shields perpendicular to it are $1\frac{7}{8}$ by $1\frac{15}{16}$ inches, with lips folded over on the bottom and one end. The isolation shield between the r.f. coils is $1\frac{3}{4}$ by $1\frac{15}{16}$ inches, and is mounted ¾ inch in from the lower edge of the cross shield.

The placing of the parts otherwise is not particularly critical, except that bypass capacitors should be connected with the shortest possible leads. Use of the smallest size disk ceramic type is recommended.

Adjustment

Tuning up the converter is a simple matter. Check the wiring to be sure that no errors have been made. Apply a.c. and see if all heaters come on. Then apply plate voltage by closing S_2 on the power supply unit. If the converter output is connected to a communications receiver tuned to the 7-Mc. range there should be a considerable increase in noise as plate voltage is applied, even with circuits out of tune.

First check the oscillator. This can be done by listening in the 43-Mc. range, if a receiver is available for that frequency, or a grid-dip meter may be used as a wavemeter. Output should appear on 43 Mc., and on that frequency only. Adjust L_7 for maximum output indication, with the grid-dip coil coupled to L_7. Check around 14.3 and 28.6 Mc. to be sure that no output is in evidence on these frequencies. Should there be energy on these frequencies it means that the crystal is oscillating on its fundamental frequency and showing output on its various harmonics. Oscillation on the fundamental indicates that the plate circuit is not properly tuned.

If the converter is wired correctly it should now be possible to receive strong signals, even before the circuits have been resonated. A calibrated signal generator is helpful, but it is by no means necessary. A test signal should be fed into the antenna connector and the core screws in all coils adjusted for maximum signal strength.

The response of the converter will not be flat across the entire 4000 kc. of the 50-Mc. band, but it will work over a wider frequency range than most directive antenna systems. The setting of the cores in L_3 and L_4 can be varied to give uniform response across the desired pass band. The input circuit should be adjusted for best signal-to-noise ratio at the middle of the desired frequency range.

The value of the small coupling capacitors, C_1 and C_2, will have some effect on the bandwidth of the r.f. portion of the converter. Few directive antennas will work over more than about 1500

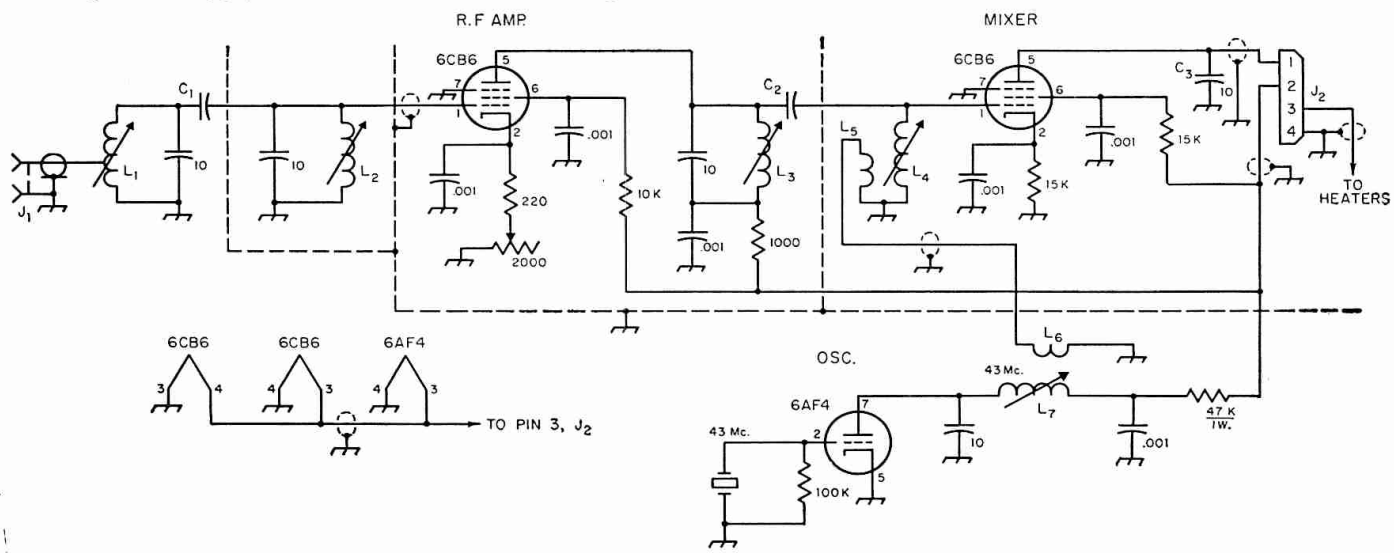


Fig. 16-10—Schematic diagram of the 50-Mc. converter. Capacitors are ceramic; values .001 and up are in μf. Resistors ½-watt unless specified.

C_1, C_2—Approx. 1 to 2 μμf. Make from two pieces of plastic-covered No. 18 wire twisted together about 1 inch.

C_3—10-μμf. ceramic. Connect at plate terminal.

L_1, L_3, L_4—11 turns No. 24 enam. at top end of ¼-inch iron-slug form (North Hills Type F-1000). L_1 tapped at 3 turns.

L_2—Same as L_3, but 9 turns.

L_5—2 turns insulated hookup wire at low end of L_4.

L_6—Same as L_5, but at low end of L_7.

L_7—Same as L_3, but 16 turns.

J_1—Coaxial connector, female.

J_2—4-pin power connector, male. Must mount flush with chassis surface.

kc. of the band, so there is seldom much point in making the front end of the converter broader than this. If optimum performance is needed at the opposite end of the band it is merely necessary to repeak the core studs for best results at the desired frequency. Adjustment of the i.f. coil in the power unit also affects the bandwidth. It can be peaked somewhat above the middle of the tuning range if it is desired to extend the coverage of the converter-antenna combination.

When the converter is tuned for best results it may be desirable to check the oscillator injection. This is best done with the aid of a noise generator, though a signal generator or weak signals may be used if care is taken to observe optimum signal-to-noise ratio, rather than mere gain. The value of the dropping resistor in series with L_7 can be varied, the idea being to use the highest value that will not affect the signal-to-noise ratio adversely.

A simple check on performance that can be made in a location free of manmade noise is as follows: Connect a 50-ohm resistor in place of the antenna coax. Observe the noise level, either by ear or as indicated on an output meter or the receiver S-meter. Now put the antenna back on. If the r.f. stage is free of regeneration, a rise in noise level when the antenna is connected shows that external noise can be heard. This noise is the limiting factor in weak-signal reception, and further reduction in receiver noise figure will serve no useful purpose.

THE 144-MC. CONVERTER

In the converter for 144 Mc., Figs. 16-11 and 16-12, triode r.f. amplifiers are used, as they give better noise figure than pentodes at this frequency and higher. The tubes shown are 6BC4s, but comparable results can be achieved with the 6AJ4, 6AM4 or 6AN4, with the necessary revision of the pin connections. Noise figure obtainable with any of these tubes is about 5 db., which is about the level at which external noise begins to limit receiver sensitivity. A noise figure of 3 db. or lower can be had with 417As, or even one 417A and one less expensive tube, but there may be no observable difference in weak-signal performance.

The cascode circuit (see beginning of chapter) is used, with the circuit of Fig. 16-2 in preference to that of 16-3. The latter, operating at lower plate voltage per stage, may be slightly more susceptible to overloading. The 6CB6 mixer is also operated under conditions designed to keep down overloading and cross-modulation troubles.

The crystal oscillator is operated at the highest frequency that is possible with simple circuitry. This holds down the number of unwanted frequencies appearing in the multiplier output, which could beat in signals from outside the intended frequency range. The crystal oscillates on 45.667 Mc., using the triode portion of a 6U8. The pentode portion is a tripler to 137 Mc.

The oscillator-tripler portion is isolated from the rest of the converter by a copper shield running down the middle of the 5 by 5-inch plate. The grid circuit of the first r.f. amplifier stage is adjacent to the tripler, but is as far away from it as possible, and the coils are positioned for minimum coupling. The lower section of the converter, as shown in Fig. 16-11, is the portion in question, the antenna connection and grid coil being at the lower right.

Above the shield may be seen the first r.f. stage, right, the second stage, with a shield down through the middle of its socket, center, and the mixer at the far left. To provide effective isolation and bypassing, feed-through capacitors are mounted in the copper shield to carry power leads from one compartment to the other. Three are used for the B-plus line and two for the heater leads.

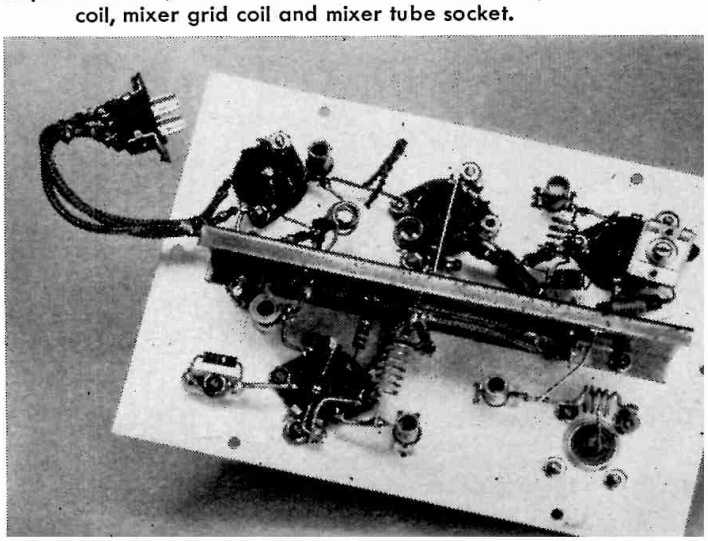
R.f. circuits and the tripler plate circuit are tuned by means of small TV-type trimmers. Four of these are shown in the photograph, but the one that is connected to the first r.f. plate coil, L_3, may be omitted, as the circuit tunes very broadly. The r.f. plate coil, L_4, and the mixer grid coil, L_5, are ¾ inch apart, center to center. Coupling between the two stages is mainly through the twisted-wire capacitor, C_{10}. The r.f. input coil, L_1, is connected to the grid pin of the V_1 by a lead that runs through a ¼-inch hole in the shield.

Both shields are made of flashing copper. The larger is 5¾ by 1¾ inches, with folded-over edges for mounting, and for rigidity. The smaller is 1½ by 1¾ inches. It is held in place by soldering to lugs under the mounting screws of the 6BC4 socket. This shield turned out to be required to prevent oscillation in the grounded-grid stage. It crosses the middle of the tube socket.

Connections for the power are made in the same manner as for the 50-Mc. converter, and leads should be long enough to permit removal of the converter from the box without unsoldering any leads. The shields are bonded together and anchored to a lug bolted to the main shield, near the left end.

Note that wafer-type sockets are used. This is

Fig. 16-11—Bottom view of the 144-Mc. converter. Crystal oscillator and tripler occupy lower left side of the assembly. Antenna input circuit is at the right. Above the partition, right to left, are the cathode trimmer, the first r.f. amplifier socket, the r.f. plate coil, the second amplifier socket, with shield across its center, the plate coil, mixer grid coil and mixer tube socket.



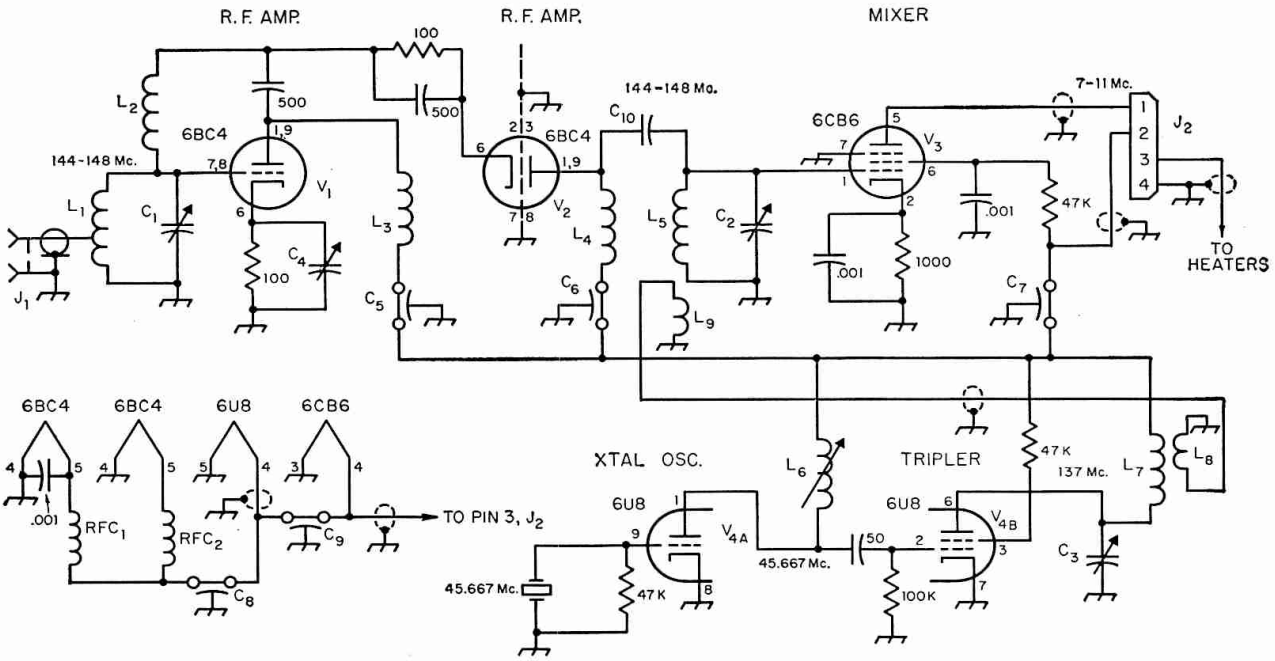


Fig. 16-12—Wiring diagram and parts information for the 144-Mc. converter. Parts specified as in Fig. 16-10.

C_1, C_2, C_3—8-$\mu\mu$f. plastic trimmer (Erie No. 532-10).
C_4—3-30-$\mu\mu$f. mica trimmer. Set at tight position initially.
C_5, C_6, C_7, C_8, C_9—500-$\mu\mu$f. feed-through bypass (Centralab MFT-500).
L_1—4½ turns No. 18 tinned, ¼-inch inside diam., ½ inch long, tapped at 1½ turns.
L_2—14 turns No. 24 enam., 3/16-inch diam., ½ inch long.
L_3—5 turns No. 18 tinned, ¼-inch diam., ¼ inch long.
L_4—5½ turns like L_3.
L_5—3½ turns like L_3.
L_6—13 turns No. 24 enam. closewound on ¼-inch diam. iron-slug form (North Hills F-1000).
L_7—8 turns like L_3, ¾ inch long.
L_8—1 turn insulated hookup wire between first two turns of L_7.
L_9—Same as L_8, inserted in L_5.
J_1—Coaxial connector, female.
J_2—4-pin power connector, male. Must mount flush with surface of chassis.
RFC_1, RFC_2—1.8 μh. solenoid r.f. choke (Ohmite Z-144).

more than an economy measure; shorter ground leads are possible with this type of socket. Where socket terminals are to be grounded, they are bent down flush with the bottom of the plate. Then a hole is drilled adjacent to the lug and it can then be secured to the plate under a washer and nut. This method of grounding is superior, at these frequencies, to the more commonly used lead-and-lug arrangement.

Adjustment

The first step in putting the 144-Mc. converter into service is to be sure that the oscillator is working correctly, as described in connection with the 50-Mc. converter. This may be done with the plate and screen voltages disconnected from the pentode portion of the 6U8, if desired, by lifting tripler plate coil and the screen resistor from the B-plus line temporarily. Be sure that the oscillator is on the right frequency, and no other, as described earlier.

Now connect the tripler plate coil and screen resistor to the B-plus line and check the tuning of the tripler capacitor, C_3. Set it for maximum output on 137 Mc., as indicated by a grid-dip meter coupled to L_7. The output required from the tripler may be checked after the r.f. section is tuned properly. It may be controlled by varying the value of the screen dropping resistor, which is 47,000 ohms in the original. The tripler may be run at the lowest input that will give satisfactory signal-to-noise ratio. Above that point the injection is not critical.

The r.f. circuits may now be adjusted. Set the trimmer, C_4, across the r.f. cathode resistor, at maximum at first. Then on a test signal tune C_1 and C_2 for maximum response. The spacing between the turns of the r.f. plate coils, L_3 and L_4, should also be adjusted for highest signal level.

If a noise generator is available, it should be used to set up the r.f. input circuit, the inductance of the neutralizing coil, and the value of the cathode bypass, C_4. If signals or a signal generator are used, the criterion should be greatest rise over noise for a given signal, rather than maximum S-meter reading or loudest volume. Adjustment of the neutralizing coil, and setting of the cathode bypass value are all but impossible without a noise generator. Lacking one, it is best to use a fixed bypass of about 100 $\mu\mu$f. for C_4, and leave the neutralizing winding at the specification given in the cut label. Changes in the neutralizing coil affect the tuning of the grid circuit. Recheck the setting of C_1 after altering L_2.

The coupling capacitor, C_{10}, is not critical, but for best rejection of i.f. signals it should be as low as will give satisfactory performance on 144-Mc. signals. Insulated wires twisted together provide a convenient adjustment method.

As the band is nearly three times as high in frequency as the 50-Mc. band, there will be less

difficulty in getting uniform response across the entire band. Tuning of the second r.f. and mixer circuits can be staggered to develop the desired bandwidth, and the value of C_{10} will have some effect on it as well.

THE 220-MC. CONVERTER

In the converter for 220 Mc., Figs. 16-13 and 16-14, an additional r.f. amplifier stage is used ahead of the cascode-and-mixer combination. This is required because the gain per stage is lower at this frequency. It is also desirable because of the added selectivity it affords. This may be very helpful in areas where interference from other services adjacent to the band may be bothersome.

The additional stage is a grounded-grid amplifier, using a modified coaxial-line plate circuit for high "Q" and selectivity. It is not a broadband device and must be retuned in covering the band. The tube shown is a 6AM4. Similar results were achieved with the 6BC4, and nearly identical performance is possible with other u.h.f. triodes. The 417A and 416B should be superior. Noise figure is about 6 db.

A series cascode using a 6BC8 dual triode follows. This type of amplifier is easily adjusted and tends to deliver superior results as the upper limit of frequency is approached. The mixer is a 6AK5. Its output circuit is, of course, the coil assembly in the power unit.

The r.f. amplifier is similar to the one described separately later in the chapter, except that the output is taken off through the bottom of the assembly, with a tuned link, instead of through a coaxial fitting on the side. In the diagram, Fig. 16-14, the plate line and coupling loop are shown as if they were coils, it being cumbersome to express a trough-line circuit schematically.

Mechanical Details

A somewhat different method of construction is employed in the 220-Mc. converter, in order to insure the most effective grounding and bypassing. A plate of aluminum is used, as in the other converters, but only for appearance and rigidity. The plate used for actual electrical grounding is a sheet of flashing copper. Wafer sockets are used, and wherever a terminal is grounded it is bent down flat and soldered directly to the copper plate. This makes for less lead and more effective grounding than where socket mounting screws and lugs are used ground connections. It also allows shield partitions of copper to be soldered directly to the base plate.

The 220-Mc. converter requires more space than the others, so a 7 by 9-inch chassis and plate are used. The lengthwise partition 1⅛ by 7 inches in size, after folding over ⅛ inch on each side for mounting and rigidity. The smaller is 1⅛ by 4 inches. The large shield is centered on the plate 2⅜ inches in from the long edge. The smaller is 4¼ inches in from the left edge.

The oscillator is similar to the 144-Mc. unit, except that an air-wound coil and a variable capacitor are used instead of a slug-tuned coil. The pentode section of the 6U8 is a quadrupler to 213 Mc. from a crystal frequency of 53.25 Mc. A series-tuned link feeds energy to the mixer grid circuit through a shielded-wire line. Oscillator-multiplier components are in the left portion of Fig. 16-13.

At the right are the mixer (upper socket) and the series cascode r.f. amplifier, below. Note that power wiring is made with shielded wire, laid close to the shields. Plate voltage is fed into the oscillator-multiplier and r.f.-mixer compartments on feed-through bypasses. Heater voltage for the r.f. amplifier goes through the plate on shielded wire at the lower left, and plate voltage at the

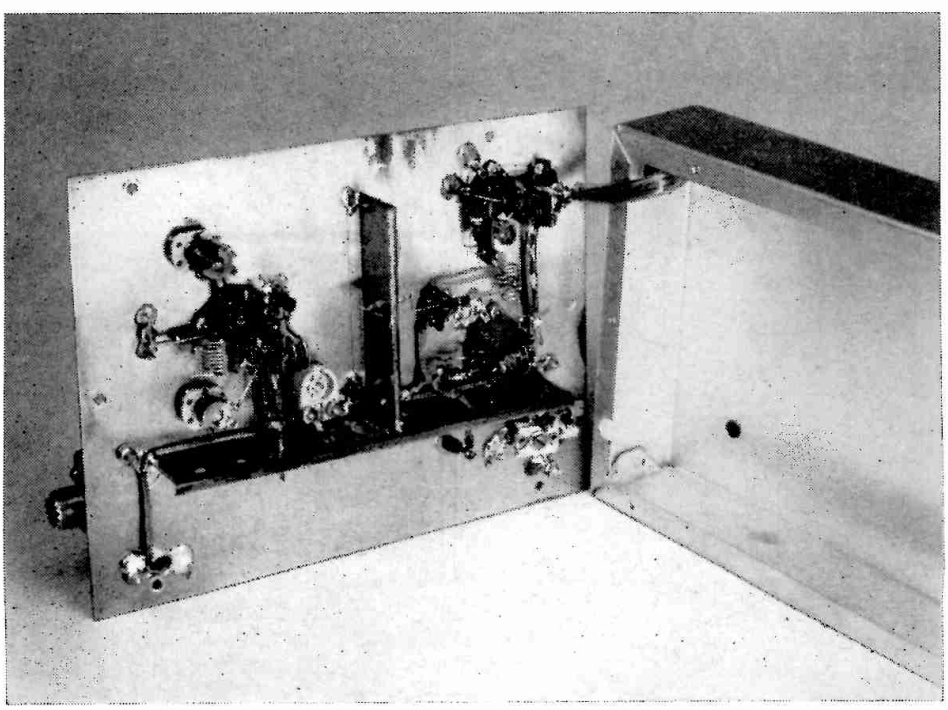


Fig. 16-13—Interior of the 220-Mc. converter. Bottom plate and partitions are of flashing copper, for effective grounding. Oscillator-multiplier circuitry is at the left; mixer and cascode r.f. amplifier at the right. Grounded-grid amplifier is above the chassis.

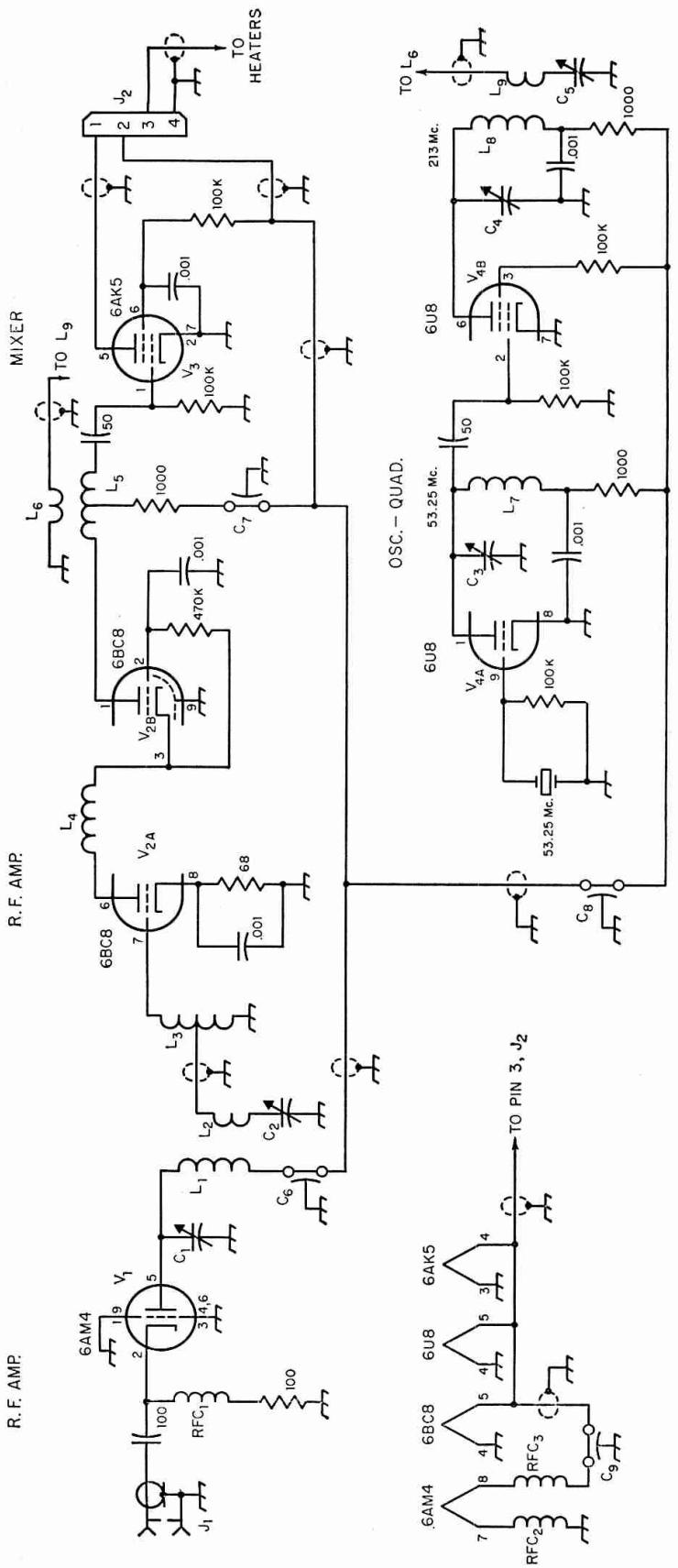


Fig. 16-14—Schematic diagram and parts information for the 220-Mc. converter.

lower right. The mica trimmer at the lower right is C_2, in series with the low side of the coupling loop, L_2. The other end of the loop comes out on a feed-through bushing, National Type TPB. Its lead to L_3 is shielded wire, running through the partition.

In working with flashing copper parts the metal work should be completed, up to the point where the parts are ready to assemble. The copper parts may then be polished with steel wool and given a fine spray coat of clear lacquer. This will help to keep them clean and bright, and it will not affect the soldering operations to be done later.

Adjustment

The oscillator and multiplier stages should be adjusted as outlined for the other converters, making sure that the

«

C_1—5-$\mu\mu$f. miniature variable (Hammarlund MAC-5).
C_2—3-30-$\mu\mu$f. mica trimmer.
C_3—20-$\mu\mu$f. miniature variable (Hammarlund MAC-20).
C_4—10-$\mu\mu$f. miniature variable (Hammarlund MAC-10).
C_5—7-45-$\mu\mu$f. ceramic trimmer (Centralab 822-BN).
C_6, C_7, C_8, C_9—500-$\mu\mu$f. feed-through bypass (Centralab MFT 500).
L_1—Inner conductor of trough line—¼-inch copper tubing, 6¼ inches long, ¼-inch diam. C_1 connects 1¾ inches from plate end. See Fig. 16-22 and text.
L_2—Coupling loop—insulated hookup wire 3 inches long. Loop portion lays close to cold end of L_1 for 2 inches. Hot end comes through chassis on National Type TPB feed-through bushing.
L_3—3 turns No. 18 tinned, ¼-inch diam., ¼ inch long, center-tapped.
L_4—4 turns like L_3, ⅜ inch long.
L_5—8½ turns like L_3, ⅝ inch long, center-tapped.
L_6—2 turns insulated hookup wire at center of L_5.
L_7—6 turns No. 20 tinned ½-inch diam., ½ inch long. (B & W No. 3003).
L_8—2 turns No. 18 tinned, ⅜-inch diam., spaced ⅛ inch.
L_9—2 turns insulated hookup wire between turns of L_8.
J_1—Coaxial fitting, female.
J_2—4-pin power connector, male. Must mount flush with surface of chassis.
RFC_1, RFC_2, RFC_3—18 turns No. 24 enam., close-wound, ⅛-inch diam.

correct frequencies are obtained. Next a signal may be fed into the 6BC8 stage through the shielded line to L_3. This may be disconnected from L_2 temporarily and coax-fed antenna or a 50-ohm signal generator termination may be connected across it. Now adjust the spacing of the turns in L_3 and L_5 for best performance. Maximum gain will be a good-enough indication here, so a noise generator is not needed.

Now the 6AM4 amplifier may be hooked up and tuned. It will be quite selective and will have to be retuned several times across the band. With the plate tuning capacitor tapped down the line as it is, the tuning range in megacycles is not great. Be sure, therefore, that it actually does tune the entire way, and does not hit maximum or minimum capacitance inside the band.

Adjustments may be made all along the line using maximum signal level as the basis for achieving the optimum setting, but only a noise generator will show if the converter is delivering the best sensitivity of which it is capable. It should be possible to get the noise figure down to about 6 db. using the 6AM4, if everything is working properly.

If any doubt exists that the coils L_3 and L_5 are tuning properly, small twisted-wire capacitors may be connected from the grid end of L_3 and the plate end of L_5 to ground, and gradually increased in value. If the gain drops when the capacitor is connected, the coil is too large. If a small amount of added capacitance increases the gain, squeeze the coil turns closer together and try again. The inductance of L_4 should not be particularly critical. It should be as large as can be used without causing instability.

Injection from the quadrupler may be controlled by varying the position of either link winding, L_6 or L_9, with respect to its coil, and by adjusting C_5. Coupling should be increased until there is no improvement in signal to noise ratio. Injection beyond that point is not critical, though it will affect the overall gain somewhat. Fairly low injection is desirable as it will keep down the level of spurious responses.

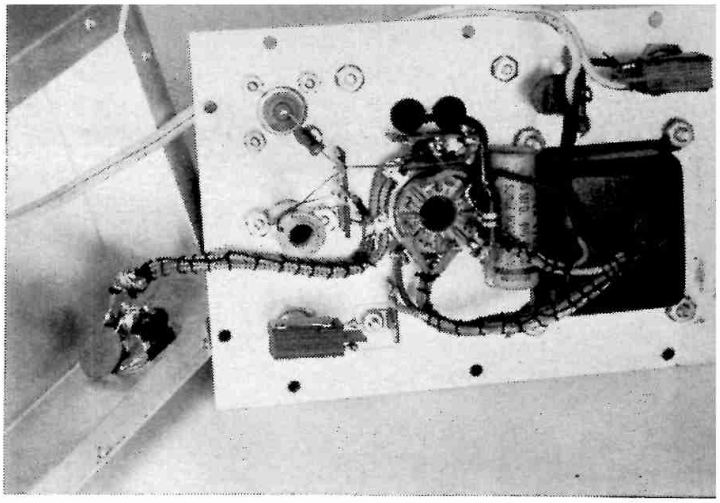


Fig. 16–15—Bottom view of the power supply and i.f. output circuitry for the v.h.f. converters. A.c. switch is above power transformer, right. Next are the filter capacitor and the rectifier socket. The switch at the lower left cuts off the high voltage. The i.f. plate coil and the output fitting are in the upper left of the picture.

● POWER SUPPLY AND I.F. OUTPUT

Though it may be possible to run a v.h.f. converter from the power supply of the receiver with which it is to be used, a supply for the converters is desirable. The one shown in Fig. 16-15 and 16-16 is inexpensive and convenient. It delivers the heater and plate power required by the converters, and in addition carries the mixer plate circuit and the provision for coupling into the receiver.

Construction is not critical. Parts are assembled on a 5 by 7-inch plate and this fastens to a similarly sized chassis that matches the converters. The 50- and 144-Mc. units plug into the

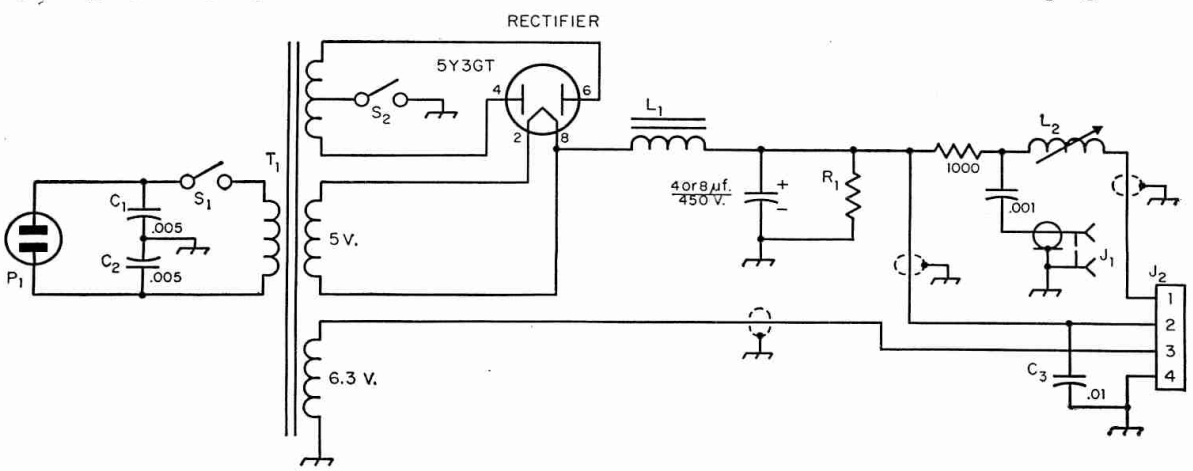


Fig. 16-16—Schematic diagram of the converter power supply and i.f. output unit. Capacitors with polarity marked are electrolytic; others ceramic.

C_1, C_2—Dual .005-μf., 125 volts a.c. disk ceramic (Sprague 125L-2D50).

C_3—.01-μf. disk ceramic. Mount at plug end of cable.

R_1—50,000 ohms, 2 watts (2 100,000-ohm 1-watt resistors in parallel).

L_1—10-hy. 50-ma. filter choke.

L_2—No. 28 enam. closewound ½ inch long on ⅜-inch iron-slug form. Wind near upper end.

J_1—Coaxial fitting, female.

J_2—4-pin power connector, female. Must mount flush with surface of chassis.

S_1, S_2—S.p.s.t. toggle switch.

T_1—Power transformer, 480 v. a.c., c.t., 40 ma., 5 v. 2 amp., 6.3 v. 2 amp. (Thordarson TS-24R00).

P_1—A.c. plug on cord.

power unit through matching fittings on the sides. The larger 220-Mc. converter has the plug mounted on the end wall of the chassis, so that its 7-inch dimension is aligned with that of the supply.

Arrangement of parts should be clear from the photographs, and parts location is in no way critical. Note that the a.c. connection is bypassed on both sides of the line. The capacitors C_1 and C_2 are a dual unit designed for this purpose. The bypass on the B-plus line, C_3, should be at the plug end of the cable, with as short leads as possible. It is important in preventing pickup of signals in the i.f. tuning range, as are C_1 and C_2.

Switches are provided for turning on the a.c., and for breaking the flow of plate current. This feature is helpful during adjustment when it may be desirable to remove the converter from its case. Plate voltage may be cut off for safety in handling, and then turned on again without loss of the time needed to warm up the tubes.

Contact between the converter case and the power supply case may be important in preventing signal pickup at 7 Mc. If i.f. signals are bothersome, try putting a spring clip under one of the screws that holds the power supply plate down. Place this so that it will make contact with the converter case or top plate when the two units are plugged together. It also may be necessary to bond the converter and power supply combination to the frame of the communications receiver with which they are to be used. This should be done with a short heavy copper strap or braid.

Connection between the i.f. unit and the receiver should be with coaxial line, and it is highly desirable to install a coaxial fitting on the receiver in place of the usual terminal strip. The connections should be removed from the back of the strip, or the terminals may still allow some i.f. pickup.

Using Other Intermediate Frequencies

The i.f. tuning range beginning at 7 Mc. was selected as the most desirable for most receivers. Other ranges may be preferred, and the i.f. can be altered easily enough. The injection frequency is lower than the signal frequency by whatever i.f. you intend to use. For example, a 50-Mc. converter with a 14-Mc. i.f. would have a crystal and injection frequency of 50–14, or 36 Mc. The 144-Mc. converter would have a 130-Mc. injection frequency, and the crystal would be one-third of this, or 43.33 Mc.

Generally speaking, single-conversion communications receivers (most inexpensive types, and all older receivers) work best with low intermediate frequencies, such as 7 Mc. or lower. Double-conversion receivers will be satisfactory in the 14-Mc. range in almost every case, and some are stable enough to do well around 30 Mc. At least one communications receiver, the NC-300, has a range designed especially for v.h.f. converter use, starting at 30.5 Mc.

Preamplifier for 220 Mc.

The amplifier shown in Figs. 16-17 to 16-19 will improve the gain and noise figure of a 220-Mc. converter that is not operating at maximum effectiveness. It also provides some additional selectivity, which may be helpful in areas where signals from outside the band are troublesome. The plate circuit has high Q, so it must be retuned in covering the band.

The schematic diagram is the same as the first stage of the 220-Mc. converter, Fig. 16-14. The signal is fed into the cathode of the grounded-grid amplifier. The plate circuit is a trough line. Any of the small u.h.f. triodes may be used, though a 6AM4 is shown. Check pin connections and cathode resistor values for other types.

Construction

The outer conductor of the line, which also serves as the chassis, is made of flashing copper. If the details of Fig. 16-18 are followed, it may be made from a single piece. A small copper shield is placed across the tube socket to isolate the input and plate circuits. Just where this shield is located depends on the tube used, as various

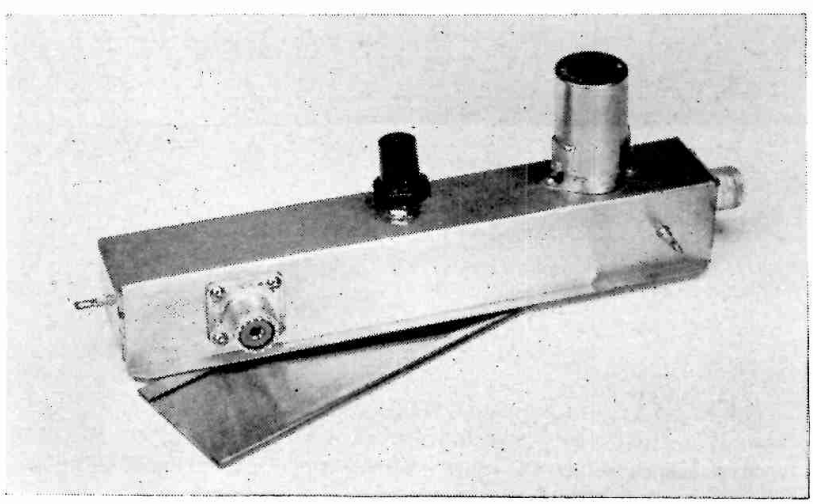


Fig. 16-17—220-Mc. trough-line preamplifier. Construction is similar to that used with the 220-Mc. converter, Fig. 16-8, except that provision is made for cable connection to a remote receiver or converter.

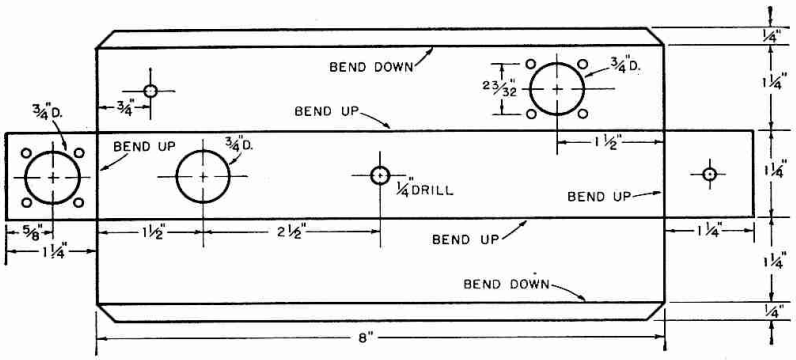


Fig. 16-18—Details of the outer conductor and chassis for the 220-Mc. preamplifier.

tubes have different grid pin arrangements. All grid terminals are bent flat against the copper case, and soldered in place.

The left end (bottom view, Fig. 16-19) contains the coaxial fitting for the antenna connection, the r.f. chokes and other components of the input circuit. The plate line, tuning capacitor, output coupling loop and coax fitting, and the B-plus feed-through capacitor mount in the large portion. A bottom cover for the line can be made of copper 8 inches long and 2¼ inches wide. Bend over a quarter inch on each side, and slip the cover over the edges of the case.

The inner conductor is ¼-inch copper tubing. Start with a piece 6¼ inches long. Saw the ends lengthwise to depths of ¼ and ½ inch. Cut off one half at each end. The remaining portions are used to make connections. The half-inch end is bent down to solder to the plate lugs of the socket. The quarter-inch end solders to the feed-through capacitor.

The tuning capacitor, C_1, is mounted with its stator bars toward the tube end of the line. The inner conductor will rest between these bars and they can be soldered to it readily. Plate voltage is fed through C_6, heater voltage through C_9. Output is taken off through the coupling loop, L_2, visible in Fig. 16-19. The series capacitor, C_2, was omitted from the preamplifier, though it might be useful if the amplifier works into a converter with an untuned input circuit.

Adjustment

The preamplifier may be connected to the converter through a coaxial line of any convenient length, but the converter input should be a coaxial fitting. To put the preamplifier into service, adjust the plate line for maximum signal strength. Then check the position of the coupling loop, adjusting for maximum response. Readjust the tuning of the line as the coupling is changed.

The tuning range of C_1 is not wide, so be sure that it actually tunes the line at both ends of the band. Some adjustment of tuning range can be had by rotating the mounting of the capacitor 180 degrees. If this does not bring the tuning within range, the mounting hole can be elongated and the position of the trimmer adjusted as required.

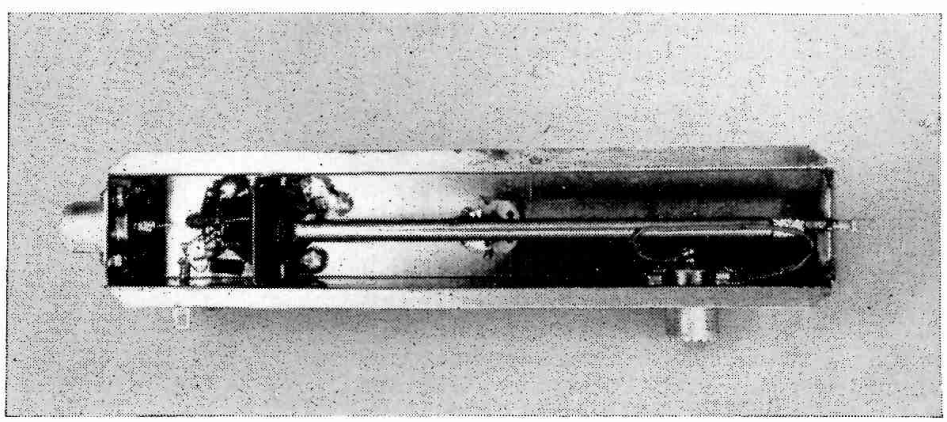


Fig. 16-19—Bottom view of the preamplifier

Receivers for 420 Mc.

For best signal-to-noise ratio, receivers for any frequency should have the highest degree of selectivity that can be used successfully at the frequency in question. With crystal control or its equivalent in stability accepted as standard practice on all bands up through 148 Mc., there is little point in using more bandwidth in receivers for these frequencies than is necessary for satisfactory voice reception, a maximum of about 10 kc. Such communication selectivity is now being used successfully by most workers on 220 and 420 Mc., too, but it imposes several problems not encountered on lower bands.

First is the matter of oscillator instability in the converter. Even the best tunable oscillator at 420 Mc. suffers from vibration and hand-capacity

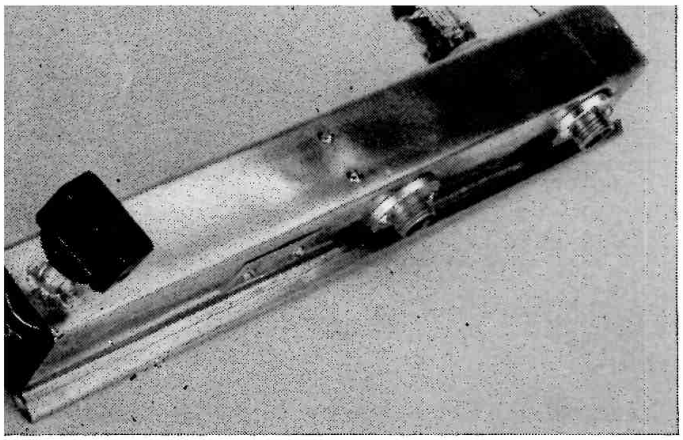


Fig. 16-20—A highly effective r.f. amplifier for 420 Mc. The tank circuit is a half-wave line made of flashing copper. Coaxial fittings are for input and output connections. Heater and plate voltages are brought in on feed-through bypass capacitors just visible on either side of the 6AJ4 tube.

effects sufficiently to make it difficult to hold the signal in a 10-kc. i.f. band width.

Then, there are still some unstable transmitters being used in work on 220 and 420 Mc. It is out of the question to copy these on a selective receiver.

Last, searching a band 30 megacycles wide is excessively time-consuming when communications-receiver selectivity is used in the i.f. system.

There is no single solution to these problems, but the best approach appears to be that of breaking up of the band into segments for different types of operation. This is being done by mutual agreement among 420-Mc. operators at present, as follows: 420 to 432 Mc. — modulated oscillators and wide-band f.m., 432 to 436 Mc. — crystal-controlled c.w., a.m. and narrow-band f.m.; 436 to 450 — television.

The first segment can be covered with a superregenerative receiver, a superheterodyne having a wideband i.f. system, or a converter used ahead of an f.m. broadcast receiver. The high selectivity required for best use of the middle portion makes a crystal-controlled or otherwise highly stable converter and communications receiver combination almost mandatory. Amateur TV is usually received with a converter ahead of a standard TV receiver, tuned to some channel that is not in use locally.

Many of the tubes used on the v.h.f. bands are useless at 420 Mc., and the performance of even the best u.h.f. tubes is down compared to lower bands. Only the lighthouse or pencil-triode tubes and a few of the miniatures are usable, and these require modifications of conventional circuit technique to produce satisfactory results.

Crystal diodes are often used as mixers in 420-Mc. receivers, as in this frequency range they work nearly as well as vacuum tubes. The over-all gain of a converter having a crystal mixer is about 10 db. lower than one using a tube, so this difference must be made up in the i.f. amplifier. The noise figure of a receiver having a crystal mixer and no r.f. stage includes the noise figure of the i.f. amplifier following the mixer, so best results require that the i.f. amplifier employ low-noise techniques discussed earlier in this chapter. If the i.f. is 50 Mc. or higher it is particularly important that a low-noise triode be used for the first i.f. stage.

Crystal diodes of the type used in radar mixers, such as the 1N21 series, are well suited to 420-Mc. mixer service, though care must be taken to avoid damage from transmitter r.f. energy. Other types of crystal diodes such as the 1N72 and CK710 will stand higher values of crystal current, and their use is recommended.

Few conventional vacuum tubes work well as mixers at 420 Mc. and higher. The 6J6 is useful where a balanced input circuit is desired, as in Fig. 16-5C. For single-ended circuitry the 6AM4 and 6AN4 are recommended. They may be used in grounded-grid or grounded-cathode circuits.

For high-selectivity coverage of the 432- to 436-Mc. segment of the band, a common practice is to use a crystal-controlled converter working into another converter for either the 50- or 144-Mc. band, tuning the latter for the four-megacycle tuning range.

● A 420-MC. R.F. AMPLIFIER

The r.f. amplifier shown in Figs. 16-20 through 16-22 is capable of a gain or more than 15 db. and its noise figure can be as low as 6 db. with careful adjustment. It will make a large improvement in the sensitivity of any converter or receiver that has no r.f. stage, or one that is working poorly.

The design shown is for either the 6AJ4 or 6AM4, but with suitable socket and pin-connection changes the 417A, 6BC4 or 6AN4 will work equally well. It is a grounded-grid amplifier with

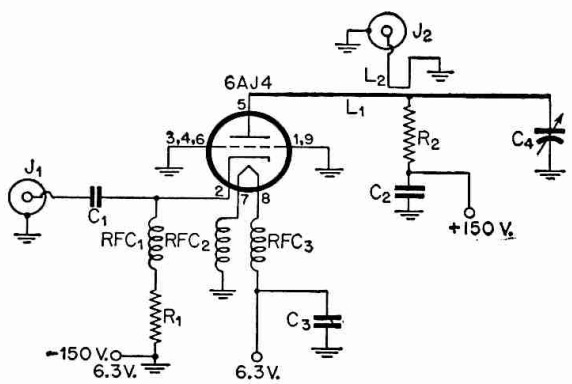


Fig. 16-21—Schematic diagram of the 420-Mc. r.f. amplifier.

C_1—500-$\mu\mu$f. ceramic.
C_2, C_3—1000-$\mu\mu$f. ceramic feedthrough (Erie style 2404).
C_4—Copper tabs, 7/8-inch diam.; see text and photographs.
R_1—150 ohms, ½ watt.
R_2—470 ohms, ½ watt.
L_1—¼-inch copper tubing, 7⅜ inches long, tapped 2⅜ inches from plate end.
L_2—Loop of insulated wire adjacent to L_1 for ¾ inch.
J_1, J_2—Coaxial fitting.
RFC_1, RFC_2, RFC_3—9 turns No. 22, ⅜-inch diam., spaced one diam.

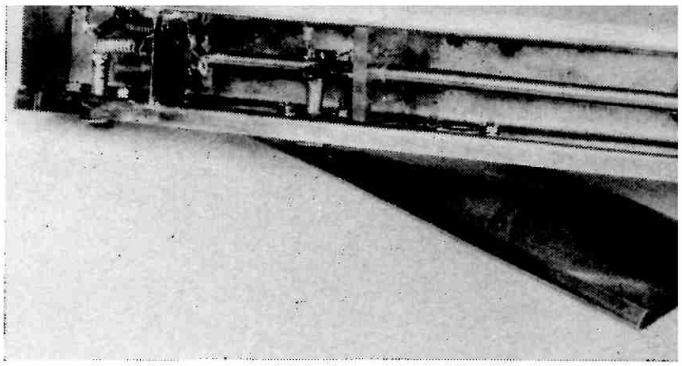


Fig. 16-22—Bottom view of the 420-Mc. r.f. amplifier, with the slip-on cover removed. The inner conductor of the tank circuit is held in place by a block of polystyrene, mounted near the low-voltage point on the line. The plate-voltage feedthrough and output coupling loop may be seen at the left of this support. Heater, cathode and antenna-circuit components are in a separate compartment at the tube end of the assembly. The line is tuned at the opposite end by a handmade copper-tab capacitor.

a half-wave line in the plate circuit. The antenna is connected to the cathode of the tube through a coupling capacitor. As the input impedance of the grounded-grid stage is low, nothing is gained by the use of a tuned circuit in the cathode lead. Output is taken off through a coupling loop at the point of lowest r.f. voltage along the line.

The amplifier is built in a frame of flashing copper that serves as the outer conductor of the tank circuit. The whole assembly is 10 inches long and 1¼ inches square, except for the bottom, which is about 1¾ inches wide. Edges are folded over with lips ¼ inch wide which slide into a bottom cover made from copper sheet 2¼ by 10 inches in size, with its edges bent up ¼ inch wide on each side.

The plate circuit is made of ¼-inch copper tubing tuned by a copper-tab capacitor at the far end from the tube. Plate voltage is fed in at the point of minimum r.f. voltage, which in this instance is about 5 inches from the open end. The antenna is connected to the cathode through a coupling capacitor. The input impedance of the grounded-grid amplifier is so low that nothing is gained by using a tuned circuit at this point. The cathode and heater are maintained above ground potential by small air-wound r.f. chokes.

The tube socket is two inches in from the end of the trough, and is so oriented that its plate connection, Pin 5, is in the proper position to connect to the line with the shortest possible lead. A copper shielding fin is mounted across the interior of the trough 2⅛ inches from the end, dividing the socket so that Pins 3, 4, 5 and 6 are on the plate side of the partition.

Minimum grid-lead inductance is important. This was insured by bending all the grid prongs down against the ceramic body of the socket, and then making the mounting hole just big enough to pass this part of the socket and the prongs. They were soldered to the wall of the trough.

Input and output connections are coaxial fittings mounted on the side wall of the trough. B-plus and heater voltage are brought into the assembly on feed-through capacitors mounted on the same side of the trough as the tube. Connection to the inner conductor of the line is made with a grid clip, so that the point of connection can be adjusted for optimum results.

The copper tubing is slotted at the plate end with a hack saw to a depth of about ¼ inch, and a strip of flashing copper soldered into this slot to make the plate connection. A copper tab about the size of a one-cent piece is soldered to the other end of the tubing to provide the stationary plate of C_4. The line is supported near the low-voltage point by a ¼-inch-thick block of polystyrene. This is centered at a point 5¼ inches in from the tube end of the trough assembly. The hole for the B-plus feedthrough is 4¼ inches from the same end.

The movable plate of C_4 is soldered to a screw running through a nut soldered to the upper surface of the trough at a point ⅜ inch in from the open end. If a fine-thread screw is available for this purpose it will make for easier tuning, though a 6-32 thread was used in this model. This made a wobbly contact, so a coil spring was installed between the top of the trough and the knob to keep some tension on the adjusting screw.

Adjustment of the 420-Mc. amplifier is made easier if a noise generator is used, though it is not as important as in the case amplifiers with tuned input circuits. If the amplifier is working properly there will be an appreciable rise in noise as the plate circuit is tuned through resonance, and it may break into oscillation if operated without load. When connected to a following stage, with a reasonably matched antenna plugged into J_1, the amplifier should not oscillate unless the coupling loop, L_2, is much too far from the inner conductor.

When the amplifier is operating stably and tuned to a test signal (or to a peak of response to a noise generator), the next step is to locate the optimum position for feeding the plate voltage into the line. This may be done by running a pencil lead slowly up and down the inner conductor, until a spot is found where touching the lead to the line has little or no effect on the operation of the amplifier. The plate voltage clip should be placed at this point and the process repeated, moving the clip slightly until it is at the minimum-voltage point precisely. This adjustment should be made at the midpoint of the tuning range over which the amplifier is to be used.

The position of the coupling loop should then be adjusted for best signal-to-noise ratio. This will probably turn out to be with the insulated wire lying against the inner conductor for a distance of about ¾ to 1 inch, starting at the minimum-voltage point just located.

● A CRYSTAL-CONTROLLED CONVERTER FOR 432 MC.

The converter shown in Figs. 16-23 through 16-26 is designed to provide high sensitivity and

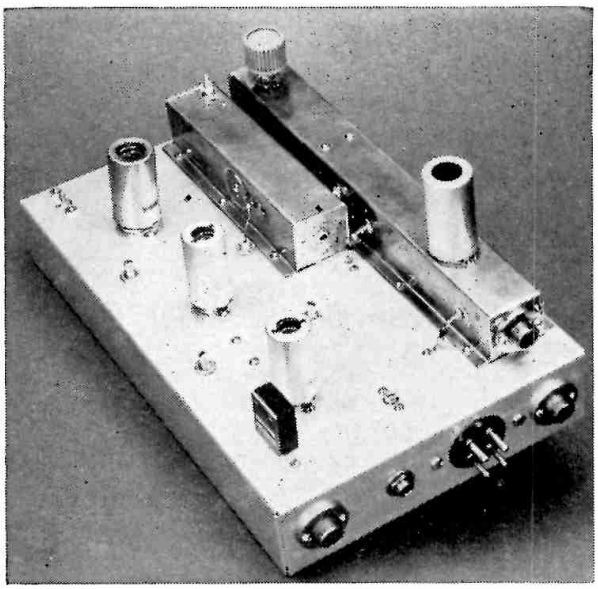


Fig. 16–23—A crystal-controlled converter for 432 to 436 Mc. R.f. and mixer stages are in copper subassemblies at the right. Oscillator, multiplier and i.f. amplifier are on the left side.

signal-to-noise ratio in reception of signals in the 432- to 436-Mc. range. It uses a grounded-grid r.f. amplifier stage similar to the one shown in Fig. 16-20, working into a crystal-diode mixer. The intermediate frequency, with the design constants given, is 50 to 54 Mc., though lower frequencies could be used by suitable modification of the injection chain.

Crystal-controlled injection on 382 Mc. is provided by two 6J6s operating as overtone oscillator-tripler and tripler-doubler, respectively. As only a small amount of r.f. is required at 382 Mc., this line-up is not difficult to build or adjust. An inexpensive 7-Mc. crystal is used. An i.f. preamplifier stage follows the crystal mixer. This may or may not be needed, depending on the performance of the receiver or converter that will serve as the tunable i.f. Low-noise amplification in the i.f. stage is a factor in the over-all performance of the system, so use of the built-in i.f. stage is recommended.

Construction

The converter is built on a 7 × 11 × 2-inch aluminum chassis, with the r.f. and mixer portions in a copper subassembly that mounts on the top of the chassis, at the right side as seen in Fig. 16-23. The oscillator-tripler and tripler-doubler 6J6s are at the left front, with the 6BQ7A i.f. amplifier at the rear. The mixer line is the short portion of the copper assembly, with the r.f. amplifier line at the right. In the bottom view, Fig. 16-25, the injection-chain and i.f. amplifier components are visible.

Fig. 16-24 is an interior view of the r.f. and mixer lines. These are made as two separate assemblies, joined by short length of copper tubing that is visible in the top view. Both tank circuits are 1¼ inches square, with ¼-inch copper tubing inner conductors. They are made from sheets of flashing copper 4¼ inches wide. The mixer compartment is 5½ inches long and the r.f. portion is 10 inches long.

The r.f. amplifier is similar structurally to the one described previously, except for the method of coupling between it and the crystal mixer. This is done with a grid clip on each line and a ceramic coupling capacitor. The lead from the capacitor, inside the amplifier line, is brought through a half-inch length of copper tubing that is soldered into the walls of both lines. The lead is insulated with spaghetti sleeving.

The B-plus feed to the r.f. stage should be at the point of minimum r.f. voltage, 1⅞ inches from the plate end of the copper tubing. The coupling tap is one inch out from the B-plus feedpoint. The coupling point on the mixer line is 1 inch from the ground end. The crystal diode is inserted in a small hole in the mixer inner conductor, 1¾ inches from the ground end. The inner conductors of the r.f. and mixer lines are 7 3/16 and 5 inches long, respectively. Mixer tuning is done with a small plastic trimmer, C_{10}, while the r.f. plate circuit is tuned with a handmade tab capacitor, C_9, similar to C_4 in Fig. 16-21.

Note the r.f. bypass, C_8, on the outside of the mixer line. This is made from a piece of copper ⅞ inch in diameter, insulated from the line housing by a piece of vinyl plastic. Two thicknesses of the material commonly used for small parts envelopes are satisfactory. The crystal, which may be any of the u.h.f. diodes, is slipped through a close-fit hole and is held in place by the wire soldered to its outside terminal.

Plate and filament voltages are fed into the assembly on feed-through bypass capacitors, visible in the top-view photograph. Antenna connection is made through a coaxial fitting on the end of the r.f. assembly. A crystal-current jack, a 4-pin power fitting and two i.f. connectors are on the end wall of the chassis. The second coaxial connector was installed so that tests could be made with and without the i.f. amplifier stage.

Wiring in the power circuits is done with shielded wire, in case that TVI might result from the oscillator or multiplier stages. The addition of a bottom plate and power-lead filtering would then be effective. Injection and i.f. coupling leads are also made of shielded wire, this serving in place of coax line that is harder to handle.

The output of the injection chain is coupled into the mixer line by means of a loop, L_8, that

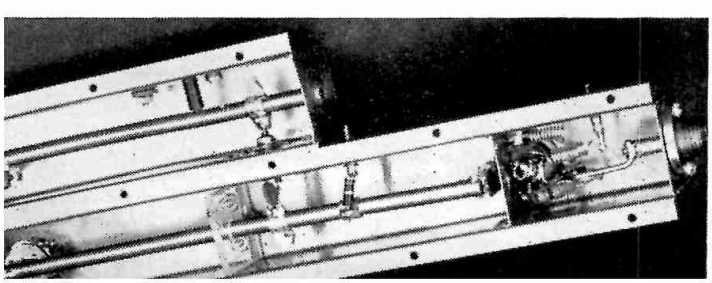


Fig. 16–24—Interior view of the r.f. amplifier and mixer assemblies. The r.f. circuit is a half-wave line. The shorter assembly is the quarter-wave line using a crystal diode mixer.

Crystal-Controlled Converter for 432 Mc.

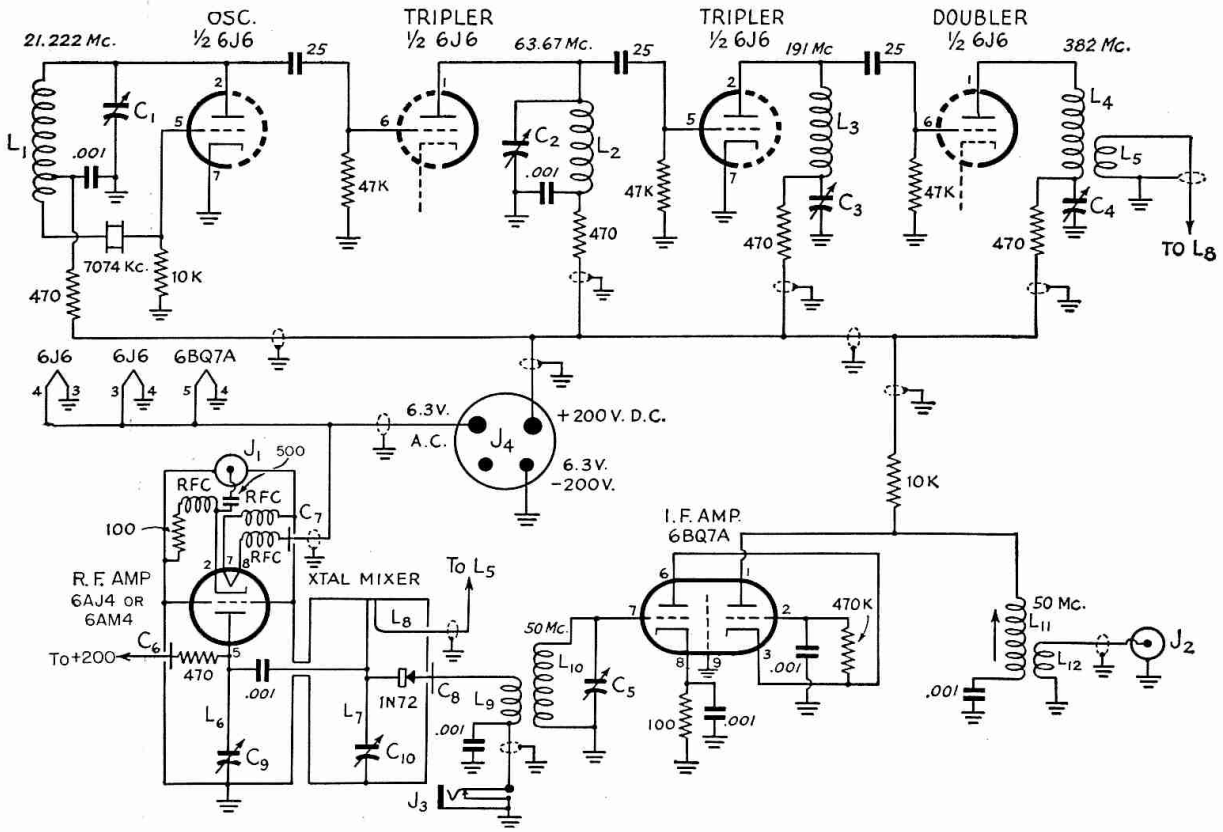


Fig. 16-26—Wiring diagram and parts list for the 432-Mc. crystal-controlled converter. Values given are for an i.f. of 50 to 54 Mc.

C_1—75-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-75).
C_2, C_3, C_4—20-$\mu\mu$f. miniature trimmer (Johnson 20M11).
C_5—25-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-25).
C_6, C_7—500-$\mu\mu$f. feed-through ceramic (Centralab MFT-500).
C_8—Handmade copper-tab bypass; see text.
C_9—Handmade copper-tab variable; see text.
C_{10}—0.5- to 5-$\mu\mu$f. plastic trimmer (Erie style 532–08–OR5).
L_1—13½ turns No. 20 tinned, ⅝-inch diam., ⅞ inch long, tapped at 4½ turns (B & W Miniductor No. 3007).
L_2—5 turns No. 20 tinned, ½-inch diam., ⅜ inch long (B & W Miniductor No. 3003).
L_3—2¾ turns similar to L_2.
L_4—2 turns No. 12 tinned, ¼-inch diam., ¼ inch long.
L_5—1 turns ins. wire between turns of L_4 May be inner conductor of shielded wire, with braid removed.
L_6—Half-wave line, ¼-inch copper tubing, 7 3/16 inches long.
L_7—Quarter-wave line, ¼-inch copper tubing, 5 inches long.
L_8—Loop of insulated wire 1 inch long and ½ inch high projecting through base plate on which line assemblies are mounted. May be made from inner conductor of shielded wire, with braid removed from last two inches.
L_9—2 turns No. 22 enam. around cold end of L_{10}.
L_{10}—6 turns similar to L_2.
L_{11}—11 turns No. 22 enam. close-wound on ⅜-inch slug-tuned form (National XR-91).
L_{12}—4 turns No. 28 silk or enamel wound over cold end of L_{11}.
J_1, J_2—Coaxial fitting.
J_3—Closed-circuit jack.
J_4—4-pin male chassis fitting.
RFC—10 turns No. 22 tinned, ⅛-inch diam. Space turns diam. of wire.

is not visible in the photographs. This loop is mounted on the copper base plate that is under the mixer and r.f. assembly. Its size and proximity to the mixer inner conductor are not particularly critical, as there is a surplus of injection under ordinary conditions of operation.

Adjustment

The first step in putting the converter into operation is to tune up the oscillator and multiplier stages. This process is similar to the adjustment of a transmitter and will not be detailed here. Check to see that the proper frequencies appear as indicated on the schematic diagram. Only enough power at 382 Mc. is needed to develop about 0.5 ma. of crystal current. Anything from 0.2 to 1.0 ma. is satisfactory. Adjustments should be made with no plate voltage on the r.f. stage.

Now connect the converter to a 50-Mc. receiver or converter and peak the i.f. amplifier circuits at about 52 Mc. on noise. Next apply plate voltage and feed a signal into the r.f. stage. Peak the r.f. and mixer capacitors for maximum response at about 434 Mc. These adjustments can be made on noise also, if the circuits were close to resonance originally. If a noise generator is not available, the margin of signal over receiver noise that is obtained on a received signal is also usable, if adjustments are made with care.

The points of connection for the B-plus and the

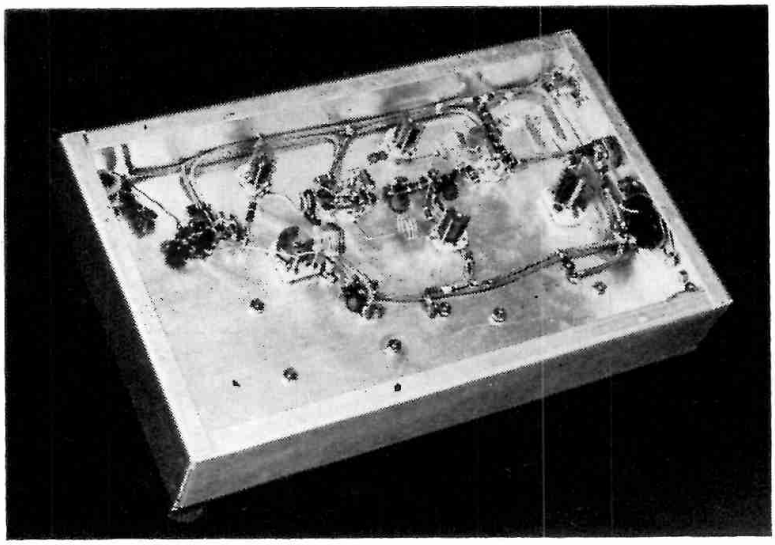


Fig. 16-25—Bottom view of the 432-Mc. converter, showing the oscillator, multiplier and i.f. amplifier circuits.

coupling taps on the r.f. and mixer lines are critical adjustments, but if the dimensions given above are followed carefully the points should be close to optimum. Adjustments can be made and checked readily if the r.f.-mixer assembly is mounted in place temporarily with a few self-tapping screws. (Originally described in January, 1954, *QST*, p. 24.)

A Crystal-Controlled Converter for 1296 Mc.

For simplicity, no r.f. amplifier stage is used ahead of the crystal mixer in the converter shown in Fig. 16-27. While a good amplifier may have advantages over a straight crystal-mixer type of superheterodyne at this frequency, much interesting work can be done with the simpler arrangement. By following certain design principles, to be discussed later, the performance of a crystal mixer can be made very nearly as good as that of the best r.f. amplifier stages.

The converter uses a 1N21E crystal diode in a radial cavity. Injection at 1280 Mc. is furnished by an oscillator-multiplier chain consisting of a 6U8, two 6J6s, and a 446A lighthouse tube. The converter layout leaves plenty of space for changes, and the various units can be modified readily. R.f. connections between stages are made with coaxial connectors and RG-58/U or RG-59/U cable.

The Cavity Mixer

Most mixer cavities described are the coaxial type, but this converter employs the radial variety. In a coaxial cavity the length is the primary frequency-determining dimension. (The diameter has a small effect.) A radial cavity resonates at a frequency almost totally dependent on its diameter. Center loading the radial cavity capacitively lowers its resonant frequency, just as does end loading a coaxial line.

The physical details of the mixer are shown in Figs. 16-28 and 16-29. The dimensions given are not critical. A first model was made by sawing a ¾-inch length off a 4½-inch diameter aluminum pipe for the main body. The ¾-inch length was chosen to accommodate the physical length of the 1N21-series crystals.

The antenna input, the local oscillator injection and the mixing crystal are all on one face of the mixer, spaced 120 degrees apart. The crystal was mounted somewhat closer to the middle of the cavity than the antenna input connector. This was because the antenna connection on an earlier model was a wire ¾ inch long, from the center conductor of the connector to the opposite end plate of the cavity. Assuming the r.f. input impedance of the crystal to be about 100 to 150 ohms, the crystal would have to be closer to the center than the input tap, for the latter to provide a good match for 50-ohm input. Laboratory

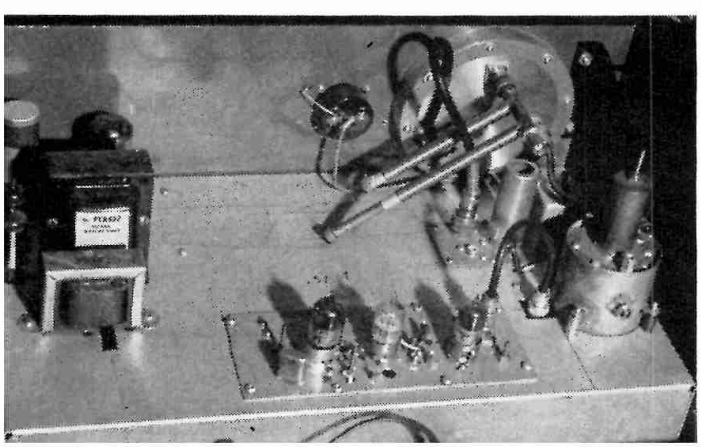


Fig. 16-27—Rear view of the 1296-Mc. crystal-controlled converter. At the back of the chassis is the injection string assembly, except for the 1280-Mc. doubler, which is at the far right. The small assembly adjacent to the doubler is the 16-Mc. i.f. preamplifier. The radial mixer cavity is fastened to the front panel. Power-supply components are at the left.

tests showed that this arrangement worked out quite well, but the loop coupling gives just as good match. The cavity is loaded quite heavily, as image rejection is no problem. So long as the image rejection is 10 db. or more the over-all noise figure will not be adversely affected. The lower Q reduces insertion loss and lessens the mechanical rigidity requirements of the mixer.

Tuning the mixer is done with a ¼-inch brass shaft passing through the end plate opposite to that containing the mixer crystal and coaxial connectors. A penny-sized copper disk is soldered to the shaft for a capacitor plate, and the shaft runs through a locking-type panel bushing adjusted to provide the necessary friction. In practice the mixer rarely requires tuning, but the red-blooded experimenter would rightfully feel cheated if something were not available for adjusting occasionally.

Spring copper wipers, shown in the drawing, were a refinement that was found to be unnecessary, but it may be just as well to add them anyway. Erratic mixer tuning was at first thought to be due to poor contact between the shaft and its bushing. Later it was found that the end plates were "oil-canning." This was cured by mounting the mixer against a heavy panel, as seen in the rear-view photograph, and changing the end plate to a heavier stock. A further refinement to reduce mixer loss and improve tuning stability was to improve contact between the main body and the end plates by undercutting the end faces of the main body,[1] as shown in Fig. 16-29A.

The mixing crystal protrudes through one end plate and contacts the opposite one. The large end is insulated by a tight-fitting piece of spaghetti sleeving. The i.f. output is brought off by a copper tab that presses down on the end of the crystal. The tab also serves as a mixer bypass capacitor. It is fastened to the end plate with two

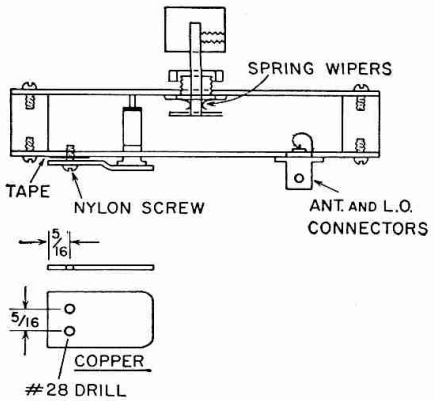


Fig. 16-28—Cut-away view of the radial cavity mixer assembly in the 1296-Mc. converter.

nylon screws, and is insulated from the plate by a strip of plastic electrical tape. (Metal screws, suitably insulated, may be used if the nylon screws are not available.) Because of 40 $\mu\mu$f. of capacitance to ground so provided, and the relatively low impedance of the circuit, about 400 ohms, there is negligible pickup at the intermediate frequency.

I.F. Preamplifier

As no communications receiver has enough gain to accommodate the low signal level from the mixer, a preamplifier is necessary. Other arrangements might give more gain and lower noise figure than the one shown in Fig. 16-30, but none would be more simple or readily adjusted. Like the injection string, the i.f. preamplifier is built as a subassembly. Experimentation with other circuits is thus made easy, but in the meantime the builder of the converter is able to receive signals.

The i.f. amplifier is a 12AT7. The first stage is grounded-grid, and has an input impedance, neglecting input capacitance, of nominally 400 ohms. This happens to be the optimum i.f. im-

[1] Cavities are available in the form shown in Fig. 16-29 from Mooradian Machine Works, 1752 E. 23rd St., Los Angeles. Price, $1.50 each.

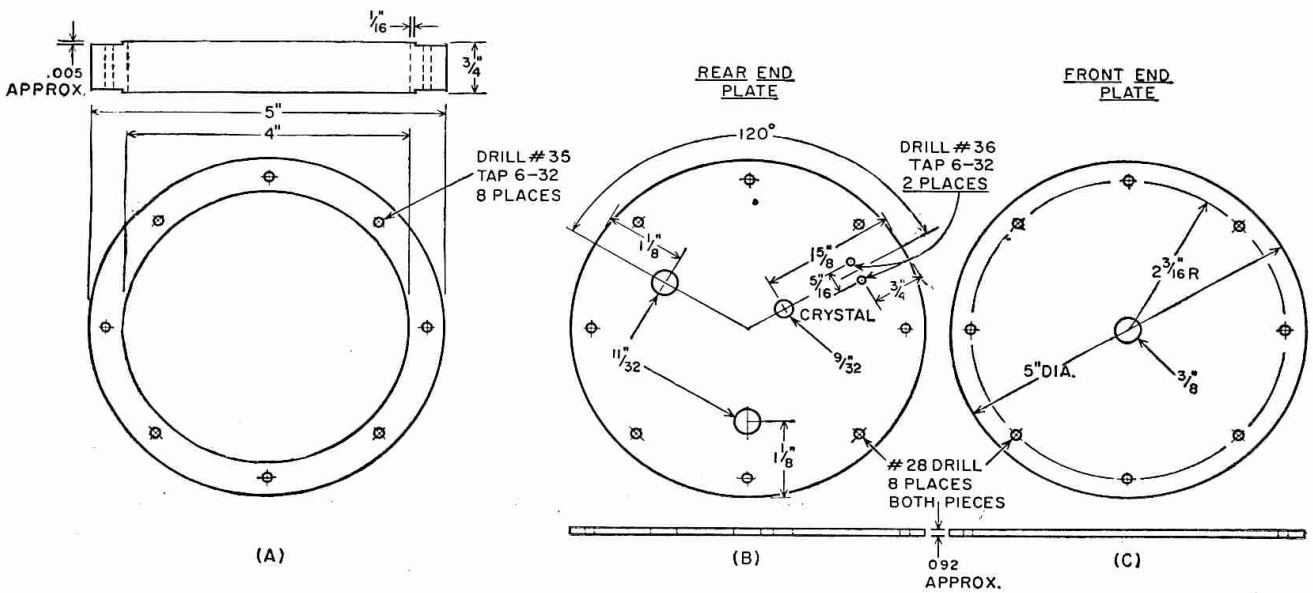


Fig. 16-29—Dimensions of the principal parts of the radial cavity. Material may be aluminum or brass.

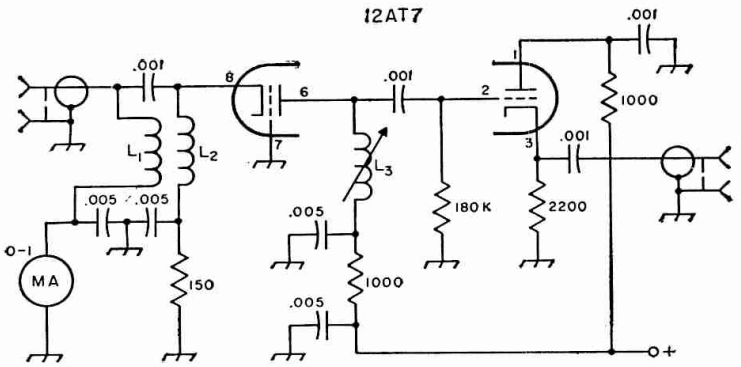


Fig. 16–30—Schematic diagram of the i.f. preamplifier stage.
L_1, L_2—Bifilar-wound choke resonating at 16 Mc. 23 double turns No. 26 enamel, ¼ diam., close-wound.
L_3—25 turns No. 22 enamel, ½-inch slug-tuned form.

pedance for the 1N21-series mixing crystal. The d.c. return for the crystal is through one half of a bifilar r.f. choke winding, the cathode return being the other half. This choke is resonated at the intermediate frequency by the combined effect of the mixer bypass, the coaxial cable and the tube input capacitances. Resonance is checked with a grid-dip meter, removing or adding turns as required. It is not at all critical, as the operating Q is low and the bandwidth consequently is large. Be sure to remove the crystal from the mixer before using the grid-dip meter; otherwise the dip may be questionable.

For maximum gain in the first stage it should be loaded with as low-capacitance and high-impedance a second stage as possible. There should be no tuning capacitance, so a slug-tuned coil is used. The second stage is a cathode follower. This has virtually infinite input resistance, and no Miller-effect capacitance, making the slug-tuned coil the sole frequency-determining element of the preamp. The pass band of the amplifier is very narrow; no more than a few hundred kilocycles. This is not a disadvantage as the coil is quickly adjusted by turning the slug and listening to the noise peak when the receiver is tuned to the desired frequency.

The casual observer may suggest at this point that the cathode follower is useless, as it has no gain and could be replaced by link coupling from the interstage coil. Nothing could be further from correct. A low-impedance link coupled to the interstage coil would load it, and the gain from the grounded-grid stage would drop rapidly. While having a *voltage* gain of only about 0.9, the cathode follower definitely provides gain compared to the conventional grounded-cathode amplifier. Furthermore, its output impedance is very nearly that of the input of most low- and medium-priced communications receivers.

Oscillator-Multiplier Chain

One problem in crystal-controlled converters operating in the higher bands is instability. Having gone to crystal control to achieve stability, one should get as much as crystals will provide. Expensive crystals and ovens can be used, but only as a last resort. Overtone oscillators are ruled out. Possibly their use would save one stage, but at some sacrifice in stability and only a slight reduction in circuit complexity. Crystal-diode multipliers have merit, but their peculiarities may not be familiar to the v.h.f. man already experienced in vacuum-tube multipliers.

As seen in Fig. 16-31, the injection chain starts with a 10.0-Mc. crystal oscillator using the pentode section of a 6U8. Check the frequency before construction progresses too far; if the oscillator is only 8 kc. off, the final injection frequency will be off by more than a megacycle. The 4th harmonic, 40 Mc., is taken from the plate circuit and fed to the triode section of the 6U8, where it is doubled to 80 Mc. Two halves of a 6J6 double twice to 320 Mc. The 320-Mc. circuit is balanced, for optimum coupling to the balanced input of the 6J6 push-push doubler to 640 Mc. The 640-Mc. plate circuit of this stage is a small loop, parallel-resonant with the 6J6 output capacitance, and tuned by a small variable series capacitor.

The final doubler is a 2C40 or 446A lighthouse tube, available at most surplus houses for 25 to 50 cents. It uses a modified ASB-5 mixer cavity, also a surplus item. Little use has been found for this unit in the past, though its companion unit, the r.f. amplifier in the ASB-5, is much in demand. Its input circuit will not quite reach 640 Mc. in its original form, so it was shortened about ¾ inch by installing a false bottom in the cathode cavity. This was made by punching a 1⅜-inch hole in the middle of the plate cover of the mixer, and slotting the rim in numerous places with a jeweler's saw or coping saw. The cavity is heated *on one end only* over a gas stove, and the original bottom and cathode conductor temporarily removed. The modified mixer plate cover is slipped over the cathode connector and soldered in place. The works is then reassembled.

The original input hole will then be covered by the false bottom. However, what was originally the ASB local oscillator injection hole is now in the exact position for the 640-Mc. drive. Under operating conditions the s.w.r. looking into this doubler was found to be 1.19. A BNC fitting is attached to take the 640-Mc. drive.

The 1280-Mc. plate circuit is quite simple, and details are more readily conveyed by drawings than by description. The inner conductor is made

Crystal-Controlled Converter for 1296 Mc.

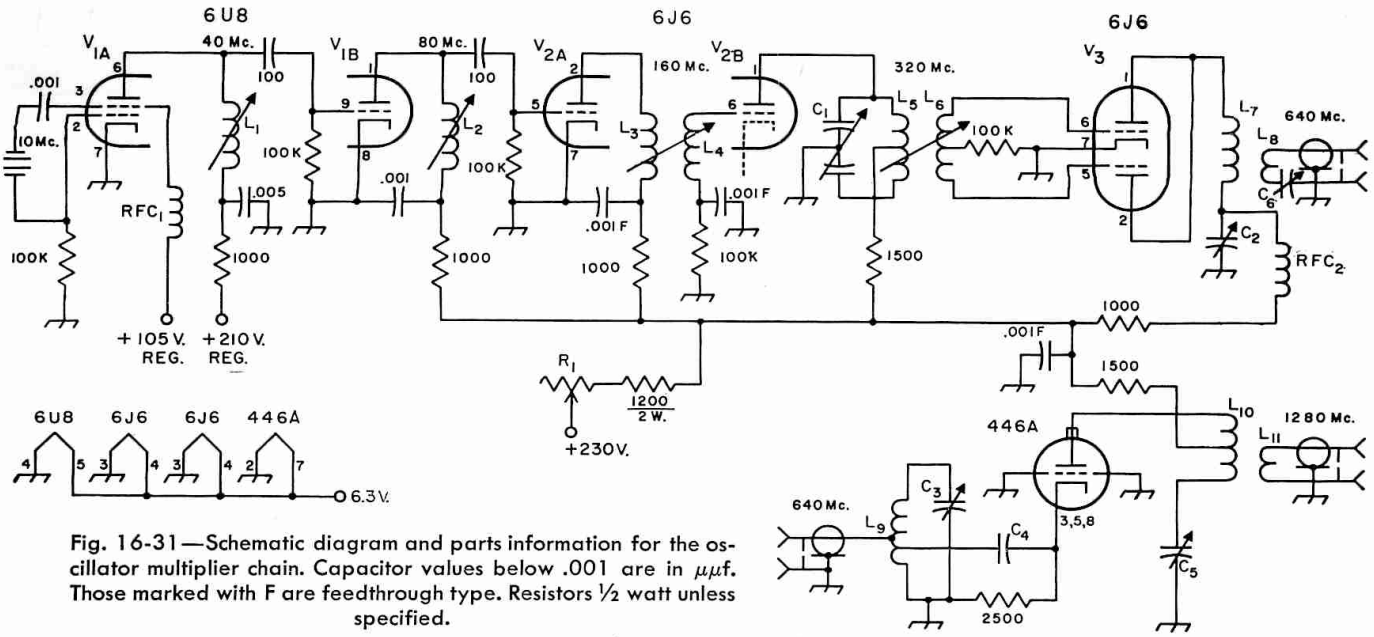


Fig. 16-31—Schematic diagram and parts information for the oscillator multiplier chain. Capacitor values below .001 are in $\mu\mu$f. Those marked with F are feedthrough type. Resistors ½ watt unless specified.

C_1—1.8 to 5.1 $\mu\mu$f.-per-section split-stator variable. (Johnson 160-205).
C_2, C_6—1.8-$\mu\mu$f. plastic trimmer (Erie 532).
C_3, C_4—Part of ASB cavity.
C_5—Piston from trimmer capacitor; slides inside end of L_{10}. See Fig. 5.
L_1—18 t. No. 22 on 9/32-inch diam. iron-slug form.
L_2—8 t. No. 22 on 9/32-inch diam. iron-slug form, spaced to ⅝ inch.
L_3—9 t. No. 22, ¼-inch diam., ¾ inch long.
L_4—8 t. like L_3. L_3 and L_4 are side by side ½ inch apart, c. to c.
L_5—3 t. No. 14, 5/16-inch diam., 7/16 inch long, center tapped. Leads ⅜ inch long.
L_6—2 t. No. 22, ½-inch diam., 7/16 inch long.
L_7—2 t. No. 14, ¼-inch diam., 7/16 inch long. One ⅜-inch lead.
L_8—2-inch length of No. 16 bent into U shape, ¾ inch wide at open end, ¼ inch at bent end.
L_9—Part of ASB cavity.
L_{10}—Final doubler plate circuit; see Fig. 5.
RFC_1—1-mh r.f. choke.
RFC_2—5 t. No. 24 enam. on 1-watt resistor.

from 5/16-inch copper fuel line, slotted at the end to receive the plate cap of the 446A. (See Fig. 16-32.) The insulating bushing, preferably of Teflon, is installed at the other end. The plunger and spring-loaded bearing were taken from a precision piston trimmer found in many pieces of surplus gear. (The ARR-26 has 30 of them.)

The ring attaching the outer conductor to the ASB cavity was cut from flat copper sheet with tin snips. It is held on the end of the ASB cavity and holes are drilled through both it and the cavity. The holes in the cavity are then tapped for 6-32 screws. The outer conductor (brass plumbing pipe) was soldered to the copper ring, and allowed to protrude through it. This makes the outer conductor press firmly against the grid ring of the ASB cavity, giving a good r.f. joint.

The output link enters through a ⅜-inch hole, about ½ inch above the plate cap, as seen in Fig. 16-32. The B-plus feed resistor enters through a 3/16-inch hole. Its cold side is connected to a feedthrough bypass, mounted on a small angle bracket, as close to the hole as possible.

Power Supply

The d.c. output voltage of the power supply is about 230. Two 0B2 regulator tubes in series provide regulated 210 volts for the oscillator plate and 105 for the oscillator screen. Other stages of the injection chain are run from the unregulated output of the supply, through a 2000-ohm wirewound potentiometer. This is useful for varying the injection electrically. It will be found that crystal currents from 50 to 500 microamperes have very little effect on the mixer operation.

The i.f. preamplifier is also operated from the unregulated supply. It will be seen from Fig. 16-31 that the oscillator voltage is left on continuously during transmit or standby periods. This keep-alive voltage eliminates the last few cycles of drift when going from transmit to receive. Stability, both long and short term, was found to be worth the effort when making straight c.w. contacts on 1296 Mc. Stability and c.w. notes were like 80 meters — a vast improvement over several previous converters used on this band. Being able to set up the receiver on a frequency precisely, and then just wait for the signal to come through, is an important factor in DX attempts on 1296 Mc.

Improvements

After the first signal is copied on the converter, ideas will come along for possible improvements.

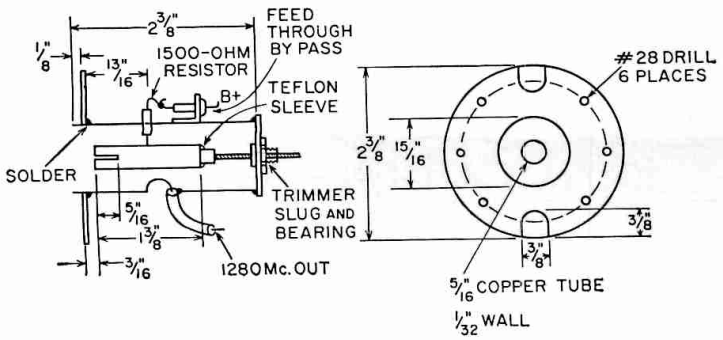


Fig. 16-32—Details of the 1280-Mc. doubler plate circuit. Circular plate fits top of an ASB cavity. See text and Fig. 16-27. The inner conductor of the line fits over the plate cap of the 2C40 or 446A tube. Output coupling loop can be made from the inner conductor of the coaxial line. Tuning capacitor is made from a piston-type trimmer and its bearing assembly.

With the subassembly construction employed ideas can be tried one at a time, with the assurance that unchanged other units will continue to operate properly. A pi-network input to the i.f. preamplifier may aid materially in matching the crystal to the grounded-grid amplifier input. This can be done by adjusting different combinations while listening to a fairly weak signal. The ninth harmonic of a 144-Mc. transmitter can be used as a signal source, though a silicon diode or vacuum-tube noise generator may give better results.

Cascode and pentode preamplifier circuits can be explored. Two problems to be expected will be instability and matching the input imedance of the amplifier. Keep in mind that mere absence of oscillation in an amplifier does not guarantee that it is free of regeneration.

Adequate injection was obtained with the setup described, but some experimenters may experience trouble due to variation in lighthouse tube condition. Many 446A and 2C40 tubes obtained on the surplus market are inferior in one way or another. Nearly all have been removed from equipment, even when they are advertised as "new."

While the ⅜-inch signal injection loop seems to work well there is no assurance at this point that some other size loop or different coupling method would not be better. Untuned mixers deserve consideration. Some experimenters report poor results with these, possibly because of mismatched antenna or feedline impedances. Poor mixer performance may also result from much of the signal being shunted into the local oscillator chain, where it is dissipated. High-*Q* tank circuit design for the injection chain output stage may be helpful here, a point that is often overlooked in 420-Mc. crystal-mixer converters, as well. The untuned mixer has the advantage that the signal may be injected at 640 or 427 Mc. with only slightly degraded performance.

In spite of the fact that most u.h.f. triodes are not supposed to work well at 1300 Mc., several experimenters have used 446As, pencil triodes and the 416B with gratifying results. The *apparent* improved performance may be due to a poorly constructed or improperly adjusted mixer, benefiting greatly from the gain of the r.f. amplifier. The 416B, particularly, was designed for commercial applications at 4000 Mc., so it should be good at 1300 Mc. when properly handled. Poor noise figures quoted for this tube are from data intended for wide-band applications. By proper cavity design the bandwidth can be kept low in amateur applications, and improved noise figures might result.

A good ready-made r.f. amplifier is the ASB-5 (CPR-46-ACJ) r.f. cavity. This was originally intended for use in the 500-Mc. region. Both input and output cavities will tune to 1300 Mc. as three-quarter wave lines. This cavity uses the 446A tube, but adapter rings have been constructed to permit the use of the 416B. The ASB-7 cavity is also useful. In many ways, this unit is more flexible than the former, and more circuit adaptations become apparent during its use. Fairly successful attempts have been made in other directions also. One such amplifier used a 6BY4 ceramic u.h.f. receiving triode on 1200 Mc. As nearly as could be determined, this tube performed as well as a 416B known to be operating properly.

Crystal mixer diodes come in a variety of types. The most common ones are 1N21 and 1N23 series. These have been made in suffixes ranging through the letter "E." The 1N21 series is intended for use from 1000 to 3000 Mc. It will work higher. The 1N23 is intended for use from 3000 to 10,000 Mc., and will work lower. As long as a mixer is operating poorly, or only fairly well, there is virtually no difference in the performance of any of these crystals. It is when a really effective mixer is coupled with an i.f. preamp of 1 to 3 db. noise figure that the amazing difference between 1N21A and the 1N21E becomes evident.

Newer crystals such as the MA421 give noise figures of 6 db. and better in standard test setups. Individually-tailored amateur circuits can be expected to perfrorm even better. Because of semiconductor progress, both in mixers and amplifiers, the vacuum-tube r.f. amplifier at 1200 Mc. appears less desirable than ever.

The rectified crystal current flowing as a result of the local oscillator injection should be measured with a milliammeter having as low a d.c. resistance as possible. Degraded performance may result from the d.c. bias developed across this resistance. The experimenter is invited to try the use of small amounts of back bias on the crystal to improve performance.

(From *QST*, September, 1959.)

V.H.F. Transmitters

Transmitter stability regulations 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 144 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 both bands very readily.

Though no stability restrictions are imposed by law on amateur operation at 144 Mc. and higher, the use of stabilized narrow-band systems pays off in improved effectiveness in both transmitter and receiver. It is this factor, more than the interference potentialities of the wide-band systems, which makes it desirable to employ advanced techniques at 144, 220 and 420 Mc.

The low-power stages of a transmitter for the v.h.f. bands need not be greatly different in design from those used for lower bands, and the techniques of Chapter Six can be used. The constructor has the choice of starting at some lower frequency, usually around 6, 8 or 12 Mc., multiplying to the operating frequency in one or more additional stages, or he can use a high initial frequency and thus reduce the number of multiplier stages. The first approach has the virtue of using low-cost crystals, but h.f. crystals may effect an economy in power consumption, an important factor in portable or emergency-powered gear.

CRYSTAL OSCILLATORS

Crystal oscillator stages for v.h.f. transmitters may make use of any of the circuits shown in Chapter Six when crystals up to 12 Mc. are used, but certain variations are helpful for higher frequencies. Crystals for 12 Mc. or higher are usually of the overtone variety. Their frequency of oscillation is an approximate odd multiple of some lower frequency, for which the crystal is actually ground. Thus 24-Mc. crystals commonly used in 144-Mc. work are 8-Mc. cuts, specially treated for overtone characteristics. The overtone crystals currently being supplied are nearly as stable as those designed for fundamental operation, and they are easy to handle in properly designed circuits.

Best results are usually obtained with overtone crystals if some regeneration is added. This makes for easy starting under load and greater output than would be obtainable in a simple triode or tetrode circuit. Regenerative circuits, with constants for 8- or 24-Mc. crystals, are shown in Figs. 17-20 and 17-24. Triodes are shown, but the same arrangement may be used with tetrode or pentolde tubes. The important point in either case is the amount of regeneration, controlled by the number of turns below the tap in L_1 of Fig. 17-20 or 17-24. There should be only enough feedback to assure easy crystal starting and satisfactory operation under load; too much will result in oscillation not under the control of the crystal.

Overtone operation is possible with standard fundamental-type crystals, using these circuits. Practically all will oscillate on their third overtones, and fifth and higher odd overtones may be possible. Adjustment of regeneration is more critical, however, if the crystals are not ground for overtone characteristics. The frequency may not be an exact multiple of that marked on the crystal holder, so care should be used in working with crystals that are near a band edge.

Crystals ground for overtone service can be made to oscillate on other overtones than the one marked on the holder. For more discussion of overtone oscillator techniques, see *QST* for April, 1951, page 56, and March, 1955, page 16.

Crystals are now available for frequencies up to around 100 Mc. They are somewhat more expensive and more critical in operation than those for 30 Mc. and lower, however. Use of 50-Mc. crystals is made occasionally as a means of preventing radiation of the harmonics from lower frequency crystals that might cause TVI.

FREQUENCY MULTIPLIERS

Frequency multiplying stages in a v.h.f. transmitter follow standard practice, the principal precaution being arrangement of components for short lead length and minimum stray capacitance. This is particularly important at 144 Mc. and higher. To reduce the possibility of radiation of oscillator harmonics on frequencies that might interfere with television orother services, the lowest satisfactory power level should be used. Low-powered stages are easier to shield or filter, in case such steps become necessary.

Common practice in v.h.f. exciter design is to make the tuned circuits capable of operation over the whole range from 48 to 54 Mc., so that the output stage can drive either an amplifier at 50 to 54 Mc. or a tripler from 48 to 144 Mc. Tripling is often done with push-pull stages, particularly when the output frequency is to be 144 Mc. or higher.

AMPLIFIERS

Most transmitting tubes now used by amateurs will work on 50 Mc., but for 144 Mc. and higher the tube types are limited to those having low input and output capacitances and compact physical structure. Leads must be as short as possible, and soldered connections should be avoided in high-powered circuits, where heating may be great enough to melt the solder.

Plug-in coils and their associated sockets or

jack bars are generally unsatisfactory for use at 144 Mc. and higher because of the stray inductance and capacitance they introduce. One way around this trouble is the use of a dual tank circuit in which the inductor for 144 Mc. is a conventional tuned line, with its shorting bar made as a removable plug. When the stage is to be used on another band the short is removed and a coil is plugged into the jack, the line then serving as a pair of plate leads. Such an arrangement will operate as efficiently on 144 Mc. as if it were designed for that band alone.

At 220 Mc. and higher it may be necessary to employ half-wave lines as tuned circuits, as shown in Fig. 17-28 (P_1 in place).

Neutralization of triode amplifiers for 50 and 144 Mc. can follow standard practice, but the stray inductance and capacitance introduced by the neutralizing circuits may be excessive for 220 Mc. and higher. In such instances grounded-grid amplifiers may be used. Driving power is applied to the cathode circuit, with the grid acting as a shield. Some of the drive appears in the output, so both the driver and amplifier must be modulated when a.m. is used. For this reason the grounded-grid amplifier is used mainly for f.m.

Instability shows up frequently in tetrode amplifiers as the result of ineffective screen bypassing. The solution lies in series-resonating the screen circuits to ground, as shown in Figs. 17-13 and 17-24. The r.f. choke and capacitor values vary with frequency, so screen neutralization is essentially a one-band device.

FREQUENCY MODULATION

Though f.m. has not enjoyed great popularity in v.h.f. operation, probably because of lack of suitable receivers in most v.h.f. stations, its possibilities should not be overlooked, particularly for the higher bands. At 420 Mc., for instance, the efficiency of most amplifiers is so low that it is often difficult to develop sufficient grid drive for proper a.m. service. With f.m. any amount of grid drive may be used without affecting the audio quality of the signal, and the modulation process adds nothing to the plate dissipation. Thus considerably higher power can be run with f.m. than with a.m. before damage to the tubes develops or the signal is of poor quality.

Frequency modulation also simplifies transmitter design. The principal obstacle to greater use of f.m. in v.h.f. work is the wide variation in selectivity of v.h.f. receivers, making it difficult for the operator to set up his deviation so that it will be satisfactory for all listeners.

V.H.F. TVI PREVENTION AND CURE

The principal causes of TVI from v.h.f. transmitters are as follows:

1) Adjacent-channel interference in Channel 2 from 50 Mc.

2) Fourth harmonic of 50 Mc. in Channels 11, 12 or 13, depending on the operating frequency.

3) Radiation of unused harmonics of the oscillator or multiplier stages. Examples are 9th harmonic of 6 Mc., and 7th harmonic of 8 Mc. in Channel 2; 10th harmonic of 8 Mc. in Channel 6; 7th harmonic of 25-Mc. stages in Channel 7; 4th harmonic of 48-Mc. stages in Channel 9 or 10; and many other combinations. This may include i.f. pickup, as in the cases of 24-Mc. interference in receivers having 21-Mc. i.f. systems, and 48-Mc. trouble in 45-Mc. i.f.'s.

4) Fundamental blocking effects, including modulation bars, usually found only in the lower channels, from 50-Mc. equipment.

5) Image interference in Channel 2 from 144 Mc., in receivers having a 45-Mc. i.f.

6) Sound interference (picture clear in some cases) resulting from r.f. pickup by the audio circuits of the TV receiver.

There are many other possibilities, and u.h.f. TV in general use will add to the list, but nearly all can be corrected completely, and the rest can be substantially reduced.

Items 1, 4 and 5 are receiver faults, and nothing can be done at the transmitter to reduce them, except to lower the power or increase separation between the transmitting and TV antenna systems. Item 6 is also a receiver fault, but it can be alleviated at the transmitter by using f.m. or c.w. instead of a.m. phone.

Treatment of the various harmonic troubles, Items 2 and 3, follows the standard methods detailed elsewhere in this *Handbook*. It is suggested that the prospective builder of new v.h.f. equipment familiarize himself with TVI prevention techniques, and incorporate them in new construction projects.

Use as high a starting frequency as possible, to reduce the number of harmonics that might cause trouble. Select crystal frequencies that do not have harmonics in TV channels in use locally. Example: The 10th harmonic of 8-Mc. crystals used for operation in the low part of the 50-Mc. band falls in Channel 6, but 6-Mc. crystals for the same band have no harmonic in that channel.

If TVI is a serious problem, use the lowest transmitter power that will do the job at hand. Much interesting work can be done on the v.h.f. bands with but a few watts output, particularly if a good antenna system is used.

Keep the power in the multiplier and driver stages at the lowest practical level, and use link coupling in preference to capacitive coupling.

Plan for complete shielding and filtering of the r.f. sections of the transmitter, should these steps become necessary.

Use coaxial line to feed the antenna system, and locate the radiating portion as far as possible from TV receivers and antenna systems.

Some v.h.f. TV tuners have removable strips that can be replaced with double-conversion inserts for u.h.f. reception. For a number of channels the first conversion frequency may then fall in or near the 144-Mc. band. Where this method is employed for u.h.f. reception the receiver is very sensitive to 144-Mc. interference. The cure is to replace the strips with others having a different conversion frequency, or use a conventional u.h.f. converter for reception of the channels from 14 up.

High-Power Transmitter for 50 and 144 Mc.

The gear described in the next several pages shows how transmitting equipment for two v.h.f. bands can be coordinated in design so as to work from a single exciter. If the builder so desires, the station may be operated from one set of power supplies and speech equipment, with a single set of meters measuring the important currents in both transmitters. Each item can be used by itself, or they combine readily to cover both 50 and 144 Mc., at a power level approaching the legal limit.

In order of their description they are an exciter capable of delivering up to 40 watts output at 48 to 54 Mc., a companion amplifier for the 50-Mc. band, a tripler-driver-amplifier for 144 Mc., and a dual antenna coupler for feeding 50- and 144-Mc. antennas having balanced lines. Their physical appearance is such that they combine neatly for rack mounting, as seen in Fig. 17-1.

THE EXCITER

Though it is shown mounted on the same panel as the 50-Mc. amplifier in Fig. 17-2, the exciter unit might well be used alone, as a versatile 50-Mc. transmitter capable of running up to about 65 watts input. Provision is made for taking off 48-Mc. output at two power levels, through J_3 or J_2, the latter being used for driving the 144-Mc. tripler to be described later.

The exciter is completely shielded, and its power leads are filtered to prevent radiation of harmonics by the power cable. In addition, there are built-in traps to absorb unwanted oscillator harmonics that might otherwise be passed on to the amplifier, or to the antenna. Harmonics of this kind are particularly troublesome when they fall in Channel 2, which is so close to the operating frequency that a filter in the antenna line is relatively ineffective against them.

The interstage coupling circuits are of band-pass design. Once they are properly adjusted they require no further tuning, when the frequency is changed over a 4-Mc. range. Thus only the crystal switch and the output plate circuit need be adjusted when changing frequency.

Circuit Details

The oscillator is a 5763, using crystals above 6, 8, 12, or 24 Mc. for 144-Mc. operation, or 6.25, 8.34, 12.5 or 25 Mc. for 50 Mc. Its plate circuit tunes 24 to 27 Mc., quadrupling, tripling or doubling the crystal frequency. (Crystals at 24 to 27 Mc. are overtone cuts that oscillate at one-third the marked frequency in this circuit.) A series-tuned trap, L_1C_1, in the oscillator plate circuit absorbs the third harmonic of 6-Mc. crystals. This 18-Mc. energy otherwise would pass on to the next stage, where it would be tripled to a frequency in Channel 2. This harmonic has been found to be a common cause of 50-Mc. TVI in Channel 2 areas.

The doubler is also a 5763. A second trap, C_4L_4, in the grid circuit, is tuned to the 7th harmonic of 8-Mc. crystals. The two traps thus prevent radiation of energy in Channel 2, the most critical transmitter problem a 6-meter man is likely to encounter in correcting TVI. They can be modified for other fre-

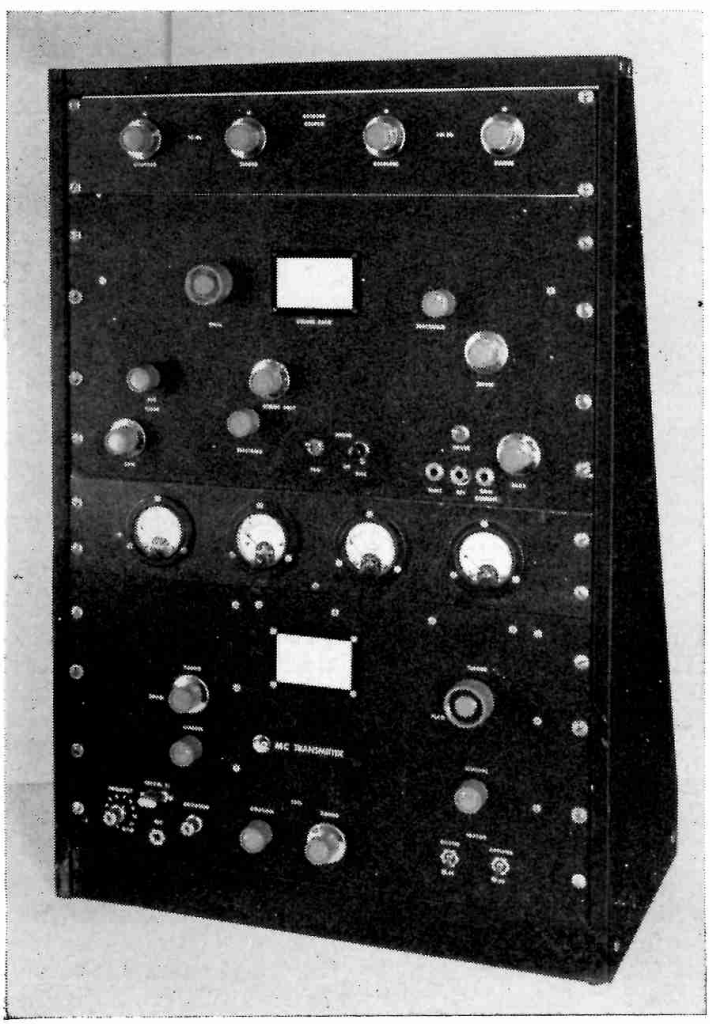


Fig. 17-1—A high-power r.f. section for a 50- and 144-Mc. station. Equipment includes a band-pass exciter for both bands, a 50-Mc. r.f. amplifier built on the same panel, a tripler-driver-amplifier for 144 Mc., and a dual antenna coupler for both frequencies. Units can be operated with a single set of power supplies, and with common speech equipment and meters.

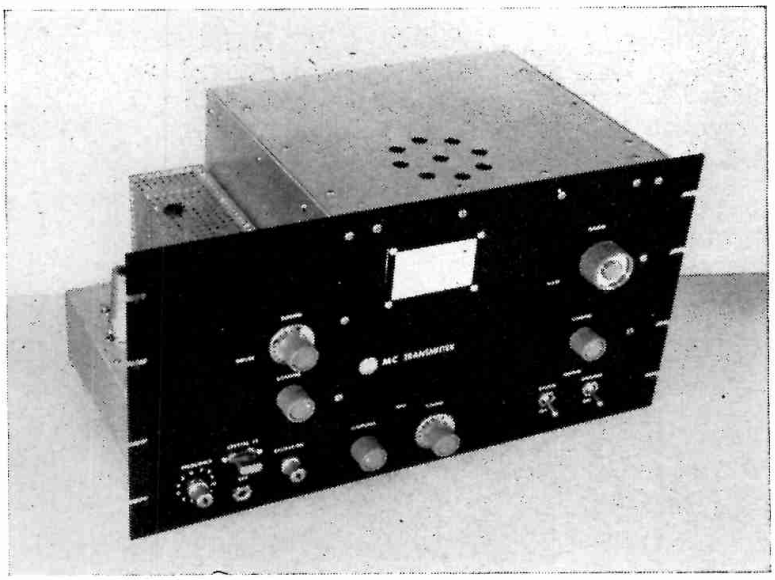


Fig. 17-2—The 50-Mc. r.f. unit. Exciter, left portion on the assembly, also serves on 144 Mc. Amplifier utilizes a 4-125A, 4-250A or 4-400A.

quencies to suit local problems. An example is the 10th harmonic of 8-Mc. crystals, that falls in Channel 6. A trap for the 5th harmonic of the crystal frequency should take care of this.

The 6146 amplifier stage has a shunt-fed pi-network plate circuit. For best stability over the entire operating range the stage is neutralized. The choke, RFC_4, is provided to short out the d.c. voltage that would appear on the output circuit if C_9 should break down. The choke in the plate lead, RFC_5, is for parasitic oscillation suppression. Note that each of the three cathode leads is bypassed separately at the socket. The exciter may be keyed in the 6146 cathode jack, J_4.

Double-tuned band-pass circuits between the oscillator and doubler, and between the doubler and final, provide essentially flat response from 48 to 52 Mc., or 50 to 54 Mc. A potentiometer in the doubler screen circuit provides excitation control for the 6146, and may be used to compensate for variations in drive that may appear at some spots in the band.

The link winding on the doubler plate circuit, L_6, is for the purpose of taking off low-level 48-Mc. output to drive the tripler in the 144-Mc. r.f. unit. Note that the keying jack in the 6146 cathode circuit is the open-circuit type. Removing the key thus disables the 6146 stage, when the first two stages are being used in this way. Separate heater and filament switches on all units allow them to be operated separately. High-voltage supplies may be left connected to all r.f. units, energizing only the filaments and heaters in the ones being used.

Construction

The exciter is built on a 5 × 10 × 3-inch aluminum chassis, with a bottom plate and a perforated aluminum cage to complete the shielding. The small knobs at the lower left of the front view are for the crystal switch and the excitation control. The crystal switch has 12 positions. Ten are for the crystals on the multiple crystal socket (Johnson No. 126–120–1). One more crystal position is provided on the front panel (a convenience if you want to use a frequency not covered by the 10 crystals in the multiple socket), and the 12th switch position is for an external v.f.o. It connects the 5763 grid to the coaxial v.f.o. input fitting, and shorts out RFC_1 and its parallel capacitor. The stage then functions as a frequency multiplier. The output frequency of the v.f.o. could thus be in the 6-, 8- or 12-Mc. range. Above the excitation control may be seen the knobs for the 6146 plate and output coupling capacitors.

Three coaxial connectors are on the rear wall of the exciter. The one at the outside edge is for v.f.o. input. The others are the doubler and 6146 output fittings. Two 4-terminal steatite strips handle the various power and metering leads. Adjacent to each terminal except the ground connection is a feed-through bypass capacitor to take the power lead through the chassis.

TVI that might result from radiation of harmonics by the power leads is prevented by filtering of each lead. The feed-through bypasses are connected to the exciter circuits through r.f. chokes, the inner ends of which are again bypassed with small disk ceramic capacitors. All power leads are made with shielded wire, bonded at intervals to the chassis.

The side view shows the multiple crystal socket at the front of the chassis. Separate crystal sockets may be used if desired. The oscillator and doubler tubes are in the foreground. The trap capacitors, C_1 and C_4, are adjacent to these tubes, while C_2 and C_3 are between them, a bit off their center line. To the rear of the 5763 doubler are C_5 and C_7. The grid tuning capacitor for the 6146, C_6, is just visible inside the amplifier compartment.

A separate lead is provided for each power circuit. Fixed bias for the 6146 is brought in from the bias supply that is part of the high-power amplifier assembly. This bias is desirable to prevent the plate current from rising too high when

the excitation is backed off. If the exciter is used alone, fixed bias is unnecessary. External meters can be connected in any of the circuits at the terminal strips.

The sides, back and top of the amplifier cage are Reynolds "Do-It-Yourself" perforated aluminum sheet, now available in many hardware stores. The pieces are joined together at the corners with lengths of ⅜-inch aluminum angle which can be bought or bent up from sheet stock. The tuning and loading capacitors are mounted on the front of the cage, so this part should be a piece of solid sheet stock rather than the perforated material. The dimensions of the cage are not critical. The original is 5¾ inches deep, 2⅝ inches across, and 4¼ inches high. Make provision for removing the top and outside sheets of perforated stock for convenience in servicing, when the exciter is mounted against the amplifier unit. Extension shafts and couplings bring out the amplifier controls to the panel.

Inside the cage, the 6146 can be seen with its socket mounted above the chassis on ½-inch metal sleeves. The cathode and screen bypasses should connect to separate ground lugs on the top of the chassis, with the shortest possible leads. This wiring can be done conveniently before the socket is mounted on the chassis if nuts are used temporarily to hold the ground lugs in place over the socket mounting screws. The neutralizing adjustment, C_8, is mounted on the rear wall of the cage, and wired to the 6146 plate clip and the feed-through bushing with ⅜-inch wide strips of thin copper. A ceramic insulator mounted on the wall near the 6146 plate cap supports the junction of RFC_5, RFC_3, and C_9. An ordinary tie point supports the other end of RFC_3 and the shielded power lead. The plate coil, L_8, can be seen in back of the 5763 doubler tube, wired between the stators of C_{10} and C_{11}. C_{12} and RFC_4 are mounted near C_{11}, and hooked between its stator bar and a ground lug. A short length of RG-58/U coax runs down through a hole in the chassis from C_{11} over to J_3.

Most of the parts visible in the chassis view can be identified from our description of the panel, rear, and top-side layouts. The oscillator cathode choke, RFC_1, can be seen mounted upright near the oscillator tube and crystal sockets. Both 5763 sockets should be oriented so that Pins 4 and 5 are adjacent to the outside chassis wall. L_1 is visible between C_1 and the oscillator tube socket. L_2 and L_3 run between this socket and that of the doubler. These coils are made from a single length of Miniductor stock with the specified number of turns removed to provide spacing between them. The same applies to L_5 and L_7. These are to the left of the 6146 socket. L_4 is between the doubler socket and C_4. The trap coils are mounted with their axes vertical, to minimize coupling to the band-pass coils. L_6 is wound around and cemented to the bypassed end of L_5.

The power lead r.f. chokes are mounted between single-terminal tie points on the rear lip of the chassis and the feed-through capacitors. The disk ceramic bypasses are then applied to the tie points. A single-terminal tie point mounted under RFC_1 holds one end of the 3300-ohm doubler screen resistor and the lead over to the terminal strip at the rear. A double tie point is mounted between the two 5763 sockets to support the bypassed ends of L_2 and L_3. Another over nearer the rear of the chassis supports the cold end of L_5 and the bottom of the doubler grid resistor.

Wiring will be simplified by the following procedure. Before mounting the crystal switch, ground one terminal of each crystal socket through a bus wire. Connect short lengths of tinned wire to the other terminal of each socket that will be under the switch. Then when the latter is installed, the wires can be run to the proper contacts and soldered in place. Note that the front wafer of the switch is used for shorting out RFC_1, while the crystal socket connections are made to the rear wafer, which is more accessible. The v.f.o. input socket is connected to the proper switch contact with a length of RG-58/U coax.

In assembling the power lead filtering compo-

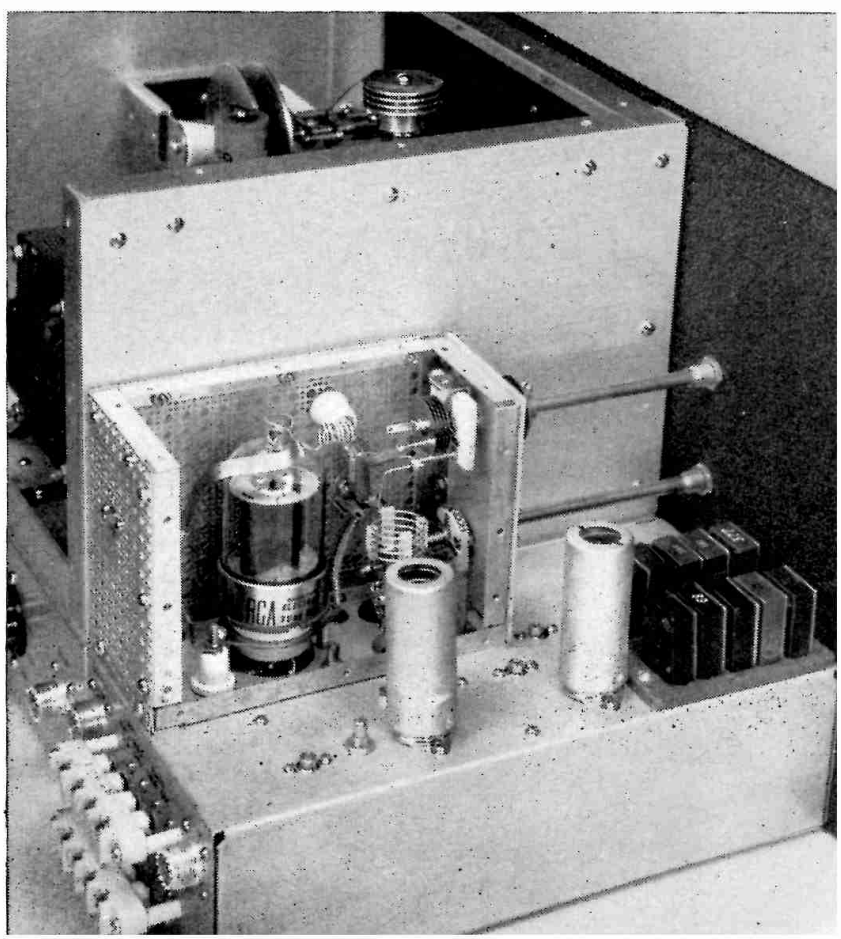


Fig. 17-3—Side view of the exciter, with cover removed. Band-pass coupling circuits eliminate front-panel tuning controls except for crystal switch and output stage tuning.

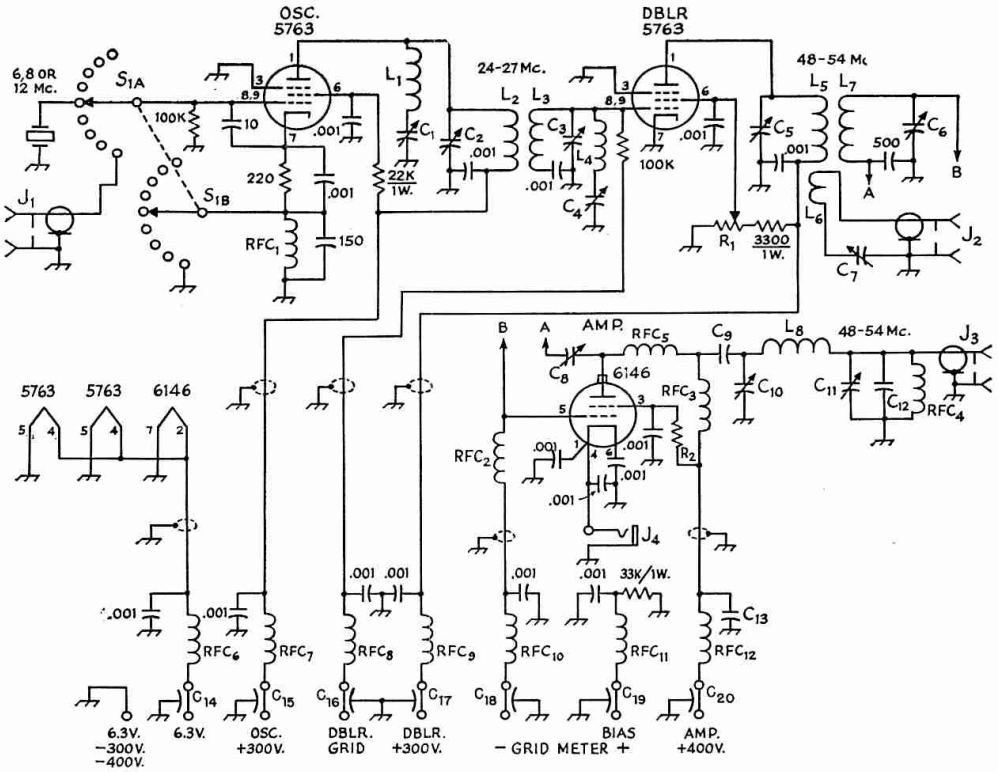


Fig. 17-4—Schematic diagram of 48–54-Mc. exciter. All capacitances less than .001 μf. are in $\mu\mu$f. All .001-μf. capacitors are disk ceramic. All resistors are ½ watt unless otherwise specified.

C_1, C_2, C_3—35-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-35).
C_4—10-$\mu\mu$f. miniature variable (Hammarlund MAC-10).
C_5, C_6—20-$\mu\mu$f. miniature variable (Hammarlund MAC-20).
C_7—50-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-50).
C_8—15-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-15).
C_9, C_{13}—.001-μf. 3000-volt disk ceramic.
C_{10}—35-$\mu\mu$f. miniature variable (Hammarlund HF-35).
C_{11}—100-$\mu\mu$f. miniature variable (Hammarlund MAPC-100B).
C_{12}—100-$\mu\mu$f. 1000-volt mica.
C_{14}–C_{20}—.001-$\mu\mu$f. feedthrough-type ceramic (Centralab FT-1000).
L_1—16 turns No. 24, ⅝-inch diam., 32 t.p.i. (B & W Miniductor No. 3008).
L_2, L_3—12 turns each No. 20, ⅝-inch diam., 16 t.p.i (B & W Miniductor No. 3007). Make from one piece of Miniductor with 5 turns removed between coils. Cold ends are adjacent.
L_4—10 turns No. 20, ½-inch diam., 16 t.p.i. (B & W Miniductor No. 3003).
L_5, L_7—6 turns No. 20, ½-inch diam., 16 t.p.i. (B & W Miniductor No. 3003). Make from one piece of Miniductor with 3 turns removed between coils.
L_6—2 turns hookup wire wound around cold end of L_5 and cemented in place.
L_8—4 turns No. 18, ¾-inch diam., 8 t.p.i. (B & W Miniductor No. 3010).
J_1, J_2, J_3—Coaxial chassis fitting (Amphenol 83-1R).
J_4—Open-circuit phone jack.
R_1—25,000-ohm 4-watt pot.
R_2—33,000-ohm 3-watt (3 100,000-ohm 1-watt in parallel).
RFC_1—2.5-mh. r.f. choke (National R-100S).
RFC_2, RFC_3, RFC_4—7-μh. solenoid v.h.f. choke (Ohmite Z-50).
RFC_5—6 turns No. 22 tinned wire, ¼-inch diam., spaced one-wire diam.
RFC_6–RFC_{12}—15 turns No. 24 enam. close-wound on high value 1-watt resistor.
S_1—2-pole 12-position miniature ceramic rotary (Centralab PA-2005).

nents at the rear of the chassis, the disk ceramic bypasses can most easily be mounted on the tie points before the latter are fastened inside the chassis. Wiring up the power leads should be done before the r.f. chokes are mounted in place.

THE 50-MC. AMPLIFIER

Though the exciter and amplifier are pictured on a single panel, the possibility of using either by itself should not be overlooked. The exciter will make a fine low-powered transmitter, and the final amplifier may be used with any exciter delivering 15 watts or more.

It will take up to the legal limit of power with a 4–400A tube, 750 watts with a 4–250A, or 400 watts with a 4–125A.

The plate circuit is a larger version of the one used in the 6146 stage of the exciter, a shunt-fed pi-network. Operation is completely stable without neutralization, probably because the natural neutralized frequency of the tubes is close to 50 Mc. Provision was originally made for neutralization, but it was found to be unnecessary. Parasitic suppression devices were not required, but if the layout is varied appreciably from that shown, the builder should check for both types of instability with great care.

The jack in the filament center-tap lead is for keying, or for insertion of a grid-bias modulator. A bias supply that delivers about 50 volts negative for the 6146 and 150 for the final amplifier is included in the final stage assembly. Filament transformers for the exciter and final are also part

of this unit. Separate filament switches are included; one for the exciter and the other for the final tube and the blower motor. Power leads except the high voltage, are brought in on an 8-pin plug.

Building the Amplifier

A 12 × 10 × 3-inch aluminum chassis is used for the amplifier unit. Thus, it may be combined with the exciter on a 10½-inch rack panel, if desired. The amplifier controls mounted near the panel bottom are, left to right, the input link reactance capacitor, C_1; the grid tuning capacitor, C_2; and S_1 and S_2. S_1 applies a.c. to the transformer for the exciter heaters and to the bias supplies. S_2 applies a.c. to the filament transformer of the amplifier and starts the cooling fan. Above the switches on the panel are the amplifier plate tuning and loading controls.

On the rear of the chassis, coaxial connectors for r.f. input and output are mounted at either end. Between them are the high-voltage connector for the plate supply, the cathode circuit jack, and a fitting for the remaining power and meter leads.

Above the chassis, the 4–250A tube is seen near the front of the chassis. Note that its socket is mounted on ½-inch sleeves. Holes ⅜ inch in diameter are drilled in the chassis directly underneath those provided in the socket for the passage of cooling air. Holes are also drilled adjacent to the cathode, grid, and screen pins to pass their leads. Bypassing of cathode and screen is done above the chassis. The heat radiating plate connector for the 4-250A was cut down to four fins to reduce the over-all height requirement. The filament transformer, T_3, and the screen modulation choke, L_4, are also topside.

The amplifier plate circuit components are to the left of the tube. The tuning capacitor, C_7, originally a neutralizing capacitor, is mounted on the side wall of the shielding assembly. Two modifications should be made to the neutralizing unit before mounting. The circular plates supplied should be replaced with larger ones, 3 inches in diameter, to increase the available tuning range. The bearing assembly of the rotor disk must be temporarily removed, and a strap of copper run between the screw holding the bearing in place and the opposite (grounded) end of the square ceramic insulating pillar, grounding the capacitor rotor. Two copper straps must be inserted between the stator disk and its insulator, to connect the stator with the blocking capacitor, C_5, and with L_3.

The blocking capacitor, the shunt-feed r.f. choke, RFC_2, and the high-voltage bypass, C_6, are assembled into one unit before mounting in the amplifier. This is done with the aid of the hardware supplied with the TV-type high voltage capacitors. The bypass capacitor, on the bottom of the stack, is equipped with one terminal threaded and one tapped. The latter is on the bottom end, for fastening the assembly to the chassis. The threaded terminal screws into the 2½-inch ceramic insulator upon which RFC_2 is wound. The ends of the choke winding are secured by lugs at each end of the insulator. C_5 should be fitted with a threaded terminal at the lower end for screwing into the top of the insulator. This also serves to fasten the ¾-inch wide strip of copper which runs up to the 4–250A plate cap. Finally, the longer of the two copper strips coming from the stator of C_7 is screwed to the top of C_5. A ½-inch feedthrough bushing brings the high-voltage up to the hot side of C_6. The loading capacitor, C_8, is mounted on the chassis directly underneath C_7. The plate coil, L_3, gets rather warm when the rig is operated at high power level, so both of its ends must be bolted in place rather than soldered. One end is bent around and fastened under a

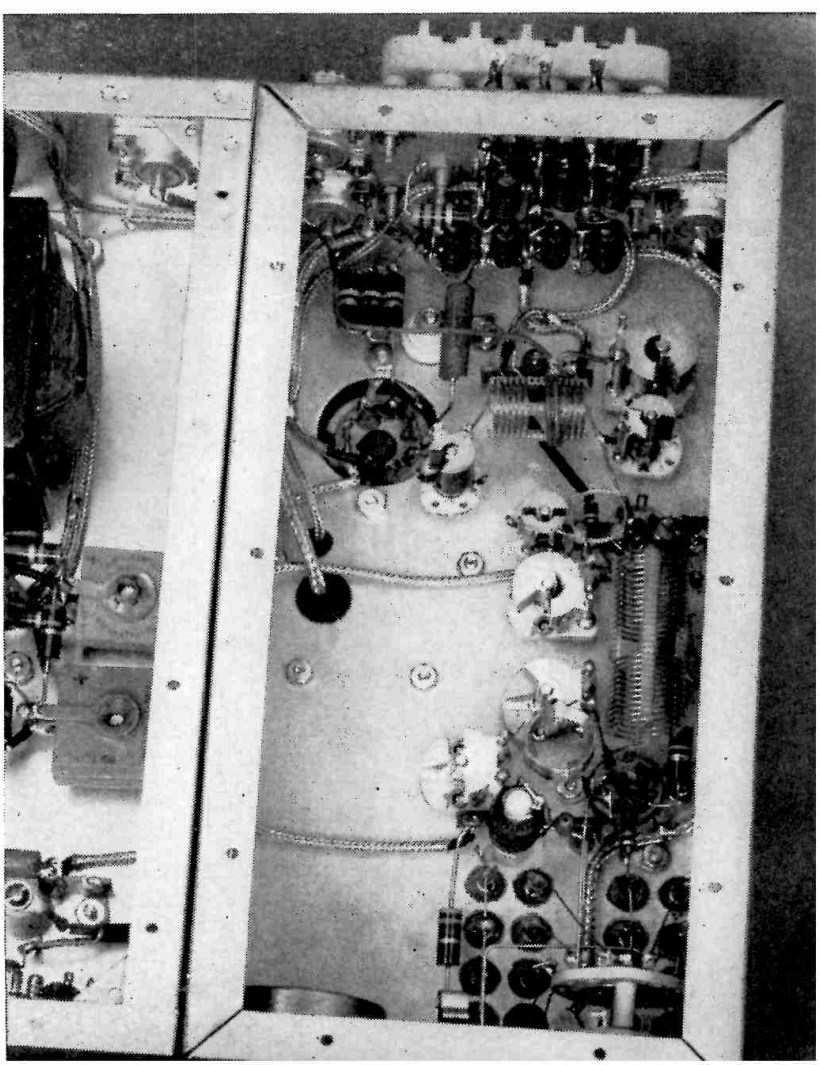


Fig. 17-5—Bottom view of the 50-Mc. exciter, showing band-pass circuits and TVI protective measures.

Fig. 17-6—Interior of the 50-Mc. final amplifier. Plate tuning capacitor is modified neutralizing unit, left.

nut provided on the stator of C_8. The other is bolted to the short length of copper strap previously fastened to the stator of C_7. A length of RG-8/U coaxial cable is run between C_8 and J_2. At the capacitor end, this cable is connected to lugs under the stator and frame mounting screws.

Solid sheet aluminum is used for the enclosure of this unit, as it must be reasonably airtight except for holes directly above the tube itself. The side that supports C_7 must be of fairly heavy stock for rigidity. Home-bent ¾-inch angle stock was used to hold the assembly together. If the over-all height of the unit is kept to just about that of the 10½-inch rack panel, there will be enough clearance above the tube plate connector.

Most of the under-chassis components are visible in the bottom view. The grid circuit is near the front edge of the chassis. Copper strap connects the tube socket grid pin with the stator of C_2. L_2 then is soldered between this strap and a tie point. L_1 is slid inside the cold end of L_2, and cemented lightly in place.

The cooling fan sucks air in from the side of the amplifier near the back corner. The motor is mounted on an aluminum bracket. The fan as supplied will blow, rather than suck, so the blades must be bent back to reverse their pitch. A small piece of aluminum window screening shields the hole cut in the chassis side for the fan.

Bias supply components occupy the lower left quarter of the bottom view. Layout and wiring of this portion of the rig is anything but critical. Shielded wire was used for all power leads. Bypassing at the power connector should be done with very short leads, and C_{14} should be mounted as close as possible to the high-voltage connector.

Adjustment and Operation

An initial setting of the exciter controls can be made before power is applied, if a grid-dip meter is available. The series traps, L_1C_1 and L_4C_4, introduce varying amounts of reactance across the tuned circuits when they are adjusted, so some further adjustment will be needed after these are set up finally, but the following procedure will result in a close approximation.

Disconnect one end of L_3, Fig. 17-4. Couple the grid-dip meter to L_2 and tune it with C_2 to about 24.5 Mc. Leaving the setting of C_2 at that position, lift one end of L_2. Reconnect L_3 and resonate C_3L_3 to about 25.5 Mc. Reconnect L_2, and the circuits should be set for operation on 48 to 52 Mc. For 50 to 54 Mc., the frequencies should be 25.5 and 26.5 Mc.

Procedure for the second band-pass circuit is similar except for the frequencies involved. For 48 to 52 Mc., disconnect L_7 and tune C_5L_5 to 49 Mc. Reconnect L_7 and disconnect L_5, tuning L_7C_6 to 51 Mc. Reconnect L_5. For the 50- to 54-Mc. range these frequencies would be about 51 and 53 Mc.

50-Mc. Amplifier

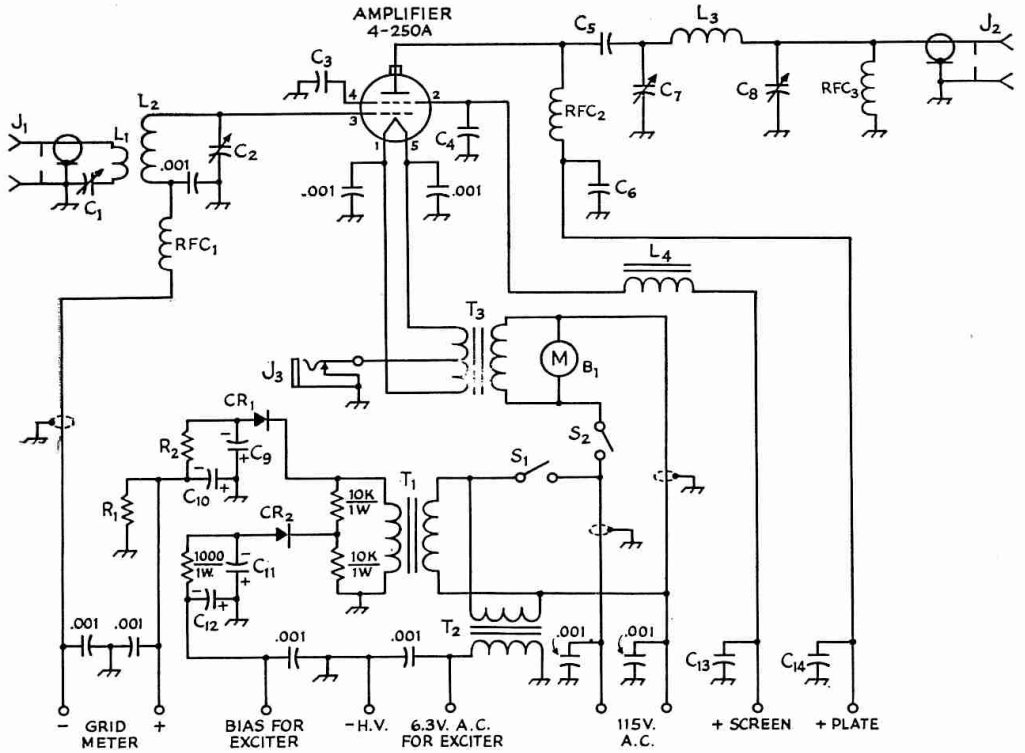


Fig. 17-7—Schematic diagram and parts list for the 4-250A amplifier. All capacitors marked .001 μf. are 600-volt disk ceramic.

C_1—50-μμf. miniature variable (Hammarlund HF-50).
C_2—15-μμf. miniature variable, double-spaced (Hammarlund HF-15X).
C_3, C_4, C_{13}—.001-μf. 1000-volt disk ceramic.
C_5, C_6, C_{14}—500-μμf. 20,000-volt ceramic (Cornell-Dubilier MM120T5).
C_7—Disk-type capacitor with 3-inch diam. plates (made from Millen 15011).
C_8—250-μμf. variable, double-spaced (Johnson 250-F20).
C_9, C_{10}, C_{11}, C_{12}—12-μf. 250-volt electrolytic.
J_1, J_2—Coaxial chassis fitting (Amphenol 83-1R).
J_3—Closed-circuit phone jack.
CR_1—65-ma. selenium rectifier (Federal 1002A).
CR_2—20-ma. selenium rectifier (Federal 1159).
L_1—5 turns No. 24, ½-inch diam., 32 t.p.i. (B & W Miniductor No. 3004).
L_2—4 turns No. 18, ¾-inch diam., 8 t.p.i. (B & W Miniductor No. 3010).
L_3—6 turns No. 12 tinned wire, 1-inch diam., spaced twice wire diam.
L_4—Filter choke, about 10-hy. 100-ma. (Triad C-10X).
B_1—Blower motor and fan (Allied cat. No. 72P715).
R_1—20,000 ohms 10 watts.
R_2—500 ohms 2 watts (2 1000-ohm 1-watt resistors in parallel).
RFC_1, RFC_3—7-μh. solenoid choke (Ohmite Z-50).
RFC_2—Solenoid choke, 42 turns No. 24 d.c.c. close-wound on ½-inch diam., 2½-inch long insulator (National GS-2).
S_1, S_3—Single-pole single-throw toggle switch.
T_1—Power transformer, 135 volts at 50 ma. (Triad R-30X).
T_2—Filament transformer, 6.3 volts at 3 amp. (Triad F-16X).
T_3—Filament transformer, 5.2 volts c.t. at 15 am. (Triad F-11U).

Connect a source of 6.3 volts a.c. at 2.5 amperes or more between the ground and heater terminals, and a low-range meter from the doubler grid return terminal to ground. Insert crystals for the desired frequency range. Apply about 200 volts d.c. to the oscillator plate-screen terminal through a 50- or 100-ma. meter. Current should be 20 to 30 ma., and grid current in the following stage should be about 0.5 ma., when the voltage is increased to the normal 300 volts. Touch up the tuning of the band-pass circuit, if necessary, to get uniform response across the desired range.

The trap circuits can be adjusted at this point, tuning for minimum signal at the frequency to be attenuated in each case. A receiver tuning to the harmonic frequencies is helpful. These will be about 18 to 20.25 Mc. for the first trap and 56 to 60 Mc. for the second, if they are for Channel 2. A TV receiver on the channels to be protected may also be used, merely tuning the traps for minimum TVI. Some slight readjustment of the band-pass circuit may be needed after the final trap tuning is done.

Now remove the grid current meter and ground the metering terminal in the doubler grid circuit. Connect a meter (0 to 5 ma. or more) between the terminals provided for measuring the 6146 grid current. Set the screen potentiometer, R_1, to about the middle of its range and apply about 200 volts to the doubler plate-screen input terminal. Adjust the band-pass circuit, L_5C_5, L_7C_6 for nearly uniform response across the desired range, using the 6146 grid current as the output indication. There should be at least 2 ma. across a 4-Mc. range when the doubler plate voltage is raised to 300. Note that the screen potentiometer controls the input to the doubler, and through it the excitation to the 6146.

The 48-Mc. output coupling adjustment, L_6C_7, may be checked at this time. The line to a 144-Mc. tripler stage should be connected to J_2, and the series capacitor, C_7, adjusted for maximum

grid current in the driven stage. Recheck the adjustment of the band-pass circuit after this is done.

The 6146 amplifier stage had to be neutralized for stable operation. Its adjustment was not critical, however, and C_8 could be set anywhere near minimum capacitance with good results. Start out with its plates meshed about ⅛ inch. With grid drive applied but no plate or screen voltage, tune the 6146 plate circuit through resonance, trying various settings of C_8 until there is no grid current dip at resonance.

A load for the 6146 output circuit is now required. This can be a 40- or 60-watt lamp, with a 50-$\mu\mu$f. capacitor in series to tune out its reactance. Adjust it for minimum reflected power, as indicated on an s.w.r. bridge. With the load connected and grid drive on, apply 300 to 400 volts to the amplifier plate and screen terminal. Tune C_{10} for maximum indicated output. Loading can be adjusted by varying C_{11}, retuning C_{10} after each movement of C_{11}.

Recheck for neutralization at this point, working for a setting of C_8 at which minimum plate current, maximum grid current, and maximum output all occur at the same setting of the plate tuning capacitor, C_{10}. The input can be run up to about 65 watts with plate modulation and 35–40 watts output should be obtained. Higher input can be run on c.w. Plate voltage should not exceed about 400 with plate modulation, though it can be somewhat more for c.w.

Now make a final check on the trap circuits, if necessary. In case TVI is experienced, adjust the traps while someone watches the TV screen, and see whether any improvement is possible. Remember that the traps shown were designed primarily to reduce Channel 2 interference. Where the trouble is with other channels, the traps can be modified to reduce the offending harmonic as required. A low-pass filter or a 4th harmonic trap will be needed if there is harmonic interference in Channels 11–13.

The amplifier as shown furnishes heater voltage and protective bias for the exciter. Hook together the 6.3-volt and ground terminals of the two units, and connect the bias output pin on the amplifier to the 6146 grid return in the exciter. Apply 115 volts a.c. to the appropriate pins on the amplifier power plug. When S_1, Fig. 17-7, is closed, the exciter heaters and the bias supplies are energized. The bias voltages are about 50 and 150 negative for the driver and amplifier, respectively. Closing S_2 lights the amplifier filament and starts the fan motor.

For the initial testing of the amplifier disconnect its fixed bias supply, by lifting the connection between R_1 and R_2, so that instability will be more evident. Connect the output of the exciter through a length of coaxial cable to J_1. Hook a 0–25- or 0–50-ma. meter to the terminals provided for measuring grid current. Turn on the exciter and adjust the driver output and amplifier input for maximum grid current. Set this current between 10 and 15 ma. with the excitation control, R_1, in the exciter. To insure proper adjustment of the amplifier grid circuit, insert an s.w.r. bridge unit such as a Micromatch in the coax connecting the driver and amplifier, and tune C_1 and C_2 in the amplifier alternately for minimum reflected power. Adjust the driver tuning for maximum forward power.

Never apply screen voltage without having the plate voltage on also, and do not operate the amplifier without load. Either will result in excessive screen dissipation, and almost certain tube failure if continued for any length of time. A usable dummy load for testing can be made by connecting two or more 100-watt lamps in parallel. A variable series capacitor, 50 $\mu\mu$f. or more, will be helpful in making the lamp load something like 50 ohms, resistive, at this frequency.

It is well to start with something less than maximum voltages in testing. If the plate voltage is under 1000 and the screen voltage about 200 to 300 volts, little harm can result if something is not quite right. With the dummy load connected, apply plate and screen voltages. Set C_8 near the middle of its range and tune C_7 for maximum output. If this occurs at or close to the end of the tuning range of C_7, adjust the spacing of the turns in the plate coil accordingly. Adjust C_8 for maximum output, returning C_7 as required. If the grid current dropped below 10 ma. under load,

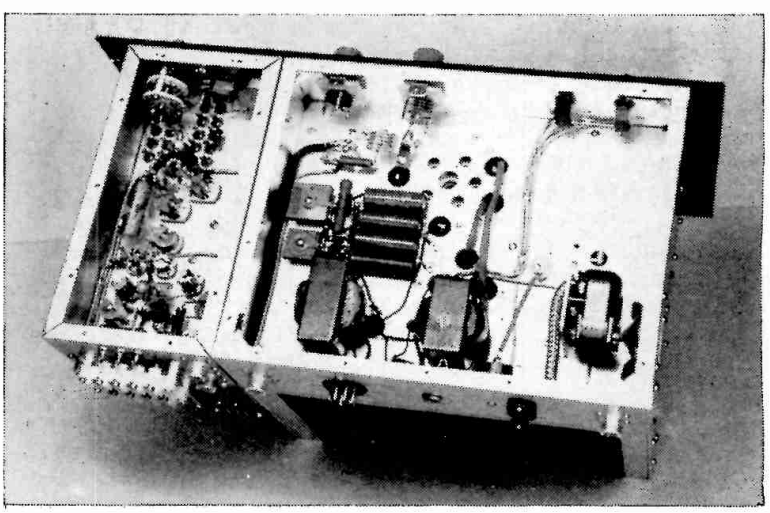


Fig. 17-8—Bottom view of 50-Mc. exciter and amplifier. Note that the two units are built separately, though they mount together on a single panel. Amplifier unit includes bias and filament supplies for both.

increase the drive with the doubler screen potentiometer in the exciter.

Check now for stability. Briefly cut off the drive and see if the amplifier grid current drops to zero. If it doesn't, the amplifier either needs neutralization, or it has a parasitic oscillation. If no grid current shows with drive removed, note whether, when drive is applied and the amplifier is tuned properly, maximum output, minimum plate current and maximum grid current all occur at the same plate tuning. If they do, the amplifier is operating satisfactorily.

If oscillation does show up, check its frequency. If it is much higher than the operating frequency (probably over 150 Mc.) v.h.f. parasitic suppression measures are in order. If it is in the 50-Mc. region, neutralization will be required. These troubles are most common in multiband designs, and unlikely in a layout of this sort. Neutralization of the capacity-bridge type, like that in the exciter, can be incorporated readily, and parasitic suppression is covered in detail elsewhere in this *Handbook*. Neutralization may require *additional* grid-plate capacitance in some layouts. Provision was made for neutralization in the original layout (explaining the plugged hole in the front panel), but it was found to be unnecessary.

When the amplifier is operating stably, the plate and screen voltages may be increased in accordance with the tube manufacturer's ratings, for the type of operation intended. Operating conditions are different for the three tubes which can be used and they should follow the manufacturer's recommendations. This is not to say that variations from the published data are unsafe or undesirable. Any of the values can be varied over quite a range if the maximum rating for each tube element concerned is not exceeded. In this connection, it is highly desirable to provide continuous metering for the grid, screen, and plate currents. This, with a knowledge of the applied voltages, will help insure proper operation and make correct adjustment a simple matter.

A 144-MC. DRIVER-AMPLIFIER

The unit shown in Figs. 17-9 through 17-14 is a three-stage tripler-driver-amplifier that may be used with the exciter just described. Driving power at 48 Mc. may be taken from the doubler stage (by connecting to J_2 in Fig. 17-4) or from the output stage, running at low power. Almost any 50-Mc. transmitter of 3 to 5 watts output could be used by substituting a suitable crystal and retuning the stages for operation at 48 to 49.3 Mc. If a small 144-Mc. transmitter is available, the tripler stage may be dispensed with, in which case about 5 watts drive on 144 Mc. is required.

This section of the station is built in two parts. The tripler and driver stages are in the small portion at the right of Fig. 17-9, with the final stage at the left. All are push-pull stages, the tripler and driver using dual tetrodes. The tripler is an Amperex 6360, followed by an RCA 6524 straight-through amplifier. This drives a pair of 4-125As in the final stage.

Input to the 4-125As can be up to 600 watts on a.m. phone, or 800 watts on c.w. or f.m. By suitable adjustment of screen and plate voltages the power can be dropped as low as 150 watts input and still maintain good efficiency. Some means of reducing power is highly desirable, as most operation on 144 Mc. can be carried on satisfactorily with low power.

The Driver Portion

The tripler and driver stages, Figs. 17-11 and 17-12, both operate well below their maximum ratings. Self-tuned grid circuits are used in each stage. This simplifies construction, and in the case of the driver stage, reduces the possibility of self-oscillation. With a surplus of drive available, the grid circuit of the 6524 may be resonated as low as 130 Mc. There is little tendency to tuned-plate tuned-grid oscillation, therefor, and neutralization is not required.

Tripler and driver are built on a standard $5 \times 10 \times 3$-inch aluminum chassis, with the tripler at the back. Its plate circuit is tuned from the front panel by an extension shaft. Omission of the screen bypass on the tripler is intentional as the stage works satisfactorily without screen bypassing.

The 6524 is easily over driven. This may be corrected by squeezing the driver grid coil turns

Fig. 17-9—The high-power 2-meter rig, with shielding enclosures in place. The small unit at the right houses the tripler and driver stages.

closer together, lowering the resonant frequency until the desired 2.5 to 3.5 ma. is obtained across the band. The farther it can be resonated below 144 Mc. the less likelihood there is of self-oscillation in the driver stage.

The 6524 is mounted horizontally, and holes are drilled in the chassis under the tube to allow for air circulation. Plate leads are made of thin phosphor bronze or copper, bent into a semicircle, connecting the butterfly capacitor and the heat-dissipating connectors. This allows the latter to be removed for changing tubes, without putting undue strain on the plate pins. The connectors have to be sawed or filed down on the insides to fit on the 6524 pins. The coupling link at the driver plate circuit is tuned, to provide efficient transfer of energy to the amplifier grids.

Small feedthrough bypasses are used in the driver screen circuit. C_5 is mounted in the aluminum plate that supports the 6524 socket, and C_6 is in the chassis surface.

Amplifier Features

Design of the 4-125A grid circuit is important in achieving efficient transfer of energy from the driver stage. The input capacitance of the large tetrodes is so high that a tuned grid circuit of conventional design cannot be used at 144 Mc., so a half-wave line is substituted, as shown in Figs. 17-13 and 17-14. The input coupling link is series tuned, permitting adjustment for minimum standing wave ratio on the coaxial line connecting it to the driver stage output link. The grid line, L_1L_2, is made of ¼-inch copper tubing, to reduce heat losses.

Maintaining the 4-125A screens and filament leads at ground potential for r.f. is necessary for stability. To this end, the tube sockets are mounted above the chassis, rather than below. They are elevated only enough to allow the socket contacts to clear the chassis, and are mounted corner to corner, with the inner corners almost touching. The grid line is brought up through ½-inch chassis holes and soldered directly to the grid contacts. This determines the line spacing, about 1½-inches center to center.

The inner filament terminals on each socket are grounded to the chassis. The others connect to feedthrough bypasses with the shortest possible leads. These are joined under the chassis with a shielded wire and tied to the filament transformer. The r.f. chokes in the screen leads are under the chassis, their wire leads coming up through Millen type 32150 feedthrough bushings inserted in chassis holes under the screen terminals. The two screen terminals on each socket are strapped together with a ⅜-inch wide strip of flashing copper. The screen neutralizing capacitor is mounted as close to the sockets as possible and still leave room for the shaft coupling on its rotor. Leads to its stators are about one half inch long.

More compact and symmetrical design is possible if a modified single-section capacitor is used for C_6. It should be the type having supports at both ends of the rotor shaft. The Millen 19140 and Hammarlund MC-140 are suitable units for the purpose. The stator bars are sawed at each side of the center stator plate. The front rotor plate is removed, making a split-stator variable with 4 plates on each stator and 8 on the rotor. This procedure may not be applicable to all 140-$\mu\mu$f. capacitors, but any method that results in a balanced unit having about 50 $\mu\mu$f. per section should do.

Construction of the final plate circuit should be clear from Fig. 17-10. Tuning is done with parts of a disk-type neutralizing capacitor (Millen 15011) mounted on ceramic stand-offs 3½ inches high. These are made of one 1-inch and one 2½-inch stand off each, fastened together with a threaded insert. Connection to the lines is made with copper or silver strap, 4½ inches from the plate end. Silver plating of all tank circuit parts is a worth-while investment, though it should not be considered a necessity. A shaft coupling designed for high-voltage service is attached to the threaded shaft of the movable plate, and this is rotated with a shaft of insulating material brought out to the front panel.

A word about the extension shafts is in order at this point. If they are of metal they may have a serious detuning effect in some circuits, even though they are connected through insulating couplings. Bakelite rod is fine, but since the insulating qualities are of no importance, ¼-inch wooden doweling will do the job just as well. Lucite or polystyrene rod will

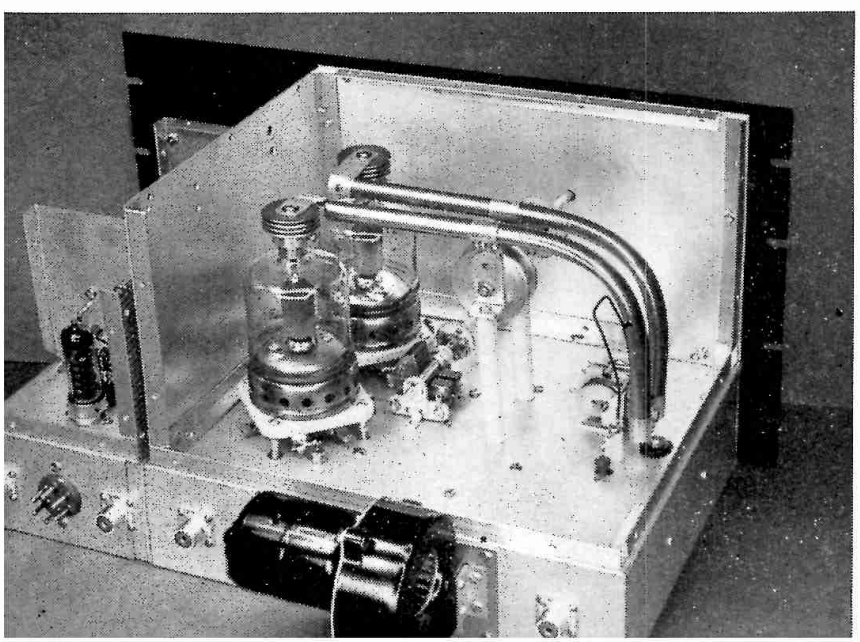


Fig. 17-10—Rear view of the 4-125A final stage. The split-stator capacitor near the middle of the picture is the screen neutralizing adjustment. The plate line is tuned with a capacitor made from parts of a neutralizing unit, mounted on ceramic stand offs.

144-Mc. Driver-Amplifier

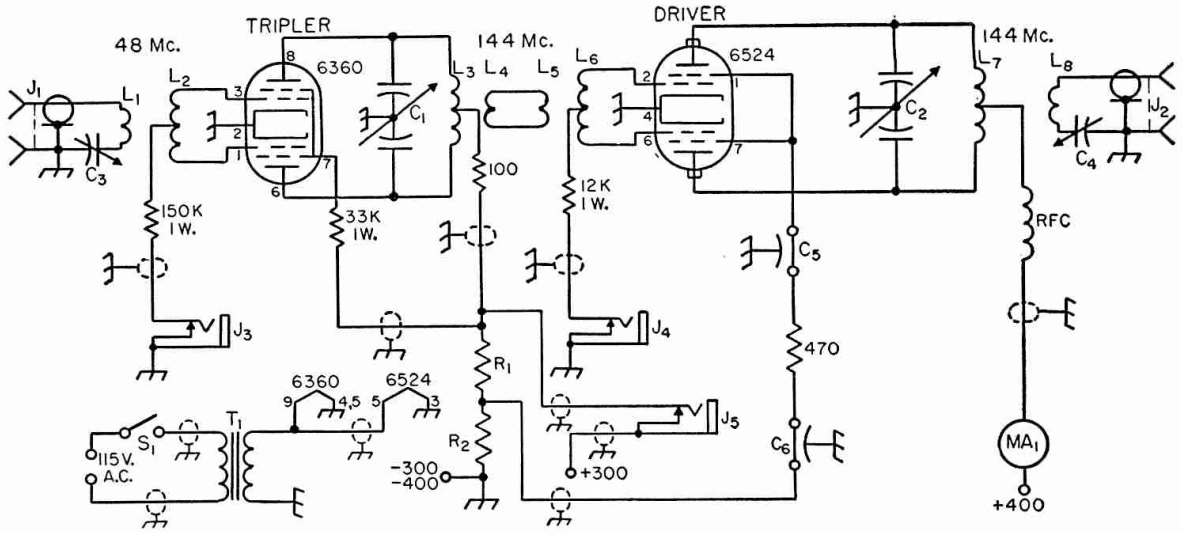


Fig. 17-11—Schematic diagram of the tripler and driver stages of the high-powered 2-meter transmitter.

C_1, C_2—10.5 $\mu\mu$f.-per-section butterfly variable (Johnson 10LB15).

C_3—25-$\mu\mu$f. screwdriver-adjustment variable (Hammarlund APC-25).

C_4—25-$\mu\mu$f. miniature variable (Bud LC-1642).

C_5, C_6—500-$\mu\mu$f. feed-through bypass (Centralab FT-500).

R_1—11,000 ohms 2 watts (two 22,000-ohm 1-watt resistors in parallel.)

R_2—50,000 ohms 2 watts (two 100,000-ohm 1-watt resistors in parallel).

L_1—2 turn insulated wire around center of L_2. Twist leads to J_1 and C_3.

L_2—13 turns No. 20, ⅝-inch diam., ⅞-inch long, center tapped (B & W Miniductor No. 3007).

L_3—3 turns No. 14 enamel, ¾-inch diam., spaced 1/16 inch center-tapped.

L_4—2 turns No. 18 enamel, same as L_3, inserted at center.

L_5—2 turns No. 18 enamel, same as L_6, inserted at center.

L_6—4 turns No. 14 enamel, ½-inch diam., turns spaced wire diameter.

L_7—2 turns No. 14 enamel, 1-inch diam., spaced ¼ inch.

L_8—1 turn No. 14 enamel between turns of L_7.

J_1, J_2—Coaxial fitting, female (Amphenol 83-1R).

J_3, J_4, J_5—Closed-circuit jack. Insulate J_5 from panel and chassis.

MA_1—External meter not shown in photo, 200 ma.

S_1—Toggle switch.

T_1—Filament transformer, 6.3 volts, 3 amp. (UTC S-55).

not stand the heat and should not be used.

The final chassis is aluminum, 10 by 12 by 3 inches, matching up with the driver chassis to fit into a standard 10½-inch rack panel. Complete enclosure is a must for TVI prevention, and it pays dividends in improved stability by providing effective isolation of circuits that tend to give trouble in open layouts.

The enclosures were made by mounting ½-inch aluminum angle stock around the edges of the chassis of both units and cutting the sides and covers to fit. It was not intended to cool the driver unit originally, so the enclosure was made of perforated aluminum. The blower for the final provided plenty of air, however, so three holes are made in the walls of the two chassis to allow some of the air flow to go through the driver enclosure as well. The chassis are bolted together where the vent holes are drilled. The main flow is up through the amplifier chassis, around the 4-125As, and out through the ¼-inch holes drilled in the top cover above the tubes. Holes in the amplifier chassis are drilled to line up with the ventilating holes in the 4-125A sockets. All other holes and cracks are sealed with household cement to confine the air to the desired paths, and bottom covers are fitted tightly to both units.

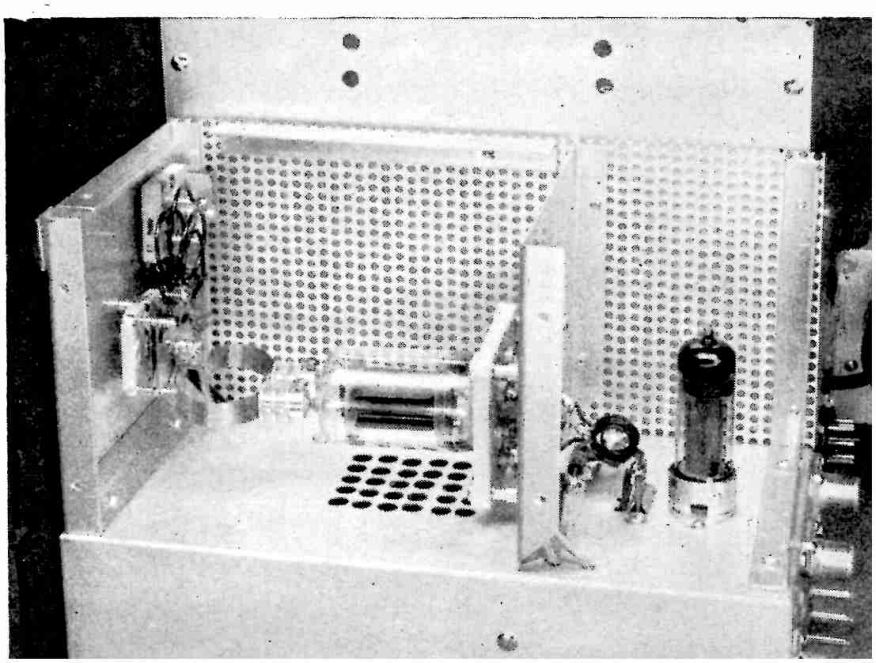


Fig. 17-12—Side view of the tripler and driver stages. Coil adjacent to the 6360 tripler tube is the grid coil for the 6524 driver. Plate leads for the driver tube are flexible copper straps, to permit removal of the tube from its socket. Screwdriver adjustment at the lower right is the reactance tuning capacitor for the tripler input link.

The somewhat random appearance of the front panel is the result of the development of the unit in experimental form. A slight rearrangement of some of the noncritical components could be made to achieve a symmetrical panel layout readily enough.

Operation

The two units have their own filament transformers. Plate supply requirements are 300 volts at 50 ma. for the tripler, 400 volts at 100 ma. for the driver, 300 to 400 volts at 75 ma. for the final screens and 1000 to 2500 volts at 400 ma. for the final plates. The driver plates and final screens may be run from the same supply, but more flexibility is possible if they are supplied separately. A variable-voltage supply for the final screens is a fine way to control the power level.

In putting the rig on the air the stages are fired up separately, beginning with the tripler. A jack (J_3, in Fig. 17-11) is provided on the front panel for measuring the 6360 grid current. About 1 ma. through the 150,000-ohm grid resistor is plenty of drive. The series capacitor, C_3, in the link can be used as a drive adjustment, if more than necessary is available.

Next plug the grid meter into the 6524 grid current jack, J_4, and tune the 6360 plate circuit for maximum grid current. If it is higher than 3 to 4 ma. increase the inductance of the grid coil, L_6, by squeezing its turns closer together. Now apply plate and screen voltage to the 6524, and check for signs of self-oscillation. If the plate circuit is tuned down to the same frequency as that at which the grid coil resonates with the tube capacitance, the stage may oscillate, but if it is stable across the intended tuning range there should be no operating difficulty resulting from a tendency to oscillate lower in frequency, and no neutralization should be needed.

Connect a coaxial line between the driver output and the final grid input preferably with a standing-wave bridge connected to indicate the standing-wave ratio on this line. Tune the driver plate circuit and its series-tuned link for maximum grid current in the final amplifier. Adjust the final grid tuning, C_1, for maximum grid current, and the series capacitor, C_3, in the link for minimum reflected power on the s.w.r. bridge. Adjust the coupling loop position for maximum transfer of power, using the least coupling that will achieve this end.

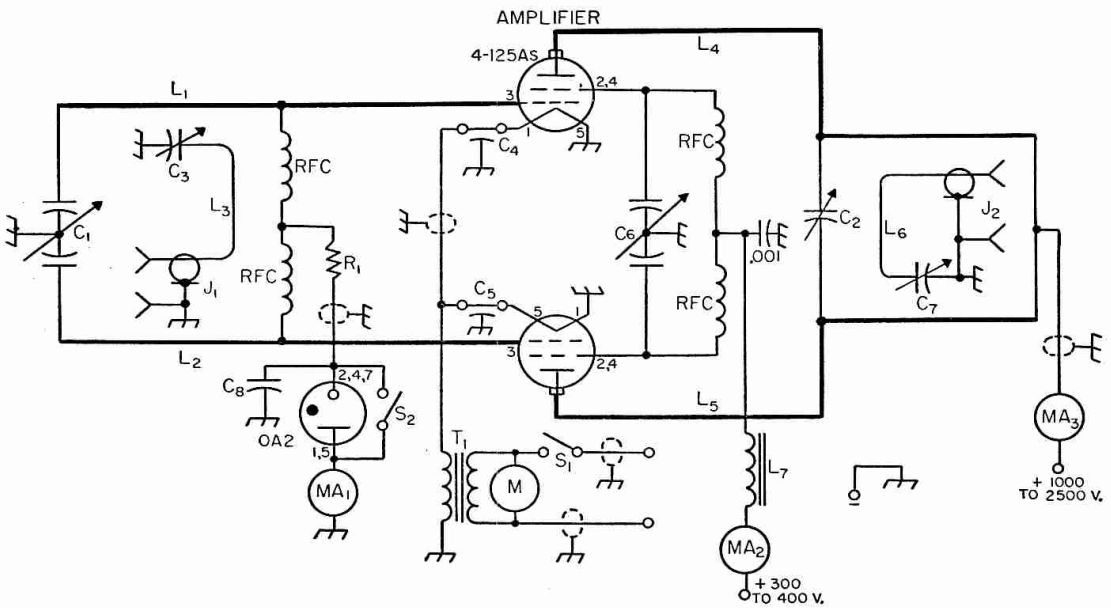


Fig. 17-13—Schematic diagram of the 4-125A amplifier for 144 Mc.

C_1—30-$\mu\mu$f.-per-section split-stator variable (Hammarlund HFD-30X).

C_2—Plate tuning capacitor made from Millen 15011 neutralizing unit; see text and photo.

C_3—25-$\mu\mu$f. miniature variable (Bud LC-1642).

C_4,C_5—500-$\mu\mu$f. feedthrough bypass Centralab FT-500).

C_6—Approx. 50-$\mu\mu$f.-per-section split-stator variable. Make from Millen 19140 or Hammarlund MC-140; see text.

C_7—25-$\mu\mu$f. variable (Johnson 25L15).

C_8—0.25-μf. tubular.

R_1—5000 ohms, 10 watts.

L_1, L_2—¼-inch copper tubing, 12 inches long, spaced 1½ inches center to center. Bend around 1½-inch radius, 1 inch from grid end.

L_3—Loop made from 5 inches No. 14 enamel. Portion coupled to line is 1 inch long each side, about ⅜ inch from line.

L_4, L_5—½-inch copper tubing 12 inches long, spaced 1½ inches center to center. Bend around 2-inch radius to make line 4 inches high. Attach C_2 4½ inches from plate end.

L_6—Loop made from 7 inches No. 14 enamel. Sides spaced 1¼ inches.

L_7—5-h. (min.) 100-ma. rating filter choke.

J_1, J_2—Coaxial fitting, female (Amphenol 83-1R).

MA_1, MA_2, MA_3—External meters, not shown; 100, 200 and 500 ma.

M—Motor-blower assembly, 17 c.f.m. (Ripley Inc., Middletown, Conn., Type 8433).

RFC—V.h.f. solenoid choke (Ohmite Z-144). Four required

S_1—Toggle switch.

S_2—Rotary jack-type switch (Mallory 720).

T_1 — Filament transformer, 5-volt 13-amp. (Chicago FO-513).

Adjust the screen neutralizing capacitor, C_6, for maximum final grid current, with the plate and screen voltages off. Do not attempt to run the final stage without load. With a fixed screen supply the screen dissipation goes very high when the plate load is removed or made too light. It is important to meter the screen current at all times. With 4-125As danger to the plates can be detected by their color, but the screen current is the only indication of possible damage to that element.

There is no suitable inexpensive dummy load for testing a v.h.f. rig of this power level. The best load is probably an antenna. This can be an indoor gamma-matched dipole, fed with coax. Its series capacitor should be adjusted for a standing-wave ratio close to 1:1. The Micromatch can be used in this operation, but adjustments should be made at less than full power. Watch for any sign of heating in the bridge unit.

The position of the coupling loop, L_6, should be adjusted for maximum transfer of energy to the antenna, keeping the coupling as loose as possible. The series capacitor, C_7, can be used as a loading adjustment thereafter. If the screen voltage is continuously variable it will be found that there is an optimum value around 325 to 350 volts.

Below are some conditions under which the rig has been operated experimentally:

Stage	E_p	I_p	E_{sc}	I_{sc}	I_g
Tripler	300 v.	35 ma.	—	—	1.5 ma.
Driver	400 v.	92 ma.	—	8 ma.	3–4 ma.
Final	1000 v.	300 ma.	400 v.	60 ma.	22 ma.
Final	2000 v.	350 ma.	350 v.	45 ma.	20 ma.
Final	2500 v.	400 ma.	320 v.	40 ma.	18 ma

The first and third conditions given for the final stage represent extremes, both exceeding the tubes' ratings in some way, so they are not recommended. At low plate voltages the screen has to be run above recommended ratings to make the tubes draw their full rated plate current and operate efficiently. At high plate voltages the screen dissipation drops markedly. The use of 4-125As at a full kilowatt input exceeds the manufacturer's maximum ratings, and is done at the user's risk. To operate safely, the maximum plate voltage for voice work at 144 Mc. should probably not go over 2000. At this level the tubes will handle 600 watts input on voice, and 750 watts on c.w. easily.

Modulation and Keying

Keying is done in the screen circuit of the driver stage, and in the screen and plate circuits of the tripler. Cathode keying of the driver was attempted, but it caused instability troubles, so was abandoned. The screen method makes the key hot, so an insulated key or a keying relay must be used in the interest of safety. The keying jack must be insulated from the panel.

Fixed bias for the final amplifier is provided by the VR-tube method. When the tube ignites at the application of drive, the capacitor C_8 charges. Removing excitation stops the flow through the VR tube and leaves the negative charge in the capacitor applied to the amplifier grids. The effectiveness of this system requires a low-leakage capacitor for C_8.

Modulation is applied to the plates only. A choke of about 10 henrys is connected in the screen lead, or the modulation can be supplied through a screen winding on the modulation transformer. The bypass value in the screen circuit should be low enough to avoid affecting the higher audio frequencies. Occasionally audio resonance in the screen choke may cause a singing effect on the modulation. If this develops, the choke may be shunted with a resistor. Use the highest value that will stop the singing.

In neutralizing the 4-125As it may be found that what appears to be the best setting of the screen capacitor will result in a very large drop in grid current when plate voltage is applied. The setting may be altered slightly, raising the full-load grid current, without adversely affecting the stability of the amplifier. The final check for neutralization is twofold. There should be no oscillation when drive is removed; and maximum grid current, minimum plate current and maximum output should all show at one setting of the plate tuning capacitor. The latter condition

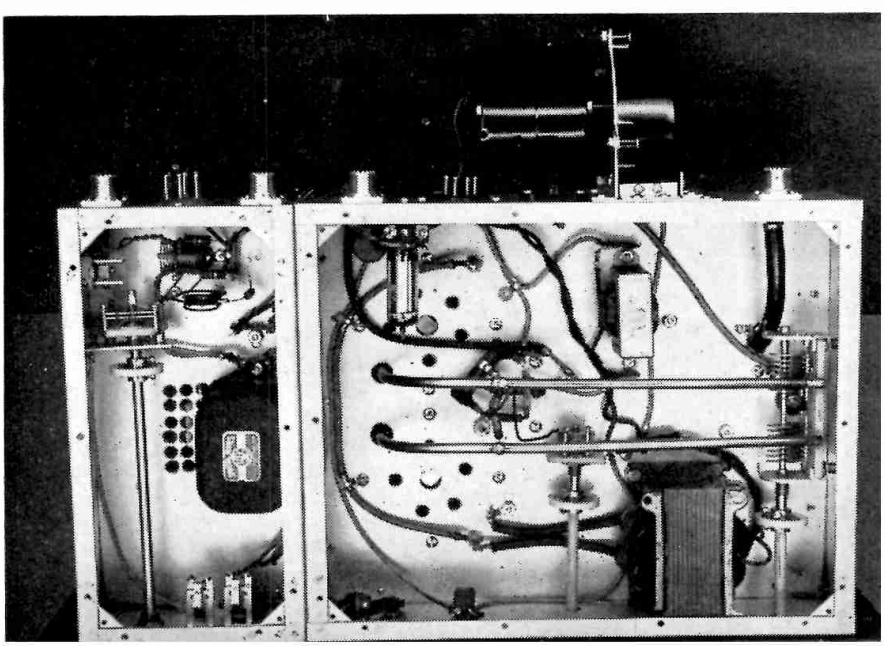


Fig. 17-14—Under-chassis view of the 2-meter transmitter. Tripler grid and plate circuits are at the upper left. Only two of the three jacks on the front panel show in the lower left. The half-wave line used in the 4-125A grid circuit is the main item of interest in the amplifier section. Both units are fitted with bottom covers, to provide shielding and confine the flow of cooling air to the desired areas.

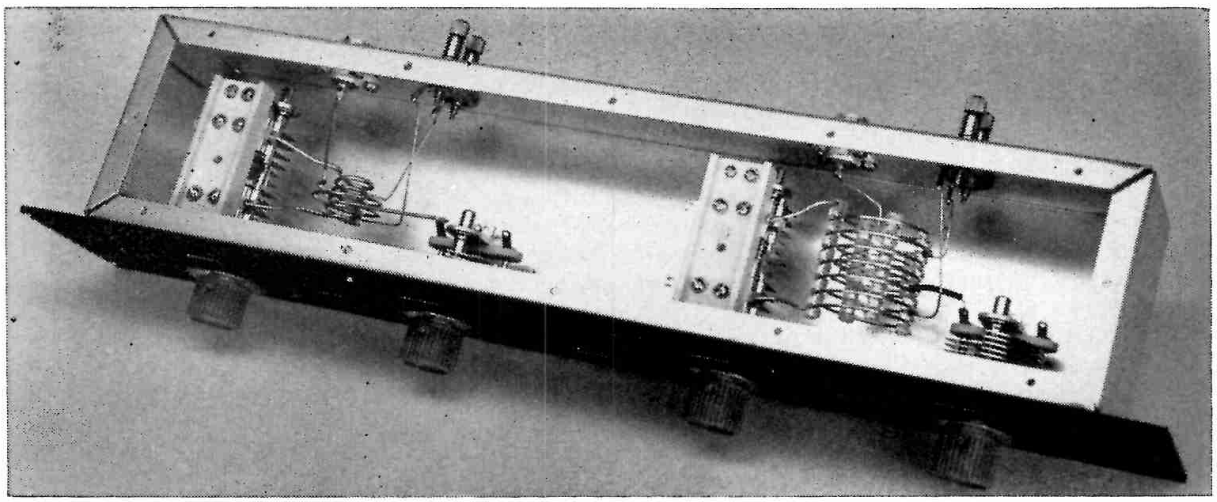


Fig. 17-15—Antenna couplers for 50 and 144 Mc. designed for use with the high-power transmitters on the previous pages.

may be observed only when the amplifier is operated without fixed bias.

ANTENNA COUPLERS FOR 50 AND 144 MC.

The antenna couplers shown in Figs. 17-15, and at the top of Fig. 17-1, can be used with 52-ohm or 75-ohm coaxial line, and with balanced lines of any impedance from 200 to 600 ohms or more. They were designed for use with the high-power transmitters described previously, but may be used at any power level.

Construction

The two couplers are identical circuitwise. They are built inside a standard 3 by 4 by 17-inch aluminum chassis, with a bottom plate to complate the shielding. The panel is 3½ inches high. If only one coupler is required, a 3 by 4 by 6-inch utility box can be used. Terminals on the back of the chassis include a coaxial input fitting and a two-post output fitting for each coupler. The circuit diagram, Fig. 17-16, serves for both.

The 50-Mc. coils are cut from commercially available stock, though they can be made by hand if desired. The coupling winding, L_1, is inserted inside the tuned circuit. The polyethylene strips on which the coils are wound keep the two coils from making electrical contact, so no support other than the wire leads is needed.

Leads to L_1 are brought out between the turns of L_2, and are insulated from them by two sleeves of spaghetti, one inside the other. Do not use the soft vinyl type of sleeving, as it will melt too readily if, through an accident to the antenna system, the coil should run hot. In the 144-Mc. coupler the positions of the coils are reversed, with the tuned circuit, L_2, at the center, and the coupling coil outside it.

Similar tuning capacitors are used in both couplers, but some of the plates are removed from the one in the 144-Mc. circuit. This provides easier tuning, though it has little effect on the minimum capacitance, and therefor on the size of the coil.

Adjusting the Couplers

An antenna coupler can be adjusted properly only if some form of standing-wave bridge is connected in the line between the transmitter and the coupler. If it is a power-indicating type, so much the better, as it then can be used for adjusting the transmitter loading, and the work can be done at normal transmitter power.

With the bridge set to read forward power, adjust the coupler capacitors and the transmitter tuning roughly for maximum indication. Now set the bridge to read reflected power, and adjust the antenna coupler capacitors, first one and then the other, until minimum reflected power is achieved. Unless the line input impedance is very highly reactive, it should be possible to get the reflected power down to zero, or very close to it. Adjustment of the coupler is now complete. Tuning for maximum transfer of power from the transmitter is done *entirely* at the transmitter.

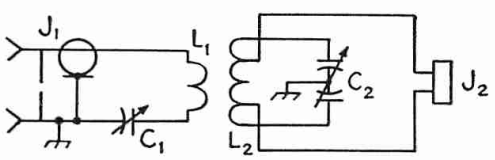


Fig. 17-16—Circuit and parts information for the v.h.f. antenna couplers.

C_1—100-$\mu\mu$f. variable for 50 Mc., 50-$\mu\mu$f. for 144 Mc. (Hammarlund MC-100 and MC-50).

C_2—35-$\mu\mu$f. per-section split-stator variable, 0.07-inch spacing (Hammarlund MCD-35SX). Reduce to 4 stator and 4 rotor plates in each section in 144-Mc. coupler for easier tuning; see text.

J_1—Coaxial fitting, female.

J_2—Two-post terminal assembly (National FWH).

L_1—50 Mc.: 4 turns No. 18 tinned, 1 inch diameter, ⅛-inch spacing (Air-Dux No. 808T).

144 Mc.: 2 turns No. 14 enam., 1 inch diameter, ⅛-inch spacing. Slip over L_2 before mounting.

L_2—50 Mc.: 7 turns No. 14 tinned, 1½ inch diameter, ¼ inch spacing (Air Dux No. 1204). Tap 1½ turns from each end.

144 Mc.: 5 turns No. 12 tinned, ½ inch diameter, ⅞ inch long. Tap 1½ turns from each end.

Simple Transmitters for 50 and 144 Mc.

The two transmitters shown in Fig. 17-17 are designed to fill several needs. They can be used as complete r.f. sections for 50 and 144 Mc., or they serve well as exciters for higher-powered amplifiers. Depending on the final amplifier tubes chosen, the power level can be anything from under 10 to as much as 50 watts input. At low power they are well suited to mobile and portable applications. Provision is included for c.w. operation. Modulation equipment for the transmitters can be found elsewhere in this *Handbook*.

The designs are as similar as possible, mechanically and electrically, the tubes and many of the parts being interchangeable. They are built on standard 5 by 10 by 3-inch aluminum chassis, with shield covers of perforated aluminum over their output stages. These shields are an aid to TVI prevention, and they provide protection for the tuned circuits mounted topside.

Circuitry

Both transmitters employ third-overtone crystal oscillators of simple design. Crystals should be in the range between 8.34 and 9 Mc. or 25 and 27 Mc. for 50-Mc. operation. For 144-Mc. work the crystals are 8 to 8.22 Mc. or 24- to 24.66 Mc. If the feedback in the oscillator circuit is adjusted to make conventional 8-Mc. crystals oscillate on their third overtone, crystals in the 24-to 27-Mc. range will also work. If only the latter (third overtone) type crystals are used, the feedback can be set at a lower level. This is controlled by the position of the tap on the coil, L_1. Crystals in the 8-Mc. range that multiply out close to a band edge should be checked carefully under actual operating conditions in the equipment, as the oscillation frequency may not be exactly three times that marked on the holder.

The oscillator is the triode portion of a 6U8 triode-pentode. The pentode section is a frequency multiplier, doubling to 50 Mc. in the 6-meter transmitter and tripling to 72 Mc. in the 144-Mc. one. The doubler section drives the output stage in the 50-Mc. rig. An extra stage is required to reach 144 Mc. This is a 12AT7 dual triode with its corresponding triode elements connected in parallel, doubling from 72 to 144 Mc. The output stage is a 2E26, where the input power is to be under 25 watts. A 6146 may be used at higher power levels. There is substantially no difference in the driving power required by these tubes, and they can be interchanged with only slight readjustment of the tuned circuits.

When the exciters are to drive an amplifier using an 829B or a 5894, the output tube should be a 2E26. The plate supply voltage need be no more than 300 volts, and as little as 200 may suffice. When the units are used alone the final plate voltage should be 300 for a 2E26, or 400 to 500 for the 6146. If the latter tube is used in exciter service the output will be sufficient to drive tetrode amplifiers of up to 1 kilowatt input.

Construction

Arrangement of parts is not particularly critical, though it would be well for the inexperienced constructor to follow the layouts shown closely in all principal details. Layout drawings, Figs. 17-21 and 17-25 are provided for those who may wish to make exact duplicates. The dimensions given apply only when identical parts to those of the original are purchased. Check sockets, particularly, for mounting dimensions before following the layouts in complete detail.

The shield covers of the two transmitters were made in slightly different ways, to illustrate differing techniques. The method used in the 50-Mc. unit may be the easier of the two for amateurs not well equipped with metal working tools. The front and back plates are 5 inches wide and 4½ inches high. The bottom half inch of each plate overlaps the main chassis, and is fastened

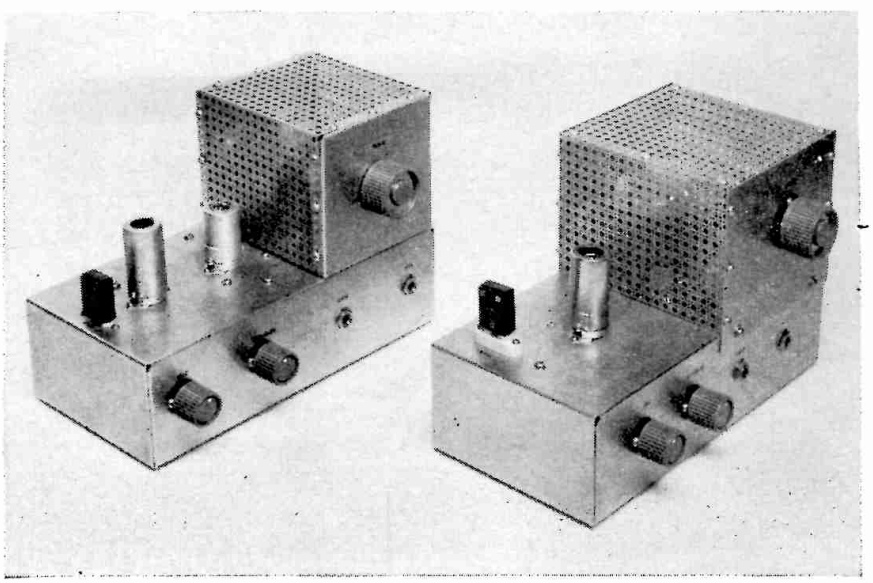


Fig. 17-17—Transmitters for 50 (right) and 144 Mc. Designs are similar and many parts are interchangeable. Power ratings may be varied from under to more than 50 watts input, depending on tube used in the output stage.

to it with self-tapping screws. The cover is made of perforated aluminum, available in many hardware stores. This can be cut and bent with simple tools. The box thus made is 4 inches high, 5 inches wide and 5 inches deep. The perforated cover is made larger than these dimensions by about ⅜ inch on all sides. The extra material is bent over so that the front and back plates can be fastened to it with self-tapping screws.

In the 144-Mc. transmitter the edges of the front and back plates are bent over, so that the cover need be only a plate bent into an inverted U. The enclosure is 4 by 4 by 5 inches in size. The bent-over edges of the front and back walls show plainly in the top view, Fig. 17-22.

Building the 50-Mc. Transmitter

Looking at the bottom view of the 50-Mc. transmitter, Fig. 17-19, we see the oscillator tuning capacitor, C_1, and the plate coil, L_1, at the right. Next to the left is the 6U8 socket. The doubler plate coil, L_2, and the amplifier grid coil, L_3, are between the tube sockets. Note that these coils are mounted side by side, with their axes vertical. Their position with respect to each other is adjusted for maximum grid drive, with the optimum spacing being about one coil diameter. The amplifier screen-dropping resistor (4 1-watt resistors in parallel) is just above the 2E26 socket. Jacks for cathode keying and grid-current measurement occupy the left side of the front wall, as seen in Fig. 17-19.

Arrangement of parts inside the shield compartment can be seen in Fig. 17-18. The amplifier tube, a 6146 in this instance, is at the left side of the box. The plate tuning capacitor, C_4, is near the middle of the front wall. The antenna loading capacitor, C_5, and the coaxial output fitting, J_1, are on the rear wall. The power connector strip is centered on the rear wall of the chassis. Note the parasitic choke, L_4, between the tube and the plate coil. This is wound on the resistor in parallel with it. The plate coil, L_5, is mounted with its axis vertical. The output coupling coil, L_6, is close against the bottom of L_5, and insulated from it by spaghetti sleeving.

The type of socket used for the amplifier tube is important. Do not use the common moulded socket with an elevated grounding ring having 4 lugs spaced around its circumference. These lugs may introduce coupling between the circuits grounded or bypassed thereto, causing instability that cannot be neutralized out. A Millen ceramic socket was used in the original, but any type that does not have the separate grounding lugs and ring is suitable. Grounding should be done to lugs under the nuts used for mounting the socket. It is imperative that bypass capacitor connections be made with virtually no leads at all, particularly in the amplifier circuits. Note that each cathode lead is bypassed separately. This is important where the cathode is keyed, as in this instance.

The neutralizing capacitor, C_3, is a type intended for mounting with one side grounded, so another mounting method must be provided in this application. A small tab of copper about ⅜ by 1 inch in size supports the capacitor, the end of the tab being soldered to a lug on the 3-lug tie-point strip nearest the socket. The 150-$\mu\mu$f. bypass at the low end of L_3 connects from that point to the ground lug at the middle of the terminal strip. The lead from the sleeve of C_3 is a stiff wire that passes up through a ⅜-inch hole in the chassis to the lower stator terminal of the plate tuning capacitor, C_4. The latter is mounted with its stator terminals one above the other.

Adjustment and Operation

For initial tests a power supply capable of delivering 200 to 300 volts d.c. at about 100 ma., and 6.3 volts a.c. or d.c. at 1.7 amperes may be used. (Only 1.25 amp. will be needed if a 2E26 is used.) The negative side of the plate supply and one side of the heater supply are connected together. The oscillator is tested first. This is done by feeding plate power to the 4700-ohm resistor in the oscillator plate lead only, disconnecting the doubler plate-screen lead temporarily.

Apply heater voltage only, and allow the tubes to warm up for 30 seconds or more. Connect a100-milliampere meter in the lead to the plate sup-

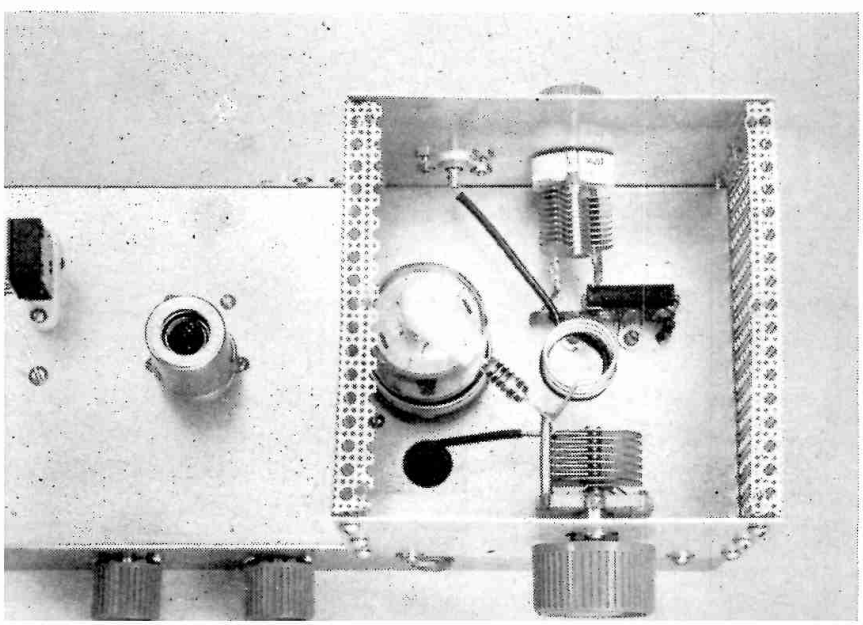


Fig. 17-18—Looking down inside the amplifier shield. The plate tuning capacitor, C_4, is on the front wall, with the loading adjustment, C_5, on the rear wall. Parasitic suppressor and plate coil connect to top stator bar of C_4. Black lead, lower left, runs through a rubber grommet to the neutralizing capacitor, below the chassis.

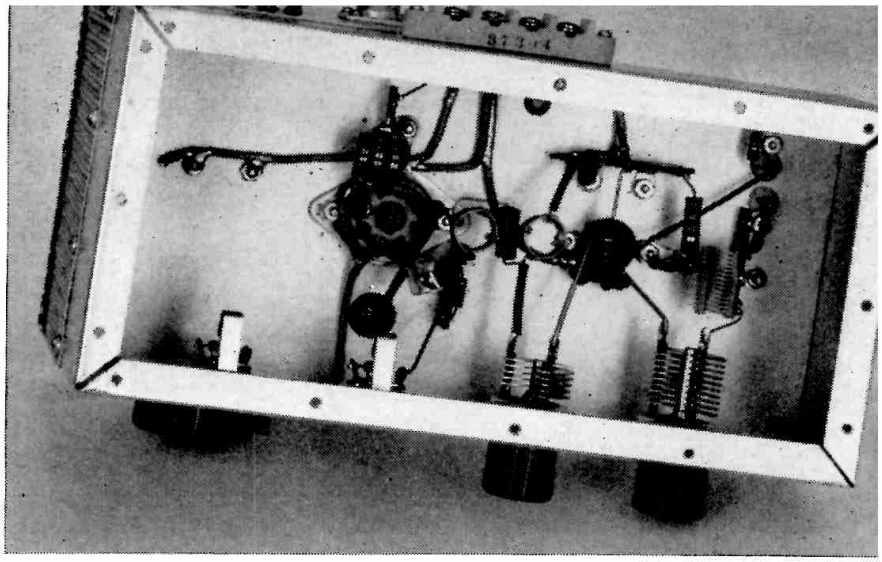


Fig. 17-19—Bottom view of the 50-Mc. transmitter. Note positions of the various coils, particularly those in the doubler plate and amplifier grid circuits, near the middle of the assembly.

ply, and apply power. Swing the oscillator tuning capacitor, C_1, through its range. There will be a sharp dip in current to about 10 ma. as the crystal starts oscillating.

Check the frequency of oscillator with a grid-dip meter or wavemeter. If you have a receiver that tunes the 25- or 50-Mc. region, listen for the oscillator to determine if it is crystal controlled. The frequency will change only slightly, if at all, when the circuit is tuned through resonance. Listen to the note with the receiver beat oscillator on, and place a screwdriver or other metal object near the tuned circuit. There should be very little change in frequency. Should the frequency change more than a few hundred cycles under these tests the oscillator may not be controlled by the crystal.

Self-oscillation is the result of too much feedback. This can be corrected by moving the tap lower on the coil. Too little feedback may prevent the oscillator from working at all, or it may drop out of oscillation when loaded appreciably by the following stage. The cure is to raise the tap position on the coil.

When the oscillator is working correctly, remove the milliammeter from its power lead and connect it between the high-voltage source and the junction of the screen resistor and 1000-ohm resistor at the low end of the plate coil. Plug a low-range milliammeter, preferably 5 or 10 ma., into the grid current jack, J_2, of the amplifier. Apply plate voltage to the first two stages and tune the doubler plate circuit for maximum grid current, as read on the meter in J_2. This should be at least 2 ma., with a 250-volt plate supply. Try varying the separation between L_2 and L_3, leaving spacing at the point that yields greatest grid current. Retune the doubler plate circuit as the

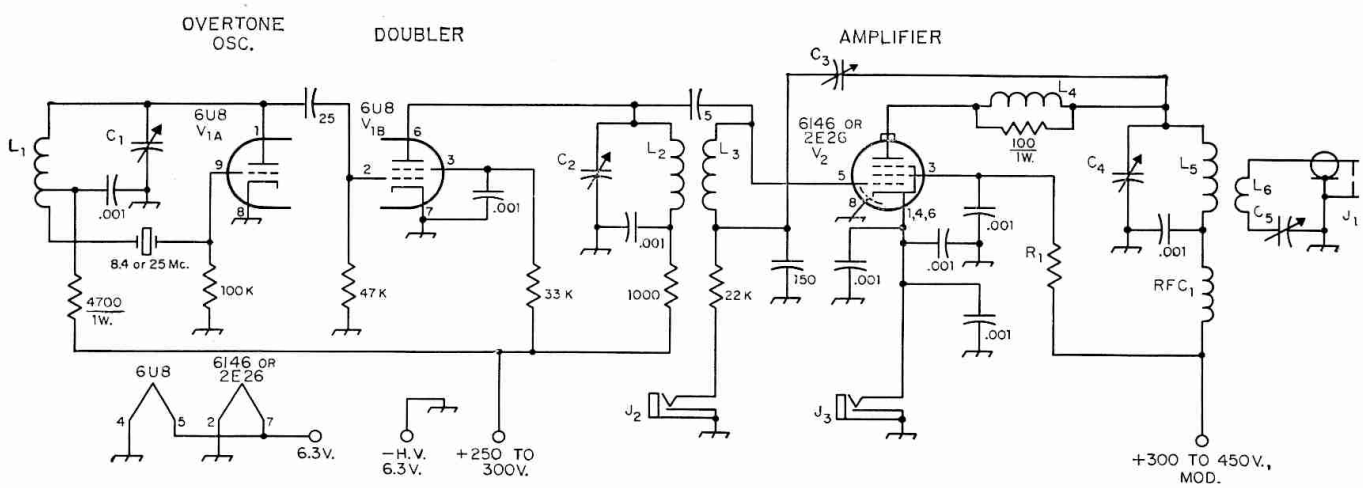


Fig. 17-20—Schematic diagram and parts information for the 50-Mc. transmitter. Capacitors are ceramic unless specified. Values under .001 are in $\mu\mu$f. Resistors ½ watt unless specified.

C_1—50-$\mu\mu$f. variable (Johnson 157-4).
C_2—25-$\mu\mu$f. variable (Johnson 157-3).
C_3—0.5 to 3 $\mu\mu$f. ceramic trimmer (Erie 3139D).
C_4—25-$\mu\mu$f. variable (Johnson 167-2).
J_1—Coaxial chassis fitting.
J_2, J_3—Closed-circuit jack.
L_1—14 t. No. 20 tinned, ½-inch diam., ⅞ inch long, tapped at 4½ t. from crystal end (B & W No. 3003).
L —6½ t., 7/16 inch long, similar to L_1.
L_3—7¼ t., ½-inch long, similar to L_2.
L_4—5 t. No. 20 wound on and spaced to fill 100-ohm 1-watt resistor.
L_5—3½ t. No. 14 tinned, ¾-inch i.d., ½-inch long.
L_6—2 t. No. 14, similar to and at cold end of L_5. Cover with spaghetti sleeving.
R_1—37,500 ohms, 4 watts (4 150,000-ohm 1-watt resistors in parallel).
RFC_1—Single-layer v.h.f. choke, 2 to 7 μh. (Ohmite Z-50 or National R-60).

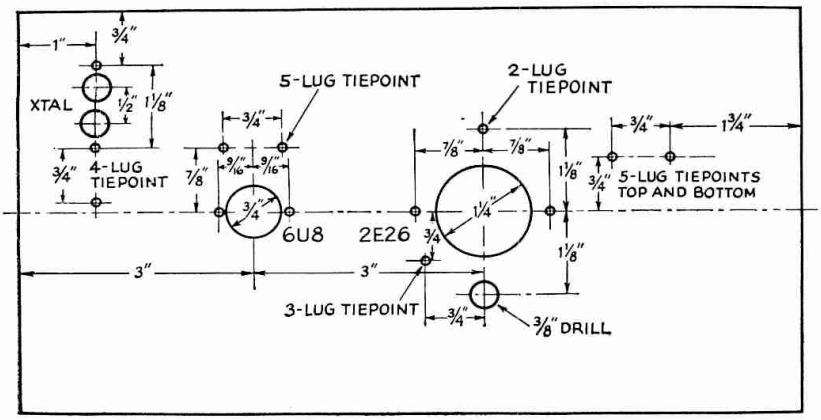


Fig. 17-21—Layout drawing of the 50-Mc. chassis top. Precise duplication is not important, though the general parts layout should be followed. Hole sizes may vary with different types of sockets.

spacing is changed.

Next comes neutralization of the amplifier. With drive on, but no plate or screen voltage, tune the amplifier plate circuit through its range, watching the grid current meter. There may be a downward dip in grid current when the plate circuit is resonated. Adjust the neutralizing capacitor, C_3, a turn or two and check the grid current dip again. If there is less change than before, the adjustment was in the right direction. Continue in this way until no downward movement can be seen in the grid current as the plate circuit is tuned through resonance.

If neutralization cannot be achieved, a different value of bypass will be required at the low end of L_3. If the neutralizing capacitor is at minimum setting when neutralization is approached, a larger value of bypass will be needed. Try 220 $\mu\mu$f. as a next step.

Power may now be applied to the final amplifier. This can be from the same source as has been used for the earlier tests, for the time being. The meter may be removed from the doubler power lead and connected between the junction of the r.f. choke, RFC_1, and screen resistor and the terminal on the back of the transmitter. This will measure the combined plate and screen current drawn by the amplifier. The meter may also be plugged into the cathode jack, where it will read combined plate, screen and grid current.

A light bulb of about 25 watts or more can be connected to a coaxial fitting and used as a dummy load in place of an antenna. This will not represent a 50-ohm load, so the tuning of the stage will not be the same as when a matched antenna system is used, but it will do for initial tests, and it will give a rough indication of power output.

Apply plate-screen power to all stages, and tune the plate circuit of the amplifier to the point where plate current dips the lowest. Now adjust the series capacitor, retuning the plate capacitor, until maximum brilliance is seen in the load lamp. Check carefully for any sign of oscillation in the amplifier. Remove the crystal from its socket briefly, while watching the amplifier grid current. This current and the amplifier output should drop to zero, and remain there regardless of the tuning of any of the transmitter circuits. Should grid *current appear* with the oscillator inoperative, recheck neutralization. The grid-current dip may be only an approximate indication of neutralization, so the adjustment may have to be touched up after power is applied to the amplifier. Turn off power as a safety measure when this is done. With perfect neutralization, maximum grid current, minimum plate current and maximum output will all occur at the same setting of the amplifier plate circuit tuning. Perfection in this respect may not be possible, but there should be no sign of oscillation (grid current in the amplifier when the drive is removed) at any setting of the tuning controls.

When the rig is operated with a properly designed antenna the settings of the amplifier plate and antenna loading adjustments may be somewhat different from those obtained with a lamp load. Both should be adjusted for maximum power delivered to the antenna. This can be recorded on a field-strength meter, giving a relative indication of the power radiated by the antenna. Better than this is a power-indicating standing-wave bridge, which may be left connected in the line to the antenna at all times.

Final operating conditions for the transmitter will depend on the supply voltage and final tube used. With a 300-volt supply the oscillator plate current will run about 10 ma. with the oscillator operating properly, and 17 ma. with the crystal out of oscillation. The doubler plate-screen current is about 12 ma. Amplifier grid current will be at least 3 ma. without plate and screen voltage, and around 2.5 ma. with the amplifier operating under load. These values will be slightly lower with a 250-volt supply. Plate-screen current to the amplifier will depend on the power level and tube. With a 2E26 at 300 volts the current will be about 20 ma. at resonance, with no load, and 95 ma. off resonance. Loaded for maximum efficiency the 2E26 plate and screen current will be about 60 ma. With a 6146 at 450 volts the loaded plate and screen current will be about 120 ma.

The 50-Mc. transmitter was described originally in *QST* for October, 1958.

The 144-Mc. Transmitter

Layout and testing of the 144-Mc. unit are very similar to the 50-Mc. model already described, so only the points of difference will be covered in this part of the text. Looking at the bottom view,

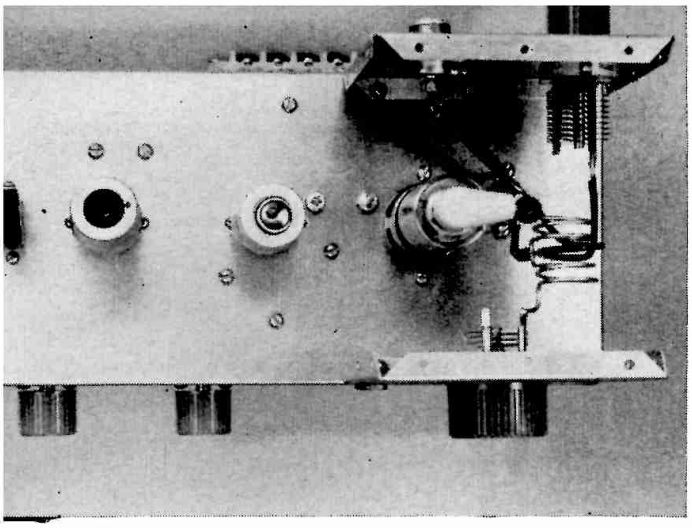


Fig. 17-22—Top view of the 144-Mc. transmitter with shield cover removed. A 2E26 is shown in the amplifier socket.

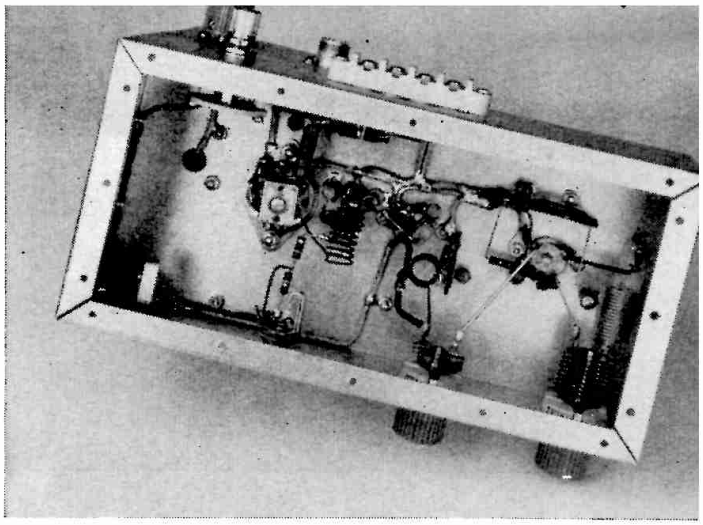


Fig. 17-23—Bottom of the 144-Mc. transmitter, with oscillator-tripler at the right. Doubler stage is near the middle of the chassis and amplifier at the left.

Fig. 17-23, the oscillator tuned circuit is at the far right. The tripler plate capacitor, C_2, is next on the front wall. The 6U8 socket is between these two capacitors, on the center line of the chassis. The 12AT7 parallel doubler socket is approximately in the middle of the chassis. The coil mounted vertically at the right and slightly below the 12AT7 socket is the tripler plate coil, L_2.

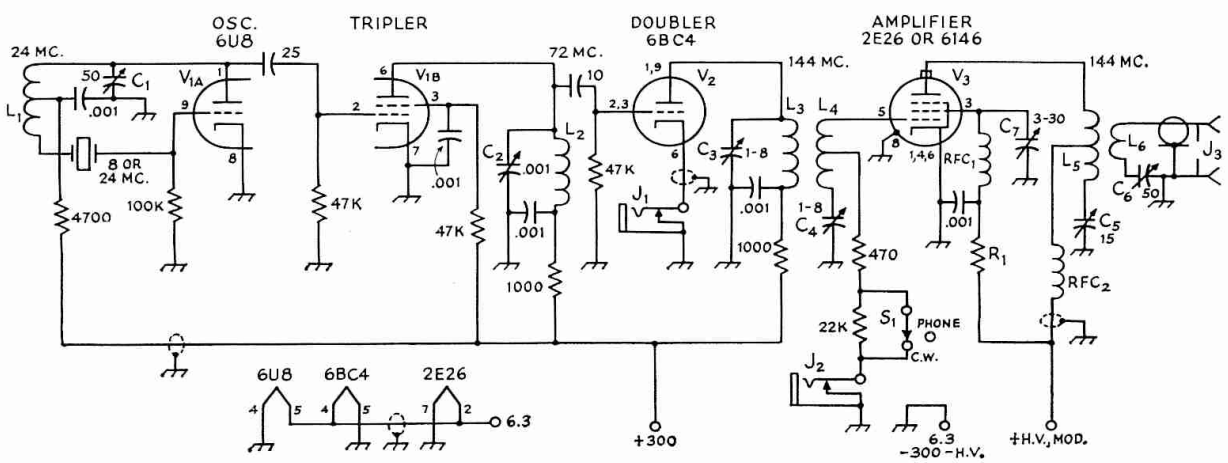
The doubler plate coil, L_3, and the amplifier grid coil, L_4, are mounted on a common center line and close together, making them appear as one coil in the photograph. The top end of L_3, as seen in the schematic diagram, Fig. 17-24, is toward the back of the chassis. The grid end of L_4 is toward the front. Capacitors C_3 and C_4 are cylindrical plastic trimmers. They are at either side of and just above the upper end of L_3.

The amplifier socket is at the left. The screen tuning capacitor, C_7, is mounted across the socket. Screen voltage is fed through the r.f. choke just above the socket. The switch for shorting out the grid leak when c.w. is used is in the upper left corner of the photograph. The two jacks on the front wall are for keying (far left) and grid current measurement.

Circuit differences between the two units, aside from the inclusion of the extra multiplier stage in the 144-Mc. model, arise mainly from the effects of tube and circuit capacitances at the higher frequency. Tube capacitances load the tuned circuits heavily, so series-tuned circuits are used in the amplifier stage. It will be seen that the keying jack is connected in the cathode of the doubler stage instead of in the amplifier cathode lead. It is difficult to bypass the amplifier cathode completely at 144 Mc., and the insertion of the keying jack in that position would cause oscilla-

Fig. 17-24—Schematic diagram and parts information for the 144-Mc. transmitter.

C_1, C_6—50-$\mu\mu$f. variable (Johnson 157-4).
C_2, C_5—15-$\mu\mu$f. variable (Johnson 157-2).
C_3, C_4—1–8-$\mu\mu$f. plastic trimmer (Erie 532-10).
J_1, J_2—Closed-circuit jack.
J_3—Coaxial chassis fitting.
L_1—14 turns No. 20 tinned, ½-inch diam., ⅞ inch long tapped at 4 turns from crystal end (B & W No. 3003).
L_2—5¾ turns No. 18 enam., 7/16-inch diam., ⅝ inch long.
L_3—2¾ turns No. 18 enam., 7/16-inch diam., ¼ inch long.
L_4—6 turns No. 18 enam., 7/16-inch diam., ⅝ inch long, center tapped.
L_5—4 turns No. 14 tinned, ¾-inch diam., turns spaced 2 diameters. Make extra space at center for L_6; see Fig. 17-22.
L_6—1 turn No. 14 enamel, ¾-inch diam. Cover with insulating sleeving and insert at center of L_5.
R_1—33,000 ohms, 3 watts (3 100,000-ohm 1-watt resistors in parallel).
RFC_1—7-μh. solenoid choke (Ohmite Z-50).
RFC_2—1.8-μh. solenoid choke (Ohmite Z-144).
S_1—S.p.s.t. switch, any type.



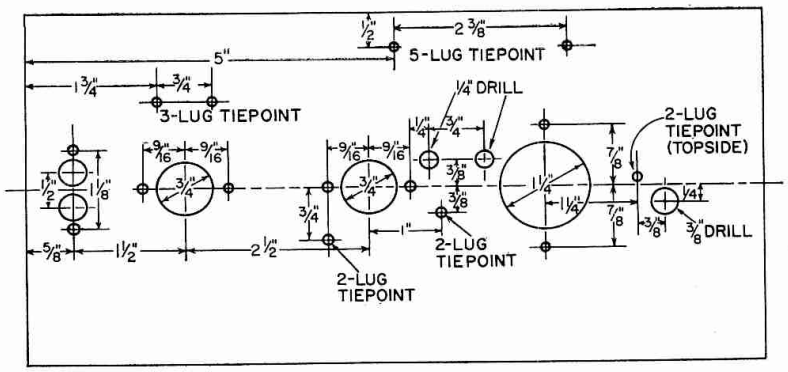


Fig. 17-25—Layout drawing of the 144-Mc. chassis.

tion. Screen bypassing is a similar problem, as conventional bypassing methods are ineffective at this and higher frequencies. Bringing the screen to ground potential requires a critical value of capacitance, so a trimmer (C_7) is connected from screen to ground.

Adjustment Procedure

The power supply for testing the 144-Mc. transmitter should deliver 6.3 volts at 1.6 amperes if a 2E26 amplifier tube is used. With a 6146 it should be capable of supplying 2 amperes. Initially 250 to 300 volts at 150 ma. will do for the plate supply. Final plate voltage for a 6146 may be as high as 500 volts.

Testing the first two stages is similar to that outlined for the 50-Mc. transmitter, except for the frequencies involved. Make sure that the oscillator is between 24 and 24.66 Mc., and that the pentode section of the 6U8 multiplies this frequency by 3. Tune the tripler plate circuit, L_2-C_2, for maximum output, as indicated by a 2-volt 60-ma. pilot lamp coupled to the cold end of L_2 with a single-turn loop of insulated wire about the diameter of the coil.

Next apply plate voltage to the 12AT7 doubler, and tune it for maximum amplifier grid current. Adjustment of C_3 and C_4 will interlock to some extent, but be sure that each is tuned for maximum grid current, as read in J_2. The switch S_1 can be in either position for this adjustment, though the grid current will be much higher if it is in the closed position.

Neutralization is done similarly to the manner outlined for the 50-Mc. transmitter, except that the setting of the screen capacitor, C_7, is the means by which it is achieved. If stability is approached as C_7 reaches maximum capacitance, a larger trimmer will be needed. Experimentation with the value of the r.f. choke in the screen lead may also be helpful. A variation of the neutralization system shown is the use of a critical value of inductance in the screen lead, and the elimination of C_7. Grid current, when neutralization is completed, should be at least 1.5 ma. with S_1 in the open position. Go over all adjustments carefully, and experiment with the spacing between L_3 and L_4 if the grid drive is low.

The balance of the testing is similar to the 50-Mc. procedure. Plate current for the 12AT7 doubler will be about 25 ma. Amplifier grid current should be all that can be obtained, but preferably not below 1.5 ma. under full load. Should it be less than 1.5 ma. running the amplifier at slightly less than full loading may make it possible to get satisfactory output and still retain good modulation characteristics.

Amplifier plate current at resonance with no load will be higher than on 50 Mc., and the output will be lower. Efficiency will be lower with a 6146 than with a 2E26, but the higher plate dissipation rating of the 6146 may make its use desirable if more output is needed than can be obtained with the 2E26. Either transmitter can be used in mobile service. For 6-volt cars the tubes can be as shown. Twelve-volt equivalents of all the tube types are now available for cars with 12-volt systems.

Modulation and Keying

For voice work a modulator is required. This should have a power output of approximately half the input to the final amplifier. Several suitable modulators are shown in other chapters of this *Handbook*. The plate and screen current of the amplifier are run through the secondary of the modulator output transformer. If the transmitter is to run at low power, a single 300-volt supply can be used for all stages, including the modulator, if it has a sufficiently high current rating.

Keying methods differ for the two r.f. units. The 50-Mc. transmitter is keyed for c.w. by breaking the cathode lead. This would cause instability if applied to the 144-Mc. transmitter, so the latter is keyed in the cathode of the doubler stage. Fixed bias must be applied to the final amplifier grid, to keep the plate current to a safe value. The voltage required will depend on the plate voltage applied to the final. The plate current need not be cut completely off, but merely held to less than the plate dissipation rating for the tube used. A 22½ volt battery is sufficient for plate voltages up to 400. The simplest way to apply bias, for occasional c.w. use, is to plug it into the grid current jack. The positive terminal of the bias battery should connect to the ground side of the plug.

The switch S_1 cuts out the 27,000-ohm grid resistor, so that the grid bias will not be excessive when fixed bias is applied. The rig can be operated in this manner (fixed bias plus the smaller of the two grid resistors) on voice, if it is desirable or convenient to do this. The grid current is so low that the bias battery will last almost indefinitely, and a small hearing aid size is suitable. It can be mounted inside the chassis and wired into the circuit permanently, where more frequent c.w. operation is expected.

Simple Transmitter for 220 and 420 Mc.

The transmitter in Figs. 17-26–17-29 is for the newcomer who wants to start with simple gear, going on to something better when he has gained construction and operating experience. It is built in two units, with the idea that the modulator can be retained when the r.f. portion is discarded.

The r.f. section is a simple oscillator with two 6AF4 or 6AT4 tubes in push-pull. Its plate circuit is changed from a quarter-wave line at 220 Mc. to a half-wave line at 420 Mc. by plugging in suitable terminations at the end of the tuned circuit.

Because the oscillator is modulated directly it will have considerable frequency modulation, and the signal will not be readable on selective receivers unless the modulation is kept at a very low level. Where a broader receiver is in use at the other end of the path a higher modulation level can be employed.

The modulator is designed for a crystal microphone. It delivers 3 to 10 watts output, depending on the plate voltage and whether a 6V6 or 6L6 tube is used. It may be considered as a long-term investment that will be suitable for use with any r.f. section of up to 20 watts input that may be constructed at a later date.

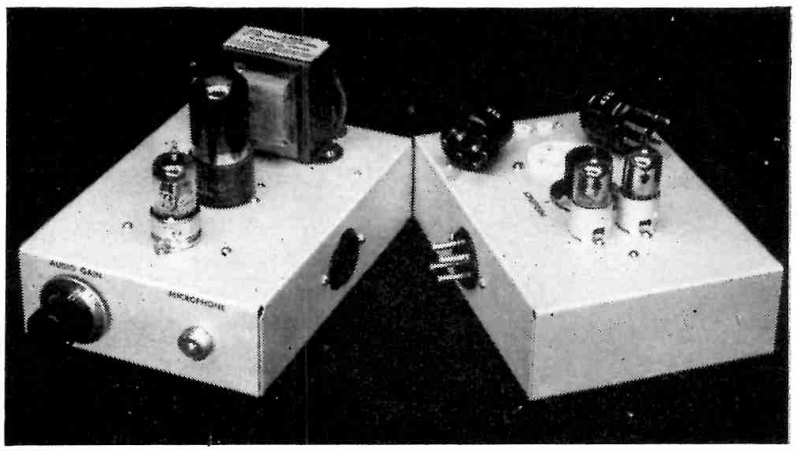


Fig. 17-26—The simple transmitter for 220 and 420 Mc. is made in two parts. The modulator, left, may be retained for use with more advanced r.f. sections than the simple oscillator shown at the right. The two units may be plugged together or connected by a cable.

Construction

The two units are built on identical 5 by 7 by 2-inch aluminum chassis, connecting by means of a plug on the oscillator and a socket on the modulator. Power is fed through a similar plug on the back of the modulator. Arrangement of parts in the modulator is not critical, but the oscillator should be exactly as shown.

Sockets for the tubes are one inch apart center to center, $2\frac{3}{16}$ inch in from the end of the chassis. C_1 is at the exact center of the chassis, with J_2 $1\frac{1}{2}$ inches to its left, as seen in Fig. 17-27. At the far left is a crystal socket, used for the antenna terminal, J_1. One-inch ceramic standoffs are mounted on the screws that hold J_2 in place. These support the antenna coupling loop, L_2.

Testing and Use

A power supply delivering about 200

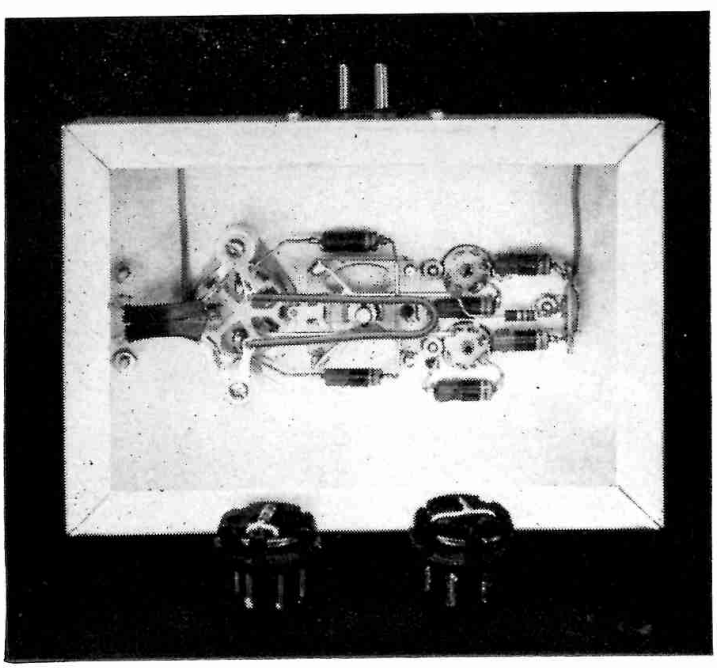


Fig. 17-27—Bottom view of the oscillator unit, showing the two-band tank circuit. The line terminations, with their protecting caps removed, are in the foreground. At the left is the 220-Mc. plug, with the 420-Mc. one at the right.

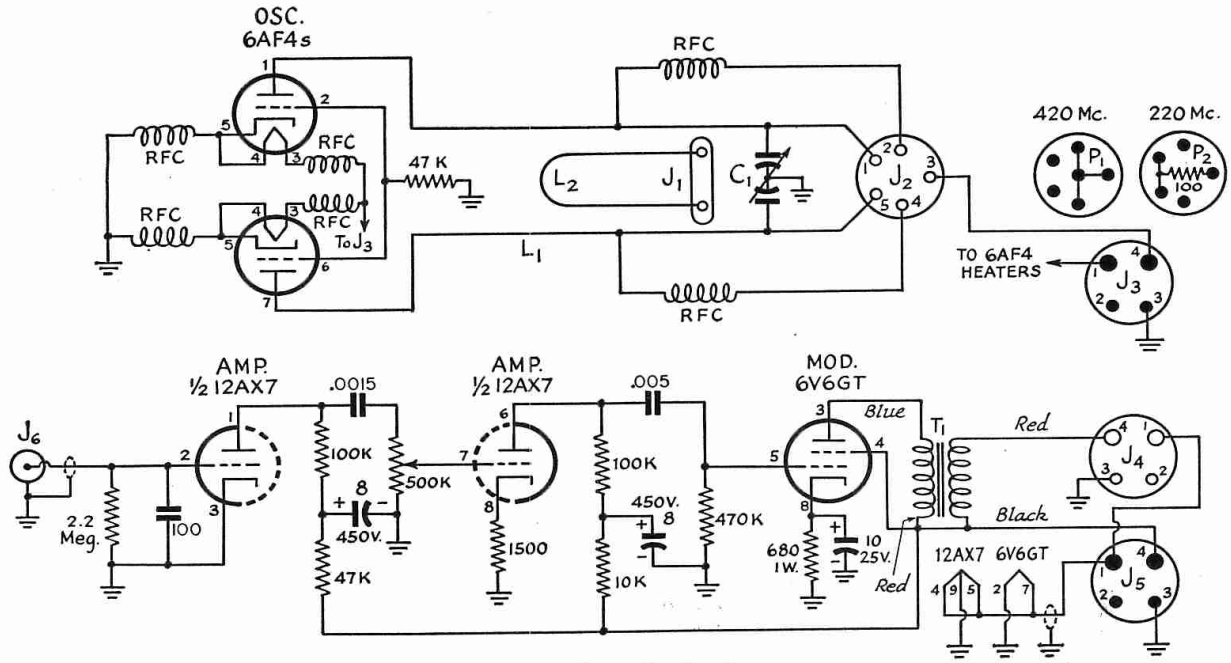


Fig. 17-28—Schematic diagram and parts information for the two-band oscillator and modulator.

C_1—10.5-$\mu\mu$f.-per-section butterfly variable (Johnson 10LB15).

L_1—2 3½ inch pieces No. 12 tinned, spaced ½ inch. Bend down ¾ inch at tube end and ½ inch at socket end. R.f. chokes connect ⅝ inch from bend at tube end. Connect C_1 at 1 inch from bend at socket end.

L_2—Hairpin loop 2¼ inches long and ½ inch wide, No. 16, covered with insulating sleeving.

J_1—Crystal socket used for antenna terminal.

J_2—5-contact ceramic socket (Amphenol 49–RSS5).

J_3, J_5—4-contact male fitting (Amphenol 86–RCP4).

J_4—4 contact female chassis fitting (Amphenol 78–S4 or RS4).

J_6—Microphone connector (Amphenol 75-PC1M).

P_1—5-contact male cable connector (Amphenol 86-PM5) with Pins 2, 3 and 4 joined together.

P_2—Same as P_1, but with Pins 1 and 5 joined. Connect 100-ohm resistor between these and Pin 3.

RFC (6 required)—12 turns No. 28 enamel close-wound on high-value 1-watt resistor.

T_1—10-watt modulation trans. (Merit A-3008).

volts d.c. at 50 ma. or more and 6.3 volts at 1 amp. or more is needed. Plug the units together or connect them by a cable. With a cable, a milliammeter may be connected between the No. 4 pins to measure the oscillator plate current. Otherwise the meter should be connected temporarily between Pin 4 of J_3 and Pin 3 of J_2, in place of the wire shown in Fig. 17-28.

Plate current should be about 25 to 30 ma. If the stage is oscillating there will be a fluctuation in current as the plate line is touched with an *insulated* metal object. Do not hold the metal in the hands for this test! The frequency is best checked by means of Lecher wires, a technique that is covered in the chapter on measurements.

With the dimensions given the range with P_1 plugged in should be about 405 to 450 Mc. With P_2 plugged in the frequency should fall within the 220-Mc. band with C_1 set in the same position as it was for the middle of the 420-Mc. band. Some alteration of the connection point for C on L_1 may be necessary to achieve this.

In using the transmitter it is well to stay between 221 and 224 Mc. to avoid out-of band operation. On 420, keep the transmitter below 432 Mc. to avoid interference with the high-selectivity work that is done between 432 and 436 Mc. (Further details on this transmitter in *QST* for December, 1954.)

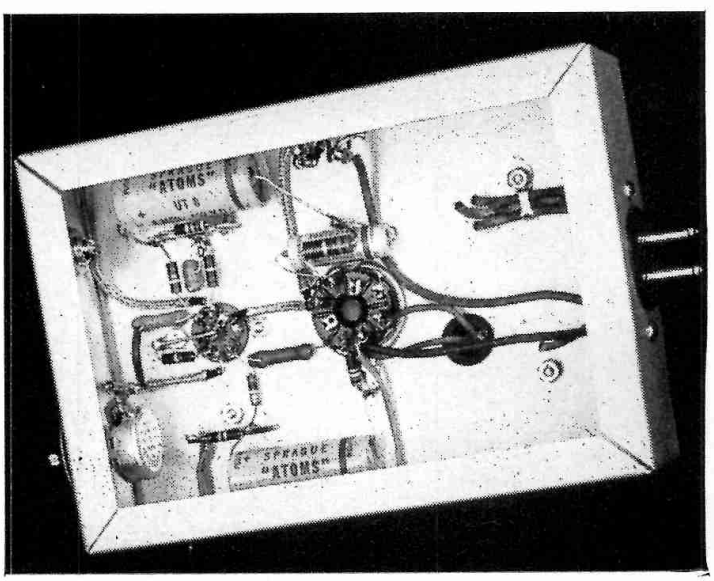


Fig. 17-29—Looking at the underside of the modulator.

A 40-Watt Transmitter for 220 Mc.

The crystal-controlled transmitter shown in Figs. 17-30 and 17-32 will run 30 to 40 watts at 220 Mc. Referring to Fig. 17-31, a simple overtone oscillator circuit uses one half of a 12AT7 dual triode. The crystal may be between 8.15 and 8.33 Mc. or 24.45 and 25 Mc. In either case, the frequency of oscillation is in the latter range, as the crystal works on its third overtone. The second half of the 12AT7 is a tripler to 73 to 75 Mc. This stage has a balanced plate circuit, so that its output may be capacitively coupled to the grids of a second 12AT7, working as a push-pull tripler to 220 Mc. The low side of the first tripler plate circuit has a balancing capacitor, C_3, so that a capacitance equal to the output capacitance of the 12AT7 can be added to that side of the circuit. Without this the two halves of the push-pull tripler may receive unequal drive, and one half of the tube will run hotter than the other.

The plate circuit of the push-pull tripler is inductively coupled to the grid circuit of an Amperex 6360 dual tetrode amplifier that runs straight through on 220 Mc. Similar inductive coupling transfers the drive to the grid circuit of the final amplifier stage, an Amperex 6252 dual tetrode. This tube is a somewhat more efficient outgrowth of the 832A, which may also be used, though with lower efficiency and output. Base connections are the same for both tubes.

The grid return of the 6252 is brought out to the terminal strip on the back of the unit, to allow for connection of a grid meter. Both this point and the tip jack in the 6360 grid return have 1000-ohm resistors completing the grid returns to ground, so that operation of the stages is unaffected if the meters are removed.

Instability in tetrode amplifiers for v.h.f. service may develop as a result of the ineffective bypassing of the screen. In the case of the 6360 stage stable operation was obtained with no bypassing at all, while on the 6252 a small mica trimmer was connected directly from the screen terminal to ground. It is operated near the minimum setting.

Construction

The transmitter is built on an aluminum plate 6 by 17 inches in size. This screws to a standard chassis of the same dimensions, which serves as both shield and case. Cut-outs about three inches square are made in the chassis and base plate, above and below the tube, to allow for ventilation. These openings are fitted with perforated aluminum or screening to preserve shielding. The case should be equipped with rubber feet, to avoid marring the surface it rests on, and to allow air circulation around the tube.

The tube sockets and all the controls except the tuning capacitor of the oscillator are mounted along the center line of the cover plate. The 220-Mc. stages are inductively coupled, using hairpin loop tank circuits the dimensions of which are given in Fig. 17-33. The tuning range of these circuits is affected by the widths of the loops as well as their length, so some variation can be had by squeezing the sides together or spreading them apart.

It is important that the method of mounting the 6252 socket be followed closely. An aluminum bracket about 2⅞ inches high and 4 inches wide supports the socket. Note that the socket and tube are on the *same* side of the plate. Holes are drilled in the plate in line with the control grid terminals to pass the grid leads. These holes are ⅜-inch diameter, and are equipped with rubber grommets to prevent accidental shorting of the grid leads to ground. The shape of the grid inductance should be such that its leads pass through the centers of the holes. The socket is supported on 5/16-inch metal pillars. It may be necessary to bend the socket lugs slightly to keep them from shorting to the mounting plate. The heater lead comes to the top of the plate, and the cathode lead bends around the bottom of it.

Power leads are made with shielded wire, and are brought out to a terminal strip on the back of the chassis. These leads and the coax to the output connector should be long enough so that the plate on which the transmitter is built can be lifted off the chassis and inverted as shown in the photograph.

Adjustment

Initial tests should be made with a power supply that delivers no more than 250 volts, and as little as 150 to 200 volts can be used. If the voltage is more than 250, insert a 5000-ohm 10-watt resistor in series with the power lead

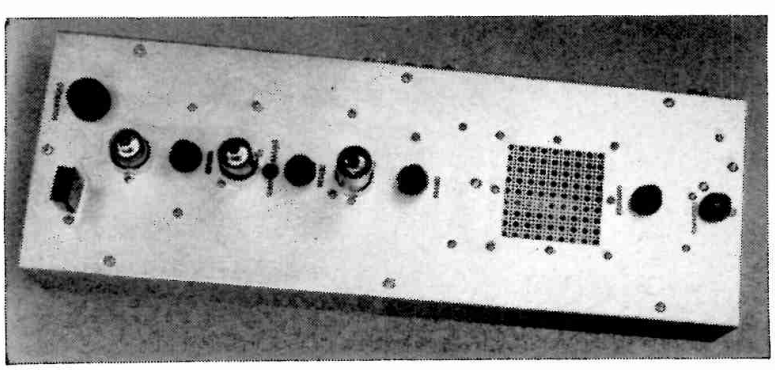


Fig. 17-30—Top view of the 220-Mc. transmitter. Final amplifier tube is inside the chassis, below the screened ventilation hole. Power connections, keying jack and output terminal are on the back of the chassis.

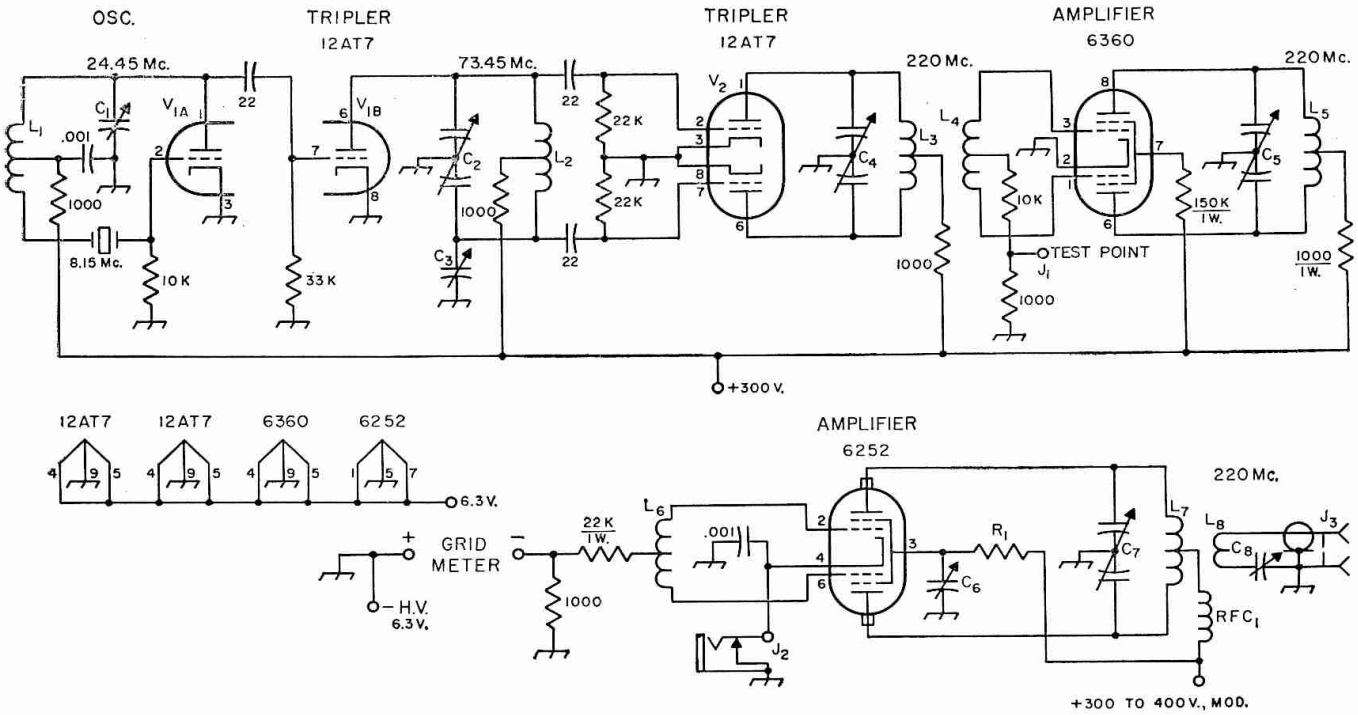


Fig. 17-31—Schematic diagram and parts information for the 220-Mc. transmitter. Capacitor values below 0.001 μf. are in μμf. Resistors ½ watt unless specified.

C_1—50-μμf. miniature variable (Hammarlund MAPC-50-B).
C_2, C_4, C_5—8-μμf. miniature butterfly variable (Johnson 160–208).
C_3, C_6—3–30-μμf. mica trimmer.
C_7—Butterfly variable, 1 stator and 1 rotor (Johnson 167–21, with plates removed).
C_8—15-μμf. miniature variable (Hammarlund MAPC-15-B).
J_1—Tip jack, insulated.
J_2—Closed-circuit phone jack.
J_3—Coaxial chassis fitting, SO-239.
L_1—15 t. No. 20 tinned, ½-inch diam., 1 inch long (B & W Miniductor No. 3003). Tap at 4 turns from crystal end; see text.
L_2—12 t. No. 18 tinned, ½-inch diam., 1 inch long, center-tapped.
L_3, L_4, L_5, L_6—U-shaped loops No. 18 enam., center-tapped. Dimensions given in Fig. 17-33.
L_7—2 t. No. 14 enam., 1-inch, 1-inch diam., leads ⅝ inch long. Center-tapped, space turns ½ inch apart.
L_8—1 t. No. 18 enam., inserted between turns of L_7. Cover with insulating sleeving.
R_1—23,500 ohms, 2 watts. (Two 47,000-ohm 1-watt resistors in parallel.)
RFC_1—25 t. No. 28 enam. on 1-watt high-value resistor.

temporarily. Plate voltage should be applied to the various stages separately, starting with the oscillator, makings ure that each stage is working correctly before proceeding to the next.

A milliammeter of 50- to 100-ma. range should be connected temporarily in series with the 1000-ohm resistor in the oscillator plate lead. When power is applied the current should be not more than about 10 ma. Rotate C_1 and note if an upward kick occurs, probably near the middle of the range of C_1. At this point the stage is oscillating. Lack of oscillation indicates too low feedback, or a defective crystal. Listen for the note on a communications receiver tuned near 24 Mc., if one is available. There should be no more than a slight change in frequency when a metallic tool is held near the tuned circuit, or when the circuit is tuned through its range. The note should be of pure crystal quality. If there is a rough sound, or if the frequency changes with mechanical vibration, the oscillator is not controlled by the crystal. This indicates too much feedback, and the tap on the coil, L_1, should be moved near the crystal end.

The proper amount of feedback is the lowest tap position that allows the oscillator to start readily under load. If 24-Mc. crystals are used the tap can be lower on the coil than with 8-Mc. crystals. When 8-Mc. crystals are operated on the third overtone, as in this case, the frequency of oscillation may not be exactly three times that marked on the crystal holder.

Now apply plate voltage to the second half of the 12AT7, again using a temporary plate meter connected in series with the 100-ohm decoupling resistor that feeds plate power to L_2. Current will be about 10 ma., as with the oscillator. Tune C_2 for maximum output. This can be determined by brilliance indication in a 2-volt 60-ma. pilot lamp connected to a 1-turn loop of insulated wire coupled to L_2. Check the frequency of this stage with a wavemeter.

Now connect a low-range milliammeter (not more than 10 ma.) between the test point, J_1, and ground. Apply power to the push-pull tripler, again using a temporary milliammeter connected in the lead to the plate coil, L_3. Tune the plate circuit for maximum indication on the grid meter. Plate current will be about 20 ma. Adjust the position of L_3 with respect to L_4 for maximum grid current. Now go back over all previous adjustments and set them carefully for maximum

220-Mc. Transmitter

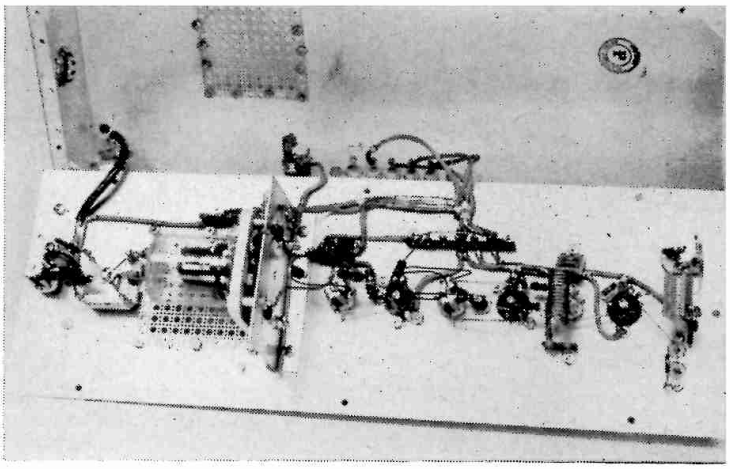


Fig. 17-32—Interior view of the 220-Mc. transmitter. All r.f. components are mounted on an aluminum plate, which is screwed to the top of a standard 6 × 17-inch chassis. Screen trimmer capacitor C_6 mounts on the tube socket mounting plate.

The crystal socket and the oscillator coil and capacitor are at the far right. Next is the first 12AT7 socket. Next to the left is the first tripler plate coil, mounted over its trimmer, with the mica balancing padder, C_3, above. The 12AT7 tripler, the test point, J_1, the tuning capacitor C_4, the tripler plate and amplifier grid loops, L_3 and L_4, the 6360 socket, the 6360 plate and amplifier grid loops, the 6252, and its tuned circuits follow in that order. The series capacitor, C_8, and the coaxial lead to the output connector, J_3, are at the far left.

grid current. Adjust the balancing padder, C_3, retuning C_2 each time this is done, until the combination of C_2 and C_3 that gives the highest grid current is found. Check the frequency to be sure that the stage is tripling to 220 Mc.

Now apply power to the 6360 plate circuit, again using the temporary meter to check the current. Connect the low-range milliammeter between the grid-metering terminal on the connector strip and ground. Set the screen trimmer, C_6, near minimum, and tune the 6360 plate circuit for maximum grid current. With 300 volts on the preceding stages, it should be possible to get at least 4 ma. Adjust the spacing between L_5 and L_6 carefully for maximum grid current, retuning C_5 each time this is done. Plate current should not exceed 55 ma.

Check for neutralization of the final amplifier by tuning C_7 through resonance while watching the grid-current meter. If there is no change, or only a slight rise as the circuit goes through resonance, the stage is near enough to neutralization to apply plate power. The 6252 has built-in cross-over capacitance, intended to provide neutralization in the v.h.f. range, so it is likely to be stable at this frequency. If there is a downward kick in the grid current at resonance, adjust the screen trimmer until it disappears. If best neutralization shows at minimum setting of the screen trimmer it may be desirable to eliminate the trimmer.

With an antenna or dummy load connected at J_3, final plate voltage can be applied. Tune the final plate circuit for maximum output, with a meter of 100 ma. or higher range connected to read the combined plate and screen current. This meter may be connected in the power lead, or it can be plugged into the cathode jack. In the latter position it will read the combined plate, screen and grid currents. Tune for maximum output and note the plate current. If it is much over 100 ma., loosen the coupling between L_7 and L_8. The input should not be over 50 watts at this frequency.

A final check for neutralization should now be made. Pull out the crystal or otherwise disable the early stages of the transmitter. The grid current and output should drop to zero. If they do not, adjust the screen trimmer until they do. Make this test only very briefly, as the tubes will draw excessive current when drive is removed. When perfect neutralization is achieved, maximum output will be found at a setting of C_7 at which plate current is at a minimum and grid current at maximum.

Operation

All stages should be run as lightly as possible, for stable operation and long tube life. No more than 300 volts should be run on the exciter stages, and if sufficient grid drive can be obtained, lower voltage is desirable. The 6360 stage runs with rather low drive, and its efficiency is consequently poor, but it delivers enough power to drive the 6252, even when run at as low as 250 volts, if all stages are operating as they should.

Observe the plates of the tubes when the transmitter is operated in a darkened room. There should be no reddening of the plates. If one side of any of the last three stages shows red and the other does not it is evidence of unbalance. This can usually be corrected by adjustment of the balancing trimmer, C_3, in the first tripler plate circuit. Lack of symmetry in lead lengths or unbalanced capacitance to ground in any of the r.f. circuits may also lead to lopsided operation.

Though the 6252 is rated for up to 600 volts on the plates, it is recommended that no more than 400 be used in this application, particularly if the stage is to be modulated for voice work.

For voice work the plate-screen current of the 6252 is run through the secondary of the output transformer on the modulator. The latter should have an output of 20 watts or so.

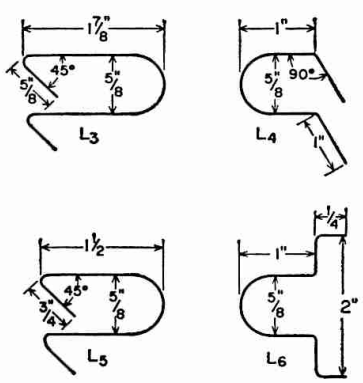


Fig. 17-33—Details of the hairpin loops used in the 220-Mc. transmitter.

A Tripler-Amplifier for 432 Mc.

Only tubes designed especially for u.h.f. service will work satisfactorily at 420 Mc. and higher. The various small receiving triodes made for u.h.f. TV use will work well in low-powered frequency multipliers and r.f. amplifiers for transmitting, but the trend is to tetrodes. Several of the latter are now available.

The tripler-amplifier shown in Figs. 17-34 to 17-37 delivers up to 20 watts output on 432 Mc.

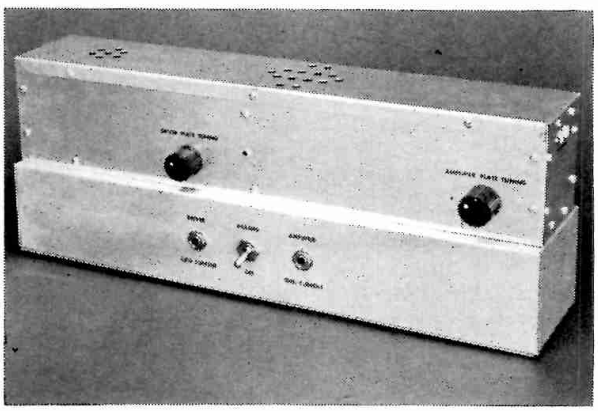


Fig. 17-34—A tripler-amplifier for 432 Mc. using dual tetrodes. Shielded construction and forced-air cooling are employed.

when driven on 144 Mc. by any 2-meter unit delivering 10 watts output or more. In plate-modulated service the output is 12 watts. Tubes are RCA 6524 dual tetrodes, but with slight modification Amperex 6252s or 5894s may be used. With 6252s the output will be about the same as with the 6524. The 5894 will deliver up to 40 watts with higher plate voltages. The 832A may also be used, but the output will be no more than 4 or 5 watts. Forced-air cooling and shielding are recommended.

The tripler tube is mounted vertically, at the left, with its socket 1½ inches below the chassis. There is just room under the socket for the self-resonant input circuit, L_2. The amplifier is horizontal, with its socket mounted in back of a plate that is 8 inches from the left edge of the 3 × 4 × 17-inch aluminum chassis. The shielding enclosure is 3¼ inches wide by 3½ inches high. A cooling fan is mounted on the rear wall of the chassis. Air circulates around the tripler tube through its 2-inch hole, flowing out through holes in the top cover. Holes are drilled in the chassis under the amplifier tube, and in the cover over it. With a bottom plate fitted to the chassis there should be enough air flowing through both top vents to lift a paper briskly when the fan is started.

Half-wave lines are used in all 432-Mc. circuits. The grid circuit of the amplifier is capacitively coupled to the tripler plate line, the two overlapping about 1¼ inches. The spacing between them must be adjusted carefully for maximum grid drive. Plate voltage is fed to the lines through small resistors. These should be connected at the point of lowest r.f. voltage on the lines. The amplifier grid r.f. chokes are connected at the tube socket.

Note that the plate line capacitors, C_1 and C_2, have their rotors floating. This is important. Grounding the rotors, or use of capacitors having metal end plates, may introduce multiple r.f. paths and circuit unbalance. The capacitors have small metal mounting brackets that are not connected directly to the rotors, but even so it was necessary to resort to polystyrene mounting plates for best circuit balance and efficiency. Holes ¾ inch in diameter are punched in the front wall to pass the rotor shafts.

Testing

The tripler-amplifier is designed to operate in conjunction with a 144-Mc. transmitter such as

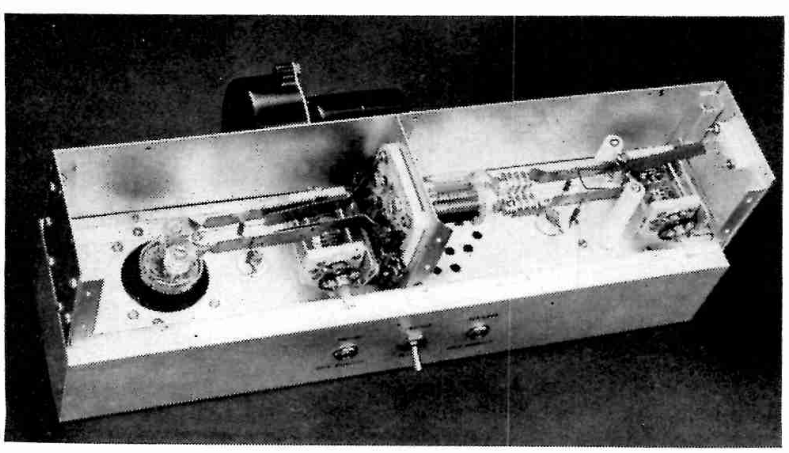


Fig. 17-35—Looking into the tripler-amplifier with the top cover and front plate removed.

A Tripler-Amplifier

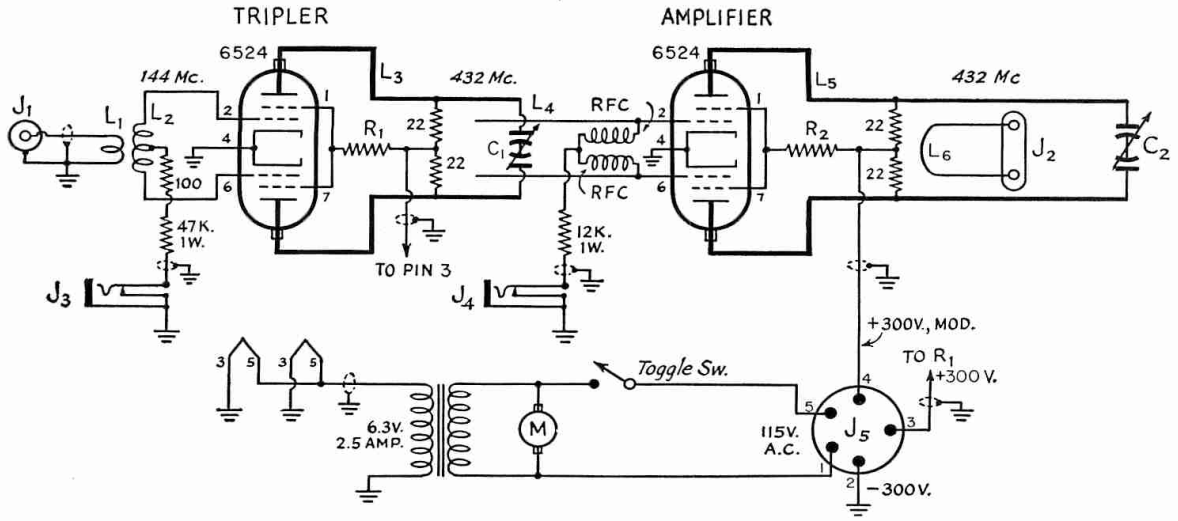


Fig. 17-36—Schematic diagram for the 432-Mc. tripler-amplifier.

C_1, C_2—10-$\mu\mu$f.-per-section split stator, double spaced (Bud LC-1664). Do not use metal end-plate or grounded-rotor types.

R_1, R_2—23,500 ohms, 2 watts (two 47,000 ohm 1-watt resistors in parallel).

L_1—2 turns No. 20 enam., ½-inch diam. Insert between turns of L_2.

L_2—4 turns No. 16 enam., ½-inch diam., ½-inch long, center-tapped.

L_3—Copper strap on heat-dissipating connectors, 3½ inches long. Twist 90 degrees ½ inch from plate end. Space ¾ inch.

L_4—Copper strap 2⅞ inches long, soldered to grid terminals. Space about ½ inch.

L_5—Copper strap 3⅞ inches long, fastened to heat-dissipating connectors. Space ¾ inch. All tank circuits of flashing copper ½ inch wide.

L_6—Coupling loop, No. 20 enam. U-shaped portion is 1 inch long and ⅝ inch wide. Mount on 3-inch ceramic stand-offs.

J_1—Coaxial input fitting (Amphenol 83-1R).

J_2—Crystal socket used for antenna terminal.

J_3, J_4—Closed-circuit jack.

J_5—5-pin male chassis connector (Amphenol 86-RCP5).

M—Motor-blower assembly, 17 c.f.m. (Ripley Inc., Middletown, Conn., Type 8433).

the 2E26 rig shown in Fig. 17-17. A plate supply of 300 volts at 200 ma. is needed (400 volts may be used with 5894s). Apply power to the 144-Mc. driver stage and adjust the spacing of the turns in L_2 and the degree of coupling between L_1 and L_2 for maximum tripler grid current. This should be about 3 ma.

Next apply plate and screen voltage to the tripler and tune C_1 for maximum grid current in the amplifier, with no plate or screen voltage to the latter. Adjust the position of the grid lines with respect to the plate circuit, readjusting C_1 whenever a change is made, until at least 4 ma. grid current is obtained.

Now connect a lamp load across the output terminal, J_2. Ordinary house lamps are not suitable. A fair load can be made by connecting 6 or more blue-bead pilot lamps in parallel. This can be done by wrapping a ¼-inch copper strap around the brass bases and soldering them all together. Then another strap should be soldered to the lead terminals. Apply plate and screen voltage and tune C_2 for maximum lamp brilliance. It should be possible to develop a very bright glow in the 6-lamp load with a plate current of about 100 ma. at 300 volts.

Cut drive very briefly to check for oscillation in the final stage. Grid current should drop to zero. The screen and grid resistors shown are for operation with plate modulation. More input can be run if the screen or grid resistance is decreased, but this should be done only when the rig is to be used for f.m. or c.w. service.

Operating conditions are about as follows: tripler grid current — 2 to 3 ma.; amplifier grid current — 3 to 4 ma.; tripler plate and screen current — 90 ma.; amplifier plate and screen current — 110 ma.; output — 12 watts.

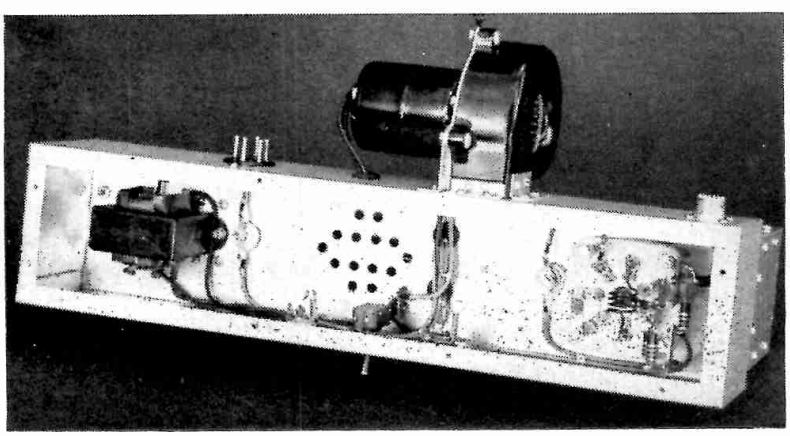


Fig. 17-37—Bottom view of the 432-Mc. transmitter.

V.H.F. Antennas

While the basic principles of antenna design remain the same at all frequencies where conventional elements and transmission lines are used, certain aspects of v.h.f. work call for changes in antenna techniques above 50 Mc. Here the physical size of arrays is reduced to the point where some form of antenna having gain over a simple halfwave dipole can be used in almost any location, and the rotatable high-gain directional array has become a standard feature of all well-equipped v.h.f. stations. The importance of antenna gain in v.h.f. work cannot be over-emphasized. 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 CONSIDERATIONS

At 50 Mc. and higher it is usually important to have the antenna work well over all or most of the band in question, and as the bands are wider than at lower frequencies the attention of the designer must be focused on broad frequency response. This may be attained in some instances through sacrificing other qualities such as high front-to-back ratio.

The loss in a given length of transmission line rises with frequency. V.h.f. feedlines should be kept as short as possible, therefor. Matching of the impedances of the antenna and transmission line should be done with care, and in open locations a high-gain antenna at relatively low height may be preferable to a low-gain system at great height. Wherever possible, however, the v.h.f. array should be well above heavy foliage, buildings, power lines or other obstructions.

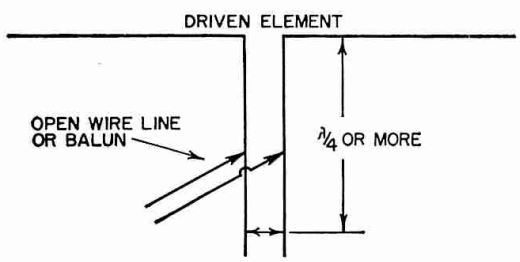


Fig. 18-1—Combination tuning and matching stub for v.h.f. arrays. Sliding short is used to tune out reactance of the driven element or phasing system. Transmission line, either balanced or coax, is connected at the point of lowest standing-wave ratio. Adjustment procedure is outlined in text.

The physical size of a v.h.f. array is usually more important than the number of elements. A 4-element array for 432 Mc. may have as much gain over a dipole as a similarly designed array for 144 Mc., but it will intercept only one-third as much energy in receiving. Thus to be equal in communication, the 432-Mc. array must equal the 144-Mc. antenna in *capture area*, requiring three times as many elements, if similar element configurations are used in both.

Polarization

Early v.h.f. work was done with simple antennas, and since the vertical dipole gave as good results in all directions as its horizontal counterpart offered in only two directions, vertical polarization became the accepted standard. Later when high-gain antennas came into use it was only natural that these, too, were put up vertical in areas where v.h.f. activity was already well established.

When the discovery of various forms of long-distance propagation stirred interest in v.h.f. operation in areas where there was no previous experience, many newcomers started in with horizontal arrays, these having been more or less standard practice on frequencies with which these operators were familiar. As use of the same polarization at both ends of the path is necessary for best results, this lack of standardization resulted in a conflict that, even now, has not been completely resolved.

Tests have shown no large difference in results over long paths though evidence points to a slight superiority for horizontal in certain kinds of terrain, but vertical has other factors in its favor. Horizontal arrays are generally easier to build and rotate. Where ignition noise and other forms of man-made interference are present, horizontal systems usually provide better signal-to-noise ratio. Simple 3- or 4-element arrays are more effective horizontal than vertical, as their radiation patterns are broad in the plane of the elements and sharp in a plane perpendicular to them.

Vertical systems can provide uniform coverage in all directions, a feature that is possible only with fairly complex horizontal arrays. Gain can be built up without introducing directivity, an important feature in net operation, or in locations where the installation of rotatable systems is not possible. Mobile operation is simpler with vertical antennas. Fear of increased TVI has kept v.h.f. men in some densely populated areas from adopting horizontal as a standard.

The factors favoring horizontal have been predominant on 50 Mc., and today we find it the standard for that band, except for emergency net operation involving mobile units. The slight advantage it offers in DX work has accelerated the trend to horizontal on 144 Mc. and higher bands, though vertical polarization is still widely used. The picture on 144, 220 and 420 Mc. is still confused, the tendency being to follow the local

trend. The newcomer should check with local amateurs to see which polarization is in general use in the area he expects to cover. Eventual standardization should be a major objective, and to this end it is recommended that horizontal polarization be established in areas where activity is developing for the first time.

IMPEDANCE MATCHING

Because line losses increase with frequency it is important that v.h.f. antenna systems be matched to their transmission lines carefully. Lines commonly used in v.h.f. work include open-wire, usually 300 to 500 ohms impedance, spaced ½ to two inches; polyethylene-insulated flexible lines, available in 300, 150 and 72 ohms impedance; and coaxial lines of 50 to 90 ohms impedance.

The various methods of matching antenna and line impedance are described in detail in Chapter 14. Matching devices commonly used in v.h.f. arrays fed with balanced lines include the folded dipole in its various forms, Fig. 14-42, the "T" Match, Fig. 14-45, the "Q" section, Fig. 14-41, and the adjustable stub, Fig. 18-1. The gamma match, useful for feeding the driven element of a parasitic array with coaxial line, is shown in schematic form in Fig. 14-45. Balanced loads such as a split dipole or a folded dipole can be fed with coax through a balun, as shown in Fig. 14-46. Practical examples of the use of these devices are shown in the following pages. The principles upon which their operation depends are explained in Chapter 14, with the exception of the adjustable stub of Fig. 18-1.

The Corrective Stub

The adjustable stub shown in Fig. 18-1 provides a means of matching the antenna to the transmission line and also tuning out reactance in the driven element. It is, in effect, a tuning device to which the transmission line may be connected at the point where impedances match. Both the shorting stub and the point of connection are made adjustable, though once the proper points are found the connections may be made permanent.

For antenna experiments the stub may be made of tubing, and the connections made with sliding clips. In a permanent installation a stub of open-wire line, with all connections soldered, may be more satisfactory mechanically. The transmission line may be open-wire or Twin-Lead, connected directly to the stub, or coaxial line of any impedance, which should be connected through a balun.

To adjust the stub start with the short at a point about a half wavelength below the antenna, moving the point of connection of the transmission line up and down the stub until the lowest standing-wave ratio is achieved. Then move the shorting stub a small amount and readjust the line connection for lowest s.w.r. again. If the minimum s.w.r. is lower than at the first point checked the short was moved in the right direction. Continue in that direction, readjusting the line connection each time, until the s.w.r. is as close to 1:1 as possible. When adjustments are completed the portion of the stub below the short can be cut off, if this is desirable mechanically.

TYPES OF V.H.F. ARRAYS

Directional antenna systems commonly used in amateur v.h.f. work are of three general types, the collinear, the Yagi, and the plane reflector

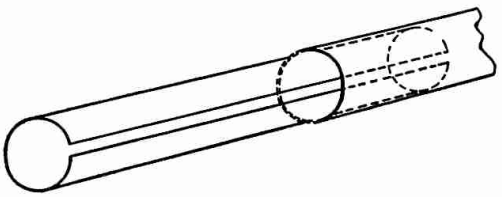


Fig. 18-2—Inserts for the ends of the elements in a v.h.f. array provide a means of adjustment of length for optimum performance. Short pieces of the element material are sawed lengthwise and compressed to fit inside the element ends.

array. Collinear systems have two or more driven elements end to end, fed in phase, usually backed up by parasitic reflectors. The Yagi has a single driven element, with one or more parasitic elements in front and in back of the driven element, all in the same plane. The plane-reflector array has a large reflecting surface in back of its driven element or elements. This may be a sheet of metal, a metal screen, or closely spaced rods or wires. The reflector may be a flat plane, or it can be bent into several forms, such as the corner and the parabola.

Examples of all three types are described, and each has points in its favor. The collinear systems such as the 12- and 16-element arrays of Figs. 18-14 and 18-15 require little or no adjustment and they present few feed problems. They work well over a wide band of frequencies. Yagi, or parasitic arrays, Figs. 18-5 to 18-10, depend on fairly precise tuning of their elements for gain, and thus work over a narrower frequency range. They are simple mechanically, however, and usually offer more gain for a given number of elements than do the collinear systems. Plane- and corner-reflector arrays are broadband devices, having broad forward lobes and high front-to-back ratio. They are easily adjusted, but somewhat cumbersome mechanically.

ELEMENT LENGTHS AND SPACINGS

Designing a v.h.f. array presents both mechanical and electrical problems. The electrical problems are basic, and their solution involves choosing the type of performance most desired. Mechanical design, on the other hand, can be subject to almost endless variations, and the form that the array will take can usually be decided by the materials and tools available. One common

TABLE 18-I

Dimensions for V.H.F. Arrays in Inches

Freq. (Mc.)	52*	146*	222.5*	435*
Driven Element	106.5	38	24⅞	12¾
Change per Mc.*	2	0.25	0.12	0.03
Reflector	111½	40	26⅛	13⅜
1st Director	101½	36	23⅝	12⅛
2nd Director	99½	35¾	23⅜	12
3rd Director	97½	35	23	11⅞
1.0 Wavelength	234	81	53	27
0.625 Wavelength	147	50½	33⅛	16¾
0.5 Wavelength	117	40½	26½	13.5
0.25 Wavelength	58½	20¼	13¼	6¾
0.2 Wavelength	47	16	10⅝	5⅜
0.15 Wavelength	35	12	8	4
Balun loop (coax)	76	26.5	17¼	8¾

* Dimensions given for element lengths are for the middle of each band. For other frequencies adjust lengths as shown in the third line of table. Example: A dipole for 50.0 Mc. would be 106.5 + 4 = 110.5 inches.

Apply change figure to parasitic elements as well.

For phasing lines or matching sections, and for spacing between elements, the midband figures are sufficiently accurate. They apply only to open-wire lines.

Parasitic-element lengths are optimum for 0.2 wavelength spacing.

source of materials for amateur arrays is commercially built TV antennas. They can often be revamped for the amateur v.h.f. bands with a minimum of effort and expense.

Dimensions for Yagi or collinear arrays and their matching devices can be taken from Table 18-I. The driven element is usually cut to the formula:

$$Length \text{ (in inches)} = \frac{5540}{Freq. \text{ (Mc.)}}.$$

This is the basis of the lengths in Table 18-I, which are suitable for the tubing or rod sizes commonly used. Arrays for 50 Mc. usually have ½ to 1-inch elements. For 144 Mc. ¼ to ½-inch stock is common. Rod or tubing ⅛ to ⅜ inch in diameter is suitable for 220 and 420 Mc. Note that the element lengths in the table are for the middle of the band concerned. For peaked performance at other frequencies the element lengths should be altered according to the figures in the third line of the table.

Reflector elements are usually about 5 per cent longer than the driven element. The director nearest the driven element is 5 per cent shorter, and others are progressively shorter, as shown in the table. Parasitic elements should also be adjusted according to Line 3 of the table, if peak performance is desired at some frequency other than midband.

Parasitic element lengths of Table 18-I are based on element spacings of 0.2 wavelength. This is most often used in v.h.f. arrays, and is suitable for up to 4 or 5 elements. Other spacings can be used, however. If the element lengths are adjusted properly there is little difference in gain with reflector spacings of 0.15 to 0.25 wavelength. The closer the reflector is to the driven element, the shorter it must be for optimum forward gain, and the greater will be its effect on the driven element impedance.

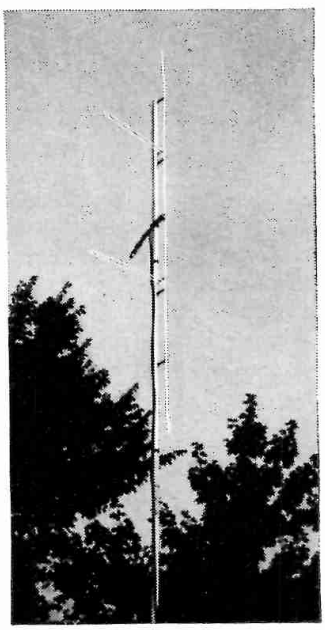


Fig. 18-3—Omnidirectional vertical array for 144 Mc. Elements of aluminum clothesline wire are mounted on ceramic standoff insulators screwed to a wooden pole. Feedline shown is 52-ohm coax, with a balun at the feedpoint. Twin-Lead or other 300-ohm balanced line may also be used, but it should be brought away horizontally from the supporting pole and elements for at least a quarter wavelength. Coax may be taped to the support.

Directors may also be spaced over a similar range. Closer spacing than 0.2 wavelength for arrays of two or three elements will require a longer director than shown in Table 18-I. Thus it can be seen that close-spaced arrays tend to work over a narrower frequency range than wide-spaced ones, when they are tuned for best performance. They also result in lower driven-element impedance, making them more difficult to feed properly. Spacings less than 0.15 wavelength are not commonly used in v.h.f. arrays for these reasons.

Practical Designs for V.H.F. Arrays

The antenna systems pictured and described herewith are examples of ways in which the information in Table 18-I can be used in arrays of proven performance. Dimensions can be taken from the table, except where otherwise noted. If the builder wishes to experiment with element adjustment, a simple method is shown in Fig. 18-2. With elements ½-inch or larger diameter a piece of the element material can be used. It is sawed lengthwise and then compressed to make

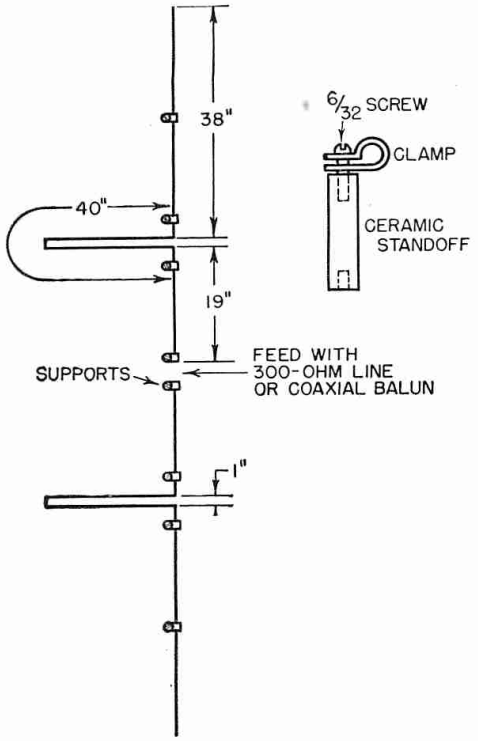


Fig. 18-4—Dimensions and supporting method for the 144-Mc. vertical array.

a tight fit inside the end of the element.

A readily available material often used for elements in arrays for 144 Mc. and higher is aluminum clothesline wire. This is a stiff hard-drawn wire about ⅛ inch in diameter. It should be used in preference to a similar-appearing wire commonly sold for TV grounding purposes. The latter is too soft to make satisfactory elements if the length is more than about two feet.

A Collinear Array for 144 Mc.

Where a vertically-polarized array having some gain over a dipole is needed, yet directivity is undesirable, collinear halfwave elements may be mounted vertically and fed in phase, as shown in Figs. 18-3 and 18-4. Such an array may have 3 elements, as shown, or 5. The impedance at the center is approximately 300 ohms, permitting it to be fed directly with TV-type line, or through a coaxial balun, as in the model shown. Either 52- or 72-ohm line may be employed without serious mismatch.

The array is made from two pieces of aluminum clothesline wire about 97 inches long overall. These are bent to provide a 38-inch top section, a folded-back 40-inch phasing loop, and a 19-inch center section. These elements are mounted on ceramic pillars, which are fastened to a round wooden pole. Small clamps of sheet aluminum are wrapped around the elements and screwed to the stand-offs. A cheaper but somewhat less desirable method of mounting is to use TV screw-eye insulators to hold the elements in place.

Feeding the array at the center with a coaxial balun makes a neat arrangement. The balun loop may be taped to the vertical support, and the coaxial line likewise taped at intervals down the mast. The same type of construction can be applied to a 220-Mc. vertical collinear array, using the lengths for that band given in Table 18-I.

PARASITIC ARRAYS

Single-bay arrays of 2 to 5 elements are widely used in 50-Mc. work. These may be built in many different ways, using the dimensions given in the table. Probably the strongest and lightest structure results from use of aluminum or dural tubing (usually 1¼ to 1½ inches in diameter) for the boom, though wood is also usable. If the elements are mounted at their midpoints there is no need to use insulating supports. Usually the elements are run through the boom and clamped in place in a manner similar to that shown in Fig. 18-12. Where a metal boom is used the joints between it and the elements must be tight, as any movement at this point will result in noisy reception.

2-Element 50-Mc. Array

The 2-element antenna of Fig. 18-5 was designed for portable use, but it is also suitable for fixed-station work with minor modification. The 2-meter array above it is described later. The elements are made in three sections, for portability, using inserts similar to that shown in Fig. 18-2. The driven element is gamma matched for coax feed, and the parasitic element is a 0.15-wavelength spaced director. Details of

Fig. 18-5—Two-element 50-Mc. and four-element 144-Mc. arrays designed for portable use. Support is sectional TV masting clamped to car door handle. Elements of 50-Mc. array are made in three sections, for stowing in back of car. Antenna for 144 Mc. is cut-down TV array. Both use gamma match, as shown in Fig. 18-6.

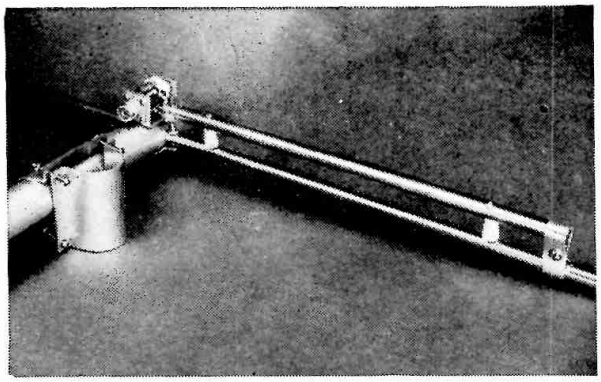


Fig. 18-6—Details of the gamma match for the 50-Mc. portable array. In a permanent installation the variable capacitor should be mounted in an inverted plastic cup or other device to protect it from the weather. The gamma arm is about 12 inches long for 50 Mc., 5 inches for 144 Mc.

the gamma section, the boom and its supporting clamp are shown in Fig. 18-6. The arm is about 12 inches long, and the capacitor is a 50-$\mu\mu$f. variable. Clean, tight connections between the arm and element are important. Where the array is to be mounted permanently outdoors the capacitor may be protected from the weather by mounting it in an inverted plastic cup. More details on this array are given in August, 1955, *QST*.

3-Element Lightweight Array

The 3-element 50-Mc. array of Fig. 18-7 weighs only 5 pounds. It uses the closest spacing that is practical for v.h.f. applications, in order to make an antenna that could be used individually or stacked in pairs without requiring a cumbersome support. The elements are half-inch aluminum tubing of 1/16-inch wall thickness, attached to the 1¼-inch dural boom with aluminum castings made for the purpose. (Dick's, RR1, Tiffin, Ohio, Type HASL.) By limiting the element spacing to 0.15 wavelength the boom is only 6 feet long. Two booms for a stacked array (Fig. 18-11) can thus be cut from a single 12-foot length of tubing.

The folded-dipole driven element has No. 12 wire for the fed portions. These are mounted on ¾-inch cone standoff insulators and joined to the outer ends of the main portion by means of metal pillars and 6-32 screws and nuts. When the wires are pulled up tightly and wrapped around the screw, solder should be sweated over the nuts and screw ends to seal the whole against weather corrosion. The same treatment should be used at each standoff. Mount a soldering lug on the ceramic cone and wrap the end of the lug around the wire and solder the whole assembly together. These joints and other portions of the array may be sprayed with clear lacquer as an additional protection.

The inner ends of the folded dipole are 1½ inches apart. Slip the dipole into its aluminum casting, and then drill through both element and casting with a No. 36 drill, and tap with 6-32 thread. Suitable inserts for mounting the stand-offs can be made by cutting the heads off 6-32 screws. Taper the cut end of the screw slightly with a file and it will screw into the standoff readily.

Cut the dipole length according to Table 18-I, for the middle of the frequency range you expect to use most. The reflector and director will be approximately 4 per cent longer and shorter, respectively. The closer spacing of the parasitic elements (0.15 wavelength) makes this deviation from the dimensions of the table desirable.

The single 3-element array has a feed impedance of about 200 ohms at its resonant frequency. Thus it may be fed with 52-ohm coax and a balun. A gamma-matched dipole may also be used, as in the 2-element array. If the gamma match and 72-ohm coax are used, a balun will convert to 300-ohm balanced feed, if Twin-Lead or 300-ohm open-wire TV line feed is desired. If the dimensions are selected for optimum performance at 50.5 Mc. the array will show good performance and fairly low standing-wave ratio over the range from 50 to 51.5 Mc.

A closeup of a mounting method for this or any other array using a round boom is shown in Fig. 18-8. Four TV-type U bolts clamp the horizontal and vertical members together. The metal plate is about 6 inches square. If ¼-inch sheet aluminum is available it may be used alone, though the photograph shows a sheet of 1/16-inch stock backed up by a piece of wood of the same size for stiffening.

High-Performance 4-Element Array

The 4-element array of Fig. 18-9 was designed for maximum forward gain, and for direct feed with 300-ohm balanced transmission line. The parasitic elements may be any diameter from ½ to 1 inch, but the driven element should be made as shown in the sketch. The same general arrangement may be used for a 3-element array, except that the solid portion of the dipole should

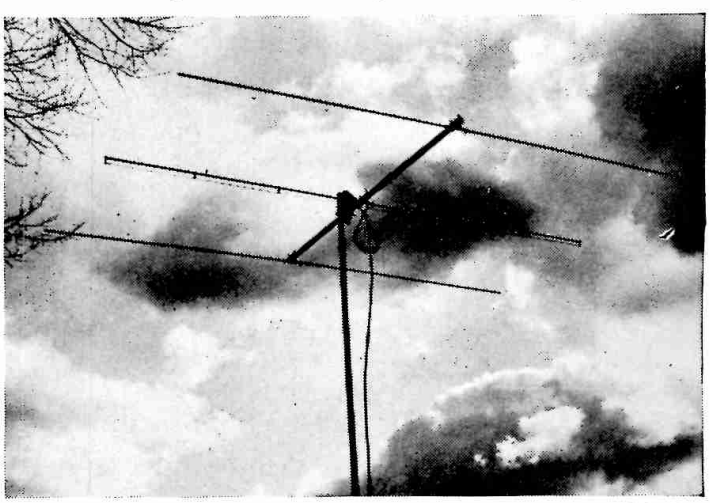


Fig. 18-7—Lightweight 3-element 50-Mc. array. Feedline is 52-ohm coax, with a balun for connection to the folded-dipole driven element. Balun may be coiled as shown or taped to supporting pipe.

be ¾-inch tubing instead of 1-inch. With the element lengths given the array will give nearly uniform response from 50 to 51.5 Mc., and usable gain to above 52 Mc. It may be peaked for any portion of the band by using the information in Table 18-I.

If a shorter boom is desired, the reflector spacing can be reduced to 0.15 wavelength and both

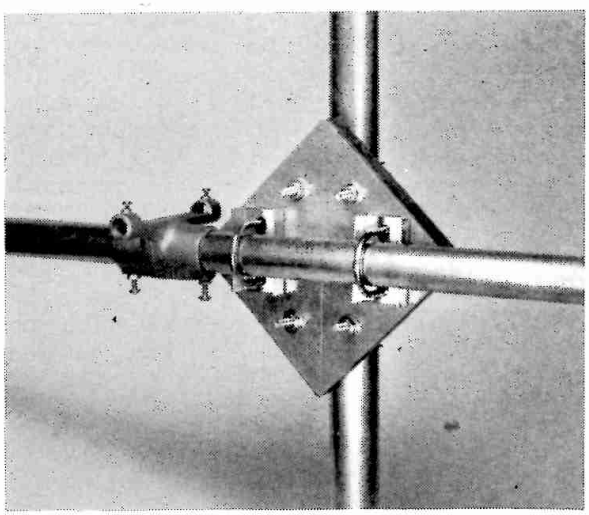


Fig. 18-8—Closeup photograph of the boom mounting for the 50-Mc. array. A sheet of aluminum 6 inches square is backed up by a piece of wood of the same size. TV-type U clamps hold the boom and vertical support together at right angles. At the left of the mounting assembly is one of the aluminum castings for holding the beam elements.

directors spaced 0.2 or even 0.15 wavelength, with only a slight reduction in forward gain and bandwidth.

5-Element 50-Mc. Array

As aluminum or dural tubing is usually sold in 12-foot lengths this dimension imposes a practical limitation on the construction of a 50-Mc. beam. A 5-element array that makes optimum use of a 12-foot boom may be built according to Table 18-I. If the aluminum casting method of mounting elements shown for the 3-element array is employed the weight of a 5-element beam can be held to under 10 pounds. The gamma match and coaxial line are recommended for feeding such an array, though a balun and 72-ohm coax can be used for the rotating portion of the line, converting to balanced feed at the anchor point.

Elements should be spaced 0.15 wavelength, or about 36 inches. With 5 or more elements, good bandwidth can be secured by tapering the element lengths properly. A dipole 110 inches long, with a 116-inch reflector, and directors of 105, 103 and 101 inches respectively will work well over the first two megacycles of the band, provided that the s.w.r. is adjusted for optimum at 51 Mc.

Long Yagis for 50 Mc.

With boom lengths greater than about 12 feet and with more elements than 4, somewhat

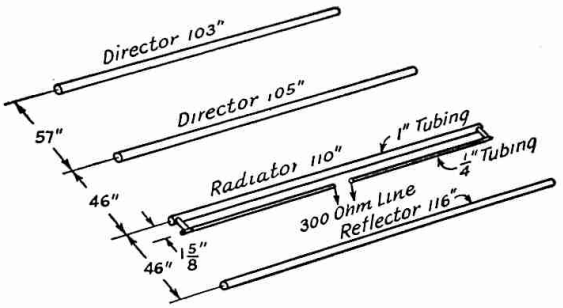
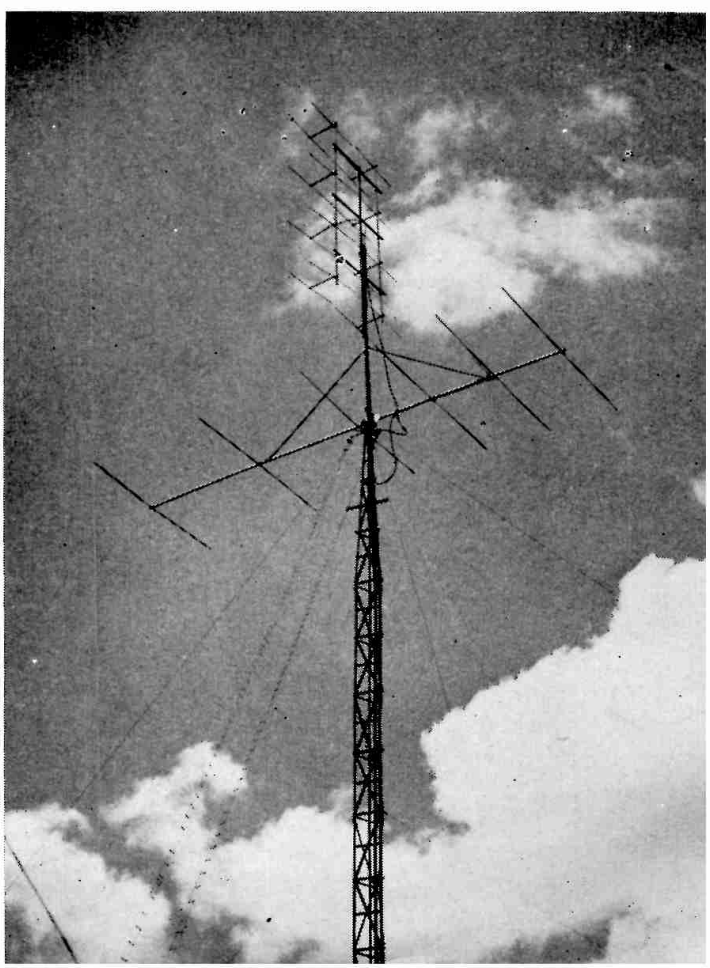


Fig. 18-9—Details of a 4-element 50-Mc. array designed for 300-ohm balanced feed. Element lengths and spacings were derived experimentally for optimum performance over the first 1.5 megacycles of the band.

better performance can be obtained by using gradually increasing spacing between the directors. The 6-element array in Fig. 18-10 is an example of this approach. It also employs a variation of the gamma match that has mechanical advantages. The long boom and wide-spaced elements give a sharpness of horizontal pattern that is not obtainable with the same number of elements in a stacked array.

The long Yagi is not a broadband device. This one works well over the first megacycle of the band with the following dimensions. Subtract 2 inches from each element for each megacycle

Fig. 18-10—A 6-element long Yagi for 50 Mc. and a 16-element collinear array for 144 Mc. Both are all-metal construction. Each has its own vertical member, which is clamped to the rotating vertical pipe that runs down through the tower bearing.



higher. Reflector — 116 inches. Driven element — 110.5. First director — 105.5. Second director — 104. Third director — 102.75. Fourth director — 101.5. Spacings are, from back forward: 36, 36, 42, 59 and 70 inches. If a longer array is to be built each additional director should be 70 inches from the last.

Construction

The long Yagi is built similar to the 3-element array of Fig. 18-7 and 18-8, using those same castings for mounting the elements. The gusset plate for fastening the boom to the vertical support is made larger, and four U bolts are used on each member instead of two. The array is mounted at its center of gravity, rather than at its physical center. The boom is braced to prevent drooping, at points about 5 feet out from the mounting point. Braces are aluminum tubing, flattened at the ends, and clamped to the boom and the vertical member. Suspension bracing, as shown in Fig. 18-10, provides strength with lightweight supports.

The dimensions given require a boom slightly more than 20 feet long. This was made up by splicing, but if a 20-foot length is available in one piece the spacings of the two forward directors can be made slightly less, in order to avoid splicing. Element spacing is not particularly critical, but lengths are fairly so.

The Gamma Match

The gamma match is ideal for matching arrays fed with coax. The arrangement shown in Fig. 18-11 combines the adjustable arm with the series capacitor, and provides a rugged assembly that can be weather-proofed readily. The main arm is cut from the same material as the elements, 15 inches long. It is supported parallel to the driven element by means of two 1-inch ceramic standoffs and sheet-aluminum clips. Its inner end is connected to the inner conductor of a coaxial fitting, mounted on a small bracket screwed to the boom.

The series capacitor, for tuning out the reactance of the matching arm and making connection to the driven element, is ¼-inch rod or tubing 14 inches long. It is maintained coaxial with the main arm by two polystyrene bushings. One is force-fitted to the end of the rod and the other is fitted tightly inside the main arm to act as a bearing. These can be made from ⅜-inch rod stock, or National Type PRC-1 forms can be adapted readily to the purpose. A clip of sheet aluminum connects the rod and the driven element. Be sure that a clean tight contact is made at this point.

Adjustment

Matching requires an s.w.r. bridge. It can be done properly in no other way. Mount the beam at least a half wavelength above ground and clear of trees and wires by at least the same distance. Set the transmitter at a frequency in the middle of the range you want to work (50.3 is a good spot for low-end operation) and adjust the position of the clip and the length of the rod outside the main arm for minimum s.w.r. Move first one variable and then the other until zero reflected power is indicated. Tighten the clip solidly, tape over the junction between the arm and the rod with waterproof tape, and the array is ready for use.

● 144-MC. PARASITIC ARRAYS

The main features of the arrays described above can be adapted to 144-Mc. antennas, but the small physical size of arrays for this frequency makes it possible to use larger numbers of elements with ease. Few 2-meter antennas have less than 4 or 5 elements, and most stations use more, either in a single bay or in stacked systems.

Parasitic arrays for 144 Mc. can be made readily from TV antennas for Channels 4, 5 or 6. The relatively close spacing normally used in TV arrays makes it possible to approximate the recommended 0.2 wavelength at 144 Mc., though the element spacing is not a critical factor. A 4-element array for 144 Mc. made from a Channel 6 TV Yagi is shown in Fig. 18-5. It is fed with a gamma match and 52-ohm coax, and was designed primarily for portable work. As most TV antennas are designed for 300-ohm feed the same feed system can be employed for the 2-meter array that is made from them.

If one wishes to build his own Yagi antennas from available tubing sizes, the boom of a 2-meter antenna should be ¾ to 1 inch aluminum

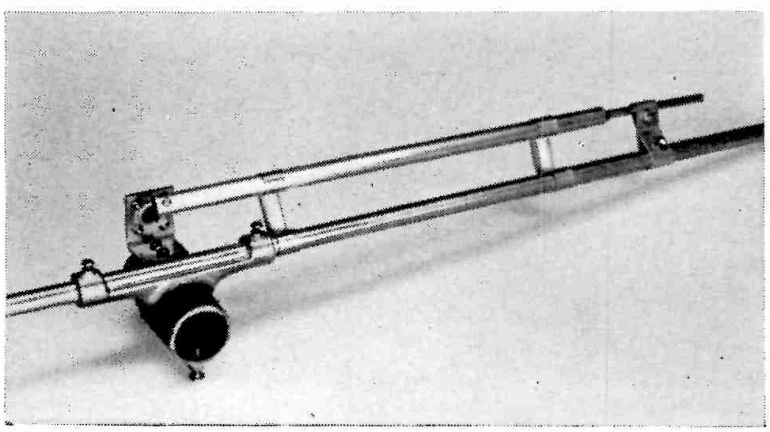


Fig. 18-11—Details of the gamma match used on the 6-element 50-Mc. array. Series capacitor is formed by sliding a rod or tube inside the main arm.

or dural. Elements can be ¼ to ½-inch stock, fastened to the boom as shown in Fig. 18-12. Recommended spacing for up to 6 elements is 0.2 wavelength, though this is not too critical. Gamma match feed is recommended for coax, or a folded dipole and balun may be used. If balanced line is to be used the folded dipole is

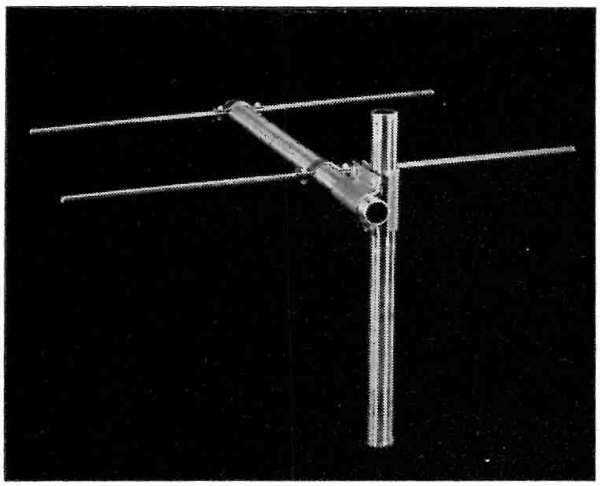


Fig. 18-12—Model showing method of assembling all-metal arrays for 144 Mc. and higher frequencies. Dimensions of clamps are given in Fig. 18-16.

recommended, the 4 to 1 ratio of conductor sizes being about right for most designs.

Very high gain can be obtained with long Yagi-type arrays for 144 Mc. and higher frequencies, though the bandwidth of such antennas is considerably narrower than for those having up to 4 or 5 elements. The first two directors in long Yagis are usually spaced about 0.1 wavelength. The third is spaced about 0.2, increasing to 0.4 wavelength or so for the forward directors. Highest gain is obtained when all directors are made the same length, but better front-to-back ratio and lower side lobe content results if the director lengths are tapered ⅛ to ¼ inch per director. Tapering the element lengths also widens the effective bandwidth. There is more on long Yagis in *QST* for January and September, 1956.

STACKED YAGI ARRAYS

The gain (in power) obtainable from a single Yagi array can be more than doubled by stacking two or more of them vertically and feeding them in phase. This refers to horizontal systems, of course. Vertically-polarized bays are usually stacked side by side. The principles to follow apply in either case.

The spacing between bays should be at least one-half wavelength, and more is desirable. For dipoles or Yagis of up to three elements optimum spacing between bays is about ⅝ wavelength, but with longer Yagis the spacing can be increased to one wavelength or more. Bays of 5 elements or more, spaced one wavelength, are commonly used in antennas for 144 Mc. and higher frequencies. Optimum spacing for long Yagis is about two wavelengths.

Where half-wave stacking is to be employed, the phasing line between bays can be treated as a double "Q" section. If two bays, each designed for 300-ohm feed, are to be stacked a half wavelength apart and fed at the midpoint between them, the phasing line should have an impedance of about 380 ohms. No. 12 wire spaced one inch will do for this purpose. The midpoint then can be fed either with 300-ohm line, or with 72-ohm coax and a balun.

When a spacing of ⅝ wavelength between bays is employed, the phasing lines can be coax. (The velocity factor of coax makes a full wavelength of line actually about ⅝ wavelength physically.) The impedance at the midpoint between two bays is slightly less than half the impedance of either bay alone, due to the coupling between bays. This effect decreases with increased spacing.

When two bays are spaced a full wavelength the coupling is relatively slight. The phasing line can be any open-wire line, and the impedance at the midpoint will be approximately half that of the individual bays. Predicting what it will be with a given set of dimensions is difficult, as many factors come into play. It will usually be of a value that can be fed through the combination of a "Q" section and a transmission line of 300 to 450 ohms impedance. An adjustable "Q" section, or an adjustable stub like the one shown in Fig. 18-1, may be used when the antenna impedance is not known.

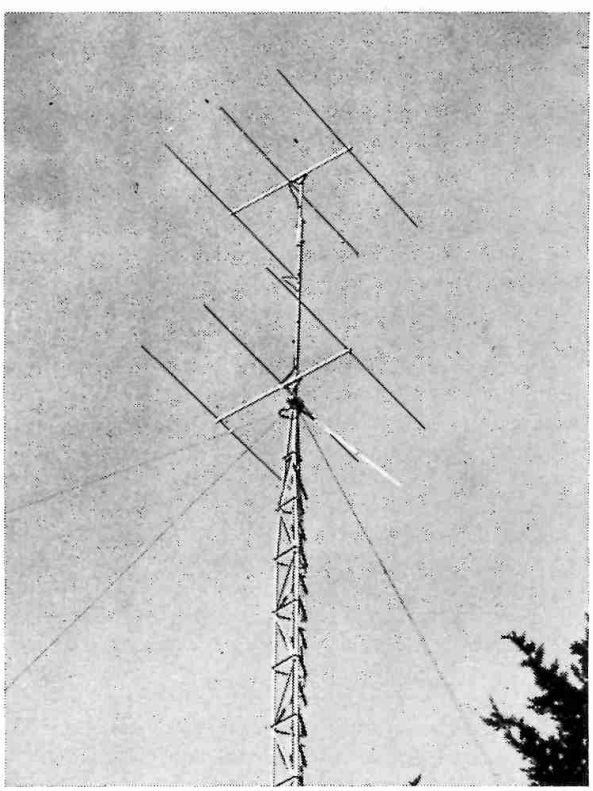


Fig. 18-13—Stacked array for 50 Mc. using two of the 3-element bays of Fig. 18-7. Phasing system and flexible section for rotation are of coaxial line. A "Q" section matches this to 450-ohm open-wire line for run to the station.

The stacked 3-over-3 for 50 Mc., Fig. 18-13, uses a coaxial phasing line and an additional section of coax to provide for the flexible portion of the feedline. Each bay is fed with a balun and halfwave section of RG-8/U cable. These are joined at the center between bays with a Tee fitting. As each bay has an impedance of 200 ohms, two 50-ohm leads are paralleled at the center, resulting in an impedance of about 20 ohms, when the coupling effect between bays is included. A flexible section of 50-ohm coax one wavelength long, with a balun at the end, steps this up to about 80 ohms. A "Q" section of ¼-inch tubing ¾ inch center to center steps this up to the point where it can be fed with 450-ohm open-wire TV line.

The "Twin-Five" for 144 Mc.

A popular stacked array for 144-Mc. work is the Twin-Five, originally developed by W2PAU [1]. In this design two 5-element arrays of standard design are stacked a full wavelength apart. If the folded-dipole driven elements are constructed so that the individual bays have a feed impedance of about 400 ohms the midpoint of the open-wire phasing line can be fed with 52-ohm coax and a balun. Where open-wire line is desired, the impedances can be matched through a "Q" section of about 300 ohms impedance. If the constructor is in doubt as to the actual feed impedance to be matched, the stub arrangement of Fig. 18-1 will take care of a wide range of impedances and lines to be matched. Dimensions can be taken from Table 18-I.

An effective 20-element array can be made by using two of these arrays side by side, with full-wave spacing horizontally also. The impedance at the midpoint of the horizontal phasing line will then be about 100 ohms, which is still well within the range of "Q" sections of practical dimensions.

● LARGE COLLINEAR ARRAYS FOR 144 MC. AND HIGHER

High gain and very broad frequency response are desirable characteristics found in curtains of half-wave elements fed in phase and backed up by reflectors. The reflector can be made up of parasitic elements, or it can be a screen extending approximately a quarter wavelength beyond the ends of the driven elements. There is not a large difference between the two types of reflectors, except that higher front-to-back ratio and somewhat broader frequency response are achieved with the plane reflector.

12- and 16-Element Arrays

Two collinear systems that may be used on 144, 220 or 420 Mc. are shown in Figs. 18-14 and 18-15. Either may be fed directly with 300-ohm transmission line, or through coaxial line and a balun. In the 12-element array, Fig. 18-14, the reflectors are spaced 0.15 wavelength in back of the driven elements, while the 16-element array, Figs. 18-15 and 18-10, uses 0.2 wavelength spacing. Dimensions may be taken from Table 18-I, and figures for the middle of the band will give good performance across either band.

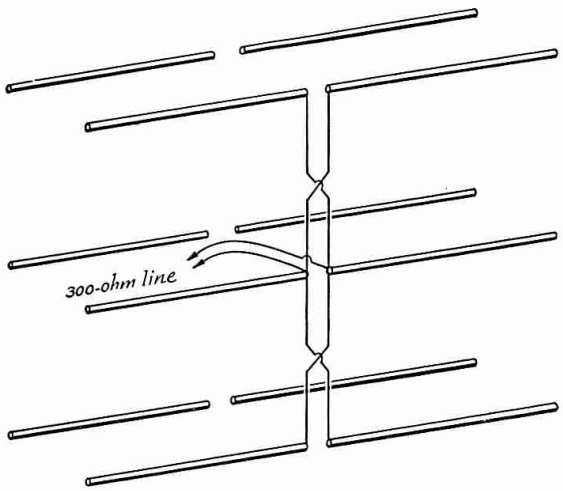


Fig. 18-14—Element arrangement and feed system of the 12-element array. Reflectors are spaced 0.15 wavelength behind the driven elements.

The supporting frame for either array may be made of wood or metal. Details of a metal support for the 12-element array are shown in Figs. 18-16 and 18-17. Note that all elements are mounted at their midpoints, and that no insulators are used. The elements are mounted in front of the supporting frame, to keep metal out of the field of the array. This method is preferable to that wherein mechanical balance is maintained

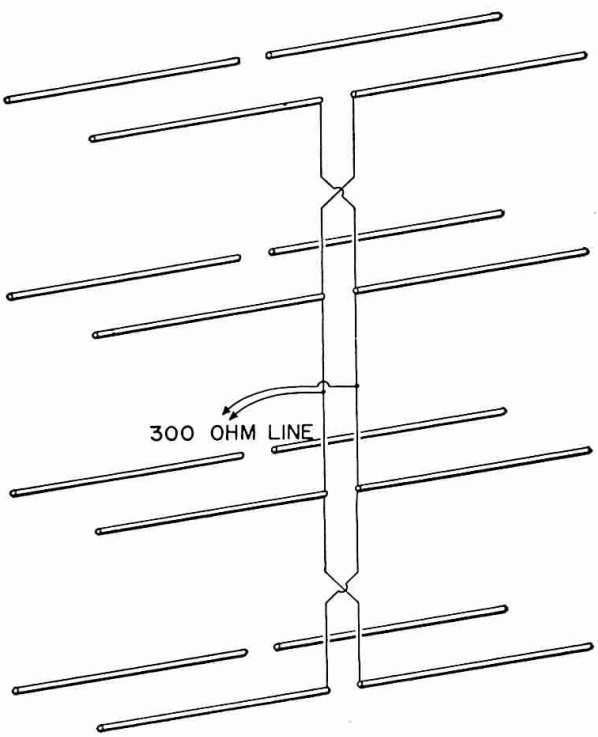


Fig. 18-15—Schematic drawing of a 16-element array. A variable "Q" section may be inserted at the feed point if accurate matching is desired. Reflector spacing is 0.2 wavelength.

[1] Brown — "The Wide-Spread Twin-Five" *CQ*, March, 1950.

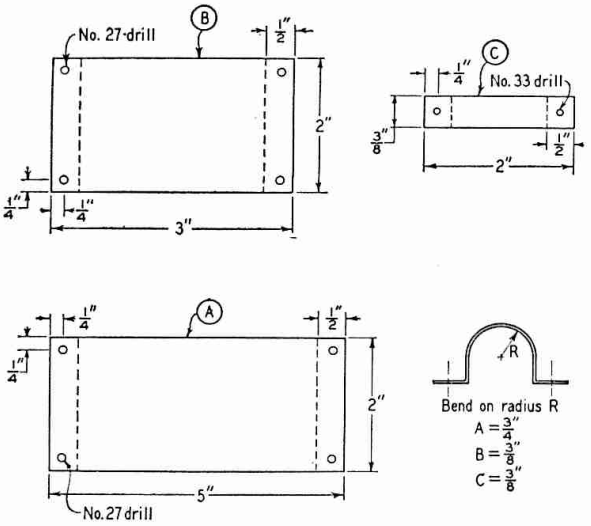


Fig. 18-16—Detail drawings of the clamps used to assemble the all-metal 2-meter array. A, B and C are before bending into "U" shape. The right-angle bends should be made first, along the dotted lines as shown, then the plates may be bent around a piece of pipe of the proper diameter. Sheet stock should be 1/16-inch or heavier aluminum.

through mounting the driven elements in front and the reflectors in back of the supporting structure.

Two 12-element arrays may be mounted one above the other and fed in phase, to form a 24-element array. This is done in the 420-Mc. array of Fig. 18-18. The two midpoints are connected through a phasing line one wavelength long, and the center of this phasing line fed through a "Q" section. The impedance at the midpoint is about 150 ohms, requiring a 255-ohm "Q" section for feeding with 450-ohm open-wire line.

Combination of collinear arrays may be carried further. Pairs of 16-element systems fed in phase are common, and even 64-element arrays (4 16-element beams fed in phase) are used in some leading stations on 144 Mc. Configurations of 32 to 64 elements are not difficult to build and support at 220 or 420 Mc. Examples of 16- and 24-element arrays for 220 and 420 Mc. are shown mounted back to back in Fig. 18-18.

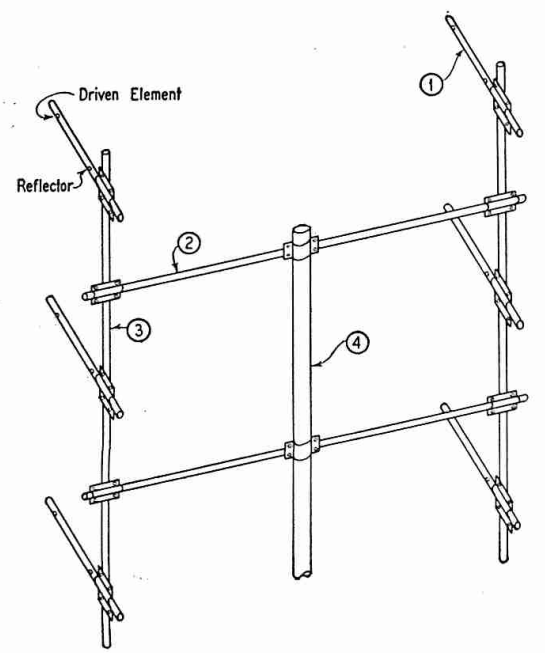


Fig. 18-17—Supporting framework for a 12-element 144-Mc. array of all-metal design. Dimensions are as follows: element supports (1) ¾ by 16 inches; horizontal members (2) ¾ by 46 inches; vertical members (3) ¾ by 86 inches; vertical support (4) 1½-inch diameter, length as required; reflector-to-driven-element spacing 12 inches. Parts not shown in sketch: driven elements ¼ by 38 inches; reflectors ¼ by 40 inches; phasing lines No. 18 spaced 1 inch, 80 inches long, fanned out to 3½ inches at driven elements (transpose each half-wave section).

ARRAYS FOR 220 AND 420 MC.

The use of high-gain antenna systems is almost a necessity if work is to be done over any great distance on 220 and 420 Mc. Experimentation with antenna arrays for these frequencies is fascinating indeed, as their size is so small as to permit trying various element arrangements and feed systems with ease. Arrays for 420 Mc., particularly, are convenient for study and demonstration of antenna principles, as even high-gain systems may be of table-top proportions.

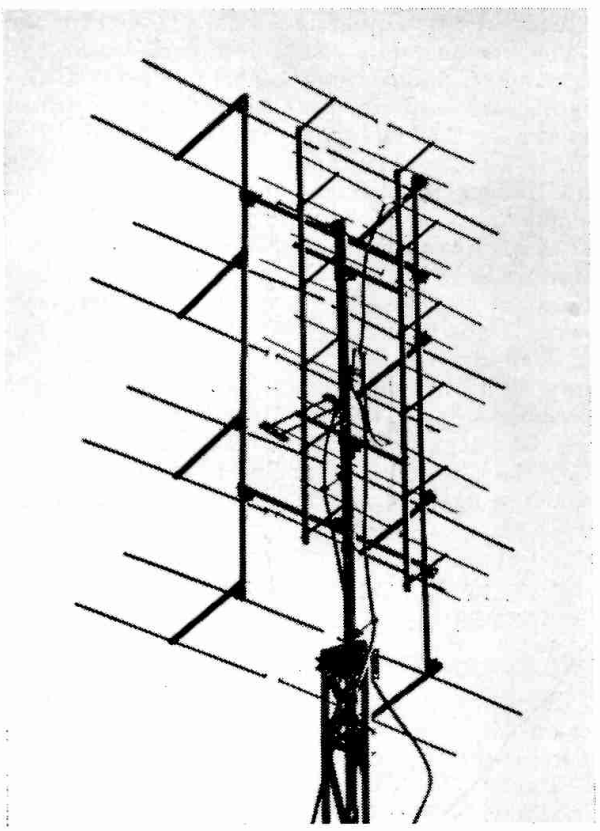


Fig. 18-18—A 24-element array for 420 Mc. and a 16-element for 220 mounted back-to-back on a single support.

Any of the arrays described previously may be used on these bands, but those having large numbers of driven elements in phase are more readily adjusted for maximum effectiveness.

A 16-element array for 220 Mc. and a 24-

element array for 420 Mc. are shown mounted back-to-back in Fig. 18-18. The 220-Mc. portion follows the 16-element design already described. It is fed at the center of the system with 300-ohm tubular Twin-Lead, matched to the center impedance of the array through a "Q" section of 7/16-inch tubing, spaced about 1½ inches center to center. This spacing was adjusted for minimum standing-wave ratio on the line.

Elements in the array shown are of 7/16-inch aluminum fuel-line tubing, which is very light in weight and easily worked. The supporting structure is dural tubing, using the clamp assembly methods of Fig. 18-16.

The 420-Mc. array uses two 12-element assemblies similar to Fig. 18-14, mounted one above the other, about one half wavelength separating the bottom of one from the top of the other. The two sets of phasing lines are joined by one-wavelength sections of Twin-Lead at the middle of the array. This junction, which has an impedance of around 150 ohms, is fed with 300-ohm tubular Twin-Lead through an adjustable "Q" section.

Elements in the 420-Mc. array are cut from thin-walled ¼-inch tubing. Their supports are the 7/16-inch stock used for the 220-Mc. elements. Slots were cut in the ends of these supports to take the elements, and a 4–40 screw was run through both pieces and drawn up tightly with a nut. The horizontal supports were fastened in holes drilled in the vertical members, and were also held in place with a 6-32 screw and nut. The small size and light weight of the 420-Mc. array require no clamps to make a strong assembly.

The two one-wavelength sections of 300-ohm line are 21¾ inches long, taking the propagation factor into account. The "Q" section may be any convenient size tubing, ¼ to ½ inch diameter. It should be made adjustable, as matching is important at this frequency. Dimensions for both arrays can be taken from Table 18-I.

(For an example of stacking several commercial 220-Mc. beams, see Tilton, "A 66-Element Stacked-Yagi Array for 220 Mc.," *QST*, January, 1959.)

MISCELLANEOUS ANTENNA SYSTEMS

Coaxial Antennas

At v.h.f. the lowest possible radiation angle is essential, and the coaxial antenna shown in Fig. 18-19 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric (coaxial) line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is non-resonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna. 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.

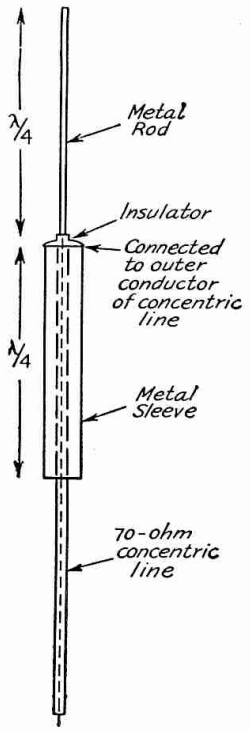


Fig. 18-19—Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

Broadband Antennas

Certain types of antennas used in television are of interest because they work across a wide band of frequencies with relatively uniform response. At very-high frequencies an 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 is 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 wideband antenna, on the other hand, will exhibit very nearly constant input impedance over several megacycles.

The simplest method of obtaining a broadband 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.

As the diameter-to-length ratio is increased, end effects also increase, with the result that the antenna must be made shorter than 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.

Plane-Reflector Arrays

At 220 Mc. and higher, where their dimensions become practicable, plane-reflector arrays are widely used. Except as it affects the impedance of the system, as shown in Fig. 18-20, the spacing between the driven elements and the reflecting plane is not particularly critical. Maximum gain occurs around 0.1 to 0.15 wavelength, which is also the region of lowest impedance. Highest impedance appears at about 0.3 wavelength. A plane reflector spaced 0.22 wavelength in back of the driven elements has no effect on their feed impedance. As the gain of a plane-reflector array is nearly constant at spacings from 0.1 to 0.25 wavelength, it may be seen that the spacing may be varied to achieve an impedance match.

An advantage of the plane reflector is that it may be used with two driven element systems, one on each side of the plane, providing for two-band operation, or the incorporation of horizontal and vertical polarization in a single structure. The gain of a plane-reflector array is slightly higher than that of a similar number of driven elements backed up by parasitic reflectors. It also has a broader frequency response and higher front-to-back ratio. To achieve these ends, the reflecting plane must be larger than the area of the driven elements, extending at least a quarter wavelength on all sides. Chicken wire on a wood or metal frame makes a good plane reflector. Closely spaced wires or rods may be substituted, with the spacing between them running up to 0.1 wavelength without appreciable reduction in effectiveness.

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 variation in the characteristic impedance with changes in frequency can be reduced to a very low value, making such an antenna suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines. A variation of this form of conical antenna is widely used in TV reception.

Corner Reflectors

In the corner reflector two plane surfaces are set at an angle, usually between 45 and 90 degrees, with the antenna on a line bisecting this angle. Maximum gain is obtained with the antenna 0.5 wavelength from the vertex, but compromise designs can be built with closer spacings. There is no focal point, as would be the case for a parabolic reflector. Corner angles greater than 90 degrees can be used at some sacrifice in gain. At less than 90 degrees the gain increases, but the size of the reflecting sheets must be increased to realize this gain.

At a spacing of 0.5 wavelength from the vertex, the impedance of the driven element is approximately twice that of the same dipole in free space. The impedance decreases with smaller spacings and corner angles, as shown in Fig. 18-20. The gain of a corner-reflector array with a 90-degree angle, 0.5 wavelength spacing and sides one wavelength long is approximately 10 db. Principal advantages of the corner reflector are broad frequency response and high front-to-back ratio.

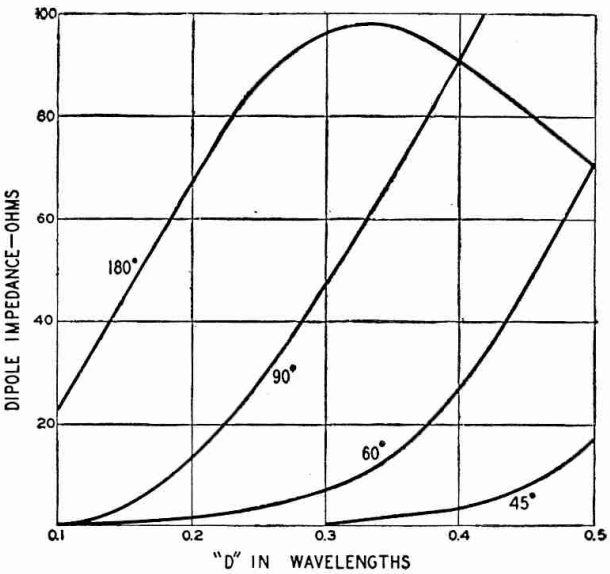


Fig. 18-20—Feed impedance of the driven element in a corner-reflector array for corner angles of 180 (flat sheet), 90, 60 and 45 degrees. "D" is the dipole-to-vertex spacing.

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, sizes that may be practical for microwave work, a beam width of approximately 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.

Mobile and Portable-Emergency Equipment

The amateur who goes in for mobile operation will find plenty of room for exercising his individuality and developing original ideas in equipment. Each installation has its special problems to be solved.

Most mobile receiving systems are designed around the use of a h.f. converter working into a standard car broadcast receiver tuned to 1500 kc. which serves as the i.f. and audio amplifiers. The car receiver is modified to take a noise limiter and provide power for the converter.

While a few mobile transmitters may run an input to the final amplifier as high as 100 watts or more, an input of about 30 watts normally is considered the practical limit unless the car is equipped with a special battery-charging system. The majority of mobile operators use phone.

In contemplating a mobile installation, the car should be studied carefully to determine the most suitable spots for mounting the equipment. Then the various units should be built in a form that will make best use of that space. The location of the converter should have first consideration. It should be placed where the controls can be operated conveniently without distracting attention from the wheel. The following list suggests spots that may be found suitable, depending upon the individual car.

On top of the instrument panel
Attached to the steering post
Under the instrument panel
In a unit made to fit between the lower lip of the instrument panel and the floor at the center of the car

The transmitter power control can be placed close to the receiver position, or included in the converter unit. This control normally operates relays, rather than to switch the power circuit directly. This permits a minimum length of heavy-current battery circuit. Frequency within any of the phone bands sometimes is changed remotely by means of a stepping-switch system that switches crystals. In most cases, however, it is necessary to stop the car to make the several changes required in changing bands.

Depending upon the size of the transmitter unit, one of the following places may be found convenient for mounting the transmitter:

In the glove compartment
Under the instrument panel
In a unit in combination with or without the converter, built to fit between the lower edge of the instrument panel and the floor at the center
On the ledge above the rear seat
In the trunk

Most mobile antennas consist of a vertical whip with some system of adjustable loading for the lower frequencies. Power supplies are of the vibrator, motor-generator, or transistor type operating from the car storage battery.

Units intended for use in mobile installations should be assembled with greater than ordinary care, since they will be subject to considerable vibration. Soldered joints should be well made and wire wrap-arounds should be used to avoid dependence upon the solder for mechanical strength. Self-tapping screws should be used wherever feasible, otherwise lock-washers should be provided. Any shafts that are normally operated at a permanent or semi-permanent setting should be provided with shaft locks so they cannot jar out of adjustment. Where wires pass through metal, the holes should be fitted with rubber grommets to prevent chafing. Any cabling or wiring between units should be securely clamped in place where it cannot work loose to interfere with the operation of the car.

Noise Elimination

Electrical-noise interference to reception in a car may arise from several different sources. As examples, trouble may be experienced with ignition noise, generator and voltage-regulator hash, or wheel and tire static.

A noise limiter added to the car broadcast receiver will go far in reducing some types, especially ignition noise from passing cars as well as your own. But for the satisfactory reception of weaker signals, some investigation and treatment of the car's electrical system will be necessary.

Ignition Interference

Fig. 19-1 indicates the measures that may be taken to suppress ignition interference. The capacitor at the primary of the ignition coil should be of the coaxial type; ordinary types are not effective. It should be placed as close to the coil terminal as possible. In stubborn cases, two

of these capacitors with an r.f. choke between them may provide additional suppression. The size of the choke must be determined experimentally. The winding should be made with wire heavy enough to carry the coil primary current. A 10,000-ohm suppressor resistor should be inserted at the center tower of the distributor, a 5000-ohm suppressor at each spark-plug tower on the distributor, and a 10,000 ohm suppressor at each spark plug. The latter may be built-in or external. A good suppressor element should be molded of material having low capacitance. Several concerns manufacture satisfactory suppressors. In extreme cases, it may be necessary to use shielded ignition wire. Suppressor ignition wire kits having the resistance distributed throughout the length of the wire are available from some automobile supply dealers. Distributed resistance of this type is somewhat superior to lumped resistance and may be used if the lead lengths are right to fit your car. They should not be cut, but used as they are sold.

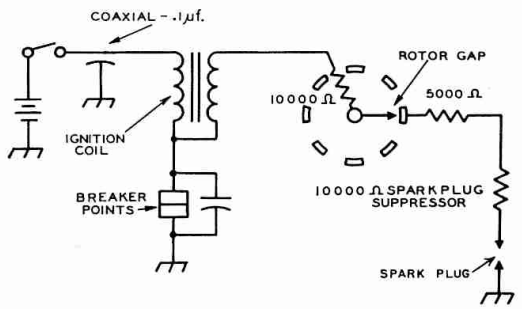


Fig. 19-1—Ignition system with recommended suppression methods.

Generator Noise

Generator hash is caused by sparking at the commutator. The pitch of the noise varies with the speed of the motor. This type of noise may be eliminated by using a 0.1- to 0.25-μf. coaxial capacitor in the generator armature circuit. This capacitor should be mounted as near the armature terminal as possible and directly on the frame of the generator.

To reduce the noise at 28 Mc., it may be necessary to insert a parallel trap, tuned to the middle of the band, in series with the generator output lead. The coil should have about 8 turns of No. 10 wire, space-wound on a 1-inch diameter and should be shunted with a 30-μμf. mica trimmer. It can be pretuned by putting it in the antenna lead to the home-station receiver tuned to the middle of the band, and adjusting the trap to the point of minimum noise. The tuning may need to be peaked up after installing in the car, since it is fairly critical.

Voltage-Regulator Interference

In eliminating voltage-regulator noise, the use of two coaxial capacitors, and a resistor-mica-capacitor combination, as shown in Fig. 19-2, are effective. A 0.1- to 0.25-μf. coaxial capacitor should be placed between the battery terminal of the regulator and the battery, with its case well grounded. Another capacitor of the same size and type should be placed between the generator terminal of the regulator and the generator. A 0.002-μf. mica capacitor with a 4-ohm carbon resistor in series should be connected between the field terminal of the regulator and ground. Never use a capacitor across the field contacts or between field and ground without the resistor in series, since this greatly reduces the life of the

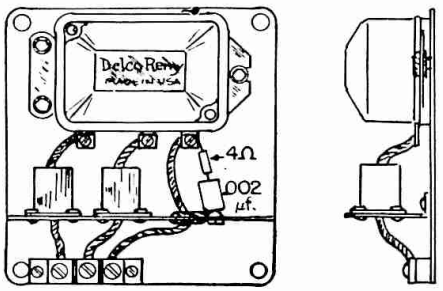


Fig. 19-2—The right way to install bypasses to reduce interference from the regulator. A capacitor should never be connected across the generator field lead without the small series resistor indicated.

regulator. In some cases, it may be necessary to pull double-braid shielding over the leads between the generator and regulator. It will be advisable to run new wires, grounding the shielding well at both ends. If regulator noise persists, it may be necessary to insulate the regulator from the car body. The wire shielding is then connected to the regulator case at one end and the generator frame at the other.

Wheel Static

Wheel static shows up as a steady popping in the receiver at speeds over about 15 m.p.h. on smooth dry streets. Front-wheel static collectors are available on the market to eliminate this variety of interference. They fit inside the dust cap and bear on the end of the axle, effectively grounding the wheel at all times. Those designated particularly for your car are preferable, since the universal type does not always fit well. They are designed to operate without lubrication and the end of the axle and dust cap should be cleaned of grease before the installation is made. These collectors require replacement about every 10,000 miles.

Rear-wheel collectors have a brush that bears against the inside of the brake drum. It may be necessary to order these from the factory through your dealer.

Tire Static

This sometimes sounds like a leaky power line and can be very troublesome even on the broadcast band. It can be remedied by injecting an antistatic powder into the inner tubes through the valve stem. The powder is marketed by General Cement and possibly others. General Cement dealers can also supply a convenient injector for inserting the powder.

Tracing Noise

To determine if the receiving antenna is picking up all of the noise, the shielded lead-in should be disconnected at the point where it connects to the antenna. The motor should be started with the receiver gain control wide open. If no noise is heard, all noise is being picked up via the antenna. If the noise is still heard with the antenna disconnected, even though it may be reduced in strength, it indicates that some signal from the ignition system is being picked up by the antenna transmission

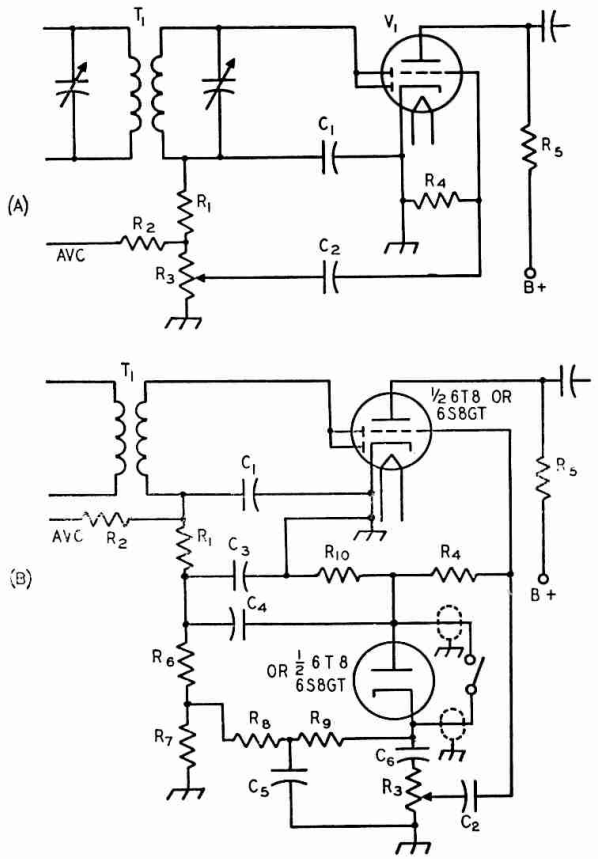


Fig. 19-3—Diagrams showing addition of noise limiter to car receiver. A—Usual circuit. B—Modification.

C_1, C_3—100-$\mu\mu$f. mica.
C_2, C_4, C_6—0.01-μf. paper.
C_5—0.1-μf. paper.
R_1—47,000 ohms.
R_2, R_{10}—1 megohm.
R_3—½ megohm.
R_7, R_8, R_9—0.47 megohm.
R_4—10 megohms.
R_5—¼ megohm.
R_6—0.1 megohm.
T_1—I.f. transformer.
V_1—Second detector.

line. The lead-in may not be sufficiently-well shielded, or the shield not properly grounded. Noise may also be picked up through the battery circuit, although this does not normally happen if the receiver is provided with the usual r.f.-choke-and-bypass capacitor filter.

In case of noise from this source, a direct wire from the "hot" battery terminal to the receiver is recommended.

Ignition noise varies in repetition rate with engine speed and usually can be recognized by that characteristic in the early stages. Later, however, it may resolve itself into a popping noise that does not always correspond with engine speed. In such a case, it is a good idea to remove all leads from the generator so that the only source left is the ignition system.

Regulator and generator noise may be detected by racing the engine and cutting the ignition switch. This eliminates the ignition noise. Generator noise is characterized by its musical whine contrasted with the ragged raspy irregular noise from the regulator.

With the motor running at idling speed, or slightly faster, checks should be made to try to determine what is bringing the noise into the field of the antenna. It should be assumed that any control rod, metal tube, steering post, etc., passing from the motor compartment through an insulated bushing in the firewall will carry noise to a point where it can be radiated to the antenna. All of these should be bonded to the firewall with heavy wire or braid. Insulated wires can be stripped of r.f. by bypassing them to ground with 0.5-μf. metal-case capacitors. The following should not be overlooked: battery lead at the ammeter, gasoline gauge, ignition switch, headlight, backup and taillight leads and the wiring of any accessories running from the motor compartment to the instrument panel or outside the car.

The firewall should be bonded to the frame of the car and also to the motor block with heavy braid. If the exhaust pipe and muffler are insulated from the frame by rubber mountings, they should likewise be grounded to the frame with flexible copper braid.

Noise Limiting

Fig. 19-3 shows the alterations that may be made in the existing car-receiver circuit to provide for a noise limiter. The usual diode-triode second detector is replaced with a type having an extra independent diode. If the car receiver uses octal-base tubes, a 6S8GT may be substituted. The 7X7 is a suitable replacement in receivers using loktal-type tubes, while the 6T8 may be used with miniatures.

The switch that cuts the limiter in and out of the circuit may be located for convenience on or near the converter panel. Regardless of its placement, however, the leads to the switch should be shielded to prevent hum pick-up.

Several other noise limiter circuits are described in ARRL's publication, *The Mobile Manual For Radio Amateurs*. The *Mobile Manual* also describes an audio squelch system. The latter is a simple circuit designed to suppress receiver background noise in the absence of a signal. It does not, however, function as a noise limiter when the receiver is tuned to a signal.

At least one manufacturer (Gonset) produces a complete noise limiter unit. The unit is mounted external to the main chassis and takes operating voltages from the receiver.

A Mobile Converter for 3.5 through 28 Mc.

Figures 19-4 through 19-7 show a crystal-controlled converter covering 3.5 through 28 Mc. without complex band switching or gang-tuned circuits. Plug-in coil assemblies provide rapid band changing and allow construction for either single-band or multiband operation. The converter uses the car broadcast receiver as a tunable i.f. amplifier.

Plate power requirements for the converter are approximately 20 milliamperes at 200 to 250 volts. This means that the unit can be supplied from the car-receiver power pack without overloading it.

The Circuit

The circuit diagram of the converter is shown in Fig. 19-5. A 6BZ6 is used in the r.f. amplifier, and a 12AT7 operates as a mixer-oscillator. The oscillator is crystal-controlled and works on the low-frequency side of the signal frequency. J_1, J_2, and J_3 are the antenna-input, mixer-output and power jacks, respectively. S_1 performs the switching in changing over from ham-band to broadcast input. S_{1A} and S_{1B} shift the antenna from the converter input circuit to the car receiver, and S_{1C} is the heater on-off switch.

Since the tuning of the converter is fixed, the circuits of the r.f. amplifier and the mixer must be broadbanded to pass all frequencies in any ham band. A slug-tuned coil, L_3, is used in the amplifier plate circuit, and RFC_1 provides a broad-band plate load for the mixer tube V_{2A}. The grid circuit of the amplifier also uses a slug-tuned coil and includes a trimmer capacitor, C_1, that permits peaking the input for the antenna in use, or in tuning completely across a band. A slug-cored coil is used at L_4 to facilitate resonating the circuit near the crystal frequency.

The frequency of the oscillator must differ from the frequency of the received signal by the frequency of the tunable i.f. amplifier. With the car broadcast receiver following the converter, the i.f. range will be from approximately 550 to 1550 kc. Since the tunable i.f. range is thus limited to a band 1000 kc. wide, the tuning range of the system with any single crystal will be restricted to 1 Mc. This is sufficient for all except the 28-Mc. band. Two crystals are required to cover the entire 10-meter band. The first of these gives a tuning range of 28 to 28.9 Mc. and the second permits tuning 28.8 to 29.7 Mc. An accompanying frequency chart lists the crystal frequencies and the ranges over which the broadcast receiver must be tuned to cover the amateur bands.

Construction

The input-tuning capacitor, C_1, the pilot lamp and the switch are in line across the panel of the converter as shown in Fig. 19-4. Each of these components is centered $\frac{3}{4}$ inch down from the top of the case and each is separated from the other in horizontal plane by $1\frac{3}{4}$ inches. The male jacks for the grid, plate and oscillator coils are below C_1, I_1 and S_1 in that order. Each jack is centered $1\frac{1}{8}$ inches up from the bottom of the cabinet.

The chassis, shown in Fig. 19-7, may be made of thin aluminum sheet and should be fastened to the side walls of the cabinet with homemade brackets, or angle stock. The sockets for V_1 (at the right as seen in the rear view) and V_2 are centered $1\frac{5}{8}$ inches in from the right and left edges of the chassis, respectively. J_3 is centered on the rear wall of the chassis with J_1 and J_2 to the right and left.

A bottom view of the converter clearly shows the components mounted below deck.

The exterior and the interior of the coil box are shown in Figs. 19-4 and 19-7. Wind the antenna coupling coils, L_1 in Fig. 19-5, around the ground ends of the grid coils before the latter are soldered in place. Wind the coupling coils rather snugly but not so tightly as to prevent adjustment of the coupling to L_2 during testing of the converter.

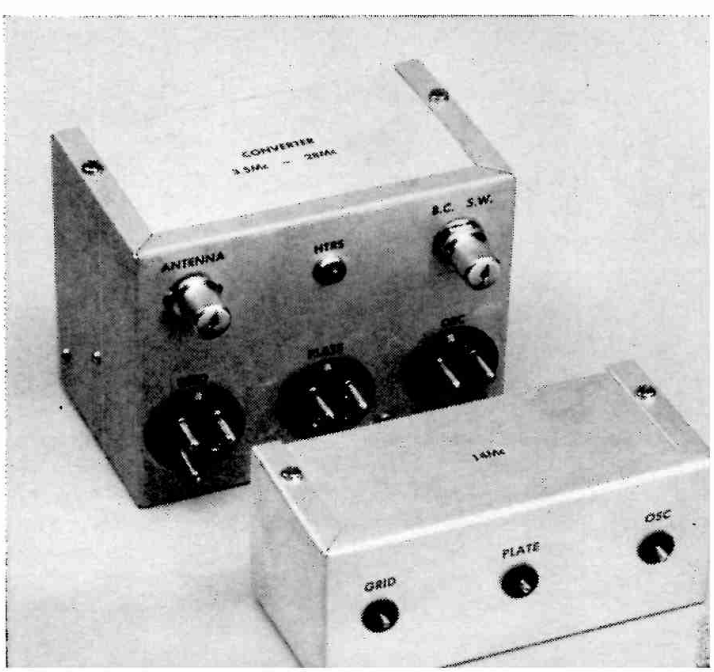


Fig. 19-4—The aluminum case for the converter measures 3 × 4 × 5 inches (Bud CU-3005 or Premier AMC-1005). Amphenol type 86-CP4 male jacks mounted on the front of the box mate with MIP 4-prong sockets mounted on the rear of the coil compartment shown in the foreground. Knobs for C_1 and S_1 are to the left and right, respectively, of the pilot lamp. The coil box measures 2¼ × 2¼ × 5 inches (Bud CU-3004 or Premier AMC-1004). Slug-adjustment screws for L_2, L_3 and L_4 protrude through rubber grommets mounted on the front wall of the plug-in coil assembly.

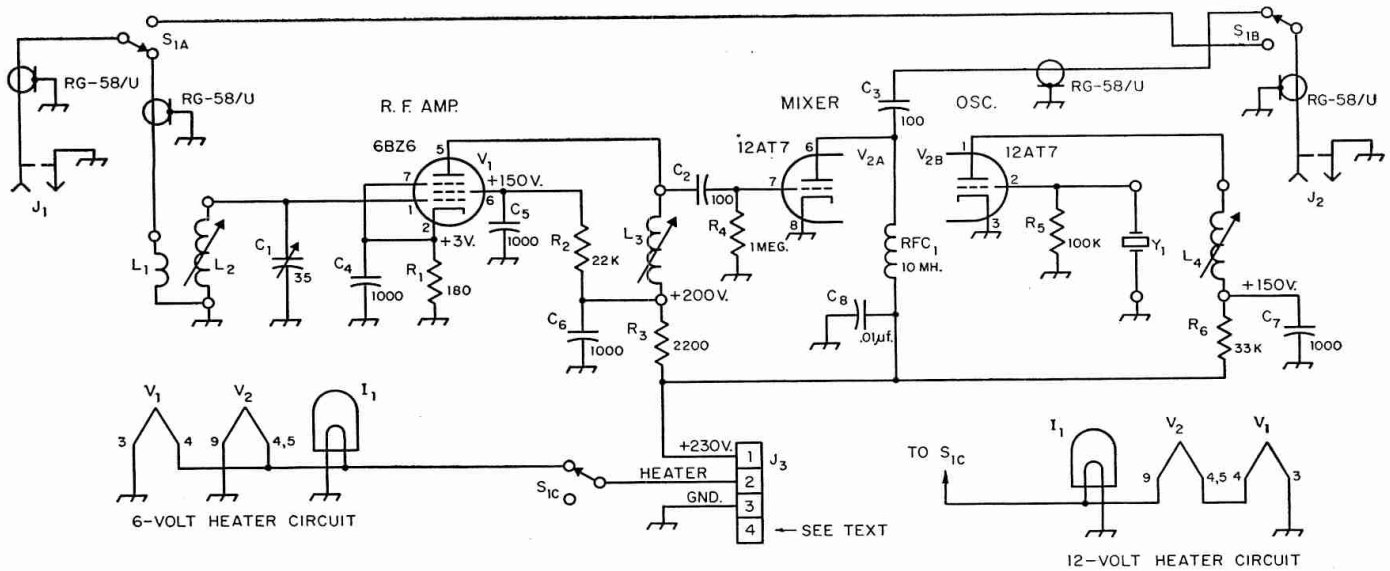


Fig. 19-5—Circuit diagram of the crystal-controlled mobile converter. Unless otherwise indicated, capacitances are in μμf., resistances are in ohms, resistors are ½ watt.

C_1—35-μμf. midget variable (Hammarlund MAPC-35-B).
C_2, C_3—100-μμf. ceramic tubular.
C_4, C_5, C_6, C_7—1000-μμf. disk ceramic.
C_8—0.01-μf. disk ceramic.
I_1—Pilot-light assembly [Johnson 147-503 with No. 44 (6-volt) or No. 1815 (12-volt) lamp].
J_1, J_2—Motorola-type shielded jack (ICA 2378).
J_3—4-prong male chassis connector (Cinch-Jones P-304-AB).
L_1, L_2, L_3, L_4—See coil chart.
R_1—180 ohms, ½ watt.
R_2—22,000 ohms, ½ watt.
R_3—2200 ohms, ½ watt.
R_4—1 megohm, ½ watt.
R_5—0.1 megohm, ½ watt.
R_6—33,000 ohms, ½ watt.
RFC_1—10-mh. r.f. choke (National R-100S).
S_1—3-pole 3-position (used as 3 p.d.t.) selector switch (Centralab PA-1007).
Y_1—See text and frequency chart (International Crystal type FA-9).

An a.c. transformer may be used for the filaments while testing the converter. The plate supply should deliver 20 milliamperes at 200 to 250 volts. A modulated-signal generator covering the bands for which the converter has been constructed is extremely helpful. To be most effective, the generator should have a 50-ohm output termination. A grid-dip meter for preliminary adjustment of the slug-tuned coils is useful, but not essential to alignment. If at all possible, the car receiver that is to be used as the tunable i.f. should be used during the testing.

Using coaxial-cable leads, connect the signal generator and the broadcast receiver to J_1 and J_2, respectively. Switch S_1 to the ham-band position, and apply heater power. The receiver need not be turned on at this time, and plate power for the converter does not have to be applied. Now, rotate C_1 to approximately half capacitance and then adjust L_2 to resonance (use the grid-dip meter as the indicator) at the low end of the band. Move the grid-dipper over to the plate circuit of the amplifier and peak L_3 at the center of the band. Next, couple the meter to L_4 of the oscillator and tune the coil to the frequency of the crystal in use.

After these initial adjustments, plate power may be applied to the converter and a frequency-indicating device used to detect oscillation of V_{2B}. If the grid-dip meter is the self-rectifying type it may be used for the check. An absorption-type wavemeter with indicator or a receiver tuned to the crystal frequency (with the b.f.o. on) may also be used for the purpose. In any

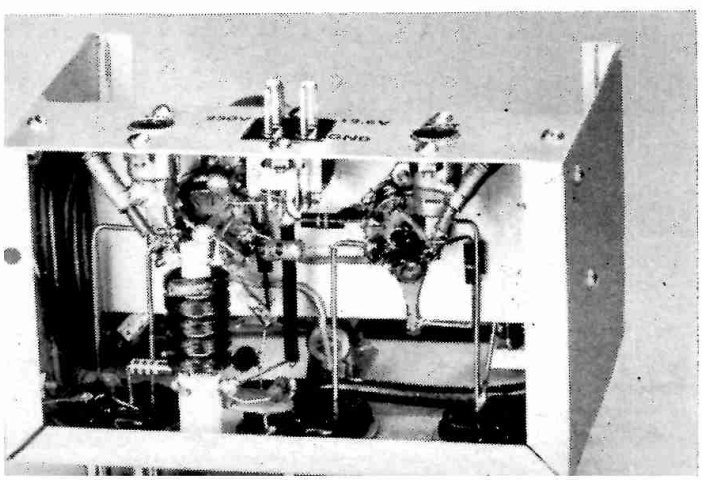


Fig. 19-6—A bottom view of the mobile converter. The amplifier tube socket at the right is mounted with Pin 7 facing toward the rear wall of the chassis. R_1 and R_2 are to the right and left of the socket, respectively. The socket for V_2 is mounted with Pins 4 and 5 facing toward the rear of the unit. C_2 is to the lower left of R_2, and RFC_1 is mounted on the front wall of the housing. C_7 and R_6 are to the left of the base of the choke. C_6, C_8 and R_3 are to the right of RFC_1. The output coupling capacitor C_3 is supported between Terminal 4 of J_3 and Pin 6 of the socket for V_2. R_4 and R_5 are partially visible to the right and left, respectively, of the V_2 socket.

Coil Chart for the Mobile Converter

Band	*Turns*	*Ind. Range, μh.*			*Type No.*		
Mc.	L_1	L_2	L_3	L_4	L_2	L_3	L_4
3.5–4	14	36–64	64–105	105–200	120-F	120-G	120-H
7–7.3	7	9–18	18–36	36–64	120-D	120-E	120-F
14–14.35	4	3–5	5–9	9–18	120-B	120-C	120-D
21–21.45	3	2–3	3–5	3–5	120-A	120-B	120-B
28–28.9	3	1–1.6	1.6–2.7	2.7–4.5	1000-A	1000-B	1000-C
28.8–29.7	3	1–1.6	1.6–2.7	2.7–4.5	1000-A	1000-B	1000-C

Note: L_1 is wound with No. 28 d.c.c. wire at grounded end of L_2. L_2, L_3 and L_4 are slug-tuned coils manufactured by North Hills Electric Co., Inc. (Mineola, L.I.)

event, L_4 should be tuned through resonance to the *high*-frequency side of the crystal frequency until the crystal oscillates reliably as indicated by rapid starting when plate power is turned on.

With the converter and the i.f. amplifier both turned on, and with the signal generator tuned to the center of the band, tune the receiver until the test signal is heard. Peak L_3 and L_4 for best response and then peak L_2 with C_1 set at half capacitance. The coupling between L_1 and L_2 may now be adjusted for optimum performance.

If the aforementioned test equipment is not available, the converter may be aligned while using a strong local of known frequency as the signal source. Of course, the signal frequency must be in the band for which the converter is to be aligned. In using this system, first set the broadcast receiver as closely as possible to the proper i.f. frequency (see the frequency chart) and then tune L_4 until the crystal oscillates. It is advisable to tune the receiver through a narrow range as the oscillator coil is being adjusted to assure that the test signal will be heard as soon as the crystal breaks into oscillation. After the signal is detected, the grid, plate and oscillator circuits may be adjusted for maximum over-all gain.

The mobile antenna should be resonant and tightly coupled to the converter. Traps for suppressing interference cause by strong local broadcast signals that feed in through the converter to the tunable i.f. have not been included in the converter because the need for them will be entirely dependent on local broadcast-station power and frequency assignments.

(Originally described in *QST*, Nov. 1957).

Frequency Chart for the Mobile Converter

Band Mc.	*Crystal Freq., Mc.*	*I.F. Range Kc.*
3.5–4	2.9	650–1100
7–7.3	6.4	600–900
14–14.35	13.4	600–950
21–21.45	20.4	600–1050
28–28.9	27.4	600–1500
28.8–29.7	28.2	600–1500

Note: I.f. range indicates broadcast receiver tuning range necessary for covering the associated amateur frequencies.

(For a description of a bandswitching crystal-controlled converter, see *QST*, January, 1955, or *The Mobile Manual for Radio Amateurs.*)

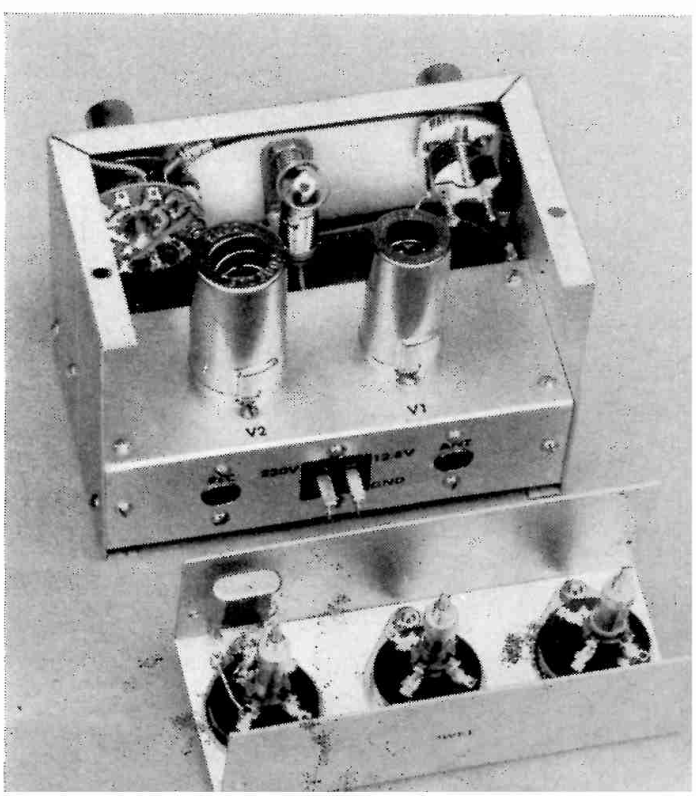


Fig. 19-7—Homemade L-shaped chassis, mounted on small brackets fastened to the side walls of the converter housing, is 4 15/16 inches long, 2 inches wide and 1½ inches deep. V_1 is mounted on the chassis to the right of V_2 as seen in this rear view. J_1, J_3 and J_2 are in line in that order from right to left across the rear wall of the chassis. An interior view of a coil compartment is shown in the foreground. Terminals of the coils are soldered directly to the socket terminals. Notice that the crystal for the oscillator is mounted adjacent to L_4.

Transistor Mobile Converter

The crystal-controlled converter shown in Fig. 19-8 is a compact, fixed-tuned converter which exhibits excellent performance when used with the automobile receiver. It is designed for one-band operation but may be constructed for any amateur band between 80 and 10 meters.

All of the components, including the power supply for the converter, are housed in a 5¼ × 3 × 2⅛-inch Minibox that can be mounted under the dashboard of the car. The unit is built in one half of the box so that it may be "dropped" for servicing or adjustment while the other half remains mounted to the dash.

Only two external connections to the converter are necessary. A coax lead from the antenna must go to the antenna input of the unit, and an output coax connection to the car radio.

The circuit for the converter is shown in Fig. 19-9. The oscillator circuit is a transistorized version of the triode Pierce. Injection for the mixer is taken from a small link wound over the cold end of the collector tank coil. The emitter of the mixer transistor is returned to ground through this link. The mixer circuit corresponds to a triode vacuum-tube mixer utilizing cathode injection from the oscillator, the major difference being the low input impedance of the transistor base as compared with the relatively high input impedance of a vacuum-tube grid.

The crystal frequency used in the oscillator portion of the converter is given in the tuned circuit data table. On 30 and 21 Mc., the crystal is operated at its third overtone and on the lower bands the fundamental mode is used.

The inductances are wound on slug-tuned forms and shunted with the capacitances shown in the tuned circuit data table.

The circuit shows a crystal diode connected from the high impedance end of L_1 to cell B_2. This gives a measure of protection for the mixer transistor in the event that an excessive amount of r.f. energy is introduced into the converter. When a signal greater in voltage than B_2 appears across L_1, the diode will conduct and short the excess r.f. to ground.

Power Supply

The converter requires about 8 volts d.c. for operation and takes on the order of 3 ma. of current. A built-in battery supply serves two important purposes. First, it eliminates one of the prime sources of ignition interference, since various noises from the electrical system of the car can be carried into the converter via the leads from the car battery. Also, with a self-

TUNED CIRCUIT DATA FOR THE TRANSISTOR CONVERTER

Band	Coil	C_1 μμf.	C_2 μμf.	Crystal Freq.	I.F. Range
28 Mc.*	L_1, 12 turns No. 20 enam. Tap at 4th turn. L_2, 2 turns No. 20 enam. L_3, 12 turns No. 20 enam. L_4, 2 turns No. 24 enam.	15	15	27.85 Mc.	650–1600 kc.
21 Mc.	L_1, 15 turns No. 20 enam. Tap at 5th turn. L_2, 3 turns No. 20 enam. L_3, 15 turns No. 20 enam. L_4, 2 turns No. 24 enam.	15	15	20.35 Mc.	650–1100 kc.
14 Mc.	L_1, 23 turns No. 24 enam. Tap at 6th turn. L_2, 5 turns No. 24 enam. L_3, 26 turns No. 24 enam. L_4, 3 turns No. 24 enam.	15	15	13.35 Mc.	650–1000 kc.
7 Mc.	L_1, 35 turns No. 28 enam. Tap at 10th turn. L_2, 6 turns No. 28 enam. L_3, 40 turns No. 28 enam. L_4, 4 turns No. 28 enam.	33	33	6350 kc.	650–950 kc.
4 Mc.	L_1, 52 turns No. 34 enam. Tap at 13th turn. L_2, 8 turns No. 34 enam. L_3, 72 turns No. 34 enam. L_4, 5 turns No. 34 enam.	40	40	2850 kc.	650–1150 kc.

* 28.5 to 29.45 Mc.

All coils close-wound on ½-inch diam. slug-tuned (iron slug) forms. Tap on L_1 to be made near cold end of coil. L_2 wound over cold end of L_1.

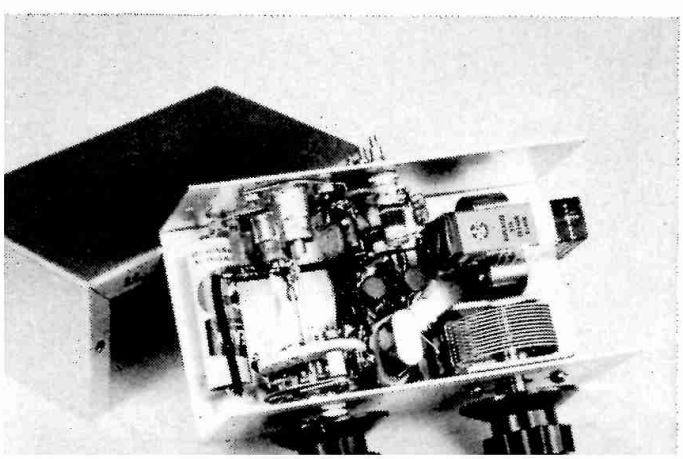


Fig. 19-8—View of the transistorized converter. The variable output capacitor C_4 is mounted on the right front panel. Directly behind C_4 is the 8.4 volt mercury battery B_1 held in place by a bracket which is sold in most hardware stores as a broom holder. The two transistors are the round black objects in the center. They are supported by their own leads which are soldered to tie points. The converter shown here operates on 10 meters.

Transistor Mobile Converter

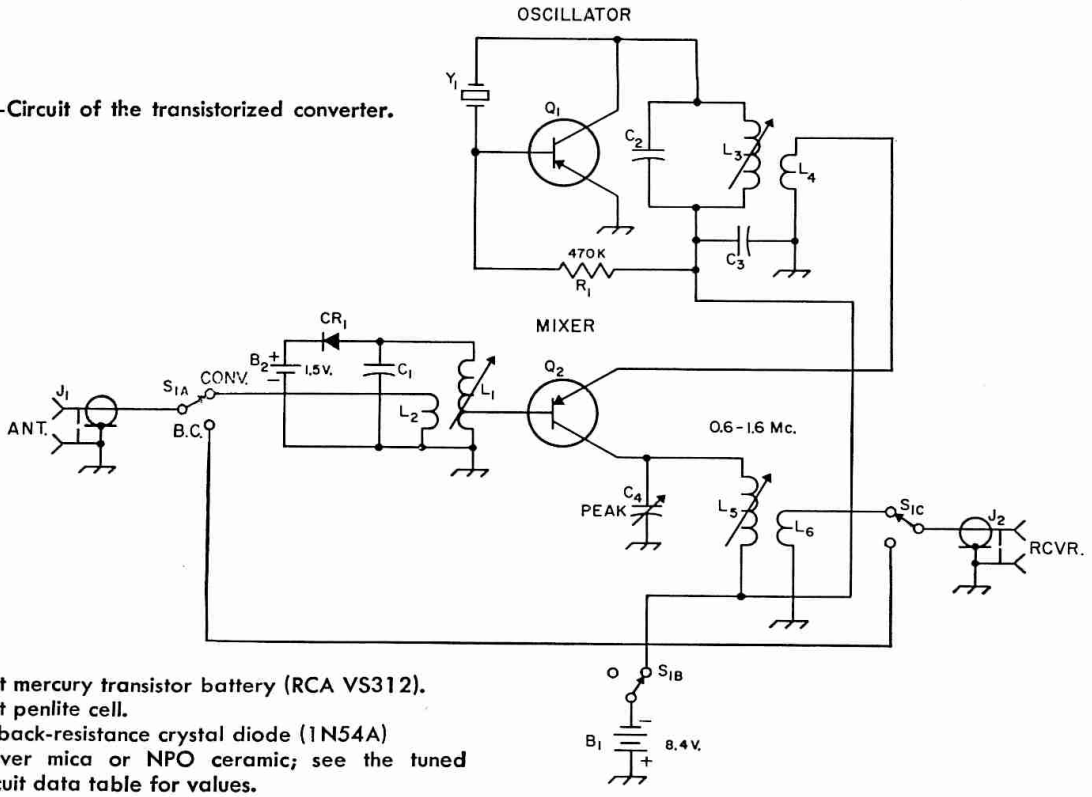


Fig. 19-9—Circuit of the transistorized converter.

B_1—8.4 volt mercury transistor battery (RCA VS312).
B_2—1.5 volt penlite cell.
CR_1—High back-resistance crystal diode (1N54A)
C_1, C_2—Silver mica or NPO ceramic; see the tuned circuit data table for values.
C_3—.005 μf. ceramic.
C_4—365 $\mu\mu$f. variable capacitor (Allied Radio Co. 61-H-009).
J_1, J_2—Automobile type antenna connectors.
L_1, L_4, inc.—See coil table.
L_5—320-500 μh. slug tuned coil (Miller 4514).
L_6—10 turns No. 30 enam. close-wound over L_5.
Q_1, Q_2—2N247 transistors.
R_1—0.47 megohms, ½ watt (value may require slight adjustment for individual transistors).
S_1—Three pole two position rotary switch (Centralab PA-2007).
Y_1—Crystal (International Crystal Co. type FA-5 for miniature socket, FA-9 for standard socket). See table for frequencies.

contained battery it is unnecessary to make any power-supply connections either to the car receiver or car battery. This saves considerable time during installation and makes the unit readily adaptable to portable operation.

Wiring

No. 30 wire is adequate for wiring because of the small current and voltage requirements of the converter. Spaghetti should be used over exposed leads that might come in contact with other parts because of the vibration that occurs in mobile operation. For the same reason, it is essential that good soldered connections be made.

The information given in the tuned-circuit data table applies to ½-inch coil forms. Ready-wound slug-tuned coils, such as the Miller 4500 series or the CTC LS3 series, can also be used with the links shown in the chart. L_1 is tapped about ⅓ up from the cold end. C_1 and C_2 should be chosen to resonate, in a given amateur band, with the inductance of the particular coil used; the L/C ratio is not critical.

Construction

The converter is assembled in one half of a 5¼ × 3 × 2⅛-inch Minibox. The box-cover (with the lips) is mounted permanently under the automobile dash. The only front-panel controls are the converter-broadcast switch S_1 and the output peaking control C_4. Mount S_1 so that the leads coming from the antenna connectors will line up with the proper switch terminals. Two 5-terminal tie points are mounted in the center of the chassis for supporting the crystal socket, transistors and other small components. The three slug-tuned inductances are supported on the rear wall of the chassis, as are the two antenna connectors.

After the major components have been installed, only a few wiring connections remain. Be sure to leave long leads on the inductances after winding them so that the leads may be directly connected to their proper points.

In the circuit, cell B_2 has its negative terminal grounded. A lug soldered to the cell case and bolted to the chassis will make a sturdy support for the cell.

Adjustment and Testing

After the unit is wired, the first test should be to make certain that the oscillator is functioning. Turn on the converter. Tune a communications receiver to the crystal frequency and adjust the slug in L_3 until the signal is heard. The oscillator will not function unless the collector tank (C_2L_3) is resonant.

After the oscillator is operating properly, install the unit in the car and turn it on. With the broadcast radio turned on, adjust the slug in L_3 for maximum background noise. Next, adjust the slug in L_1 for maximum noise, or select a weak signal and peak it up for maximum gain. Then set the car radio at the high end of the i.f. band and adjust the slug in L_5 for maximum gain with C_4 at minimum capacity. The low end of the i.f. band should peak when C_4 is set near maximum. If only one segment of a particular band is going to be used, additional gain can be had by peaking the coils for that portion of the band. If, for example, 75-meter phone operation is desired, peak the converter for 3800 to 4000 kc. rather than 3500 to 4000 kc.

Crystal-Controlled Converters for 50 and 144 Mc.

The mobile converters shown in Figs. 19-10 through 19-13 combine simplicity with good v.h.f. design practice. Although only two tubes are used in each, the converters include a stage of r.f. amplification plus crystal-controlled oscillators. Ten meters was chosen as the i.f. because when the broadcast receiver is used as the tunable i.f. for v.h.f. converters images are a problem, and only 1 Mc. at a time could be tuned. The converters described here, therefore, are designed to work into a 10-meter converter or receiver. This can be a tunable converter which in turn works into the broadcast receiver, or a complete self-contained 10-meter receiver.

The 50 Mc. Unit

The circuit diagram for the 50-Mc. unit is shown in Fig. 19-11. A 6AK5 is used as an r.f. amplifier. The same gain with lower noise can be obtained with a cascode-type dual-triode amplifier, but the performance of this pentode stage is satisfactory and its design is considerably simpler than the triode amplifier.

The crystal oscillator makes use of a 22-Mc. overtone crystal. A crystal on the required injection frequency eliminates the need for multiplier stages, and makes possible the use of a simple oscillator circuit. The 10-meter receiver or converter is tuned from 28 to 30 Mc. in covering 50 to 52 Mc. If a general coverage receiver covering 26 to 30 Mc. is used, a 24-Mc. crystal in the oscillator will allow tuning 50 to 54 Mc. However, any injection frequency may be used to cover a desired portion of the band.

The pentode half of the 6U8 tube is used as a mixer. The oscillator and mixer sections are in the same tube envelope so there is enough stray coupling between the two for adequate oscillator injection.

The diagram shows the heaters connected for 12 volts. If 6-volt operation is desired, the heaters are connected in parallel and R_1 is disregarded.

The converters are built in a $5\frac{1}{4} \times 3 \times 2\frac{1}{8}$-inch Minibox. All of the parts are mounted on the bottom half of the box while the upper half (the one with lips) is fastened under the car dash. The bottom half containing all the components can be slid in and out for easy servicing.

Fig. 19-10 shows the placement of most of the components. The output peaking control C_1 and switch S_1 are mounted on one side of the chassis to form the front panel. The tubes, slug-tuned inductances, crystal socket and antenna connectors are mounted directly opposite on the back wall. Two tie-points are bolted to the base of the box for connecting and supporting leads and components. When wiring, make the r.f. leads as short and direct as possible.

The 144-Mc. Unit

The circuit diagram for the 144-Mc. converter is shown in Fig. 19-13. Two 6U8 tubes are used with the pentode section of one tube acting as the r.f. amplifier followed by the triode-section mixer. The other 6U8 is used as an overtone crystal oscillator and pentode frequency multiplier. By combining all the features of a 4-tube crystal-controlled converter in a two-tube model space-saving simplicity is achieved.

The same basic circuit used in the 50-Mc.

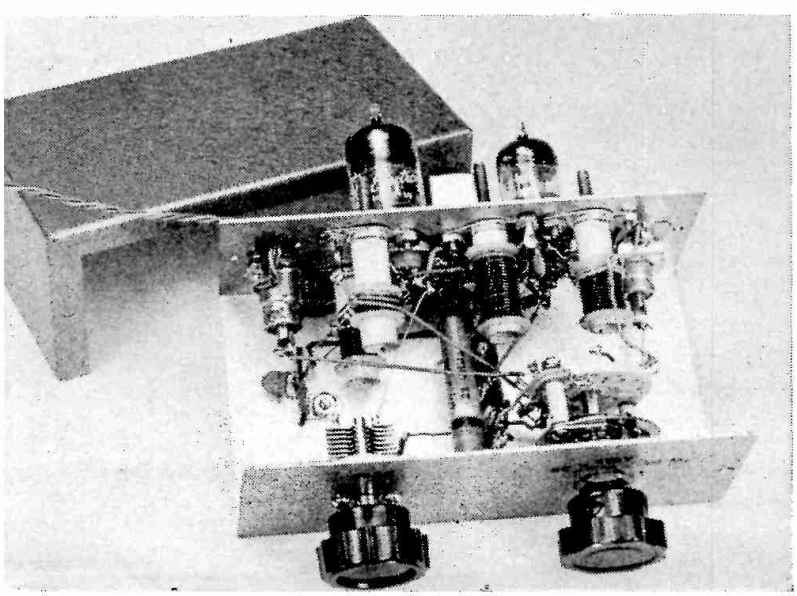


Fig. 19-10—View of the 50-Mc. converter. The inductances are from left to right: (bottom) L_7, (top) L_5L_6, L_3L_4, L_1L_2. The top of crystal Y_1 can be seen between the tubes. The 22-ohm 2-watt resistor in the center of the chassis is the heater current compensating resistor, used for 12-volt operation. Input and output antenna connectors are mounted on opposite ends of the back wall. Power is fed to the unit through the twisted power cable running in from the left side of the photograph.

Crystal-Controlled Converters

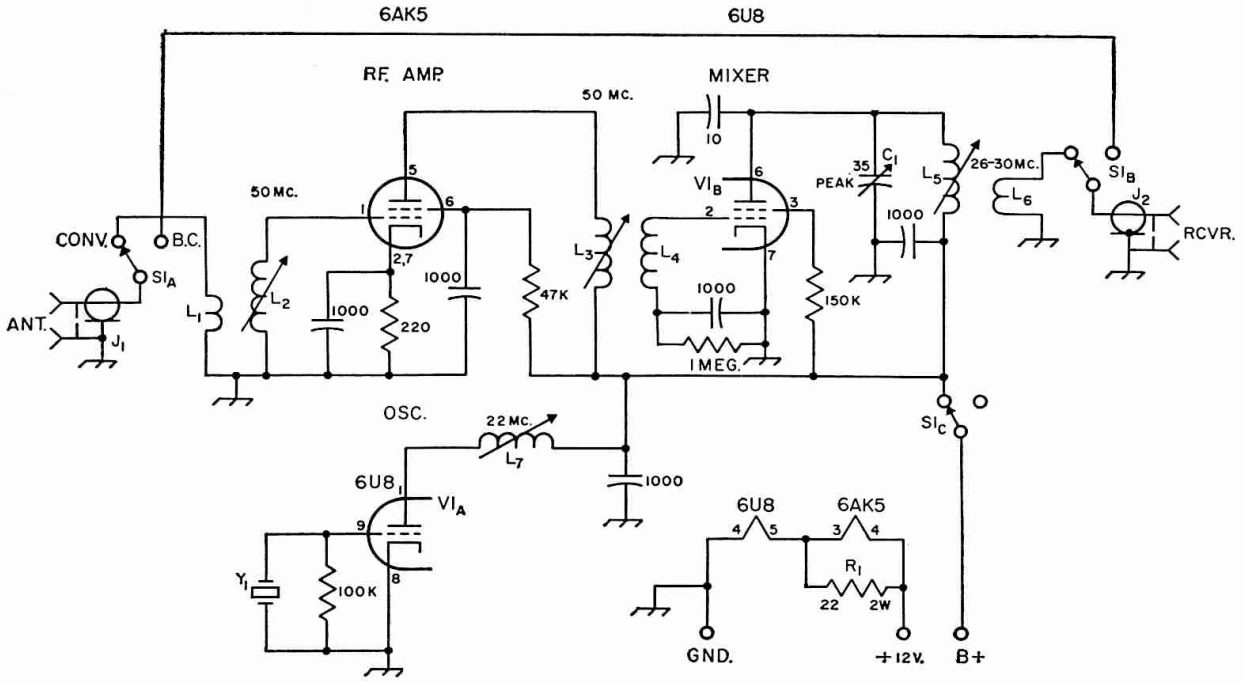


Fig. 19-11—Schematic diagram for the 50-Mc. mobile converter. All resistors ½ watt unless otherwise specified. Capacitor values below 0.001 μf. are in $\mu\mu$f. All 0.001 μf. capacitors are disk ceramic. Other fixed capacitors are tubular ceramic.

C_1—35-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-35-B).
J_1, J_2—Automobile type antenna connectors.
L_1—3 turns No. 20 insulated wire, close-wound over cold end of L_2.
L_2—9 turns No. 20 enam. wire, close-wound on ½ inch slug tuned coil.
L_3—16 turns No. 20 enam. wire, close-wound on ½ inch slug tuned coil form.
L_4—6 turns No. 20 insulated wire, close-wound over cold end of L_3.
L_5—14 turns No. 20 enam. wire, close-wound on ½ inch slug tuned coil form.
L_6—2 turns No. 20 insulated wire, close-wound over cold end of L_2.
L_7—28 turns No. 30 enam. wire, close-wound on ½ inch slug tuned coil.
R_1—22-ohm 2-watt resistor (used for 12-volt heater operation only).
S_1—Three-pole two-position rotary switch (Centralab PA-2007).
Y_1—22 Mc. overtone crystal. (International Crystal type FA-5 for miniature socket, FA-9 for standard socket).

model is followed in the 144-Mc. unit except for the addition of a multiplier stage following the crystal oscillator. The oscillator operates at 38.666 Mc. and is multiplied to 116 Mc. in the tripler stage. As in the 50-Mc. converter, this unit is designed to work into a 10-meter receiver or converter. If the i.f. tunes from 27 to 30 Mc., the converter will tune from 144 to 147 Mc. However, any segment of the band may be tuned by choosing the proper crystal frequency.

Unlike the 50-Mc. converter, the oscillator-multiplier stages of the 144-Mc. converter are physically separated from the mixer stage. It is necessary, therefore, to couple the 116-Mc. energy from the multiplier stage to the grid of the mixer. Capacitor C_2 is used for this purpose. It consists of a pair of twisted hook-up wires with one end of one lead connected to the mixer

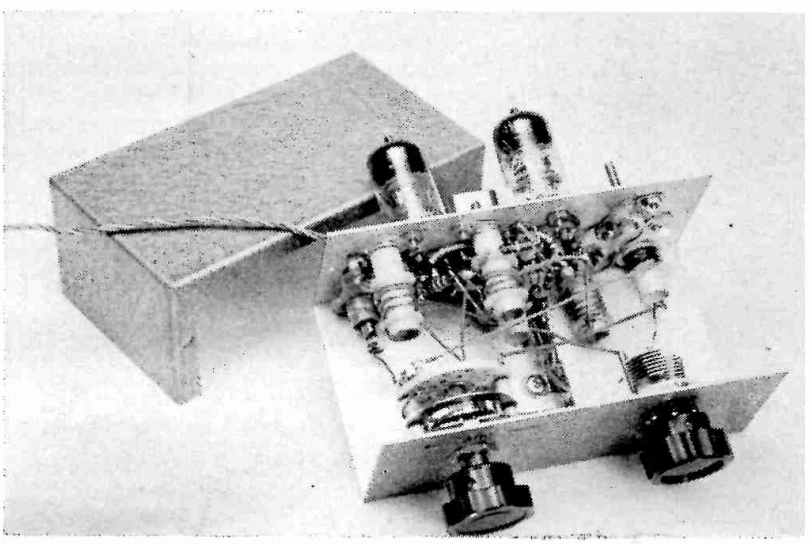


Fig. 19-12—View of the 144-Mc. converter. The inductances from left to right are: (top) L_1L_2, L_3L_4, L_5L_6, (bottom) L_7 and L_8. All components except S_1 and C_1 are mounted on the back wall of the chassis. A single tie point in the bottom of the channel supports various leads and provides junctions for sundry connections. The input and output antenna connectors are placed near the bottom right and left of the back panel. The crystal Y_1 is between the two tubes. Converter power is fed through the twisted cable which passes through a hole and grommet in the back wall of the chassis.

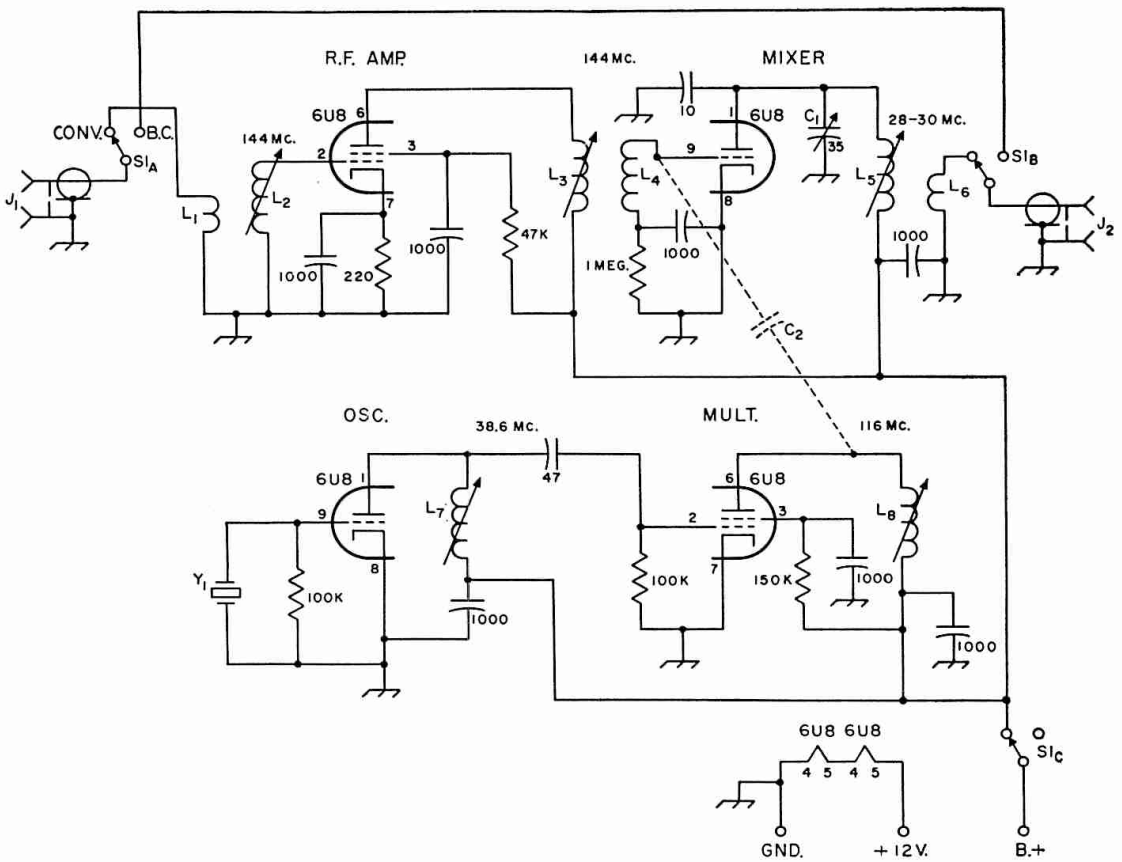


Fig. 19-13—Schematic diagram for the 144-Mc. converter. All resistors ½ watt unless otherwise specified. Capacitor values below 0.001 μf. are in $\mu\mu$f. All 1000-$\mu\mu$f. capacitors are disk ceramic. Other fixed capacitors are tubular ceramic.

C_1—35-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-35-B).

C_2—Oscillator injection capacitor (see text).

J_1, J_2—Automobile type antenna connectors.

L_1—2 turns No. 18 enam., ⅜ inches long, on ½ inch slug tuned coil form.

L_2—2 turns No. 20 insulated wire, close wound over cold end of L_1.

L_3—2 turns No. 18 enam., ⅜ inches long, on ½ inch slug tuned coil form.

L_4—2 turns No. 20 insulated wire, close wound over cold end of L_1.

L_5—9 turns No. 24 enam., close wound on ½ inch slug tuned coil form.

L_6—2 turns No. 20 insulated wire, close wound over cold end of L_5.

L_7—10 turns No. 24 enam., close wound on ½ inch slug tuned coil form.

L_8—5 turns No. 18 enam., ½ inches long, on ½ inch slug tuned coil form.

S_1—Three-pole two-position rotary switch (Centralab PA-2007).

Y_1—38.666 Mc. overtone crystal (International Crystal Co. type FA-5 for miniature socket, FA-9 for standard socket).

grid and the end of the other lead connected to the multiplier plate.

The circuit diagram shows the heaters connected for 12-volt operation. For 6 volts, the heaters should be connected in parallel.

The same basic outline of construction used in the 50-Mc. converter is followed in the 144-Mc. unit. Fig. 19-12 shows how output peaking control C_1 and the control switch S_1 are mounted on the front wall of the chassis while most of the remaining parts are secured to the rear surface. A single tie point is mounted on the bottom of the chassis for connecting and supporting various leads. The input and output antenna connectors are mounted at opposite ends of the back wall of the chassis.

Testing the Converters

The 50-Mc. converter requires 0.625 ampere at 6 volts (or 0.45 ampere at 12 volts) for the heaters, and approximately 17 ma. at 150 volts for the plate supply. If the car radio delivers in excess of 180 volts, the plate voltage on the converter should be limited by a dropping resistor.

The 144-Mc. converter requires 0.9 ampere at 6 volts (or 0.45 ampere at 12 volts) for the heaters. A plate voltage of 150 volts is required at about 30 ma.

All tuned circuits should be checked for resonance with a grid-dipper. The proper frequency for each circuit is given in Figs. 19-11 and 19-13. Apply power to the converter under test, and adjust the oscillator circuit until it goes into oscillation. This can be confirmed by tuning the home receiver to the oscillator frequency. Tune the oscillator inductance until the maximum oscillator signal is obtained. Now feed a 50 or 144-Mc. signal into the converter under test. This signal may come from a signal generator

or a grid-dip meter, or may be an actual signal from the antenna. Go through the converter stage by stage, adjusting the inductances for peak output. After the first run of peaking is completed the converter should be spot-checked through the entire band to make sure the over-all response is fairly flat. Output capacitor C_1 is used to peak the output circuit. L_5 is adjusted so that C_1 peaks at mid-capacitance in the center of the i.f. tuning range.

A 20-Watt High-Frequency Mobile Transmitter

Figures 19-14 through 19-17 illustrate a complete 20-watt transmitter that may be operated on any band from 80 to 10 meters. The design avoids the complication, expense and difficult construction associated with the average multiband transmitter, but does not confine its application to any one band. Changing from one band to another as operating interest varies is a simple matter of unsoldering a pair of readily-accessible coils and replacing them with others for the new band.

Circuits

The circuit of the transmitter is shown in Fig. 19-15. A 5763 crystal oscillator drives a 2E26 final amplifier. Quadrupling frequency in the output of the grid-plate oscillator from a 7-Mc. crystal will provide adequate drive for the final on 10 meters. Sufficient capacitance is provided in the plate tank of the 2E26 for a Q of 10 or more on all bands except 80 meters. On 80 meters, the tank Q will drop to about 6, but there is little danger of appreciable harmonic output when feeding a high-Q antenna such as the usual loaded whip. Adequate output coupling on this band is assured by tuning the output link line. Parallel plate feed is used in both stages.

The audio circuit is equally simple. One triode unit of a 12AU7 is used as a grounded-grid amplifier. This provides low-impedance input for a carbon microphone without the need for a microphone transformer. The second triode unit of the 12AU7 is used in conventional fashion to drive a 1635 Class B modulator. This tube operates at zero bias with an idling current of only 10 ma. D. c. voltage for operating the carbon microphone is obtained by connecting the microphone in series with the two speech-amplifier cathodes and ground.

The 1-ma. meter M_1 may be switched across appropriate multiplier shunts to read amplifier grid or plate current, or modulator plate current. A d.p.d.t. change-over relay, K_1, actuated by the microphone push-to-talk switch, is also provided. One pole shifts the antenna from receiver to transmitter, while the other mutes the receiver by shorting the voice coil of the speaker. S_1 removes screen voltage from the 2E26 and disables the relay so that the oscillator may be tuned up before the amplifier is put on the air.

Construction

A 5 × 6 × 9-inch steel utility box (Middletown Mfg. Co., Middletown, Conn.) is used as the cabinet for the transmitter. The chassis is bent up from aluminum sheet approximately 1/16 inch thick. The chassis is 8¾ inches wide, 6 inches deep and has 2-inch lips along the front and rear edges.

C_3 and C_4 are mounted on the front wall of the partition with their shaft centers 1⅜ inches above the chassis. The shaft of C_4 is centered 1¼ inches from the open edge of the shield, while the shaft of C_3 is centered 3 inches in. The shafts of these capacitors are connected to panel-bearing units by rigid metal shaft couplers.

The socket for the 2E26 is submounted on ¾-inch spacers, beneath a 1¼-inch clearance hole centered 1 inch from the rear edge of the chassis and 2 inches in from the side. RFC_4 is mounted horizontally from the front wall of the partition, below and between C_3 and C_4.

The output tank coil, L_2, is cemented to a 1-inch cone insulator and soldered between a rear stator terminal of C_3 and a grounding lug on the chassis. The bottom end of L_3 is connected to a rear stator terminal of C_4, while the other end goes through a small feed-through point in the chassis to a relay terminal immediately below. The 5763 is centered between the partition and the front panel, and between the shafts of C_3 and C_4.

Fig. 19-17 shows the modulation transformer in the upper right-hand corner of the chassis. The secondary taps of T_2 should be set for 7500 ohms. The 12AU7 and 1635 sockets are centered

Fig. 19-14—A panel-illuminating lamp is mounted to the right of the meter, along with the amplifier-tank and antenna-link tuning controls. Along the bottom, from left to right, are the microphone jack, meter switch, filament switch, tune-operate switch, oscillator tuning control and the crystal.

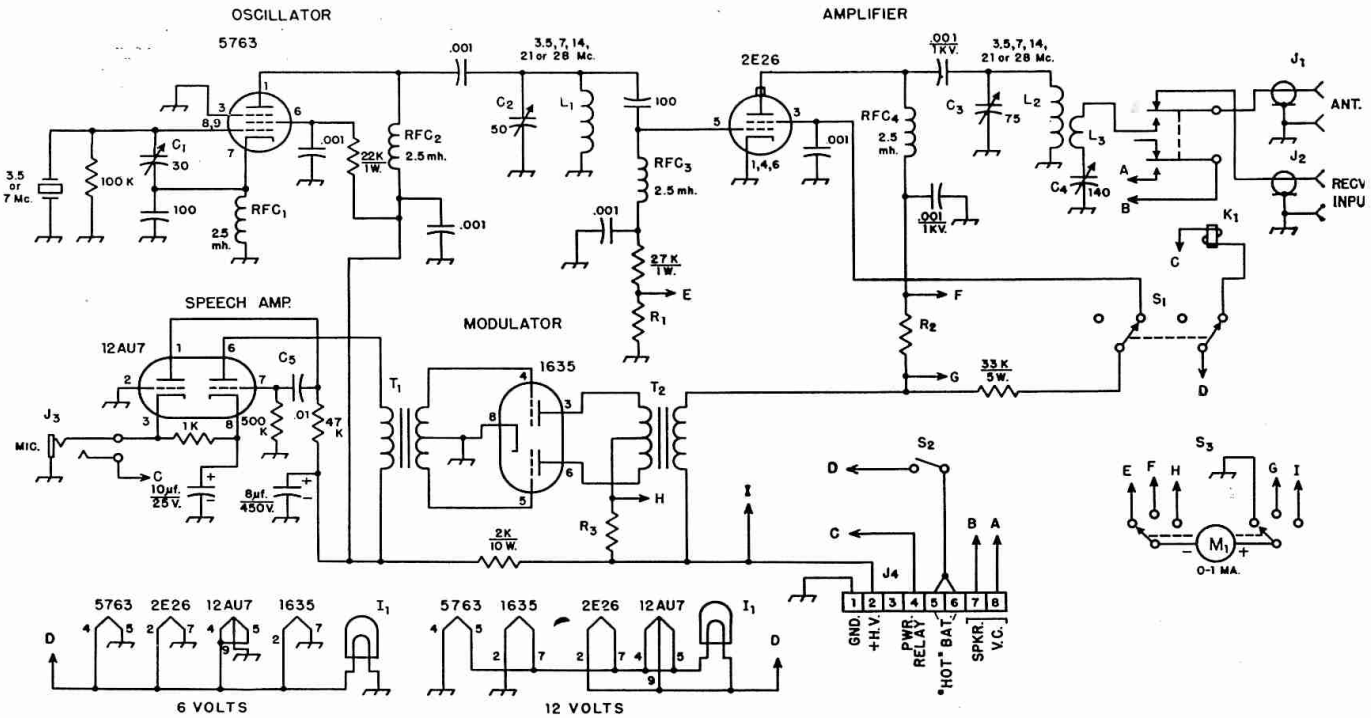


Fig. 19-15—Circuit of the single-band mobile transmitter. All resistors are ½ watt unless otherwise specified. All capacitances less than 0.001 μf. are in $\mu\mu$f. All 0.001-μf. capacitors are disk ceramic. Fixed capacitors of smaller value may be mica or NP0 ceramic. Capacitors marked with polarity are electrolytic.

C_1—Mica or ceramic trimmer.
C_2—Air variable (Hammarlund HF-50).
C_3—Air variable (Johnson 167-4).
C_4—Air variable (Hammarlund HF-140).
C_5—Paper ceramic.
I_1—6.3-volt 250-ma. dial lamp.
J_1, J_2—Coaxial connector (SO-239).
J_3—Push-to-talk microphone jack.
J_4—Power connector (octal tube socket).
K_1—D.p.d.t. 6-volt or 12-volt d.c. relay (Guardian Series 200).
L_1, L_2, L_3—See coil table.
M_1—0–1 d.c. milliammeter, 2⅜-in. (Triplett 227-T).
R_1—10-times shunt for M_1 (6.1 ohms for 55-ohm meter.)
R_2, R_3—100-times shunt for M_1. (0.5 ohm for 55-ohm meter.)
S_1—D.p.d.t. rotary switch (Centralab PA-1002).
S_2—S.p.s.t. toggle switch.
S_3—2-pole 3-position rotary switch (Centralab PA-1003).
T_1—Driver transformer, 2.5:1 primary to ½ secondary (Merit A-2920).
T_2—10-watt modulation transformer (Merit A-3008).

on a line about halfway between the rear of the meter and the modulation transformer. The socket for the 12AU7 is centered ⅞ inch from the end of the chassis. Then the socket for the 1635 is spaced sufficiently from the 12AU7 socket so that the driver transformer, T_1, can be mounted between the two sockets, underneath the chassis.

The two coaxial connectors, J_1 and J_2, are mounted on the rear lip of the chassis, spaced to avoid the 2E26 socket. An octal socket serves as the power-supply connector J_4, and the changeover relay is centered between this socket and the nearest coaxial connector.

Testing

The unit will operate from any supply delivering 300 to 400 volts at 125 ma. or more.

While the 2E26 might be used as a doubler if necessary, straight-through operation is recommended. Crystals in the 80-meter band will provide adequate drive for the final on all bands up to and including the 14-Mc. band. Crystals in the 7-Mc. band are needed for 21- and 28-Mc. output. Coils should be selected from the coil table to suit the band desired.

The oscillator is adjusted with S_1 in the tune position, and the meter switch turned to read amplifier grid current. With power supplied, C_2

Table of Coil Dimensions

Band	*Lμh.*	*Turns*	*Diam. In.*	*Length In.*	*Wire Size*	*B&W No.*	*Airdux No.*
			L_1				
80	29	44	1	1⅜	24	3016	832
40	6.3	28	⅝	⅞	24	3008	532
20	2.8	16	⅝	1	20	3007	516
15	0.9	9	⅝	9/16	20	3007	516
10	0.5	6	⅝	⅜	20	3007	516
			L_2				
80	32	80	¾	2½	24	3012	632
40	8	41	¾	2½	20	3011	616
20	3.5	20	¾	1¼	20	3011	616
15	1.6	16	¾	2	18	3010	608
10	1.1	12	¾	1½	18	3010	608

L_3 — 3 turns No. 20, 1-inch diam., 3/16 inch long, over ground end of L_2 (B&W 3015, Airdux 816) for 80, 40 and 20 meters; 2 similar turns for 15 and 10 meters.

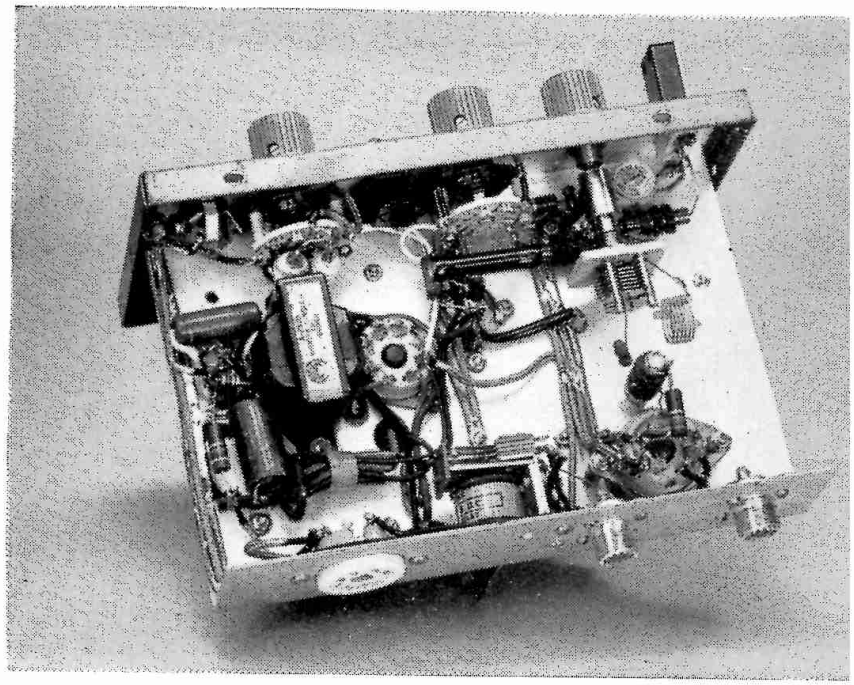


Fig. 19-16—Bottom view of the 20-watt mobile transmitter. The driver transformer is placed between the two audio-tube sockets. Along the front lip of the chassis, from left to right, are the microphone jack, meter switch, filament switch S_2, tune-up switch S_1, oscillator tank capacitor C_2 and the crystal socket. C_2 is spaced back of the panel, and mounted behind the 5763 socket. L_1 is soldered across the terminals of the capacitor. All power and control wiring is done with shielded wire.

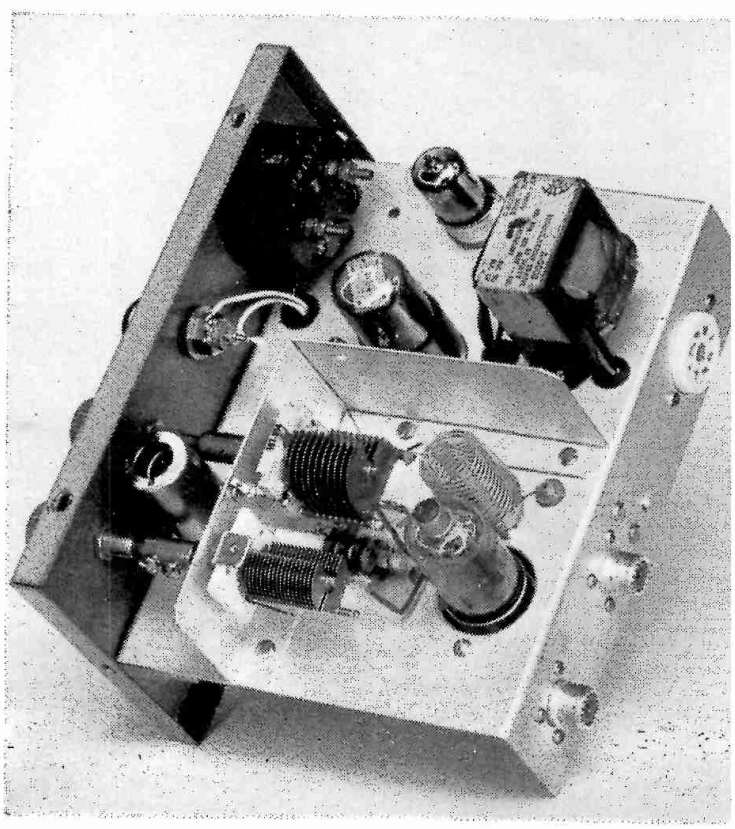
should be adjusted for maximum grid current. The tuning should be checked with a wave-meter to make sure that the oscillator output circuit is tuned to the desired frequency. Then C_1 should be adjusted for maximum grid current. The reading should be at least 3 or 4 ma.

A pair of G.E. type 1820, 28-volt, 1-amp, miniature lamps connected in series makes a good dummy load for testing the final. With S_1 thrown to the operate position, the meter switched to read 2E26 plate current, and power applied, adjust C_3 for a dip in plate current. Check the frequency with a wavemeter coupled to the output tank. Then adjust C_4 until the meter reads 50 ma. Retune C_3 for the plate-current dip. It may take a little juggling back and forth between C_3 and C_4 before an adjustment is reached where the meter reads 50 ma. at the plate-current dip. The load lamps will not light to full brilliance, but it should be possible to determine the adjustment that gives maximum output. With the amplifier fully loaded, the grid current should still remain at 3 to 4 ma.

The meter should now be turned to read modulator plate current. Without voice, the meter should read about 10 ma. When speaking into the microphone, a kick of the meter reading up to 40 or 50 ma. on peaks should indicate 100 per cent modulation. The r.f. amplifier plate current should remain essentially steady under modulation, but the lamps in the dummy load should show some increase in brilliance.

Adjustment when an antenna is substituted for the dummy load should be done in a similar manner. The antenna must, of course, be checked for resonance in advance with a g.d.o. or by other means. (Originally described in *QST*, Jan., 1957). (For a description of a bandswitching mobile transmitter with v.f.o., see *QST*, August and Sept., 1957).

Fig. 19-17—Interior view of the single-band mobile transmitter. The output components are separated from the other components by an L-shaped aluminum partition which measures 4½ inches along the front and 4 inches along the side. It is 2¼ inches high with ½-inch lips along the bottom edges for fastening to the chassis.



Mobile Transmitters for 50 and 144 Mc.

Figs. 19-18 through 19-23 show circuits and constructional details of compact transmitters covering the 6- and 2-meter bands. The units are only 3 inches deep and therefore are suitable for under-the-dash mounting.

Output on 50-Mc. is obtained by using crystals in the 50-Mc. range. This eliminates any necessity for multiplier stages and greatly simplifies the circuit. In the two-meter unit, a 48-Mc. crystal is used which is multiplied to 144 Mc. by a tripler stage.

Although the r.f. amplifier used in the transmitters will operate at higher voltages, the units are designed primarily to work from a 300-volt, 100-ma. supply. A transistor modulator can be used with the units with a saving in total current drain.

The 50-Mc. Unit

The circuit of the 50-Mc. transmitter is shown in Fig. 19-20. A 5763 (6417 when using 12-volt heaters) is triode-connected in an overtone-type crystal oscillator. Feedback winding L_2 helps to sustain 3rd-overtone oscillation and may require some slight adjustment for optimum output in its placement with respect to L_1. The 50-Mc. signal from the oscillator is capacitively coupled to the grid of the 2E26 (6893 when using 12-volt heaters) amplifier. A jack J_1 on the rear of the transmitter allows the grid current to be measured.

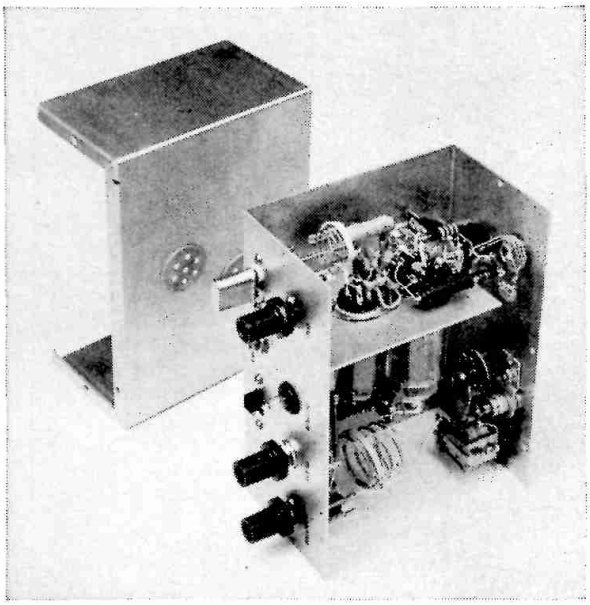


Fig. 19-19—The 50 Mc. mobile transmitter is built into a 7 × 5 × 3-inch aluminum Minibox (Bud CU-3008). Oscillator coil L_1L_2 is near the top left. The jack on the right rear panel is the grid-current meter jack. One-inch holes are punched in both halves of the Minibox for ventilation. Perforated hole plugs can be used for neater appearance. In actual use, the transmitter would sit with the tubes horizontal. The half of the box at left is mounted under the car dash so that the transmitter half can be easily pulled in and out of position for servicing or adjustment.

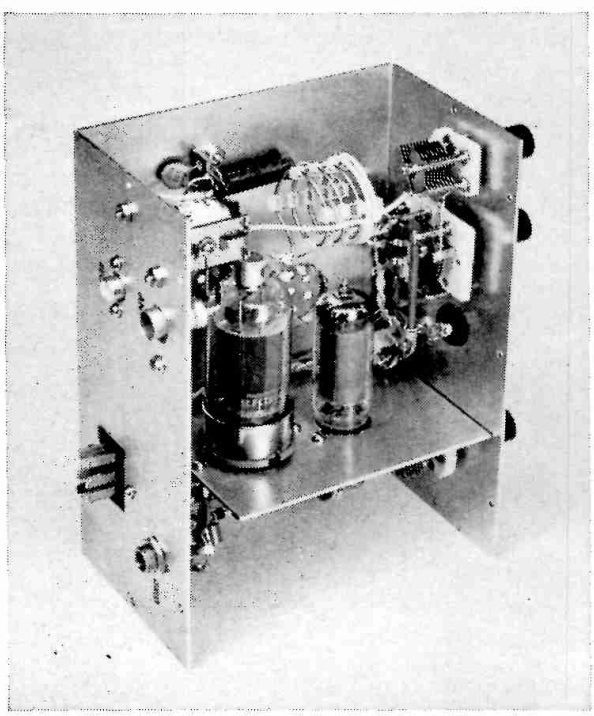


Fig. 19-18—View of the 50-Mc. transmitter showing the r.f. amplifier tank circuits and output loading control. C_3 is on the top right of the panel with C_2 just below it. Output indicator I_1 is below C_2. This view also shows the two antenna connectors, power plug and grid current jack which are mounted on the rear surface.

The amplifier plate tank circuit, C_2L_3, is tuned to resonance by variable capacitor C_2.

The 144-Mc. Unit

The 144-Mc. circuit is shown in Fig. 19-22. The oscillator is similar to the one used in the 50-Mc. transmitter. The 48-Mc. signal from the oscillator is capacitively coupled to the pentode multiplier which is operated as a frequency tripler. From the tripler, the signal is inductively coupled to the grid of the r.f. amplifier. Since this stage contains a fixed capacitor, it is tuned by "pinching" or "spreading" the turns of L_4. As in the 50-Mc. unit, provision is made for measuring grid current (jack J_1).

The amplifier tank circuit in the 144-Mc. model is series tuned. Output coupling is through a single-turn link, L_6. Neutralization is required in this unit; the neutralizing capacitor consists of a 2½-inch length of No. 12 wire with one end connected to pin 5 (control grid) of the amplifier tube, and with the other end run up beside the amplifier tube after passing through the chassis (see the photograph in Fig. 19-21). A piece of spaghetti is used to insulate the neutralizing wire from the chassis.

Construction

A 7 × 5 × 3-inch Minibox is used as the

6- and 2-Meter Mobile Transmitters

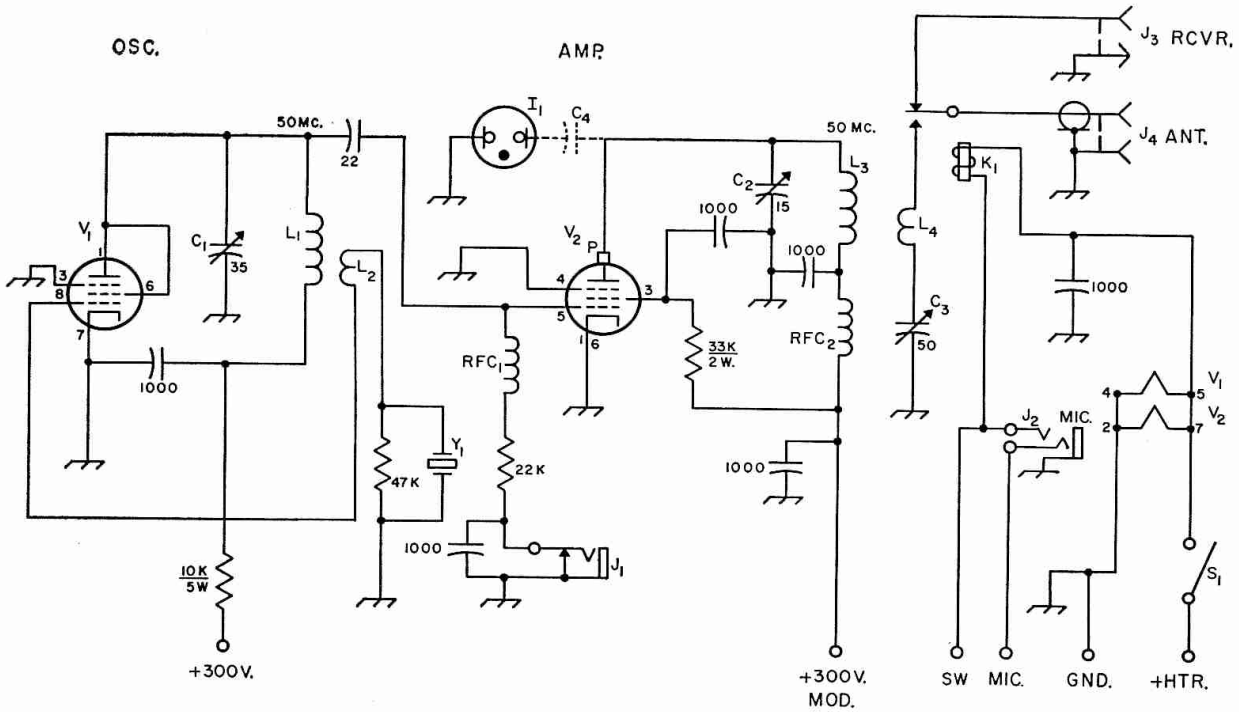


Fig. 19-20—Schematic diagram of the 50-Mc. mobile transmitter. Unless otherwise indicated, capacitances are in $\mu\mu$f., resistances are in ohms, resistors are ½ watt unless specified otherwise.

C_1—35-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-35-B).
C_2—15-$\mu\mu$f. midget variable capacitor (Hammarlund HF-15).
C_3—50-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-50-B).
C_4—Coupling capacitor for output indicator (see text).
K_1—Midget antenna relay s.p.d.t. (Advance AM/2C/-12VD. Note: the last four figures in the number indicate the coil voltage. For 6 volts d.c. it should read /6VD).
L_1—3 turns No. 20, ⅝-inch dia., 5/16 inches long (B & W 3006).
L_2—2 turn link No. 20 insulated wire, close wound over cold end of L_1.
L_3—4 turns No. 16, 1-inch dia., 1-inch long (B & W 3013)
L_4—2 turn link No. 20 insulated wire, close-wound over cold end of L_3.
I_1—Neon bulb (NE-2).
J_1—Circuit closing jack.
J_2—3 conductor mike jack.
J_3, J_4—Automobile type antenna connectors.
RFC_1, RFC_2—Single-layer v.h.f. choke, 2 to 7 μh. (Ohmite Z-50 or National R-60).
S_1—S.p.s.t. slide switch.
V_1—5763 for 6 volts, 6417 for 12 volts.
V_2—2E26 for 6 volts, 5893 for 12 volts.
Y_1—50-Mc. 3rd overtone crystal (International Crystal Co. type FA-9).

chassis for the transmitters. A single bracket supports the tubes and associated parts. The bracket has a single bend and is fastened to the Minibox with machine screws.

The 6- and 2-meter transmitters are almost identical mechanically. The only real difference between the two is that the 2-meter model has an additional multiplier tube, mounted in line with the oscillator tube on the bracket.

All parts should be mounted before wiring is begun. Since both ends of the chassis are open, wiring and mounting of parts is a simple job. The photographs show the relative position of most of the components. Try to keep r.f. leads as short as possible. The relay, antenna connectors, power plug and grid current jack are all mounted on the rear panel.

The output indicator I_1 is coupled to the final tank circuit through capacitor C_4. This capacitor is actually a few turns of hook-up wire wound over a piece of insulated wire that is

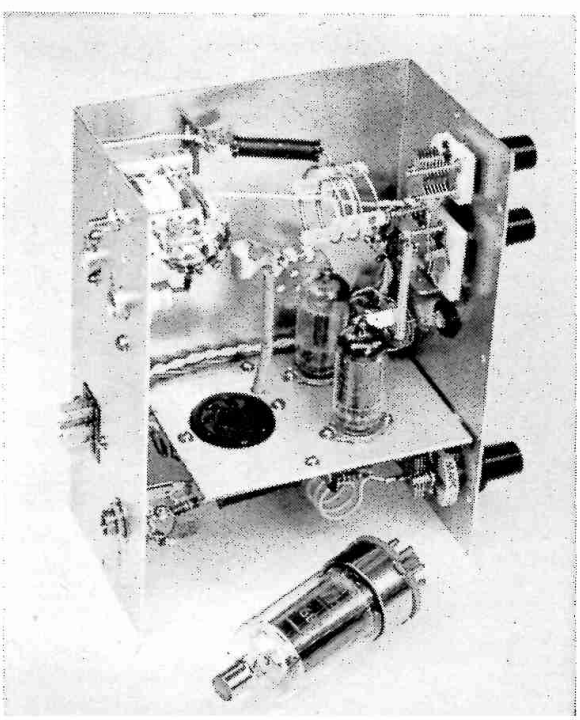


Fig. 19-21—The 144 Mc. transmitter with the r.f. amplifier tube removed to show the neutralizing lead C_N. Except for the 6BJ6 multiplier tube in the foreground, the same basic layout is used here as in the 50-Mc. unit.

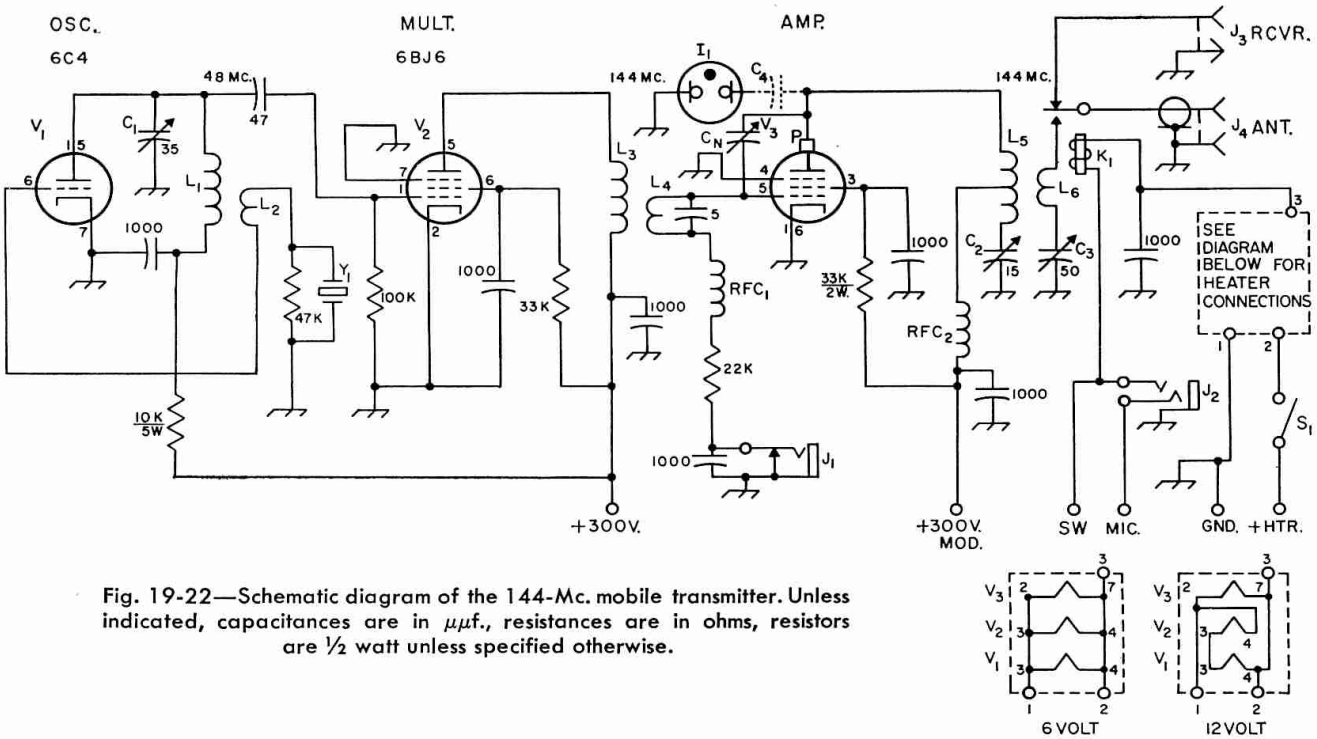


Fig. 19-22—Schematic diagram of the 144-Mc. mobile transmitter. Unless indicated, capacitances are in $\mu\mu$f., resistances are in ohms, resistors are ½ watt unless specified otherwise.

C_1—35-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-35-B).
C_2—15-$\mu\mu$f. midget variable capacitor (Hammarlund HF-15).
C_3—50-$\mu\mu$f. midget variable capacitor (Hammarlund MAPC-50-B).
C_4—Coupling capacitor for output indicator (see text).
C_n—Neutralizing capacitor (see text)
K_1—Midget antenna relay s.p.d.t. (Advance AM/2C/-12VD. Note: the last four figures in the number indicate the coil voltage. For 6 volts it should read 6VD.)
L_1—4 turns No. 20, ⅝ inch diam., 5/16 inches long (B & W 3006).
L_2—2 turn link No. 20 insulated wire, close wound over cold end of L_1.
L_3—1 turn No. 20 insulated wire ½-inch diam.
L_4—2 turns No. 20 insulated wire ½-inch diam.
L_5—3 turns No. 16, 1-inch diam., ¾ inches long, center tapped (B & W 3013).
L_6—1 turn link No. 20 insulated wire wound in the center of L_5.
I_1—Neon bulb (NE-2).
J_1—Circuit closing jack.
J_2—3 conductor mike jack.
J_3, J_4—Automobile type antenna connector.
RFC1, RFC2—Single-layer v.h.f. choke, 2 to 7 μh. (Ohmite Z-50 or National R-60).
S_1—S.p.s.t. slide switch.
V_1—6C4.
V_2—6BJ6.
V_3—2E26 for 6 volts, 5893 for 12 volts.
Y_1—48 Mc. 3rd overtone crystal. Crystal frequency found by dividing desired output frequency by 3 (International Crystal Co. type FA-9).

connected to the final tank circuit. If the lamp fails to ignite, a few more turns may be needed.

Testing Notes

An a.c. power supply delivering 300 volts at 100 ma. can be used during testing of the transmitter. Heater-current requirements for the 50-Mc. unit are 1.55 ampere for 6-volt operation and 0.775 ampere for 12 volts. The 144-Mc. unit requires 1.1 ampere at 6 volts and 0.55 ampere at 12 volts. Do not connect the plate supply to the r.f. amplifier power terminal (marked "300 mod." in the circuit diagram) at this time. The correct crystal and a dummy load should be kept on hand for the test.

To test the driver stage, plug a grid-current meter (0-5 ma.) in J_1, and apply heater voltage. Plug in the proper crystal and turn on the plate voltage (exciter stages only). As quickly as possible adjust capacitor C_1 until the oscillator goes into oscillation. This will be indicated by a downward kick in the plate current. Grid current should begin to show when oscillation occurs.

In the 144-Mc. unit, adjust for maximum grid current by "pinch-tuning" L_3, L_4 once oscillation has begun. Adjust C_1 for maximum grid current. If there is difficulty in obtaining grid drive, try adjusting the position of L_2 with respect to L_1. In the 2-meter model, some rearrangement of L_3 and L_4 may be needed in order to achieve maximum grid drive.

Before testing the 144-Mc. amplifier it will be necessary to neutralize it. With power applied to the exciter portion, slowly rotate the output tuning control C_2 through its full range. If the amplifier is neutralized, there will be no fluctuation in the grid current. If there is such a fluctuation, adjust the neutralizing wire to a new position with respect to the amplifier tube and swing the plate-tuning control again. Repeat until the grid current remains steady, showing that the amplifier is neutralized.

Connect a dummy load to the output antenna connector, close the antenna relay and apply plate power to the entire transmitter. As quickly as possible, tune C_2 for minimum plate current.

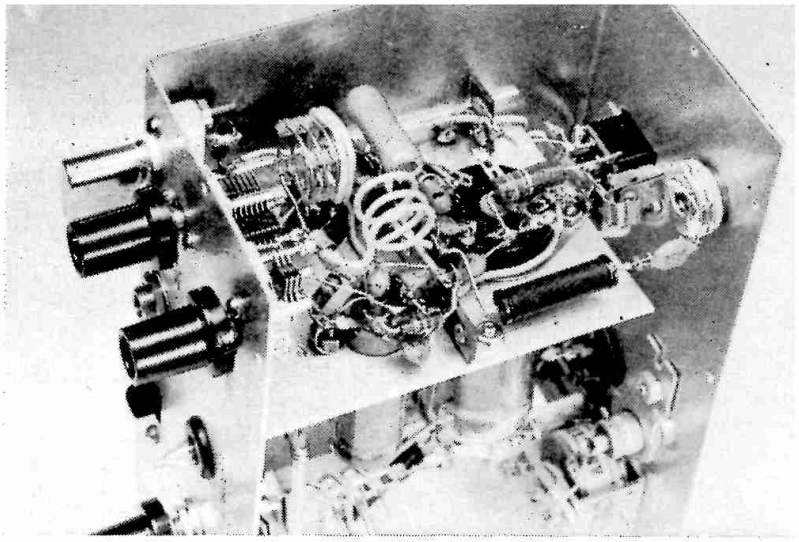


Fig. 19-23—View of the 144-Mc. transmitter. The coil and link near the top left rear are L_1L_2. In the foreground are coils L_3L_4.

It is necessary to perform this operation rapidly because the amplifier may draw excessive plate current when not tuned to resonance. When tuned to resonance, the output indicator bulb I_1 will light. This r.f. indicator is not only a tuning aid in the car but also acts as a continuous monitor to show that the transmitter is in operation. Capacitor C_3 is the loading control and should be adjusted for maximum plate current after the amplifier is resonated.

A microphone jack J_2 is included on the transmitter chassis to simplify the control circuits. Leads from the microphone (marked "sw" and "mic" in the diagram) go to the power connector at the rear of the transmitter.

Mobile Modulators

Vacuum-tube modulators for mobile operation are in general similar to those used in fixed-station installations. Equipment shown in the section on modulators may be modified for use with almost any mobile transmitter. As in fixed station work, the mobile modulator must be capable of supplying to the plate modulated r.f. stage sine-wave audio power equal to 50 per cent of the d.c. plate input for 100 percent modulation.

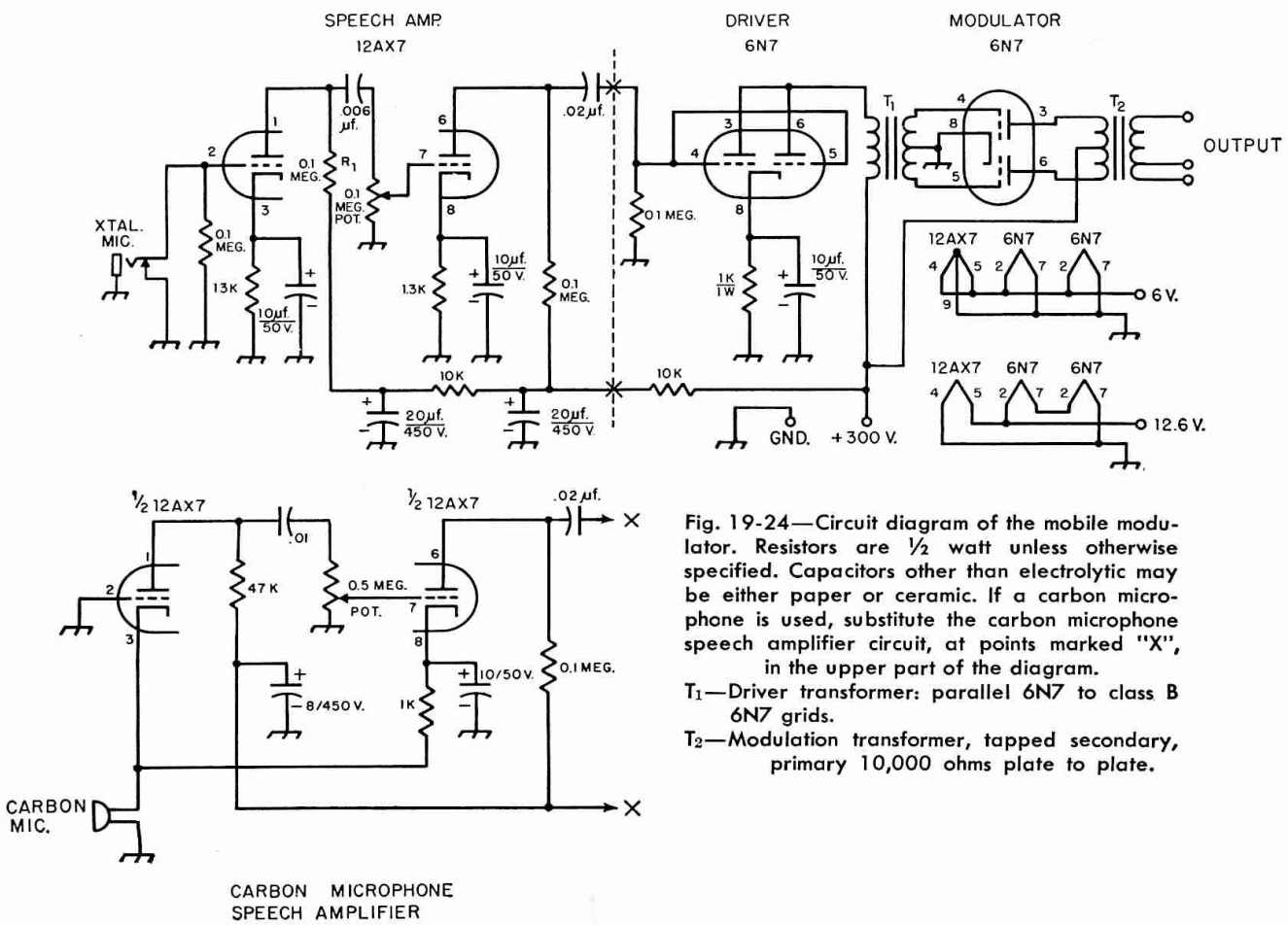


Fig. 19-24—Circuit diagram of the mobile modulator. Resistors are ½ watt unless otherwise specified. Capacitors other than electrolytic may be either paper or ceramic. If a carbon microphone is used, substitute the carbon microphone speech amplifier circuit, at points marked "X", in the upper part of the diagram.

T_1—Driver transformer: parallel 6N7 to class B 6N7 grids.

T_2—Modulation transformer, tapped secondary, primary 10,000 ohms plate to plate.

A 10-WATT MOBILE MODULATOR

Fig. 19-24 shows a modulator that can be used with any mobile a.m. transmitter whose input does not exceed 20 to 25 watts. A resistance-coupled speech amplifier using a single 12AX7 drives a Class A 6N7 which in turn drives a Class B 6N7. The 6N7 uses the two triode sections in parallel, to obtain sufficient driving power.

Also shown in Fig. 19-24 are the changes in the speech-amplifier circuit necessary to adapt it for use with a carbon microphone. D.c. voltage for the carbon microphone is obtained by connecting the microphone in series with the speech-amplifier cathodes.

The modulator requires 300 volts at about 90 ma. for plate power, and 6 volts at 1.9 amperes or 12 volts at .95 amperes for the heaters. Heater connections are given for both voltages. The plate supply should use a large capacitance (100 μf. or more) in the output, to serve as a reservoir for the heavy peak-current demands.

The main constructional precaution to be observed when building the modulator is that the output transformer T_2 should not be mounted too close to the speech amplifier circuits. Separation will reduce the chance of feedback through stray coupling. A tube shield over the 12AX7 will serve to hold it in the socket over bumpy roads; good octal sockets will normally need no tube clamps to retain the 6N7s.

In any mobile installation, the modulator may be separated from the r.f. assembly by any convenient distance. The cable connecting the modulator to the r.f. section should be made with individually shielded leads.

A 25-WATT TRANSISTOR MODULATOR

Figs. 19-25 through 19-27 show a complete transistor modulator that obtains its power directly from the automobile's 12-volt storage battery. It requires only a fraction of the space required by a comparable vacuum-tube unit, and it allows full use of the high-voltage power supply for the r.f. section.

The unit is based on a design orignally published by Delco Radio [1]; it is a 12-volt 25-watt Class B modulator. Among the advantages of a modulator of this type are the compactness (25 watts of audio in approximately 90 cubic inches), high over-all efficiency, no warm-up time, and low idling current when not modulating. It will modulate an r.f. stage input of between 45 and 50 watts, at an impedance level of 4000 ohms with the output transformer listed (about 450 volts and 110 ma.). Suitable 12-volt heater tubes for the modulated output stage include the 1625

[1] *Transistor Application Note 6-B*, Delco Radio Division, General Motors Corp, Kokomo, Indiana.

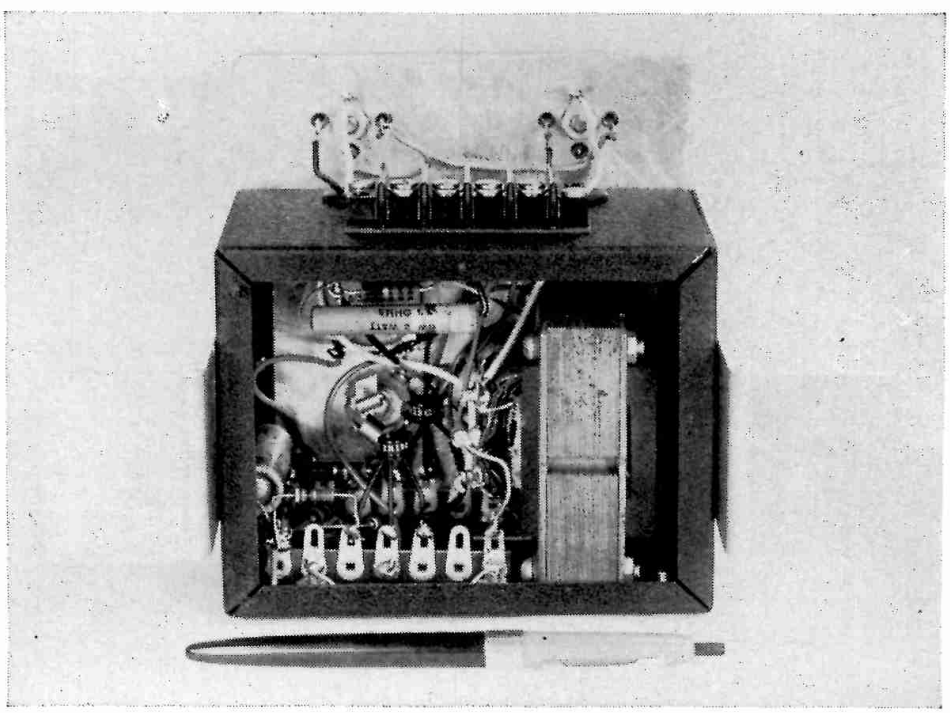


Fig. 19-25—A 3 × 4 × 5-inch utility box is sufficient to house the modulation transformer and all of the smaller components of the 25-watt transistor modulator.

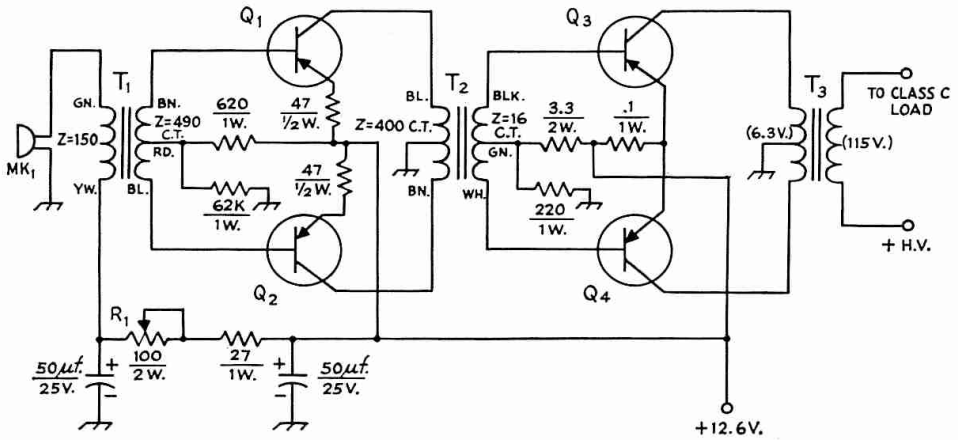


Fig. 19-26—Circuit of the 25-watt transistor modulator. Resistances are in ohms. Capacitors are electrolytic.

MK_1—Single-button carbon microphone.
Q_1, Q_2—2N190 (GE) or 2N109 (RCA).
Q_3, Q_4—DS-501 (Delco).
R_1—100-ohm 2-watt potentiometer.
T_1—150 ohms c.t. (c.t. not used) to 490 ohms c.t. (Thordarson TR-5).
T_2—400 ohms c.t. to 16 ohms, c.t. (see text), Stancor TA-41).
T_3—6.3-volt c.t., 3-amp, filament transformer used as modulation transformer (see text) (Stancor P-5014).

(similar to 807) and the 6883 (similar to the 6146). The exciter portion of the transmitter can be made up of 6417s (similar to the 5763) or of 12V6-GTs or 12L6-GTs (12-volt-heater versions of the 6V6 and 6L6). Maximum economy will be obtained with a transistorized power supply, similar to the unit described later in this chapter.

Construction

The unit is constructed on a 4 × 5 × 3-inch utility box on which a ⅛-inch aluminum cover 5 × 6 inches is substituted. This provides a 2-inch overhang on one edge for mounting the power transistors, and it also serves as a heat sink. Two transformers, plus gain control and mike jack, are also mounted on the cover (see Fig. 19-27).

For a modulation transformer the unit uses a 6.3-volt filament transformer turned backwards; that is, with the 6.3-volt 3-ampere winding toward the collectors. This transformer is mounted inside the utility box. Ample room is left for the input transformers, resistors and capacitors. It was found necessary to add an input filter on the 12-volt line to prevent hash from getting into the microphone circuit and adding noise.

To obtain a true center tap for the driver transformer, a transformer having taps at 4 and 16 ohms is used. Since the impedance varies as the square of the turns ratio, the 4-ohm tap provides a center tap.

Transistor Mounting

Because the collector connection is common with the case of the transistor, mica spacers must be used between the transistor cases and ground. (Insulator package No. 1221264). These can be obtained in a special mounting kit from Delco distributors.

A four-lug terminal box is located on top of the utility box to provide for the 12-volt and output connections of the modulator. Although wiring of the unit may appear difficult, it becomes a relatively simple job if the internal wiring is done separately, before putting on the front cover.

Be careful to apply as little heat as possible when soldering any transistor connections. Either G.E. 2N190 or RCA 2N109 can be used for the input transistors. Although several other types could be used for the output transistors, the specified DS-501 should be easier to obtain than some since it is sold as a replacement in car-radio service.

It is not likely that a 0.1-ohm 1-watt resistor (see Fig. 19-26) can be purchased at any radio store. A satisfactory substitute is to wind a suitable length of resistance wire over a 2-watt resistor used as a form, or three 0.33-ohm ½-watt resistors can be wired in parallel to obtain a value sufficiently close.

Testing

After wiring and construction of the unit is completed, testing for proper operation can be done in several ways. One method is simply to connect a 4000-ohm 10-watt resistor across the modulation transformer output connections and then place a d.c. ammeter in series with the 12-volt line, and watch the current variation while talking into the microphone. The idling current should be around 700 ma., kicking up to above 2 amperes on peaks. Do not, under any circumstances, try to operate the unit without a load of some sort on the output terminals as this may damage the output transistors.

Another method of testing is to place another

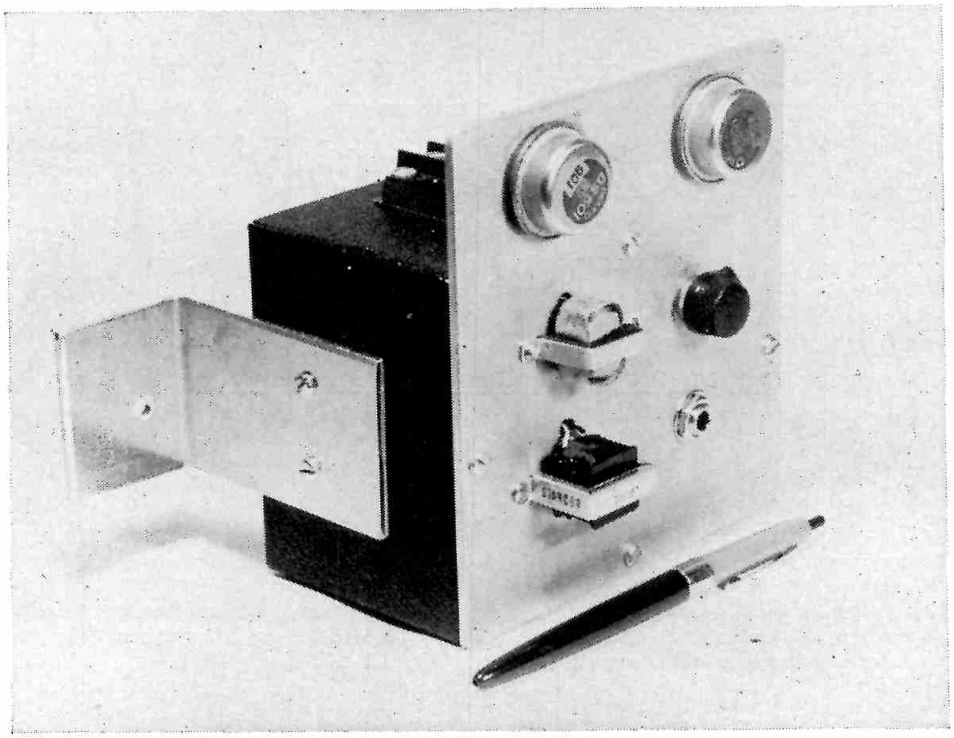


Fig. 19-27—The front cover of the modulator unit serves as a heat sink. The driver transformer and microphone jack are at the bottom, the microphone transformer and potentiometer control at the center, and the two power transistors at the top.

6.3-volt filament transformer back-to-back with the modulation transformer, to bring the impedance down to a low level, and then connect a p.m. speaker to the 6.3-volt winding.

A 'scope test can be made after the unit is connected to the transmitter. The Class C load level can be adjusted for impedance matching.

An F1 carbon microphone is suitable for use with this unit. Although not shown in Fig. 19-26, the unit should be connected so that it is turned on only while the transmit-receive switch is in the transmit position. An inexpenisve 12-volt automobile-horn relay (e.g., Echlin HR 101), available at most filling stations or automobile parts distributors, should be used to close and open the circuit. The relay arm and contact should be connected in the +12.6-volt lead from the battery and fuse. If excessive sparking is noted at the relay contacts it may be reduced by moving the 50-μf. 25-volt capacitor to the fuse side of the relay contacting circuit.

Concerning placement of the unit in the car: Try to find a location away from high-temperature spots and in a well-ventilated area. The trunk is not recommended since there is little ventilation; this area can become quite hot in the summertime and damage to the transistors could result. The engine compartment makes a convenient place to mount the unit but this space is not adequately ventilated except possibly while the car is in motion. The most favorable spot is on the fire wall in the passenger compartment, or under the front seat. These areas are usually well ventilated, or at least cooler than any other enclosed section of the car. As in any mobile installation where the modulator is some distance from the r.f. section, the audio leads from the secondary of the modulation transformer to the modulated r.f. stage should be made with individually-shielded leads.

(Original description appeared in *QST* for November, 1959.)

The Mobile Antenna

For mobile operation in the range between 1.8 and 30 Mc., the vertical whip antenna is almost universally used. Since longer whips present mechanical difficulties, the length is usually limited to a dimension that will resonate as a quarter-wave antenna in the 10-meter band. The car body serves as the ground connection. This antenna length is approximately 8 feet.

With the whip length adjusted to resonance in the 10-meter band, the impedance at the feed point, *X*, Fig. 19-28, will appear as a pure resistance at the resonant frequency. This resistance will be composed almost entirely of radiation resistance (see index), and the efficiency will be high. However, at frequencies lower than the resonant frequency, the antenna will show an increasingly large capacitive reactance and a decreasingly small radiation resistance.

The equivalent circuit is shown in Fig. 19-29. For the average 8-ft. whip, the reactance of the

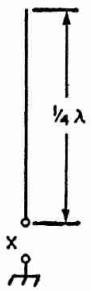


Fig. 19-28—The quarterwave whip at resonance will show a pure resistance at the feed point X.

capacitance, C_A, may range from about 150 ohms at 21 Mc. to as high as 8000 ohms at 1.8 Mc., while the radiation resistance, R_R, varies from about 15 ohms at 21 Mc. to as low as 0.1 ohm at 1.8 Mc. Since the resistance is low, considerable current must flow in the circuit if any appreciable power is to be dissipated as radiation in the resistance. Yet it is apparent that little current can be made to flow in the circuit so long as the comparatively high series reactance remains.

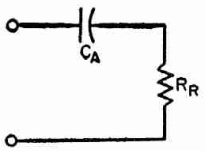


Fig. 19-29—At frequencies below the resonant frequency, the whip antenna will show capacitive reactance as well as resistance. R_R is the radiation resistance, and C_A represents the capacitive reactance.

Eliminating Reactance

The capacitive reactance can be canceled out by connecting an equivalent inductive reactance, L_L, in series, as shown in Fig. 19-30, thus tuning the system to resonance.

Fig. 19-30—The capacitive reactance at frequencies lower than the resonant frequency of the whip can be canceled out by adding an equivalent inductive reactance in the form of a loading coil in series with the antenna.

Unfortunately, all coils have resistance, and this resistance will be added in series, as indicated at R_C in Fig. 19-31. While a large coil may radiate some energy, thus adding to the radiation resistance, the latter will usually be negligible

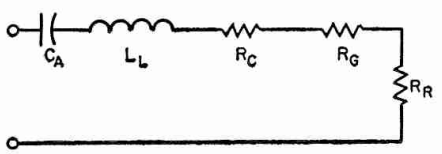


Fig. 19-31—Equivalent circuit of a loaded whip antenna. C_A represents the capacitive reactance of the antenna, L_L an equivalent inductive reactance. R_C is the loading-coil resistance, R_G the ground-loss resistance, and R_R the radiation resistance.

compared to the loss resistance introduced. However, adding the coil makes it possible to feed power to the circuit.

Ground Loss

Another element in the circuit dissipating power is the ground-loss resistance. Fundamentally, this is related to the nature of the soil in the area under the antenna. Little information is available on the values of resistance to be expected in practice, but some measurements have shown that it may amount to as much as 10 or 12 ohms at 4 Mc. At the lower frequencies, it may constitute the major resistance in the circuit.

Fig. 19-31 shows the circuit including all of the elements mentioned above. Assuming C_A lossless

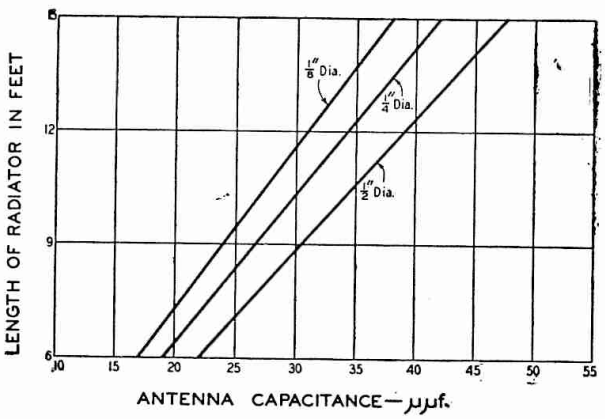


Fig. 19-32—Graph showing the approximate capacitance of short vertical antennas for various diameters and lengths, at 3.9 Mc. These values should be approximately halved for a center-loaded antenna.

and the loss resistance of the coil to be represented by R_C, it is seen that the power output of the transmitter is divided among three resistances — R_C, the coil resistance; R_G, the ground-loss resistance; and R_R, the radiation resistance. Only the power dissipated in R_R is radiated. The power developed in R_C and R_G is dissipated in heat. Therefore, it is important that the latter two resistances be minimized.

MINIMIZING LOSSES

There is little that can be done about the nature of the soil. However, poor electrical contact between large surfaces of the car body, and especially between the point where the feed line is grounded and the rest of the body, can add materially to the ground-loss resistance. For example, the feed line, which should be grounded as close to the base of the antenna as possible, may be connected to the bumper, while the bumper may have poor contact with the rest of the body because of rust or paint.

Loading Coils

The accompanying tables show the approximate loading-coil inductance required for the various bands. The graph of Fig. 19-32 shows the approximate capacitance of whip antennas of

TABLE 19-I

Approximate Values for 8-ft. Mobile Whip						
Base Loading						
$f_{kc.}$	*Loading* $L_{\mu h.}$	R_C (Q50) *Ohms*	R_C (Q300) *Ohms*	R_R *Ohms*	*Feed R* Ohms*	*Matching* $L_{\mu h}$ *
1800	345	77	13	0.1	23	3
3800	77	37	6.1	0.35	16	1.2
7200	20	18	3	1.35	15	0.6
14,200	4.5	7.7	1.3	5.7	12	0.28
21,250	1.25	3.4	0.5	14.8	16	0.28
29,000	....	....	....	....	36	0.23
Center Loading						
1800	700	158	23	0.2	34	3.7
3800	150	72	12	0.8	22	1.4
7200	40	36	6	3	19	0.7
14,200	8.6	15	2.5	11	19	0.35
21,250	2.5	6.6	1.1	27	29	0.29

R_C = Loading-coil resistance; R_R = Radiation resistance.
* Assuming loading coil Q = 300, and including estimated ground-loss resistance.
Suggested coil dimensions for the required loading inductances are shown in a following table.

various average diameters and lengths. For 1.8, 4 and 7 Mc., the loading-coil inductance required (when the loading coil is at the base) will be approximately the inductance required to resonate in the desired band with the whip capacitance taken from the graph. For 14 and 21 Mc., this rough calculation will give more than the required inductance, but it will serve as a starting point for final experimental adjustment that must always be made.

Also shown in table 19-I are approximate values of radiation resistance to be expected with an 8-ft. whip, and the resistances of loading coils — one group having a Q of 50, the other a Q of 300. A comparison of radiation and coil resistances will show the importance of reducing the coil resistance to a minimum, especially on the three lower-frequency bands.

To minimize loading-coil loss, the coil should have a high ratio of reactance to resistance, i.e., high Q. A 4-Mc. loading coil wound with small wire on a small-diameter solid form of poor quality, and enclosed in a metal protector, may have a Q as low as 50, with a resistance of 50 ohms or more. High-Q coils require a large conductor, "air-wound" construction, turns spaced, the best insulating material available, a diameter not less than half the length of the coil (not always mechanically feasible), and a minimum of metal in the field. Such a coil for 4 Mc. may show a Q of 300 or more, with a resistance of 12 ohms or less. This reduction in loading-coil resistance may be equivalent to increasing the transmitter power by 3 times or more. Most low-loss transmitter plug-in coils of the 100-watt size or larger, commercially produced, show a Q of this order. Where larger inductance values are required, lengths of low-loss space-wound coils are available.

TABLE 19-II

Suggested Loading-Coil Dimensions					
Req'd $L_{\mu h.}$	*Turns*	*Wire Size*	*Diam. In.*	*Length In.*	*Form or B & W Type*
700	190	22	3	10	Polystyrene
345	135	18	3	10	Polystyrene
150	100	16	2½	10	Polystyrene
77	75	14	2½	10	Polystyrene
77	29	12	5	4¼	160T
40	28	16	2½	2	80B less 7 t.
40	34	12	2½	4¼	80T
20	17	16	2½	1¼	80B less 18 t.
20	22	12	2½	2¾	80T less 12 t.
8.6	16	14	2	2	40B less 4 t.
8.6	15	12	2½	3	40T less 5 t.
4.5	10	14	2	1¼	40B less 10 t.
4.5	12	12	2½	4	40T
2.5	8	12	2	2	15B
2.5	8	6	2⅜	4½	15T
1.25	6	12	1¾	2	10B
1.25	6	6	2⅜	4½	10T

Mobile Antennas

Center Loading

The radiation resistance of the whip can be approximately doubled by placing the loading coil at the center of the whip, rather than at the base, as shown in Fig. 19-33. (The optimum position varies with ground resistance. The center is optimum for average ground resistance.) However, the inductance of the loading coil must be

Fig. 19-33—Placing the loading coil at the center of the whip antenna, instead of at the base, increases the radiation resistance, although a larger coil must be used.

approximately doubled over the value required at the base to tune the system to resonance. For a coil of the same Q, the coil resistance will also be doubled. But, even if this is the case, center loading represents a gain in antenna efficiency, especially at the lower frequencies. This is because the ground-loss resistance remains the same, and the increased radiation resistance becomes a larger portion of the total circuit resistance, even though the coil resistance also increases. However, as turns are added to a loading coil (other factors being equal) the inductance (and therefore the reactance) increases at a greater rate than the resistance, and the larger coil will usually have a higher Q.

Top Loading Capacitance

Since the coil resistance varies with the inductance of the loading coil, the coil resistance can be reduced by reducing the number of turns. This can be done, while still maintaining resonance, by adding capacitance to the portion of the antenna above the coil. This capacitance can be provided by attaching a capacitive surface as high up on the antenna as is mechanically feasible. Capacitive "hats," as they are usually called, may consist of a light-weight metal ball, cylinder, disk, or wheel structure as shown in Fig. 19-34. This should be added to the capacitance of the whip above the loading coil (from Fig. 19-32) in determining the approximate inductance of the loading coil.

When center loading is used, the amount of capacitance to be added to permit the use of the same loading inductance required for base loading is not great, and should be seriously considered, since the total gain made by moving the coil to the center of the antenna may be quite marked.

Tuning the Band

Especially at the lower frequencies, where the resistance in the circuit is low compared to the coil reactance, the antenna will represent a very

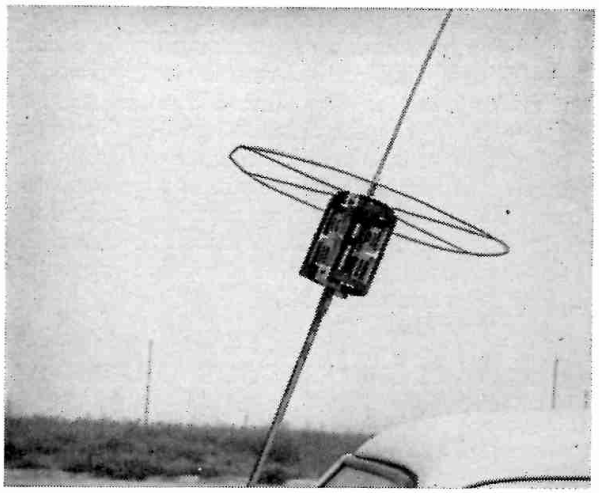
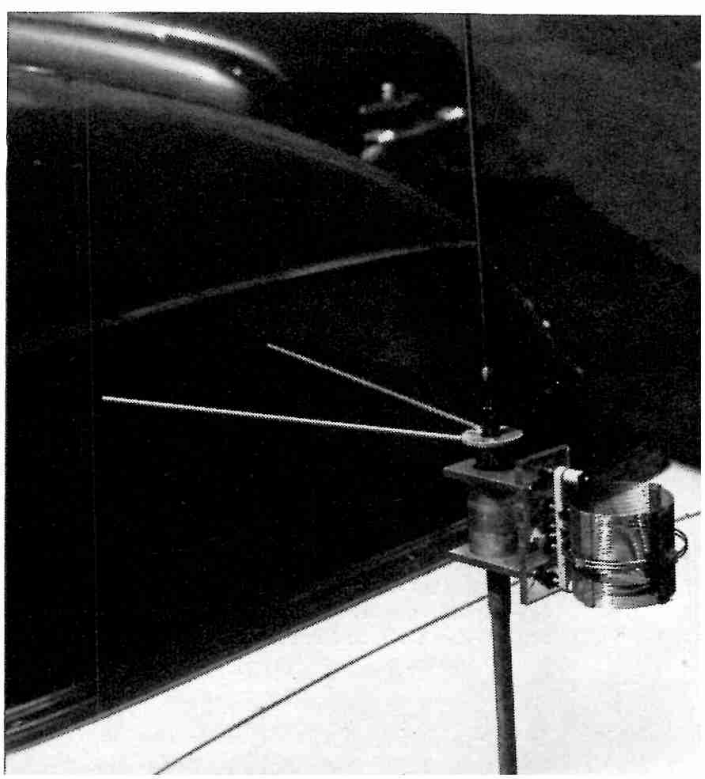


Fig. 19-34—The top-loaded 4-Mc. antenna designed by W6SCX. The loading coil is a B & W transmitting coil. The coil can be tuned by the variable link which is connected in series with the two halves of the coil.

high-Q circuit, making it necessary to retune for relatively small changes in frequency. While many methods have been devised for tuning the whip over a band, one of the simplest is shown in Fig. 19-35. In this case, a standard B & W plug-in coil is used as the loading coil. A length of large-diameter polystyrene rod is drilled and tapped to fit between the upper and lower sections of the antenna. The assembly also serves to clamp a pair of metal brackets on each side

Fig. 19-35—W8AUN's adjustable capacity hat for tuning the whip antenna over a band. The coil is a B & W type B 160-meter coil, with a turn or two removed. Spreading the rods apart increases the capacitance. This simple top loader has sufficient capacitance to permit the use of approximately the same loading-coil inductance at the center of the antenna as would normally be required for base loading.



of the polystyrene block that serve both as support and connections to the loading-coil jack bar.

A ⅛-inch steel rod, about 15 inches long, is brazed to each of two large-diameter washers with holes to pass the threaded end of the upper section. The rods form a loading capacitance that varies as the upper rod is swung away from the lower one, the latter being stationary. Enough variation in tuning can be obtained to cover the 80-meter band. (Original description appeared in *QST*, September, 1953.)

REMOTE ANTENNA RESONATING

Fig. 19-36 shows circuits of two remote-control resonating systems for mobile antennas. As shown, they make use of surplus d.c. motors driving a loading coil removed from a surplus ARC-5 transmitter. A standard coil and motor may be used in either installation at increased expense.

The control circuit shown in Fig. 19-36-A is a three-wire system (the car frame is the fourth conductor) with a double-pole double-throw switch and a momentary (normally off) single-pole single-throw switch. S_2 is the motor reversing switch. The motor runs so long as S_1 is closed.

The circuit shown in Fig. 19-36B uses a latching relay, in conjunction with microswitches, to automatically reverse the motor when the roller reaches the end of the coil. S_3 and S_5 operate the relay, K_1, which reverses the motor. S_4 is the motor on-off switch. When the tuning coil roller reaches one end or the other of the coil, it closes S_6 or S_7, as the case may be, operating the relay and reversing the motor.

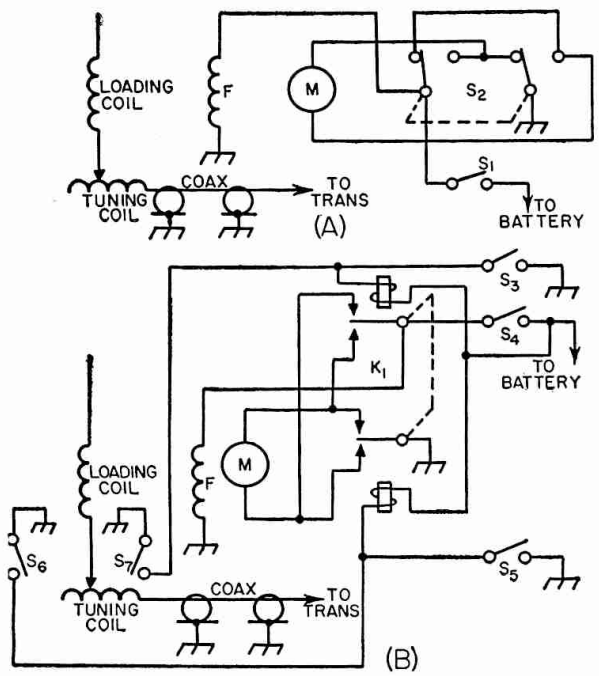


Fig. 19-36—Circuits of the remote mobile-whip tuning systems.

K_1—D.p.d.t. latching relay.

S_1, S_3, S_4, S_5—Momentary-contact s.p.s.t., normally open.

S_2—D.p.d.t. toggle.

S_6, S_7—S.p.s.t. momentary-contact microswitch, normally open.

The procedure in setting up the system is to prune the center loading coil to resonate the antenna on the highest frequency used without the base loading coil. Then, the base loading coil is used to resonate at the lower frequencies. When the circuit shown in Fig. 19-36A is used for control, S_1 is used to start and stop the motor, and S_2, set at the "up" or "down" position, will determine whether the resonant frequency is raised or lowered. In the circuit shown in Fig. 19-36B, S_4 is used to control the motor. S_3 or S_5 is momentarily closed (to activate the latching relay) for raising or lowering the resonant frequency. The broadcast antenna is used with a wavemeter to indicate resonance.

(Originally described in *QST*, Dec., 1953.)

Several companies offer motor tuning for getting optimum performance over a low-frequency band. (For a complete description of the commercially available remotely-tuned systems, see Goodman, "Frequency Changing and Mobile Antennas," *QST*, Dec., 1957.)

Automatic Mobile Antenna Tuning

A somewhat more complex antenna tuning system for 75 and 40 meters is one that automatically tunes the antenna as the transmitter frequency is shifted. After initial adjustments, the radiator is kept in resonance without attention from the operator. (For a description of the automatic system, see Hargrave, "Automatic Mobile Antenna Tuning, *QST*, May, 1955.)

FEEDING THE ANTENNA

It is usually found most convenient to feed the whip antenna with coax line. Unless very low-*Q* loading coils are used, the feed-point impedance will always be appreciably lower than 52 ohms — the characteristic impedance of the commonly-used coax line, RG-8/U or RG-58/U. Since the length of the transmission line will seldom exceed 10 ft., the losses involved will be negligible, even at 29 Mc., with a fairly-high s.w.r. However, unless a line of this length is made reasonably flat, difficulty may be encountered in obtaining sufficient coupling with a link to load the transmitter output stage.

One method of obtaining a match is shown in Fig. 19-37. A small inductance, L_M, is inserted at the base of the antenna, the loading-coil inductance being reduced correspondingly to maintain resonance. The line is then tapped on the coil at a point where the desired loading is obtained. Table 19-I shows the approximate inductance to be used between the line tap and ground. It is advisable to make the experimental matching coil larger than the value shown, so that there will be provision for varying either side of the proper position. The matching coil can also be of the plug-in type for changing bands.

Adjustment

For operation in the bands from 29 to 1.8 Mc.,

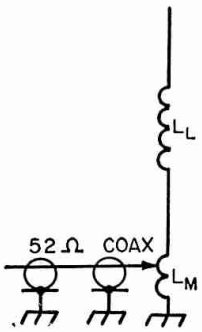


Fig. 19-37—A method of matching the loaded whip to 52-ohm coax cable. L_L is the loading coil and L_M the matching coil.

the whip should first be resonated at 29 Mc. with the matching coil inserted, but the line disconnected, using a grid-dip oscillator coupled to the matching coil. Then the line should be attached, and the tap varied to give proper loading, using a link at the transmitter end of the line whose reactance is approximately 52 ohms at the operating frequency, tightly coupled to the output tank circuit. After the proper position for the tap has been found, it may be necessary to readjust the antenna length slightly for resonance. This can be checked on a field-strength meter several feet away from the car.

The same procedure should be followed for each of the other bands, first resonating, with the g.d.o. coupled to the matching coil, by adjusting the loading coil.

After the position of the matching tap has been found, the size of the matching coil can be reduced to only that portion between the tap and ground, if desired. If turns are removed here, it will be necessary to reresonate with the loading coil.

If an entirely flat line is desired, a s.w.r. indicator should be used while adjusting the line tap. With a good match, it should not be necessary to readjust for resonance after the line tap has been set.

It should be emphasized that the figures shown in the table are only approximate and may be altered considerably depending on the type of car on which the antenna is mounted and the spot at which the antenna is placed.

● ANTENNAS FOR 50 AND 144 MC.

A Simple Vertical Antenna

The most convenient type of antenna for mobile v.h.f. work is the quarter-wave vertical radiator, fed with 50-ohm coaxial line. The antenna, which may be a flexible telescoping "fish pole," can be mounted in any of several places on the car. An ideal mounting spot is on top of the car, though rear-deck mounting presents a better spot for esthetic reasons. Tests have shown that with the car in motion there is no observable difference in average performance of the antennas, regardless of their mounting positions. There may be more in the way of directional effects with the rear-deck mount, but the over-all advantage of the roof mount is slight.

A good match may be obtained by feeding the simple vertical with 50-ohm line. However, it is well to provide some means for tuning the system, so that all variables can be taken care of. The simplest tuning arrangement consists of a variable capacitor connected between the low side of the transmitter coupling coil and ground, as shown in Fig. 19-38. This capacitor should

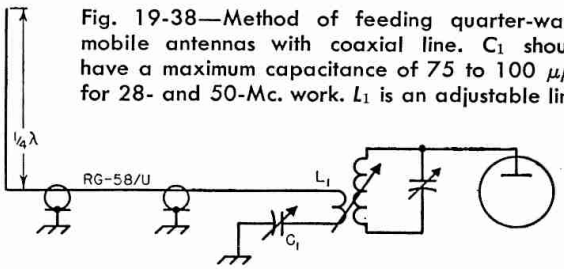


Fig. 19-38—Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacitance of 75 to 100 $\mu\mu$f. for 28- and 50-Mc. work. L_1 is an adjustable link.

have a maximum capacitance of 75 to 100 $\mu\mu$f. 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.

Horizontal Polarization

Horizontally polarized antennas have a considerable advantage over the vertical whip under usual conditions of mobile operation. This is particularly true when horizontal polarization is used at both ends of a line-of-sight circuit, or on a longer circuit over reasonably flat terrain. An additional advantage, especially on 6 meters, is a marked reduction in ignition noise from neighboring cars as well as from the station car.

A Horizontally Polarized Two-Band Antenna for V.H.F.

One type of horizontally-polarized antenna, called the "halo," is shown in Fig. 19-39. It is a dipole bent into a circle, with the ends capacitively loaded to reduce the circumference. Since the 50- and 144-Mc. bands are almost in third harmonic relationship, it is possible to build a single halo that will work on both bands. The antenna is changed from one band to another by changing the spacing between the end loading plates and adjusting the matching mechanism.

Mechanical Details

The halo is made of $\frac{7}{16}$-inch aluminum fuel-line tubing. This material is both strong and very light, but any tubing of about $\frac{1}{2}$-inch diameter could be used equally well. The loop is 67 inches in circumference and the capacitor plates are $2\frac{1}{4}$ inches square, with the corners rounded off.

To fasten the capacitor plates to the ends of the tubing, aluminum rod stock is turned down on a lathe to make a tight fit into the ends. This is tapped for 6–32 thread, and then forced into the tubing ends. Holes are drilled through tubing and inserts, at each end of the halo, and a screw run through each to keep the inserts from turning around or slipping out. The binding-head screws that hold the plates to the inserts are equipped

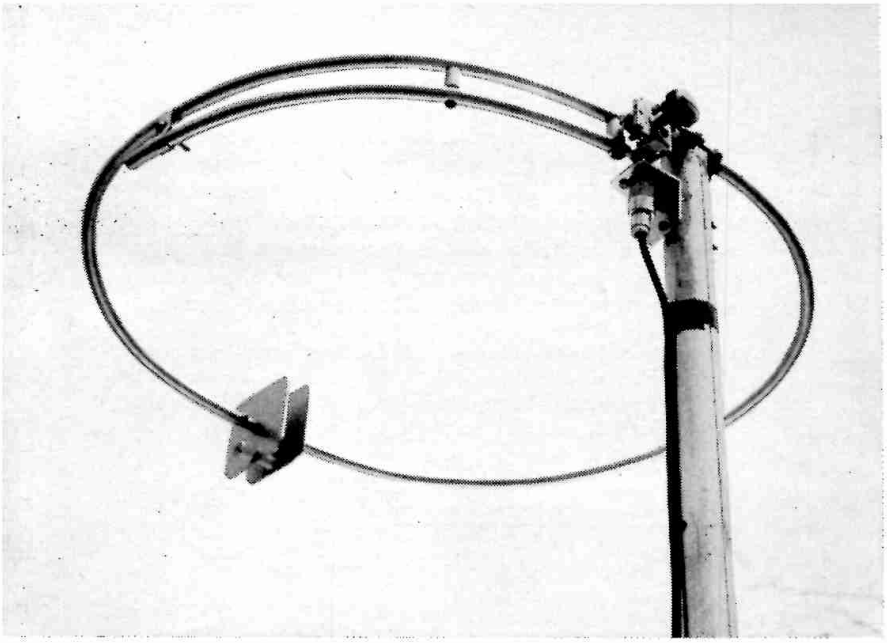


Fig. 19-39—The 2-band halo as it appears when set up for 50-Mc. operation. Changing to 144 Mc. involves decreasing the plate spacing by swapping cone insulators, and resetting the gamma matching clip and series capacitor.

with lock washers. The holes for mounting the ceramic cone spacer are drilled directly below the center, midway between the center and the edge of the capacitor plates.

The halo is set into a slot cut in the vertical support. This slot should be just big enough to permit the halo to be forced into it. The halo has to be stiffened, so cut it at the center and insert about 2 inches of aluminum rod, again turned down on a lathe to fit tightly inside the tubing. The two pieces of tubing are then pushed together, over the insert, and drilled each side of center to pass 6–32 screws. The halo and insert are also drilled at the midpoint, to pass the mounting screw. This is an 8–32 screw, 1¼ inches long. If lathe facilities are not available, the mounting of the capacitor plates and the securing of the halo to the vertical support can be handled with angle brackets.

Mechanical stability is important so straps of aluminum ½ inch wide are wrapped around the halo either side of the mounting post. These are bent at right angles and the ends pulled together with a bolt.

The matching arm is 14½ inches long, of the same material as the halo itself. It is mounted below the halo on two ¾-inch cone standoffs. For convenience in detaching the feed line a coaxial fitting is mounted on an L bracket bolted to the vertical support. The stator bar of the 25-$\mu\mu$f. variable capacitor (Johnson 167-2) is soldered directly to the coaxial fitting. The rotor of the capacitor is connected to the gamma arm through a piece of stiff wire. For further stiffening an aluminum angle bracket is screwed to the lower mounting stud of the capacitor and the other end mounted under the screw that holds the first cone standoff in place. Contact between the arm and the halo proper is made through a strap of ½-inch wide aluminum bent to form a sliding clip. Be sure that a clean tight contact is made between the tubing and the clip, as high current flows at this point. A poor or varying contact will ruin the effectiveness of the antenna.

Adjustment

The capacity-loaded halo is a high-Q device so it must be tuned on-the-nose, or it will not work properly. The only reliable method for adjusting a halo is to use a standing-wave bridge, making tuning and matching adjustments for minimum reflected power. Using a field-strength meter and attempting to adjust for maximum radiated power can give confusing indications, and is almost certain to result in something less than maximum effectiveness.

The adjustment process with this design can be simplified if the halo is first resonated approximately to the desired frequency ranges with the aid of a grid-dip meter. Set the clip at about one inch in from the end of the arm, and the series capacitor at the middle of its range. Check the resonant frequency of the loop with the grid-dip meter, with the ¾-inch spacer between the capacitor plates. It should be close to 50 Mc. If the frequency is too low, trimming the corners of the plates or putting shims under the ceramic spacer will raise it somewhat. If the frequency is too high already, make new and slightly larger capacitor plates.

Next, insert an s.w.r. bridge between the antenna and the transmission line. Apply power and swing the capacitor through its range, noting whether there is a dip in reflected power at any point. If the reflected power will not drop to zero, slide the clip along the gamma arm and retune the capacitor, until the lowest reading possible is obtained. If this is still not zero, the halo is not resonant. If the halo capacitance is on the low side, moving the hands near the plates will cause the reflected power to drop. Closer spacing of the plates, larger plates or a longer halo loop are possible solutions.

These adjustments should be made on a frequency near the middle of the range you expect to use. Adjusting for optimum at 50.25 Mc., for example, will result in usable operation over the first 500 kc. of the band, and a good match (below 1.5 to 1) from 50.1 to 50.4. The s.w.r. will rise rapidly either side of this range.

To tune up on 144 Mc., insert the ½-inch cone between the capacitor plates. Slide the clip back on the gamma arm about 3 to 4 inches and repeat the adjustment for minimum reflected power,

using a frequency at the middle of a 2-Mc. range. Tuning up at 145 Mc., for example, will give quite satisfactory operation from the low end to 146 Mc., the halo being much broader in frequency response when it is operated on its third harmonic. In this model the series capacitor in the gamma arm was at about the middle of its range for 50 Mc., and near minimum for 144 Mc. Slight differences in mechanical construction may change the value of capacitance required, so these settings should not be taken as important.

The photograph, Fig. 19-39, shows a method used to avoid running the chance that the second ceramic cone would be missing when a band change was to be made. The head was cut from a 6–32 screw, leaving a threaded stud about ½ inch long. This is screwed into one of the ceramic cones. The other cone then serves as a nut, to tighten down the capacitor plate. In changing bands merely swap cones. (Original description appeared in *QST*, Sept., 1958.)

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A Field-Strength Meter for Portable-Mobile Use

The field-strength meter of Figs. 19-40 through 19-42 can be used in a mobile station as an antenna-resonance indicator or as a continuous output indicator showing that the transmitting system is actually radiating. It is designed to be inserted between the automobile broadcast receiving antenna, which acts as the r.f. pick-up, and the broadcast receiver. Small magnets or rubber suction cups on the back plate will hold the meter securely on top of the car dash. Although in this position the meter will be face up in most cases, it can nevertheless usually be read from the driver position.

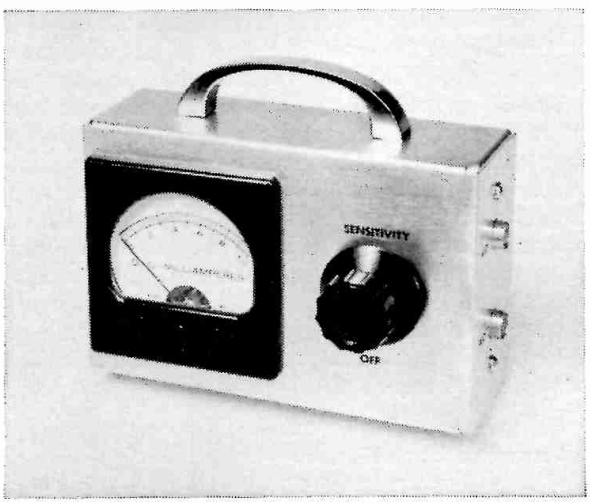


Fig. 19-40—A front view of the field-strength meter. Sensitivity control R_1 is to the right of the 0–1 indicating meter. Antenna input and output connectors are mounted on the right end of the box.

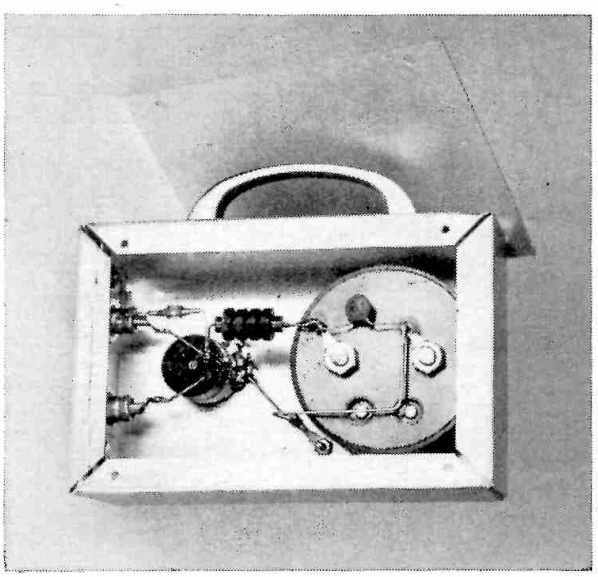


Fig. 19-42—Inside view of the meter. The back plate shown in the photograph is used as a cover for the box.

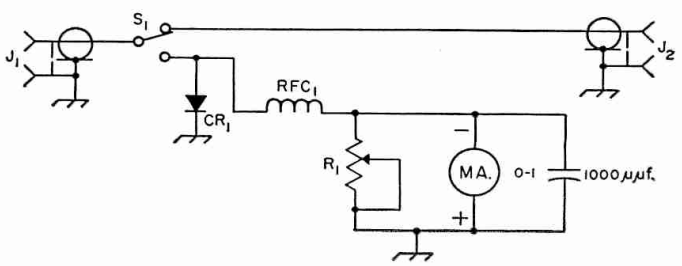


Fig. 19-41—Circuit of the field-strength meter.
CR_1—Crystal diode (1N34A).
J_1, J_2—Automobile type antenna connectors.
RFC_1—2.5 mh. r.f. choke.
R_1—500 ohm potentiometer (Mallory U-2).
S_1—S.p.d.t. switch for above potentiometer.

A handle can be mounted on the meter box so that the meter can easily be carried about for portable measurements. The same basic layout less the handle can be used if the box is to be mounted under the dash or in the glove compartment.

The circuit for the field-strength meter is shown in Fig. 19-41. The values shown are not critical. Nearly any type of crystal detector can be used and the meter movement can be anything from 100 μa. to 2 ma. or more, depending upon the size and placement of the antenna and the power output of the transmitter. All components, including the 3-inch indicating meter, are housed in a 2 × 6 × 4-inch aluminum chassis. If a smaller meter is used, the box could be reduced in size accordingly. However, in mobile operation a large meter is more convenient to read while in motion. An illuminated meter could be substituted for the one shown in the photograph for use at night. A switch, S_1, is used in the circuit to switch the antenna to the field-strength meter position or straight through to the broadcast set. For portable or temporary mobile operation, a short pick-up wire can be used instead of the automobile receiving antenna. The pick-up antenna lead comes into a connector mounted on one end of the box. There is a second connector for attaching the lead to the broadcast receiver.

Conelrad Monitoring

The conelrad rules discussed in the chapters on high-frequency receivers and operating a station must be observed by amateurs who operate mobile. One convenient form of compliance is by means of a separate tunable converter covering the broadcast band, and converting to the same i.f. as the i.f. used by the ham-band converter. This type of converter may also be used when the car radio is used as the tunable i.f. for a broad-band converter, providing that the receiver is tuned to the converter i.f. at ten-minute intervals. This can be accomplished most conveniently by setting one of the push buttons to tune the receiver to the monitor output frequency.

The circuit of a broadcast-band converter is shown in Fig. 19-43. The input circuit $C_{1A}L_2$ covers the broadcast band. The oscillator circuit $C_{1B}L_3$ tunes the range of 2050 to 3000 kc. to produce an i.f. of 1500 kc. A type 6SA7 may be used in the circuit and, of course, either a 12BE6 or a 12SA7 should be used for 12-volt operation.

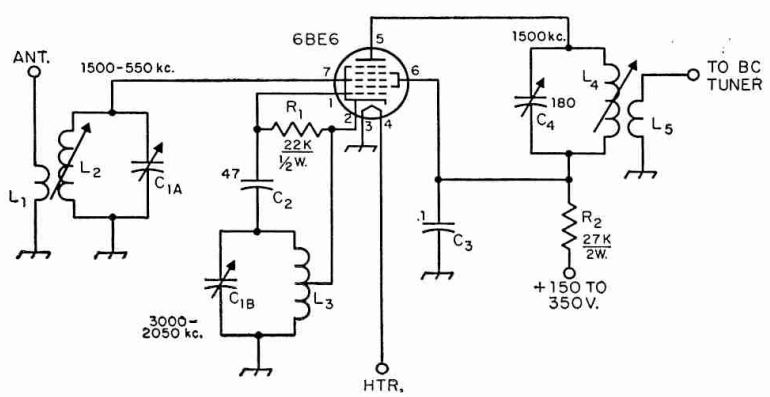
Plates must be removed from C_{1B} to provide the required tuning range. The oscillator section of the dual unit is the one having the smaller number of plates. Starting at the rear, all rotor plates except five should be removed. It isn't necessary to remove the unused stators. Be very careful to make sure that there are no shorted plates after the modification is complete.

L_2 is a ferrite-core loopstick. This coil usually comes with a length of wire attached to the ungrounded end and wound around the loopstick. When unwound, the short length of wire is intended to provide additional pickup if needed. Disconnect this wire from L_2 and, without unwinding it, use it for L_1.

L_3 is close-wound with 60 turns No. 30 enameled, and either tapped at about one third of the way up from the ground end, or with a separate cathode coil consisting of about one third the number of turns on L_3, wound over the ground end of L_3, and wound in the same direction. The bottom end of this winding should be grounded.

Power for the converter may be taken from the car radio supply since the current requirement is negligible. With 150 volts at the positive B terminal of the converter, the converter draws approximately 4 ma. and the drop across R_2 is about 100 volts. The converter will work well at supply voltages up to 350 or more without change in the resistance value of R_2. The current drain will, of course, be higher at the higher supply voltages, and the wattage rating of the resistor may have to be increased. If current drain is an important consideration, the resistance value of R_2 can be increased in proportion to the increase in supply voltage.

Fig. 19-43—Circuit of the conelrad converter for mobile use.



C1—Dual variable capacitor, broadcast-replacement type for superhet receivers, C_{1B} altered as described in the text (approx. 90 μμf.).
C2—47-μμf. mica.
C3—0.1-μf. 400-volt paper.
C4—180-μμf. mica trimmer (Arco type 463).
L1—See text.
L2—BC ferrite core loopstick (approx. 230 μh.).
L3—See text (approx. 65 μh.).
L4—National XR-50 iron-slug form wound full with No. 32 enam. wire (approx. 85 μh.).
L5—15 turns No. 28 wound over cold end of L_4.