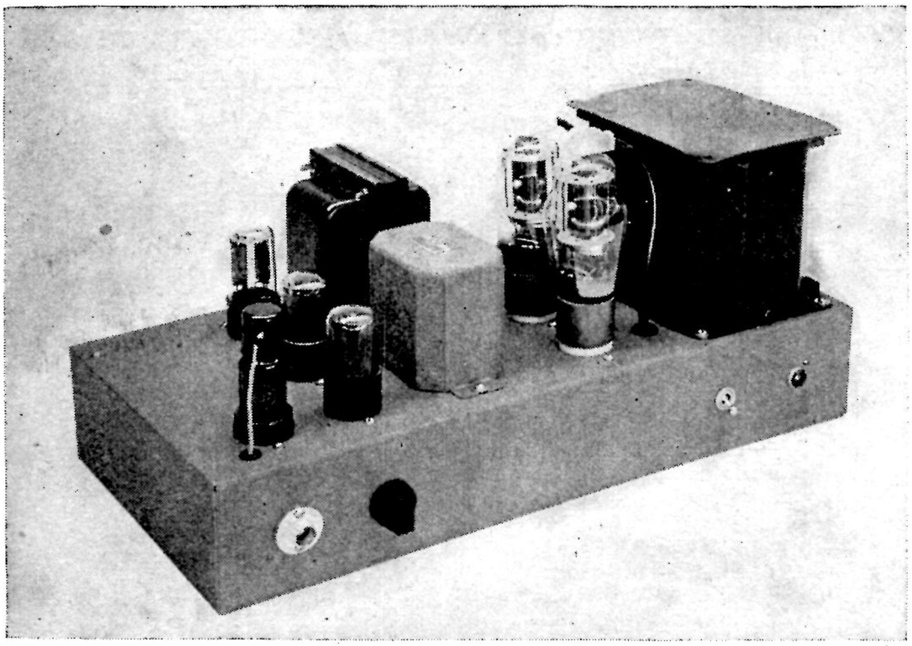


Fig. 9-33 — A speech amplifier and 807 modulator for plate modulation of transmitters up to 200 watts input. The microphone jack and the gain control are at the left end of the chassis. The audio components and tubes occupy the front section and the power supply for the driver tubes is laid out along the rear edge.

◆

cess of the rating of the 6.3-volt winding on the ordinary small power transformer.

Resistors R_{14} and R_{15} and condenser C_8 are placed in the 807 screen circuit to suppress the r.f. parasitic oscillations that sometimes occur with these tubes. Their use is principally a precautionary measure, and they may not be required in most installations.

The frequency response of this unit is maximum in the range from about 200 to 2500 cycles, for greatest voice effectiveness and mini-

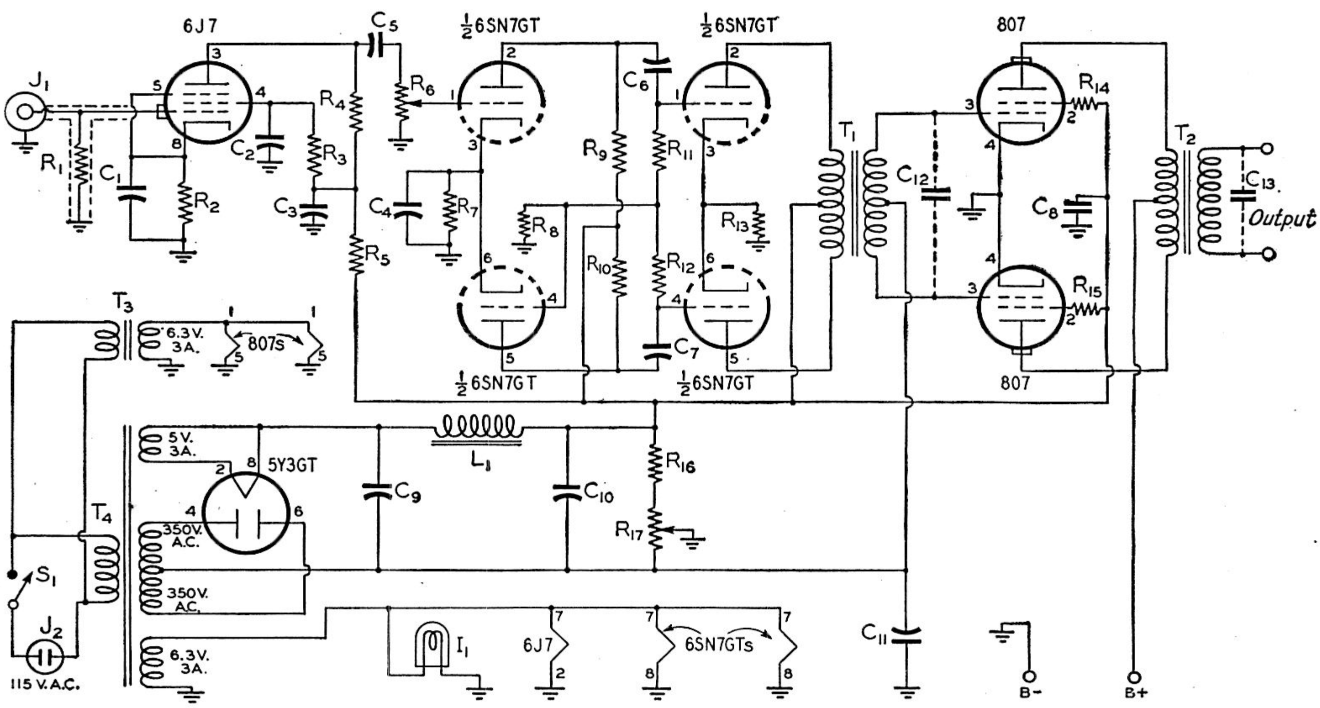


Fig. 9-34 — Circuit diagram of the p.p. 807 speech amplifier-modulator.

C_1 — 10-μfd. 50-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_9, C_{10} — 8-μfd. 450-volt electrolytic.
C_4, C_{11} — 50-μfd. 50-volt electrolytic.
C_5, C_6, C_7 — 0.01-μfd. 400-volt paper.
C_8 — 0.0068-μfd. mica.
C_{12} — 0.001-μfd. mica (see text).
C_{13} — 0.02-μfd. mica (see text).
R_1 — 1 megohm.
R_2, R_7 — 1500 ohms.
R_3 — 1.5 megohms.
R_4, R_8, R_{11}, R_{12} — 0.22 megohm.
R_5 — 47,000 ohms.
R_6 — 1-megohm volume control.
R_9, R_{10} — 0.1 megohm.
R_{13} — 470 ohms.
R_{14}, R_{15} — 100 ohms.
R_{16} — 15,000 ohms, 10 watts.
R_{17} — 1000-ohm wire-wound potentiometer.
(All resistors ½ watt unless otherwise noted.)
L_1 — Smoothing choke, 30 hy., 75 ma., 340-ohm d.c. resistance (Utah 4002).
I_1 — 6.3-volt a.c. pilot-lamp-and-socket assembly.
J_1 — Microphone-cable jack.
J_2 — Panel-mounting a.c. plug (Amphenol 61-M1).
S_1 — S.p.s.t. switch.
T_1 — Push-pull plates to push-pull grids (UTC S-9).
T_2 — Output transformer, type depending on requirements. A multitap transformer (UTC VM-3) is shown in photos.
T_3 — Filament transformer, 6.3 volts, 3 amp. (Thordarson T-21F10).
T_4 — Power transformer, 350 volts a.c. each side of center-tap, 70-ma. rating. Filament windings: 5 v., 3 amp.; 6.3 v., 3 amp. (Stancor P-4078).

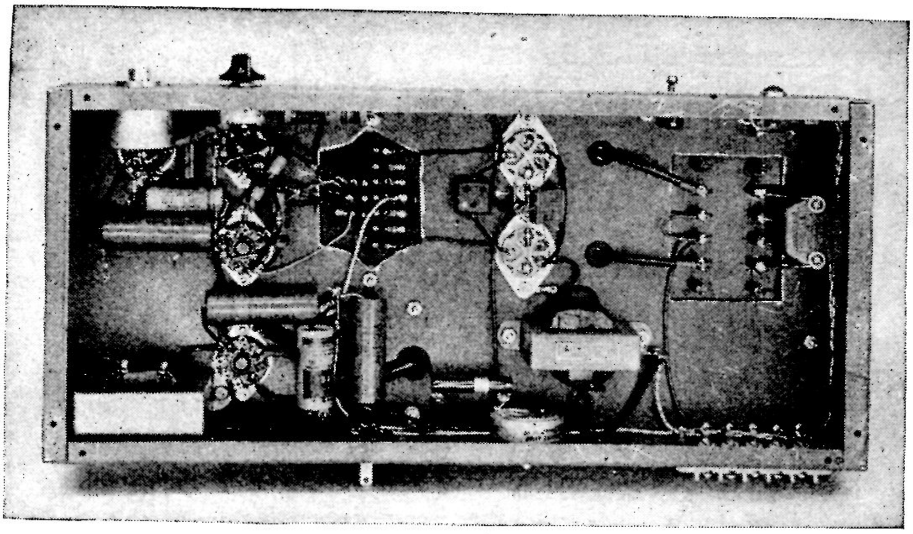


Fig. 9-35 — Below-chassis view of the 807 modulator. The shielded microphone jack is in the upper left-hand corner. The filter choke is mounted in the lower left-hand corner and the 807 filament transformer is to the rear and slightly to the right of the 807 tube sockets. The condenser for attenuating the high audio frequencies, shown at the right-hand end of the chassis, is supported by No. 12 wire leads which connect to the output terminals of the modulation transformer.

mum width of the r.f. channel. Frequencies above 2500 cycles are attenuated by condensers C_{12} and C_{13}, the former across the secondary of the driver transformer and the latter across the secondary of the output transformer. The capacitance values given are about optimum for the types of transformers specified and should be close to optimum for other transformers of similar ratings. The voltage rating of C_{13} should be at least equal to the d.c. voltage on the modulated r.f. amplifier.

Construction

The photographs show the general layout of components. The 6J7 and 6SN7GT phase inverter are in line at the left-hand front edge of the chassis. The 6SN7GT driver and 5Y3GT rectifier are to the rear of the phase inverter. The driver transformer, T_1, is at the front and the power transformer, T_4, is at the rear. Plate leads for the 807s run through rubber grommets in the chassis.

The bottom view shows the by-pass condensers and resistors grouped around the sockets to which they connect. The bias-control potentiometer, R_{17}, is mounted on the rear edge of the chassis, and the power-supply bleeder, R_{16}, is mounted between R_{17} and an insulated tie-point on the chassis near the a.c. input socket (also on the rear chassis edge). A jack shield (National JS-1) covers the microphone jack, and the first-stage grid resistor, R_1, is mounted inside this shield. The lead to the 6J7 grid cap must be shielded and the shield grounded.

The No. 1 terminals of the driver transformer specified should be connected to the grids of the 807s. If a different transformer is used, it should have a primary-to-secondary ratio (total) of about 1-to-1 to couple the 6SN7GT and 807 grids properly. The output-transformer turns ratio will depend on the type of operation selected and the modulating impedance of the Class C amplifier. Operated at ICAS ratings, the 807s will deliver a tube output of 120 watts into a plate-to-plate load of 6950 ohms. This requires a plate supply capable of delivering 240 ma. at 750 volts. At CCS ratings the tubes will deliver 80 watts into a 6400-ohm load and require a 600-volt 200-ma. plate supply. The bias should be set, by means of R_{17}, to give -32 volts between the potentiometer arm and chassis for ICAS operation, and to -30 volts for CCS operation.

Drivers for Class-B Modulators

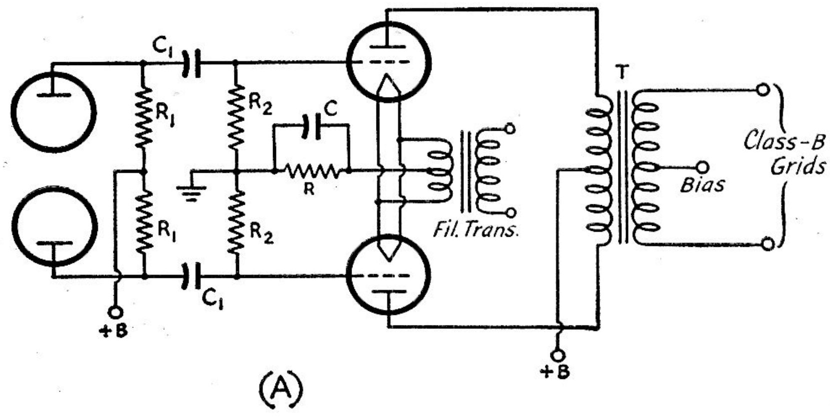
Driving Power

Class B amplifiers are driven into the grid-current region, so power is consumed in the grid circuit. The preceding stage or driver must be capable of supplying this power at the required peak audio-frequency grid-to-grid voltage. Both of these quantities are given in the manufacturer's tube ratings. The grids of the Class B tubes represent a variable load resistance over the audio-frequency cycle, because the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source that will maintain the waveform of the signal without distortion even though the load varies. That is, the driver stage must have good regulation. To this end, it should be capable of delivering somewhat more power than is consumed by the Class B grids, as previously described in the discussion on speech amplifiers. It is also desirable to use an input coupling transformer having a turns ratio giving the largest step-down in the voltage between the driver plate or plates and the Class B grids that will permit obtaining the specified grid-to-grid a.f. voltage.

The peak output voltage at the primary of the driver transformer is

$$E_o = 1.4\sqrt{PR}$$

where E_o = Peak value of a.f. output voltage
P = Rated power output of driver
R = Rated load impedance for driver



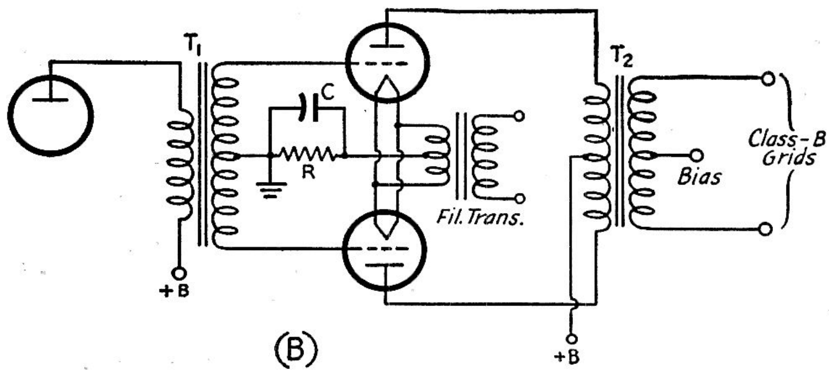


Fig. 9-36 — **Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling condenser and grid resistor, respectively; values also may be taken from Table 9-I.**

In both circuits the output transformer, T, T_2, should have the proper turns ratio to couple between the driver tubes and the Class B grids. T_1 in B is usually a 2:1 transformer, secondary to primary. R, the cathode resistor, should be calculated for the particular tubes used. The value of C, the cathode by-pass, is determined as described in the text.

This is correct when the driver is working into its rated load. When the load is the grid circuit of a Class B modulator, the impedance varies over the audio-frequency cycle. It reaches a minimum value when the signal reaches a positive peak and the grid current has its maximum instantaneous value. The best that can be done is to assume that, if the driver has more than enough power output, the turns ratio of the driver transformer should be

$$N = \frac{E_o}{E}$$

where N = Driver-transformer turns ratio, primary to secondary

E_o = Peak output voltage of driver

E = Peak grid-to-grid voltage required by Class B tubes.

Example: A pair of Class B tubes requires a driving power of 7 watts and a peak a.f. grid-to-grid voltage of 200 for maximum output. The driver has a maximum output of 10 watts when working into a 5000-ohm load. The peak output voltage at the primary of the driver transformer is

$$E_o = 1.4\sqrt{PR} = 1.4\sqrt{10 \times 5000}$$
$$= 1.4 \times 224 = 314 \text{ volts}$$

The driver-transformer turns ratio, primary to secondary, should be

$$N = \frac{E_o}{E} = \frac{314}{200} = 1.57 \text{ to } 1$$

Commercial transformers frequently are designed for specific driver-modulator combinations, and the turns ratio is chosen to give as good driver regulation as the conditions will permit.

The driver transformer, T or T_2 in Fig. 9-36, may couple directly between the driver tube and the modulator grids or may be designed to work into a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the output of the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator; the second transformer not only introduces additional losses but also impairs the voltage regulation of the driver stage.

Driver Tubes

The variation in grid resistance of a Class B amplifier over the audio-frequency cycle poses a special problem in the driver stage. To avoid distortion, the driver output *voltage* (not power) must stay constant (for a fixed signal voltage on its grid) regardless of the variations in load resistance.

The fundamental requirement for good voltage regulation in any electrical generator is that the internal resistance must be low. In a vacuum-tube amplifier, this means that the tubes must have a low value of plate resistance. The best tubes in this respect are low-μ triodes (the 6A3 is an example) and the worst are tetrodes and pentodes as represented by the 6V6 and 6L6. This does not mean that tetrodes (or pentodes) cannot be used, but it does mean that they should not be used without taking measures to reduce the effective plate resistance (see next section).

In selecting a driver stage always choose Class A or AB_1 operation in preference to Class AB_2. This not only simplifies the speech-amplifier design but also makes it easier to apply negative feed-back to tetrodes for reduction of plate resistance. It is possible to obtain a tube power output of approximately 25 watts (from 6L6s) without going beyond Class AB_1 operation; this is ample driving power for the popular Class B modulator tubes, even when a kilowatt transmitter is to be modulated.

The rated tube output (as shown by the tube tables) should be reduced by about 20 per cent to allow for losses in the Class B input transformer. If two transformers are used, tube-to-line and line-to-grids, allow about 35 per cent for transformer losses. Another 25 per cent should be allowed, if possible, as a safety factor and to improve the voltage regulation.

Fig. 9-36 shows representative circuits for a push-pull triode driver using cathode bias. If the amplifier operates Class A, the cathode resistor need not be by-passed, because the a.f. currents from each tube flowing in the cathode resistor are out of phase and cancel each other. However, in Class AB operation this is not true; considerable distortion will be generated at high signal levels if the cathode resistor is not by-passed. The by-pass capacitance required can be calculated by a simple rule: the cathode resistance in ohms multiplied by the by-pass capacitance in microfarads should equal at least 25,000. The voltage rating of the condenser should be equal to the maximum bias voltage. This can be found from the maximum-signal plate current and the cathode resistance.

Example: A pair of 6A3s is to be used in Class AB_1, self-biased. From the tube tables, the cathode resistance should be 780 ohms and the maximum-signal plate current 120 ma. From Ohm's Law,

$$E = RI = 780 \times 0.12 = 93.6 \text{ volts}$$

From the rule mentioned previously, the by-pass capacitance required is

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

A 40- or 50-μfd. 100-volt electrolytic condenser would be satisfactory.

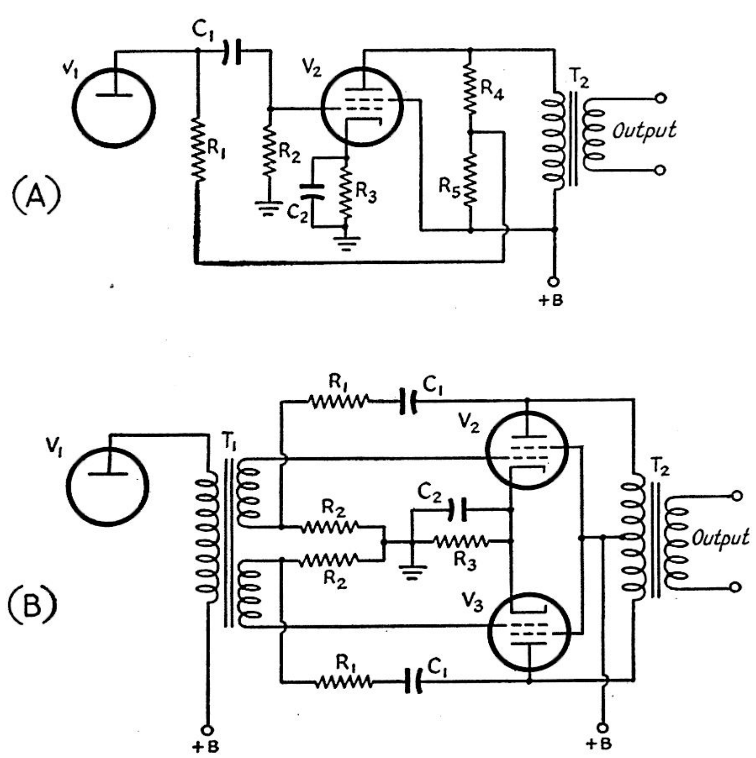


Fig. 9-37 — Negative feed-back circuits for drivers for Class B modulators. A — Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μfd.; C_2, 50 μfd.
B — Push-pull beam-tetrode driver. If V_1 is a 6J5 and V_2 and V_3 6L6s, the following values are suggested: R_1, 0.1 megohm; R_2, 22,000 ohms; R_3, 250 ohms; C_1, 0.1 μfd.; C_2, 100 μfd.

Negative Feed-Back

Whenever tetrodes or pentodes are used as drivers for Class B modulators, negative feed-back should be used in the driver stage. This will reduce the distortion caused by the variable load resistance represented by the Class B grids. It also reduces the distortion inherent in the driver stage itself, when properly applied. The effect of feed-back is to reduce the apparent plate resistance of the driver, and this in turn helps to maintain the a.f. output voltage at a more constant level (for a constant signal on the grid) when the load resistance varies. It is readily possible to reduce the plate resistance to a value comparable to or lower than that of low-μ triodes such as the 6A3.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-37. Fig. 9-37A shows resistance coupling between the preceding stage and a single tetrode, such as the 6V6, that operates at the same plate voltage as the preceding stage. Part of the a.f. voltage across the primary of the output transformer is fed back to the grid of the tetrode, V_2, through the plate resistor of the preceding tube, V_1. The amount of voltage so fed back is determined by the voltage divider, R_4R_5. The total resistance of R_4 and R_5 in series should be large compared to the rated load resistance of V_2. Instead of the voltage divider, a tap on the transformer primary can be used to supply the feed-back voltage, if such a tap is available.

The amount of feed-back voltage that appears *at the grid* of tube V_2 is determined by R_1, R_2 and the plate resistance of V_1, as well as by the relationship between R_4 and R_5. Calculation of the feed-back voltage, although not mathematically difficult, is not ordinarily practicable because the plate resistance of V_1 is seldom known at the particular operating conditions used. Circuit values for a typical tube combination are given in detail in Fig. 9-37.

The push-pull circuit in Fig. 9-37B requires an audio transformer with a split secondary. The feed-back voltage is obtained from the plate of each output tube by means of the voltage divider, R_1R_2. The blocking condenser, C_1, prevents the d.c. plate voltage from being applied to R_1R_2; the reactance of this condenser should be low, compared to the sum of R_1 and R_2, at the lowest audio frequency to be amplified. Also, the sum of R_1 and R_2 should be high compared to the rated load resistance for V_2 and V_3.

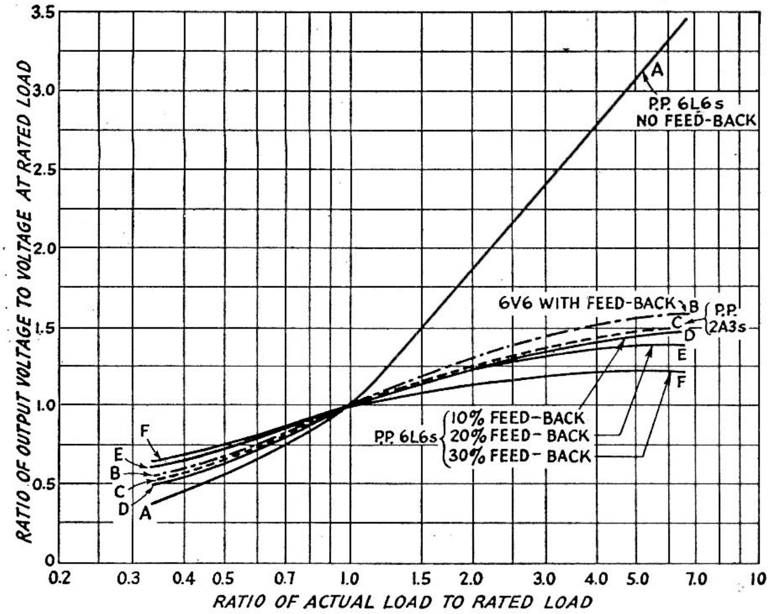


Fig. 9-38 — Output voltage regulation of two types of beam-tetrode drivers with negative feed-back. For comparison, the regulation with a pair of 2A3s (no feed-back) also is shown.

where n is the feed-back percentage, and R_1 and R_2 are connected as shown in the diagram. The higher the feed-back percentage, the lower the effective plate resistance. However, if the percentage is made too high the preceding tube, V_1, may not be able to develop enough voltage, through T_1, to drive the push-pull stage to maximum output without itself generating harmonic distortion. Distortion in V_1 is not compensated for by the feed-back circuit. If V_2 and V_3 are 6L6s operated self-biased in Class AB_1 with a load resistance of 9000 ohms, V_1 is a 6J5, and T_1 has a turns ratio of 2-to-1, total secondary to primary, it is possible to use over 30-per-cent feed-back without going beyond the output-voltage capabilities of the 6J5. Actually, it is unnecessary to use more than about 20-per-cent feed-back. This value reduces the effective plate resistance to the point where the output voltage regulation is better than that of 6A3s or 2A3s without feed-back.

In this circuit the feed-back voltage that is developed across R_2 also appears at the grid of V_2 (or V_3) because there is no appreciable current flow (in the usual audio range) through the transformer secondary and grid-cathode circuit of the tube, provided the tubes are not driven to grid current. The circuit should not be used with tubes that are operated Class AB_2. The per cent feed-back is

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

Instead of the voltage-divider arrangement shown in Fig. 9-37B for obtaining feed-back voltage, a separate winding on the output transformer can be used, provided it has the proper number of turns to give the desired feed-back percentage. Special transformers are available for this purpose.

The improvement in constancy of output voltage resulting from the use of negative feed-back is shown graphically in Fig. 9-38.

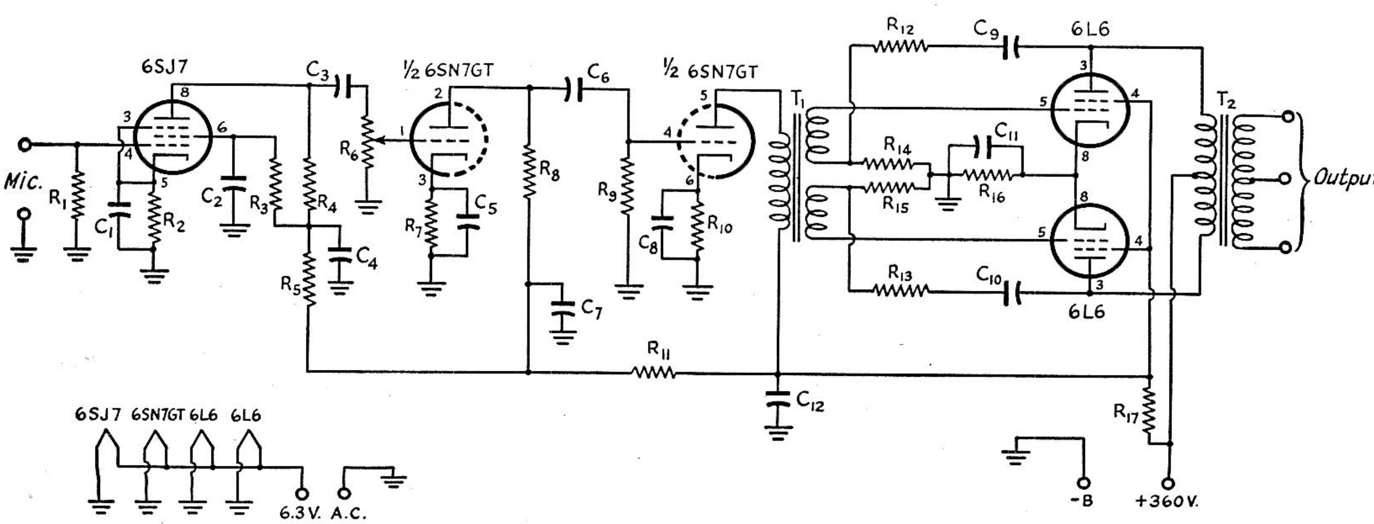


Fig. 9-39 — Circuit diagram of speech amplifier using 6L6s with negative feed-back, suitable for driving Class B modulators up to 500 watts output.

C_1, C_5, C_8 — 20-μfd. 25-volt electrolytic.
C_2, C_9, C_{10} — 0.1-μfd. 400-volt paper.
C_3, C_6 — 0.01-μfd. 600-volt paper.
C_4, C_7, C_{12} — 10-μfd. 450-volt electrolytic.
C_{11} — 100-μfd. 50-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_8 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 10,000 ohms, ½ watt.
R_{12}, R_{13} — 0.1 megohm, 1 watt.
R_{14}, R_{15} — 22,000 ohms, ½ watt.
R_{16} — 250 ohms, 10 watts.
R_{17} — 2000 ohms, 10 watts.
T_1 — Interstage audio, 2:1 secondary (total) to primary, with split secondary winding.
T_2 — Class B input transformer to suit modulator tubes.

In order to compare the various types of tubes, the variation in output voltage is shown as a percentage of the output voltage when the tubes are working into the rated load. The load resistance also is expressed as a percentage of the rated load resistance for the particular tube, or pair of tubes, used.

A circuit for a speech-amplifier suitable for driving a Class B modulator is given in Fig. 9-39. In this amplifier the 6L6s are operated Class AB_1 and will deliver up to 20 watts to the grids of the Class B amplifier. The feedback circuit requires no adjustment, but does require an interstage transformer with two separate secondary windings (split secondary).

This amplifier may be constructed along the same lines as in Fig. 9-29, observing the same precautions with respect to shielding the 6SJ7 grid circuit. Although the power output is the same as from the amplifier of Fig. 9-32, an additional voltage-amplifier stage is incorporated in the circuit. This is necessary because the voltage fed back from the plates to the grids of the 6L6s opposes the voltage from the preceding stage, so the latter must be increased in order to maintain the same power output from the 6L6s. In turn, this necessitates more over-all voltage gain than is required to drive Class AB_1 push-pull 6L6s without feedback.

The output transformer, T_2, should be selected to work between a 9000-ohm plate-to-plate load and the grids of whatever Class B tubes will be used. The power-supply requirements for this amplifier are essentially the same as for the amplifier of Fig. 9-32.

Class-B Modulators

Plate modulation of all but low-power transmitters requires so much audio power that the Class B amplifier is the only practical type to use. (Included in the Class B category are high-power modulators of the Class AB_2 type; whether the operation is in one class or the other is principally a matter of degree.)

Class B modulator circuits are practically identical no matter what the power output of the modulator. The diagrams of Fig. 9-40 therefore will serve for any modulator of this type that the amateur may elect to build. The triode circuit is given at A and the circuit for tetrodes at B. When small tubes with indirectly-heated cathodes are used, the cathodes should be connected to ground.

Modulator Tubes

Class B audio ratings of various types of transmitting tubes are given in the tube tables of Chapter Twenty-Five. Choose a pair of tubes that is capable of delivering sine-wave audio power equal to somewhat more than half the d.c. input to the modulated Class C amplifier. It is sometimes convenient to use tubes that will operate at the same plate voltage as that applied to the Class C stage, because one power supply of adequate current capacity may then suffice for both stages. However, available power-supply components do not always permit this, and better over-all performance and economy may result from the use of separate power supplies.

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. The efficiency of the output transformer will vary with its construction, and may be assumed to be in the vicinity of 80 per cent for the less-expensive units and somewhat higher for higher-priced transformers. To be adequate for modulating the transmitter, therefore, the modulator should have a theoretical power capability about 25 per cent greater than the actual power needed for modulation.

It is always desirable to provide more power *capability* in the modulator than actually is

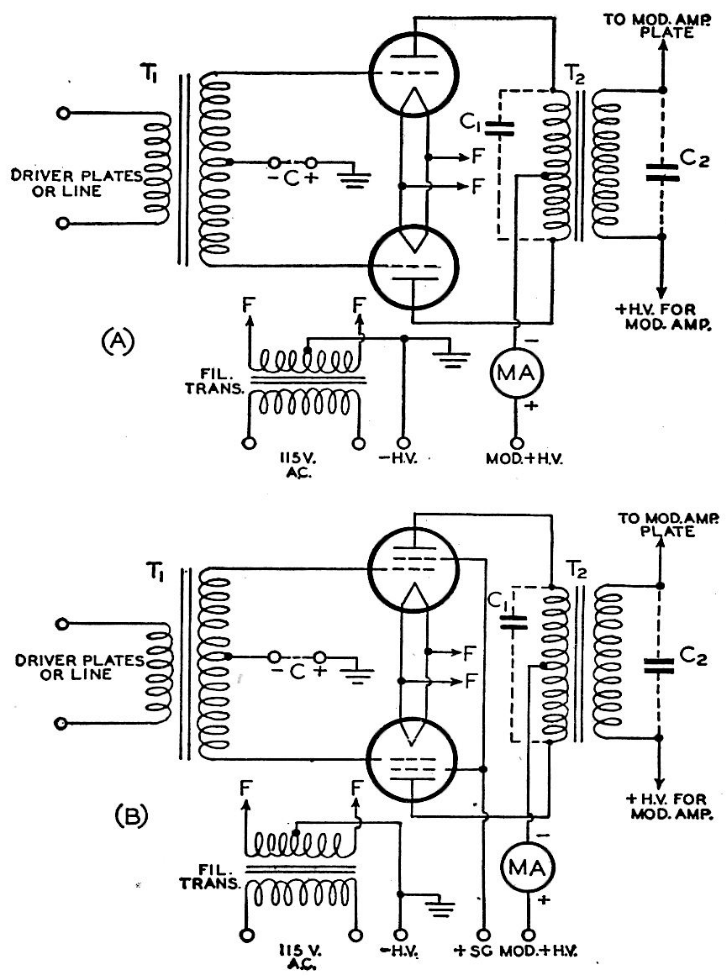


Fig. 9-40 — **Class B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.**

needed. Then the tubes need not be worked to the limit of their capacity, with the result that there is minimum distortion and therefore no audio harmonics — and no consequent broadening of the r.f. channel.

Matching to Load

In giving Class B ratings on power tubes, manufacturers specify the plate-to-plate load impedance into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance of the Class C r.f. stage, so a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

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

where N = Turns ratio, primary to secondary
Z_m = Modulating impedance of Class C r.f. amplifier
Z_p = Plate-to-plate load impedance for Class B tubes

Example: The modulated r.f. amplifier is to operate at 1250 volts and 250 ma. The power input is

$$P = EI = 1250 \times 0.25 = 312 \text{ watts}$$

so the modulating power required is 312/2 = 156 watts. Increasing this by 25% to allow for losses and a reasonable operating margin gives 156 × 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

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

Commercial Class B output transformers usually are rated to work between specified primary and secondary impedances and frequently are designed for specific Class B tubes. In such a case, it will be unnecessary to calculate the turns ratio when the recommended tube combination is used. Many transformers are provided with primary and secondary taps, so that various turns ratios can be obtained to meet the requirements of various tube combinations.

It may be that the exact turns ratio required by a particular tube combination cannot be secured, even with a tapped modulation transformer. *Small* departures from the proper turns ratio will have no serious effect if the modulator is operating well within its capabilities; if the actual turns ratio is within 10 per cent of the ideal value the system will operate satisfactorily. Where the discrepancy is larger, it is always possible to choose a new set of operating conditions for the Class C stage to give a modulating impedance that can be matched by the turns ratio of the available transformer. This may require operating the Class C amplifier at higher voltage and less plate current, if the modulating impedance must be increased, or at lower voltage and higher current if the modulating impedance must be decreased. However, this process cannot be carried too far without exceeding the ratings of the Class C tubes for either plate voltage or current, even though the power input is kept at the same figure. In such a case the only solution is to operate at reduced input and use less of the power available from the modulator.

Suppressing Audio Harmonics

Distortion in either the driver or Class B modulator itself will cause a.f. harmonics that may lie outside the frequency band needed for intelligible speech transmission. While it is almost impossible to avoid some distortion, it *is* possible to cut down the amplitude of the higher-frequency harmonics. The purpose of condensers C_1 and C_2 across the primary and secondary, respectively, of the Class B output transformer in Fig. 9-40 is to

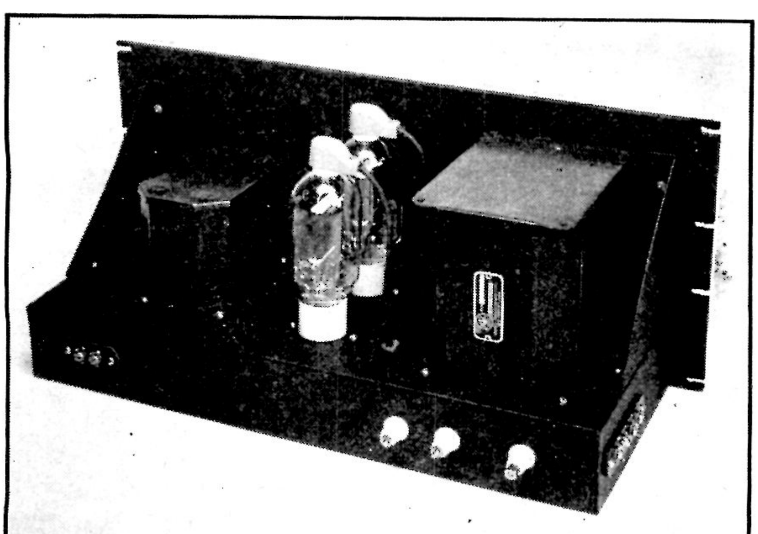


Fig. 9-41 **— A conventional chassis arrangement for low- and medium-power Class B modulator stages. The mechanical layout in general follows the typical circuit diagrams given in Fig. 9-40.**

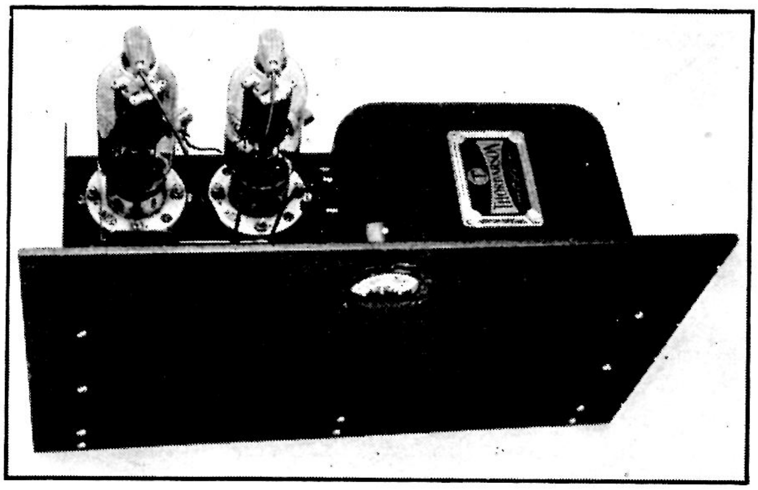


Fig. 9-42 — **A chassis arrangement for a higher-power Class B modulator. This unit has the filament transformer for the tubes mounted on the chassis. Where the input transformer is included with the speech amplifier, less chassis space will be needed. The tubes are placed near the rear, where the ventilation is good. The plate milliammeter is provided with a small plate over the adjusting screw, to prevent touching the screw accidentally. A Presdwood panel was used for this modulator; with a metal panel, the meter should be mounted behind glass on a well-insulated mount (the meter insulation is not intended for voltages above a few hundred) or connected in the filament center-tap rather than in the high-voltage lead.**

♦

reduce the strength of harmonics and unnecessary high-frequency components existing in the modulation.

The condensers act with the leakage inductance of the transformer winding to form a rudimentary low-pass filter. The values of capacitance required will depend on the load resistance (modulating impedance of the Class C amplifier) and the leakage inductance of the particular transformer used. In general, capacitances between about 0.001 and 0.006 μfd. will be required; the larger values are necessary with the lower values of load resistance. A test set-up for measuring frequency response (described in a later section in this chapter) will quickly show the optimum values to use, if a small assortment of condensers is on hand for experimenting. The object is to find the combination of C_1 and C_2 that will give the most rapid reduction in response as the signal frequency is raised above about 2500 cycles.

The voltage rating of each condenser should at least be equal to the d.c. voltage at the transformer winding with which it is associated. In the case of C_2, part of the total capacitance required usually is supplied by the plate by-pass or blocking condenser of the modulated amplifier, so C_2 need only be large enough to make up the difference.

Grid Bias

Many modern transmitting tubes designed for Class B audio work can be operated without grid bias. Besides eliminating the need for a grid-bias supply, this reduces the variation in grid impedance over the audio-frequency cycle and thus gives the driver a more constant load into which to work. With these tubes, the grid return lead from the center-tap of the driver transformer secondary is simply connected to the filament center-tap or cathode ground.

When the tubes require bias, it should always be supplied from a *fixed* voltage source. Neither cathode bias nor grid-leak bias can be used with a Class B amplifier; with both types the bias changes with the amplitude of the signal voltage, whereas proper operation demands that the bias voltage be unvarying no matter what the strength of the signal. When only a small amount of bias is required it can be obtained conveniently from a few dry cells. When greater values of bias are required, a heavy-duty "B" battery may be used if the grid current does not exceed 40 or 50 milliamperes on voice peaks. Even though the batteries are charged by the grid current rather than discharged, a battery will deteriorate with time and its internal resistance will increase. When the increase in internal resistance becomes appreciable, the battery tends to act like a grid-leak resistor and the bias varies with the applied signal. Batteries should be checked with a voltmeter occasionally while the amplifier is operating. If the bias varies more than 10 per cent or so with voice excitation the battery should be replaced.

As an alternative to batteries, a regulated bias supply may be used. This type of supply is described in Chapter Seven.

Plate Supply

The plate supply for a Class B modulator should be sufficiently well filtered to prevent hum modulation of the r.f. stage. An additional requirement is that the output condenser of the supply should have low reactance, at 100 cycles or less, compared to the load into which each tube is working. (This load is one-fourth of the plate-to-plate load resistance.) A 4-μfd. output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply, usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

To keep distortion at a minimum, the voltage regulation of the plate supply should be as good as it can be made. If the d.c. output voltage of the supply varies with the amount of current taken, it should be kept in mind that the voltage at *maximum* current determines the amount of power that can be taken from the modulator without distortion. A supply whose voltage drops from 1500 at no load to 1250 at the full modulator plate current is a 1250-volt supply, so far as the modulator is concerned, and any estimate of the power output available should be based on the lower figure.

It is particularly important, in the case of a tetrode Class B stage, that the screen-voltage power-supply source have excellent regulation, to prevent distortion. The screen voltage should be set as exactly as possible to the recommended value for the tube.

● IMPROPER OPERATION

Overexcitation

When a Class B amplifier is overdriven in an attempt to secure more than the rated power, distortion increases rapidly. The high-frequency harmonics which result from the distortion modulate the transmitter, producing spurious sidebands which can cause serious interference over a band of frequencies several times the channel-width required for speech. This may happen even though the transmitter is not being overmodulated. It *will* happen if the modulator is incapable of delivering the power required to modulate the transmitter fully, or if the Class C amplifier is not adjusted to give the proper modulating impedance.

As previously stated, the tubes used in the Class B modulator should be capable of somewhat more than the power output nominally required. In addition, the Class C amplifier should be adjusted to give the proper modulating impedance and the correct output transformer turns ratio should be used. Even though means may be incorporated in the speech amplifier to attenuate frequencies above those necessary for intelligible speech, it is still possible for high-frequency sidebands to be radiated if distortion occurs in the modulator, or if the transmitter is overmodulated. Such high-frequency harmonics as may be generated in the modulator can be reduced by connecting condensers across both the primary and secondary of the output transformer as previously described.

Operation Without Load

Excitation should never be applied to a Class B modulator until after the Class C amplifier is turned on and is drawing the value of plate current required to present the rated load to the modulator. With no load to absorb the power, the primary impedance of the transformer rises to a high value and excessive audio voltages are developed across it — frequently high enough to break down the transformer insulation. If the modulator is to be tested separately from the transmitter, a resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the transformer secondary.

Checking Speech Equipment

One way to check the performance of speech equipment is to put the complete transmitter on the air and depend on the comments of other amateurs. If it turns out that anything is wrong, fixing it becomes a slow and rather painful process not only to those concerned but to those who have to put up with the interference it causes. It is also a not altogether reliable method, since the reports are necessarily biased by the receiving operator's opinions of what is good or bad, to say nothing of the reluctance of most operators to be wholly frank — they don't want to hurt your feelings or appear to be casting doubt on your abilities.

The other method is to check it yourself, with the help of some measuring gear. An adequate job can be done with equipment that is neither elaborate nor expensive. A simple set-up is shown in Fig. 9-43. The only equipment that is not likely to be already at hand is the audio oscillator (the construction of a very simple one is described in Chapter Sixteen). The voltmeter — one that operates at audio frequencies is necessary — is available in any multirange volt-ohm-milliammeter that has a rectifier-type a.c. range. The headset is included for aural checking of the amplifier performance.

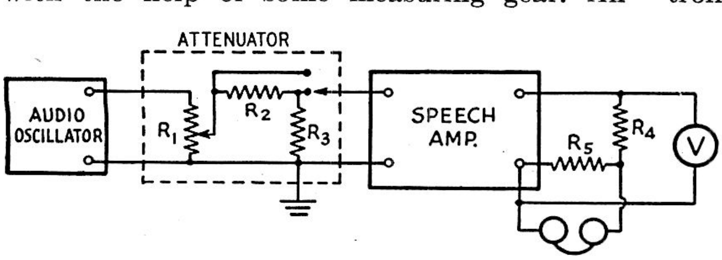


Fig. 9-43 — Simple test set-up for checking a speech amplifier. The audio-oscillator frequency range should be from about 100 to 5000 cycles. It is not necessary that it be continuously variable; a number of "spot" frequencies will be satisfactory. Suitable resistor values are: R_1, 50,000-ohm potentiometer; R_2, 4700 ohms; R_3, 470 ohms; R_4, rated load resistance for amplifier output stage; R_5, determine by trial for comfortable headphone level (25 to 100 ohms, ordinarily). V is a high-resistance a.c. voltmeter, multirange rectifier type.

A two-step attenuator for the output of the audio oscillator is recommended so that a wide range of output voltages can be smoothly controlled. Also, R_3 should have relatively low resistance — 500 ohms or less; operating at low impedance will minimize stray hum pick-up, which might cause false results when the amplifier gain is high.

As a preliminary check, cover the microphone input terminals with a metal shield (with the audio oscillator and attenuator disconnected) and, while listening in the headset, note the hum level with the amplifier gain control in the off position. The hum should be very low under these conditions. Then increase the gain-control setting to maximum and observe the hum; it will no doubt increase. Then connect the audio oscillator and attenuator and, starting from minimum

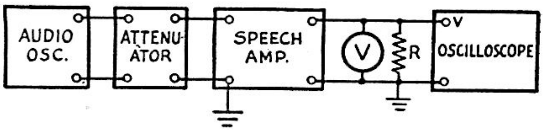


Fig. 9-44 — Test set-up using an oscilloscope for checking waveform.

signal, increase the setting of R_1 until the voltmeter indicates full power output. (The voltage should equal $\sqrt{PR}$, where P is the expected power output in watts and R is the load resistance — R_4 in the diagram.) Listen carefully to the tone while increasing R_1 to see if there is any change in its character. When it begins to sound like a musical octave instead of a single tone, distortion is beginning. Assuming that the output is substantially without audible distortion at full output, substitute the microphone for the audio oscillator and speak into it in a normal tone while watching the voltmeter. Reduce the gain-control setting until the meter "kicks" nearly up to the full-power reading on voice peaks. Note the hum level, as read on the voltmeter, at this point; the hum level should not exceed one or two per cent of the voltage at full output.

If the hum level is too high, the amplifier stage that is causing the trouble can be located by temporarily short-circuiting the grid of each tube, in turn, to ground. When shorting a particular grid makes a marked decrease in hum, the hum presumably is coming from a *preceding* stage, although it is possible that it is getting its start in that particular grid circuit. If shorting a grid does *not* decrease the hum, the hum is originating either in the plate circuit of that tube or the grid circuit of the next. Aside from wiring errors or a defective tube, objectionable hum usually originates in the first stage of the amplifier.

If distortion occurs below the point at which the expected power output is secured, the stage in which it is occurring can be located by working from the last stage toward the front end of the amplifier, applying a signal to each grid in turn from the audio oscillator and adjusting the signal voltage for maximum output. In the case of push-pull stages, the signal may be applied to the primary of the interstage transformer — *after* disconnecting it from the plate-voltage source. Assuming that normal design principles have been followed and that all stages are theoretically working within their capabilities, the probable causes of distortion are wiring errors (such as accidental short-circuit of a cathode resistor), defective components, or use of wrong values of resistance in cathode and plate circuits.

An oscilloscope having amplifiers and a linear sweep circuit is a useful instrument in testing audio amplifiers because it provides a ready check on waveform and thus shows distortion instantly. It may be connected across the output circuit as shown in Fig. 9-44, and also may be moved from stage to stage to check the waveform at the grid as well as at the plate. When connected to circuits that are not at ground potential for d.c., a condenser (about 0.1 μfd.) should be connected in series with the "hot" oscilloscope lead. The hot lead preferably should be shielded so that it will not pick up stray hum and introduce it into the amplifier.

● CLASS-B MODULATORS

Once the speech amplifier is in satisfactory working condition, a Class B modulator can be checked by similar means. A circuit is shown in Fig. 9-45. The resistance of R_1 should be equal to the modulating impedance of the Class C amplifier to be modulated, and the resistor should have a power rating equal to the rated power output of the modulator. Calculate the voltage to be expected across R_1 at full output; if it exceeds the range of the meter the meter may be connected across say half or one-fourth of R_1 and the readings multiplied by 2 or 4, respectively. Only a few ohms will be needed at R_2, in the average case, to give a good signal in the headphones. As a safety precaution, ground the output terminal to which the headphones are connected and use a resistor at R_2 that has ample current-carrying capacity.

Hum will seldom be a problem in the modulator. Distortion may be checked as described

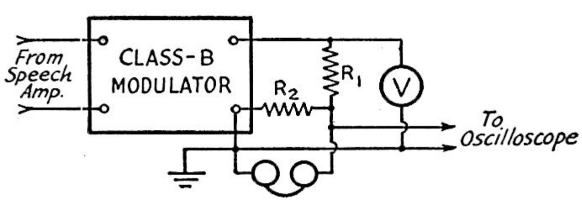


Fig. 9-45 — Set-up for checking a Class B modulator.

previously; the oscilloscope is excellent for this purpose. If a variable-frequency audio oscillator is used, a check on the frequency response of the over-all system can be obtained by varying its frequency (check its output voltage at each frequency change) and observing the variation in the modulator output voltage. The high-frequency response of the system can be attenuated by trying condensers of various values across the primary and secondary of the output transformer, as pointed out in the discussion on Class B modulators. The response above 3000 cycles should be small compared to the response in the 200- to 2500-cycle region so that the channel occupied by the transmitter will not be excessive. A simple way to check this is to apply a sine-wave signal of about 1500 cycles and increase its amplitude until distortion becomes noticeable; when this occurs the tone no longer sounds pure but sounds like a musical octave. The condenser values should then be adjusted until the tone sounds pure again at the same signal amplitude.

Checking 'Phone-Transmitter Operation

Proper adjustment of a 'phone transmitter is aided immeasurably by the oscilloscope; it will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary. They do, however, give a different type of pattern than is obtained without them.

When using the tube without a sweep circuit, radio-frequency voltage from the modulated amplifier is applied directly to the vertical deflection plates of the tube, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the amplitude of the horizontal signal varies, the r.f. output of the transmitter also varies, and this produces a wedge-shaped pattern or **trapezoid** on the screen. If the oscilloscope has a horizontal sweep, the r.f. voltage is applied to the vertical plates as before (never through an amplifier) and the sweep produces

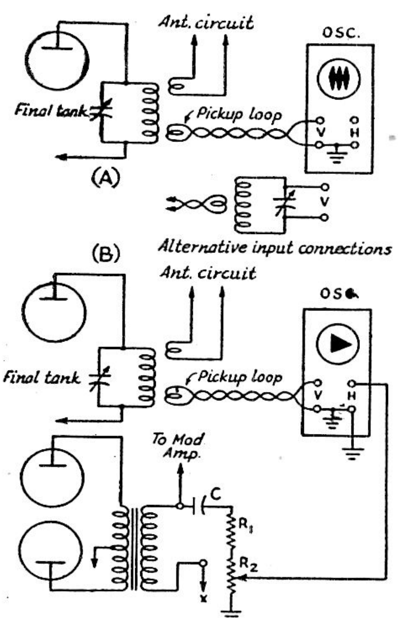


Fig. 9-46 — **Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation.**

a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a **wave-envelope** modulation pattern.

Oscilloscope connections for both types of patterns are shown in Fig. 9-46. The connections for the wave-envelope pattern are somewhat simpler than those for the trapezoidal figure. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a twisted-pair line and pick-up coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate r.f. harmonics, and the tuning control provides a means for adjustment of the pattern height.

To get a wave-envelope pattern the position of the pick-up coil should be varied until a carrier pattern, Fig. 9-47B, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, a rapidly-changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100 per cent. This is illustrated by Fig. 9-47D, where the point X represents the sweep line (reference line) alone, YZ is the carrier height, and PQ is the maximum height of the modulated wave. If the height is greater than the distance PQ, as illustrated in E, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 per cent.

Assuming that the modulation is symmetrical, any modulation percentage can be measured directly from the screen by measuring the maximum height with modulation and the height of the carrier alone; calling these two heights h_1 and h_2 respectively, the modulation percentage is

$$\frac{h_1 - h_2}{h_2} \times 100$$

Connections for the trapezoidal pattern are shown in Fig. 9-46B. The vertical plates are coupled to the transmitter tank circuit through a pick-up loop; alternatively, the tuned input circuit to the oscilloscope may be used. The horizontal plates are coupled to the output of the modulator through a voltage divider, R_1R_2. R_2 should be a potentiometer so the audio voltage can be adjusted to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control. The value of R_1 will depend upon the audio output voltage of the modulator. This voltage is equal to $\sqrt{PR}$, where P is the audio power output of the modulator and R is the modulating impedance of the modulated r.f. amplifier. In the case of grid-bias modulation with a 1:1 output transformer, it will be satisfactory to assume that the a.c. output voltage of the modulator is equal to $0.7E$ for a single tube, or to $1.4E$ for a push-pull stage, where E is the d.c. plate voltage on the modulator. If the transformer ratio is other than 1:1, the voltage so calculated should be multiplied by the actual secondary-to-primary turns ratio.

The total resistance of R_1 and R_2 in series should be 0.25 megohm for every 150 volts of modulator output; for example, if the modulator output voltage is 600, the total resistance should be four (600/150) times 0.25 megohm, or 1 megohm. Then, with 0.25 megohm at R_2, R_1 should be 0.75 megohm. The blocking condenser, C, should be 0.1 μfd. or more, and its voltage rating should be greater than the maximum voltage in the circuit. With plate modulation, this is twice the d.c. voltage applied to the plate of the modulated amplifier.

Trapezoidal patterns for various conditions of modulation are shown in Fig. 9-47 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pick-up coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher the modulation percentage, the wider and more pointed the wedge becomes. At 100-per-cent modulation it just makes a point on the axis, X, at one end, and the height, PQ, at the other end is equal to twice the carrier height, YZ. Overmodulation in the upward direction is indicated by increased height over PQ, and in the downward direction by an extension along the axis X at the pointed end. The modulation percentage may be found by measuring the modulated and unmodulated carrier heights, in the same way as with the wave-envelope pattern.

Nonsymmetrical Waveforms

In voice waveforms the maximum amplitude in one direction from the axis frequently is greater than in the other direction (although the average *energy* on both sides is the same). For this reason the percentage of modulation in the "up" direction frequently differs from that in the "down" direction. With a given voice and microphone, this difference in modulation percentage is usually always in the same direction. Overmodulation in the downward direction causes more out-of-channel interference than overmodulation upward, because of the sharp break — generating high-order harmonics — when the carrier goes to zero.

It is therefore advisable to "phase" the modulation so that the side of the voice waveform having the *larger* excursions causes the instantaneous carrier power to *increase* — that is, modulate upward. This reduces the likelihood of overmodulation on the "down" peak. The direction of the larger excursions can readily be found by careful observation of the oscilloscope pattern. The phase can be reversed by reversing the connections of one winding of any transformer in the speech amplifier or modulator.

Modulation Monitoring

It is always desirable to modulate as fully as possible, but 100-per-cent modulation should not be exceeded — particularly in the downward direction — because harmonic distortion will be introduced and the channel-width increased. This causes unnecessary interference to other stations. The oscilloscope is the best instrument for continuously checking the modulation. However, simpler indicators may be used for the purpose, once calibrated.

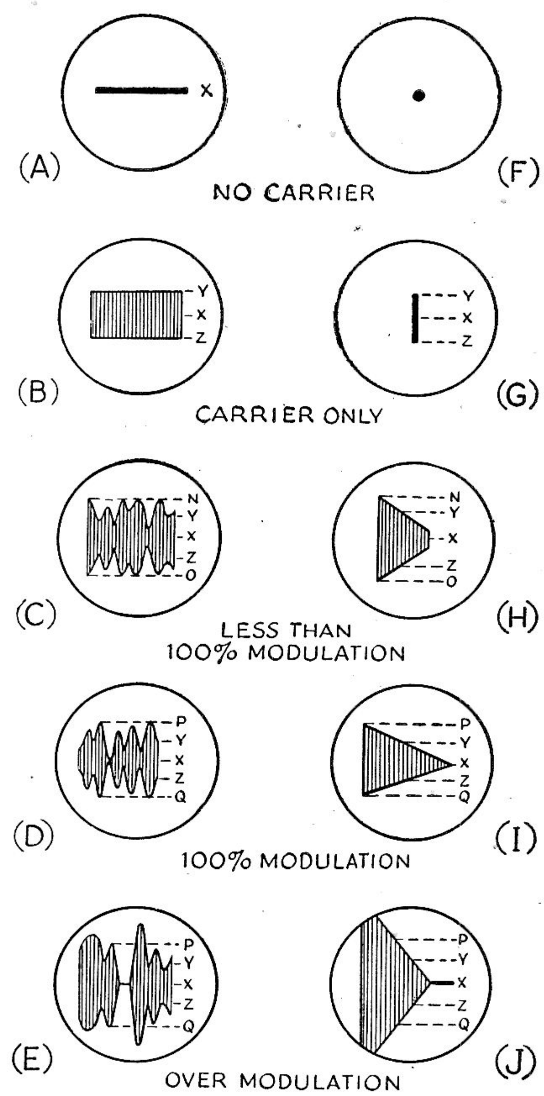
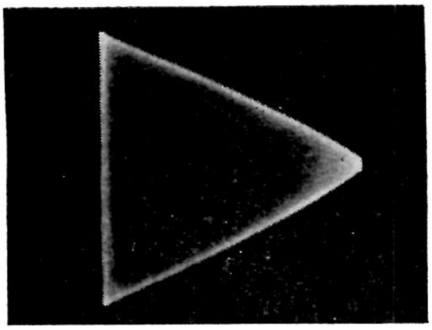
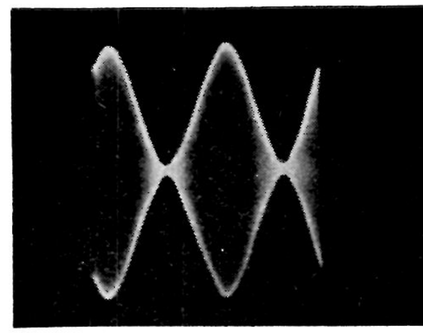


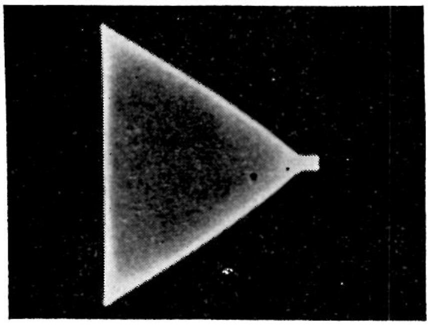
Fig. 9-47 — Wave-envelope and trapezoidal patterns representing different conditions of modulation.

A convenient indicator, when a Class B modulator is used, is the plate milliammeter in the Class B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity that give 100-per-cent modulation on voice peaks, and simultaneously observe the maximum Class B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice to adjust the gain so that it is not exceeded.

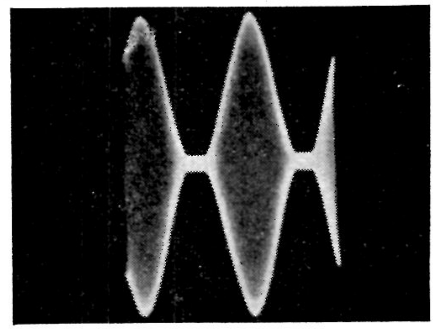


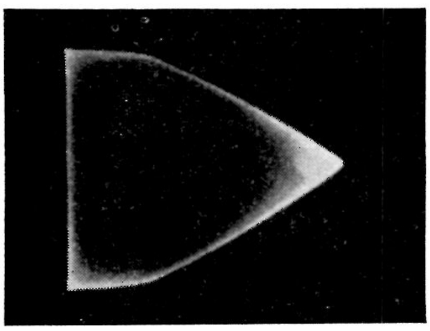
Properly-operated 'phone transmitter modulated 100 per cent.



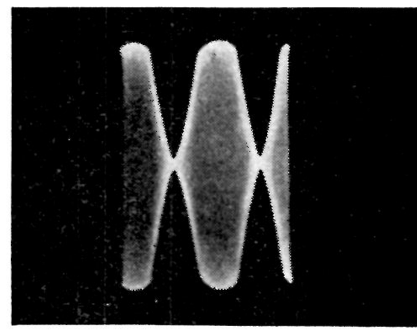


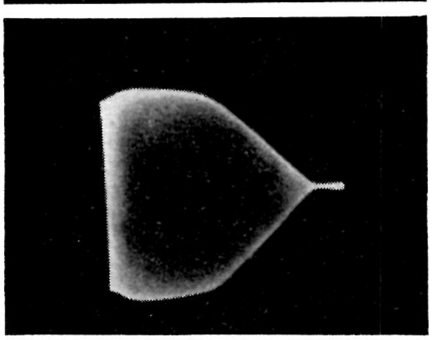
Overmodulation of a transmitter having high modulation capability. Distortion occurs only on the down-peaks.



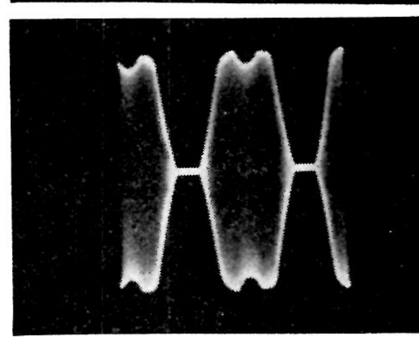


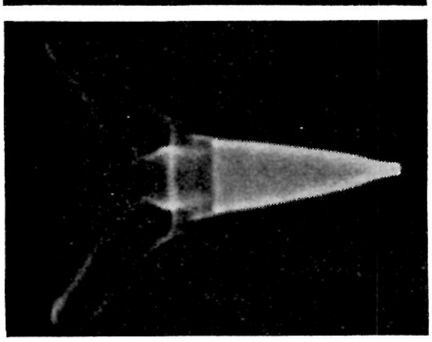
Nonlinearity in modulated r.f. stage, frequently caused by insufficient excitation of a plate-modulated amplifier or overexcitation of a grid-bias modulated amplifier. The amplifier modulates linearly in the downward direction but the up-peaks are flattened.



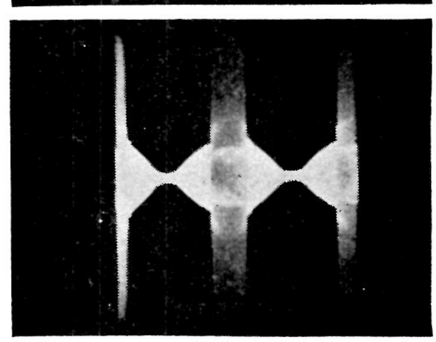


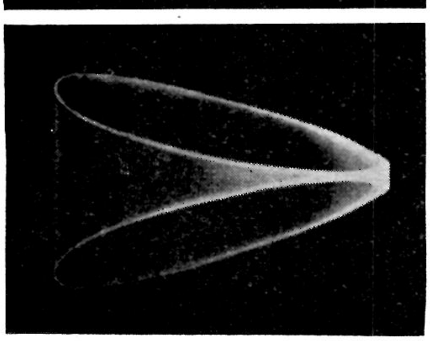
Overmodulation and nonlinear operation (insufficient modulation capability). These patterns are similar to those directly above, but with the modulation carried beyond 100 per cent in the downward direction.



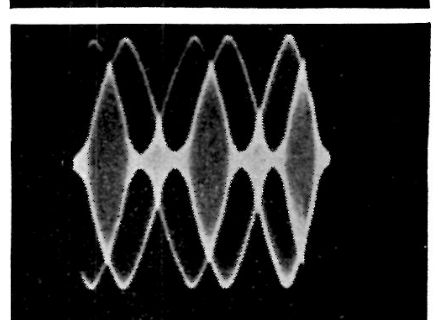


Overmodulation and parasitic oscillations in the modulated amplifier. The trapezoidal pattern also shows phase distortion caused by incorrect coupling between the oscilloscope and audio system.





Left — Phase distortion caused by incorrect coupling between audio system and oscilloscope. Right — Multiple pattern caused by incorrect setting of oscilloscope time-base control. In both cases the wave is modulated 100 per cent.



PHOTOGRAPHS OF TYPICAL OSCILLOSCOPE PATTERNS

These photographs show various conditions of modulation as displayed by the wedge or trapezoidal patterns in the left-hand column and the wave-envelope patterns in the right-hand column.

(Photographs reproduced through courtesy of the Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

A sensitive rectifier-type voltmeter (copper-oxide type) also can be used for modulation monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading that represents 100-per-cent modulation.

The plate milliammeter of the modulated r.f. stage also is of some value as an indicator of overmodulation. The average plate current stays constant if the amplifier is linear, so the reading will be the same whether or not the transmitter is modulated. When the amplifier is overmodulated, especially in the downward direction, the operation is no longer linear and the average plate current will change. A flicker of the pointer may therefore be taken as an indication of overmodulation or nonlinearity. However, it is possible that under some operating conditions the average plate current will remain constant even though the amplifier is considerably overmodulated. Therefore an indicator of this type is not wholly reliable unless it has been checked previously against an oscilloscope.

Linearity

The linearity of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave-envelope pattern, and less auxiliary equipment is required. The connections are the same as for measuring modulation percentage (Fig. 9-46B). If the amplifier is perfectly linear, the sloping sides of the trapezoid will be perfectly straight from the point at the axis up to at least 100-per-cent modulation in the upward direction. Nonlinearity will be shown by curvature of the sides. Curvature near the point, causing it to approach the axis more slowly than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may also be caused by imperfect neutralization (a push-pull amplifier is recommended because better neutralization is possible than with single-ended amplifiers) or by r.f. leakage from the exciter through the final stage. The latter condition can be checked by removing the plate voltage from the modulated stage, when the carrier should disappear, leaving only the beam spot remaining on the screen (Fig. 9-47F). If a small vertical line remains, the amplifier should be reneutralized; if this does not eliminate the line, it is an indication that r.f. is being picked up from lower-power stages, either by coupling through the final tank or via the oscilloscope pick-up loop.

Inward curvature at the large end of the pattern is caused by improper operating conditions of the modulated amplifier — usually improper bias or insufficient excitation, or both, with plate modulation. In grid-bias and cathode-modulated systems, the bias, excitation and plate loading are not correctly proportioned when such curvature occurs. The usual reason is that the amplifier has been adjusted to have too-high carrier efficiency without modulation.

For the wave-envelope pattern, it is necessary to have a linear horizontal-sweep circuit in the oscilloscope and a source of sine-wave audio signal voltage (such as an audio oscillator or signal generator) that can be synchronized with the sweep circuit. The linearity can be judged by comparing the wave envelope with a true sine wave. Distortion in the audio circuits will affect the pattern in this case (such distortion has no effect on the trapezoidal pattern, which shows the modulation characteristic of the r.f. amplifier alone), and it is also readily possible to misjudge the shape of the modulation envelope, so that the wave envelope is less useful than the trapezoid for checking linearity of the modulated amplifier.

Fig. 9-48 shows typical patterns of both types. The cause of the distortion is indicated for grid-bias and suppressor modulation. The patterns at A, although not truly linear, are representative of properly-operated grid-bias modulation systems. Better linearity can be obtained with plate modulation of a Class C amplifier.

Faulty Patterns

The drawings of Figs. 9-47 and 9-48 show what is normally to be expected in the way of pattern shapes when the oscilloscope is used to check modulation. If the actual patterns differ considerably from those shown, it may be that the pattern is faulty rather than the transmitter. It is important that only r.f. from the modulated stage be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding paragraph. If r.f. is present also on the horizontal plates, the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a spot where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$fd.) should be connected across the horizontal plates as close to the cathode-ray tube as possible. An r.f. choke (2.5 mh. or smaller) may also be connected in series with the ungrounded horizontal plate.

"Folded" trapezoidal patterns, and patterns in which the sides of the trapezoid are elliptical instead of straight, occur when the audio sweep voltage is taken from some point in the audio system other than that where the a.f. power is applied to the modulated stage. Such patterns are caused by a phase difference between the sweep voltage and the modulating voltage. The connections should always be as shown in Fig. 9-46B.

Plate-Current Shift

As mentioned above, the d.c. plate current of a modulated amplifier will be the same with and without modulation so long as the amplifier operation is perfectly linear and other conditions remain unchanged. This also assumes that the modulator is working within its capabilities. Because there is usually some curvature of the modulation characteristic with grid-bias modulation there is normally a slight upward change in plate current of a stage so modulated, but this occurs only at high modulation percentages and is barely detectable under the usual conditions of voice modulation.

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1) Insufficient excitation to the modulated r.f. amplifier.
2) Insufficient grid bias on the modulated stage.
3) Wrong load resistance for the Class C r.f. amplifier.
4) Insufficient output capacitance in the filter of the modulated-amplifier plate supply.
5) Heavy overloading of the Class C r.f. amplifier tube or tubes.

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

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

When a common plate supply is used for both a Class B (or Class AB) modulator and a modulated r.f. amplifier, the plate current of the latter may "kick" downward because of poor power-supply voltage regulation with the

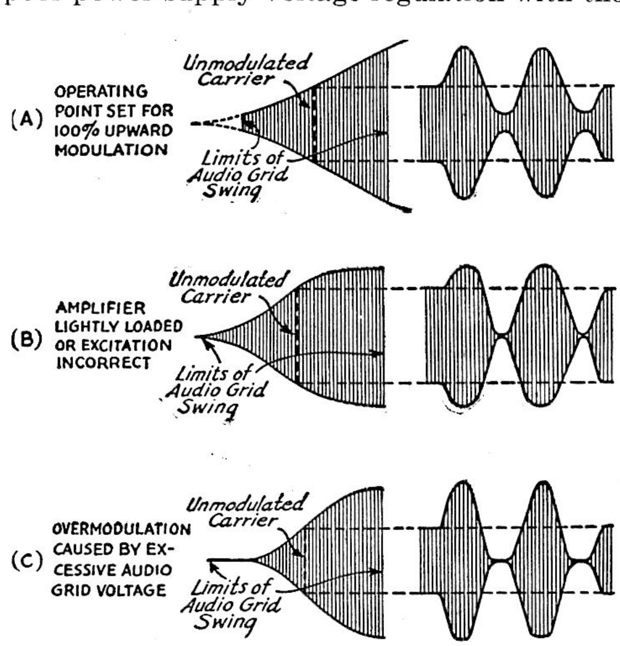


Fig. 9-48 — Oscilloscope patterns representing proper and improper adjustments for grid-bias or cathode modulation. The pattern obtained with a correctly-adjusted amplifier is shown at A. The other drawings indicate nonlinear modulation from typical causes.

varying additional load of the modulator on the supply. The same effect may occur with high-power transmitters because of poor regulation of the a.c. supply mains, even when a separate power-supply unit is used for the Class B modulator. Either condition may be detected by measuring the plate voltage applied to the modulated stage; in addition, poor line regulation also may be detected by observing if there is any downward shift in filament or line voltage.

With grid-bias modulation, any of the following may be the cause of a plate current shift greater than the normal mentioned above:

Downward kick: Too much r.f. excitation; insufficient operating bias; distortion in modulator or speech amplifier; too-high resistance in bias supply; insufficient output capacitance in plate-supply filter to modulated amplifier; amplifier plate circuit not loaded heavily enough; plate-circuit efficiency too high under carrier conditions.

Upward kick: Overmodulation (excessive audio voltage); distortion in audio system; regeneration because of incomplete neutralization; operating grid bias too high.

A downward kick in plate current will accompany an oscilloscope pattern like that of Fig. 9-48B; the pattern with an upward kick will look like Fig. 9-48A, with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and Hum on Carrier

Noise and hum may be detected by listening to the signal on a receiver, provided the receiver is far enough away from the transmitter to avoid overloading. The hum level should be low compared to the voice at 100-per-cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing. With a hum-free carrier, hum introduced by the modulator can be checked by turning on the modulator but leaving the speech amplifier off; power-supply filtering is the likely source of such hum. If carrier and modulator are both clean, connect the speech amplifier and observe the increase in hum level. If the hum disappears with the gain control at minimum, the hum is being introduced in the stage or stages preceding the gain control. The microphone also may pick up hum, a condition that can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground on the microphone and speech system usually is essential to hum-free operation.

Hum can be checked with the oscilloscope, where it has the same appearance as ordinary modulation on the carrier. While the percentage usually is rather small, if the carrier shows modulation with no speech input hum is the likely cause. The various parts of the transmitter may be checked through as described above.

Spurious Sidebands

A superheterodyne receiver having a crystal filter is needed for checking spurious sidebands outside the normal communication channel. The r.f. input to the receiver must be kept low enough, by removing the antenna or by adequate separation from the transmitter, to avoid overloading and consequent spurious receiver responses. With the crystal filter in its sharpest position and the beat oscillator turned on, tune through the region outside the normal channel limits (3 to 4 kilocycles each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent beat-notes coinciding with voice peaks — or, in bad cases of distortion or overmodulation, as "clicks" or crackles well away from the carrier frequency. Sidebands more than 3 to 4 kilocycles from the carrier should be of negligible strength in a properly-modulated 'phone transmitter. The causes are overmodulation or nonlinear operation.

R.F. in Speech Amplifier

A small amount of r.f. current in the speech amplifier — particularly in the first stage, which is most susceptible to such r.f. pick-up — will cause overloading and distortion in the low-level stages. Frequently also there is a regenerative effect which causes an audio-frequency oscillation or "howl" to be set up in the audio system. In such cases the gain control cannot be advanced very far before the howl builds up, even though the amplifier may be perfectly stable when the r.f. section of the transmitter is not turned on.

Complete shielding of the microphone, microphone cord, and speech amplifier is necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Direct coupling or unsymmetrical coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible because these systems sometimes cause the transmitter chassis to take an r.f. potential above ground. Inductive coupling to a two-wire transmission line is advisable. This antenna effect can be checked by disconnecting the antenna and dissipating the r.f. power in a dummy antenna, when it usually will be found that the r.f. feed-back disappears. If it does not, the speech amplifier and microphone shielding are at fault.

Overmodulation Indicators

The most positive method of *preventing* overmodulation is the clipper-filter system described earlier, when properly set up and adjusted. In the absence of such a system — or even with it, just to be safe — some form of overmodulation indicator should be in constant use when the transmitter is on the air.

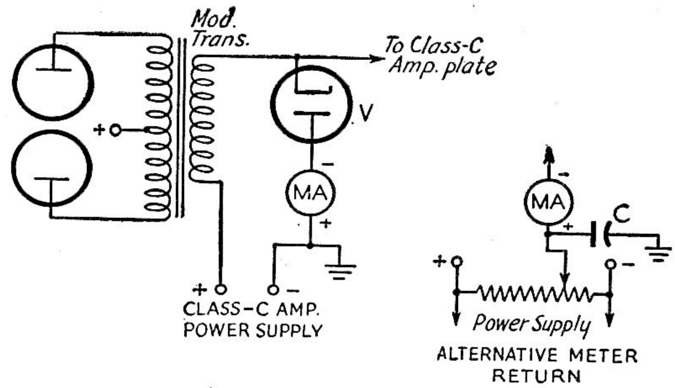


Fig. 9-49 — A negative-peak overmodulation indicator. Milliammeter *MA* may be any low-range instrument (up to 0–50 ma. or so). The inverse peak-voltage rating of the rectifier, *V*, must be at least equal to the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit can be used to indicate modulation in excess of any desired value below 100 per cent.

The best device for this purpose is the cathode-ray oscilloscope. The trapezoidal and wave-envelope patterns are equally useful. A 60-cycle sinusoidal sweep will be quite satisfactory for the wave-envelope pattern. Either pattern should be watched particularly for the bright spots at the axis that accompany overmodulation in the downward direction. The speaking-voice intensity should be kept below the level that shows 100-per-cent modulation on the 'scope.

Overmodulation on negative peaks is more likely to result in spurious sidebands than overmodulation in the upward direction because of the sharp break that occurs when the carrier is suddenly cut off and on. The milliammeter in the negative-peak indicator of Fig. 9-49 will show a reading on each overmodulation peak that carries the instantaneous voltage on the plate of the Class C modulated amplifier "below zero" — that is, negative. The rectifier, *V*, cannot conduct so long as the negative half-cycle of audio output voltage is less than the d.c. voltage applied to the r.f. tube. The rectifier tube must be of a type suitable for the Class C plate voltage employed, and its filament transformer must have similarly-rated insulation.

The effectiveness of the monitor is improved if it indicates at somewhat less than 100-percent modulation, as it will then warn of the danger of overmodulation before it actually occurs. It can be adjusted to indicate at any desired modulation percentage by making the meter return to a point on the power-supply bleeder as shown in the alternative diagram. The by-pass condenser, *C*, insures that the full audio voltage appears across the indicator circuit. The modulation percentage at which the system indicates is determined by the ratio of the d.c. voltage between the meter tap and the positive terminal to the total d.c. voltage.

Frequency and Phase Modulation

The primary advantage of frequency or phase modulation over amplitude modulation comes from the fact that noise or "static," whether natural or set up by electrical machines, is fundamentally an amplitude effect. An AM detector responds to noise just as readily as to the desired modulation on a signal. However, if the receiving system responds only to frequency or phase changes and is insensitive to amplitude variations, it will give normal reception of an FM or PM signal but will not receive noise.

This statement, although an oversimplification, conveys the basic idea. In practice it is only partially accurate; the improvement that can be realized by using FM or PM instead of AM depends on the strength of the received signal, the character of the noise, and the way the noise is distributed over the receiver passband. In general, the wider the channel occupied by the signal the better the noise suppression — if the signal strength is above a certain **threshold** value. The wider the channel occupied by the signal, the stronger the signal required to reach the threshold. The noise suppression in the receiver is most effective when the noise is evenly distributed over the receiver passband and least so when the noise appears on one side or the other of the incoming carrier. (The noise itself usually is properly distributed, but misalignment in receiver circuits will cause uneven response over the passband.) The noise suppression also is most marked when the noise is of the "impulse" type, having a high peak amplitude but short duration.

In amateur work, FM and PM have been used not so much because of the possibility of an improved signal-to-noise ratio but because of more-or-less incidental advantages. For example, in the ultrahigh and superhigh frequency ranges some tubes do not lend themselves well to amplitude modulation, but can easily be frequency-modulated. On the lower frequencies FM and PM are often used because they cause less interference than AM in unshielded broadcast receivers in the vicinity.

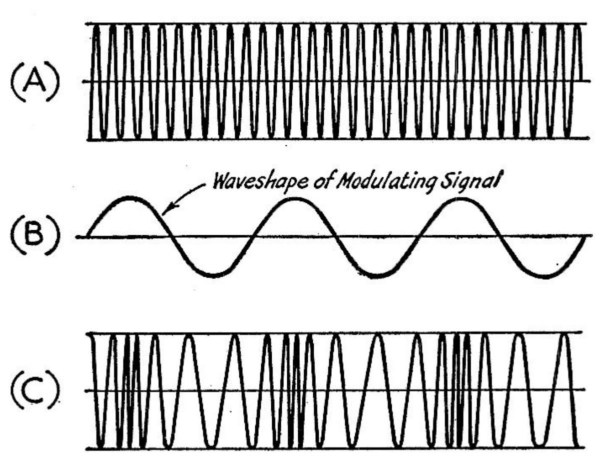


Fig. 9-50 — Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

Frequency Modulation

The fundamental principle of frequency modulation is easy to understand. Suppose we have an oscillator operating at a frequency of, say, 3900 kc. Further suppose that we vary the oscillator tuning control back and forth so that at one extreme the frequency is 3905 kc. and at the other, 3895 kc.; that is, plus and minus 5 kc. on either side of the carrier frequency. Imagine that the tuning is varied back and forth in that fashion at 1000 times per second. Then we are *frequency modulating* the oscillator at an audio frequency of 1000 cycles.

The **frequency deviation** is the maximum change in frequency from the carrier frequency; in this example it is 5 kc. So long as the tuning control is varied between the same two extremes, the frequency deviation is the same no matter how rapidly the control is varied; i.e., no matter what the modulating frequency. In other words, we can make the deviation any reasonable figure we want, whether it is a few hundred cycles or tens of kilocycles, and it is not affected by the modulating frequency.

Fig. 9-50 is a representation of frequency modulation. In the unmodulated carrier each cycle occupies the same time as the preceding one. When a modulating signal is applied, the carrier frequency is increased during one half-cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the drawing by the fact that the r.f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency is proportional to the instantaneous amplitude of the modulating signal — or, to use the analogy above, to the position of the oscillator tuning control — so the frequency deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative. That is, the frequency deviation follows the changes in the amplitude of the modulating signal.

Phase and Frequency

Phase modulation is a little more difficult. To understand the difference between FM and PM it is necessary to appreciate that the frequency of an alternating current is determined by the *rate at which its phase changes*. A current in which the phase changes rapidly has a higher frequency than one in which the phase changes slowly. For example, if the phase moves through 360 degrees in one second the frequency is one cycle per second, but if the phase moves through 1080 degrees in one second (3 × 360 degrees) there are three complete cycles in one second.

If the phase moves along at a constant speed the frequency also is constant. But if the rate of phase change is speeded up or slowed down there is an accompanying shift in the frequency. If the speed is increased the frequency becomes higher; if the speed is decreased the frequency becomes lower.

Now suppose we have a transmitter operating at a fixed frequency; a frequency that is unaffected by tuning an amplifier that is a few stages removed from the frequency-controlling oscillator. We cannot change the *frequency*, but we can *shift the phase* of the r.f. current by adjusting the tuning control of the amplifier. We might, for example, shift the phase of the current in that circuit 10 degrees by detuning the tank circuit from resonance. Once the detuning is finished the phase shift is permanent, but there is still just exactly the same number of cycles per second as before — so the frequency is exactly the same as it was in the first place. *But during the time that the phase shift is taking place there is a change in frequency.* If the phase is advanced (moved forward) the frequency increases; if it is retarded (slowed down) the frequency decreases.

This "instantaneous" frequency change would never be noticed in tuning an amplifier tank circuit, because the frequency deviation depends on the *speed* with which the phase is shifted. Any manual adjustment would be too slow to make an observable frequency change. But when the phase is shifted back and forth at an audio-frequency rate the frequency deviation *is* observable, and it is directly proportional to the rate at which the phase is shifted. The rate of phase shift is naturally proportional to the total number of degrees through which the phase is shifted; it is also proportional to the amplitude of the modulating signal (a large signal will shift the phase more, in the same time, than a small signal), and to the frequency of the modulating signal because the phase shift is more rapid the greater the number of times it is shifted per second.

To summarize, then, in FM the carrier frequency deviation is proportional to the amplitude of the modulating signal but not to its frequency. In PM the deviation is proportional to *both* the amplitude and frequency of the modulating signal. Fig. 9-50 is just as representative of PM as it is of FM, because it is impossible to tell the two apart when there is only one modulating frequency.

Modulation Depth

In FM or PM there is no condition that corresponds exactly to overmodulation in AM. "Percentage of modulation" has to be defined a little differently for these systems. Practically, "100-per-cent modulation" is reached when the transmitted signal occupies a channel just equal to the bandwidth for which the *receiver* is designed. If the channel occupied is wider than the receiver can accept, the receiver distorts the signal and the end effect is much the same as overmodulation in AM. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25-per-cent modulation. Until the maximum width is set for the channel, percentage of modulation has no meaning.

In amateur work no specifications have been set up for channel-width except in the case of "narrow-band" FM or PM (frequently abbreviated NFM), where the channel-width is defined as being the same as that of a properly-modulated AM signal. That is, the channel-width for NFM does not exceed twice the highest audio frequency in the modulating signal. NFM transmissions based on an upper audio limit of 3000 cycles therefore should occupy a channel no wider than 6 kc.

FM and PM Sidebands

From the descriptions given above of the fundamentals of frequency and phase modulation it might be concluded that the channel occupied by the transmission would be no greater than the frequency deviation on each side of the carrier. However, if we applied the same line of reasoning to amplitude modulation we should reach the conclusion that an AM signal takes up no more space than the carrier alone, since only the *amplitude* of the carrier varies. Both conclusions would be wrong; the fact is that both FM and PM set up sidebands, just as AM does. In the case of FM and PM, single-tone modulation sets up a whole series of pairs of sidebands that are harmonically related to the modulating frequency, whereas in AM there is only one pair of sidebands.

The number of "extra" sidebands that occur in FM and PM depends on the relationship between the modulating frequency and the carrier frequency deviation. The ratio between the frequency deviation, in cycles per second, and the modulating frequency, also in cycles per second, is called the **modulation index.** That is,

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

Example: The maximum frequency deviation in an FM transmitter is 3000 cycles either side of the carrier frequency. The modulation index when the modulating frequency is 1000 cycles is

$$\text{Modulation index} = \frac{3000}{1000} = 3$$

At the same deviation with 3000-cycle modulation the index would be 1; at 100 cycles it would be 30, and so on.

The modulation index is also equal to the phase shift (in radians). In PM the index is constant regardless of the modulating frequency; in FM it varies with the modulating frequency, as shown in the previous example. To identify any particular FM system, the limiting modulation index — that is, the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used — is called the **deviation ratio.**

Fig. 9-51 shows how the amplitudes of the carrier and the various sidebands vary with the modulation index. This is for single-tone modulation; the first sideband (actually a pair, one above and one below the carrier) is displaced from the carrier by an amount equal to the modulating frequency, the second is twice the modulating frequency away from the carrier, and so on. For example, if the modulating frequency is 2000 cycles and the carrier frequency is 29,500 kc., the first sideband pair is at 29,498 kc. and 29,502 kc., the second pair

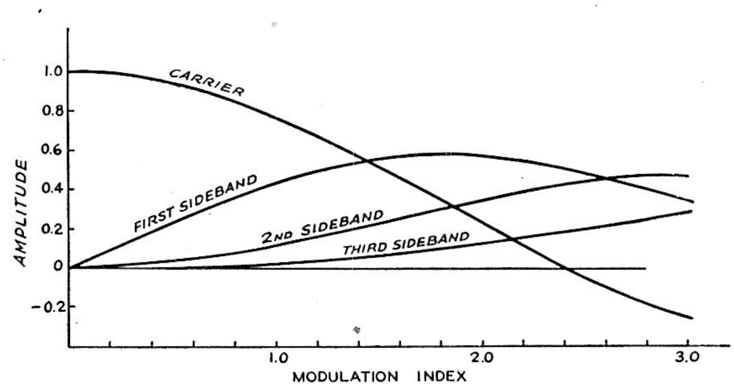


Fig. 9-51 — How the amplitude of the pairs of sidebands varies with the modulation index in an FM or PM signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

is at 29,496 kc. and 29,504 kc., the third at 29,494 kc. and 29,506 kc., etc. The amplitudes of these sidebands depend on the modulation index, not on the frequency deviation. In AM, regardless of the percentage of modulation (so long as it does not exceed 100 per cent) the sidebands would appear *only* at 29,498 and 29,502 kc. under the same conditions.

Note that, as shown by Fig. 9-51, the carrier strength varies with the modulation index. (In amplitude modulation the carrier strength is constant; only the sideband amplitude varies.) At a modulation index of approximately 2.4 the carrier disappears entirely and then becomes "negative" at a higher index. This simply means that its phase is reversed as compared to the phase without modulation. In FM and PM the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index. In AM the sideband power is supplied by the modulator in the case of plate modulation, and by changing the power input and efficiency in the case of grid-bias modulation.

Fig. 9-51 can be carried out to considerably higher modulation indexes, in which case it will be found that more and more additional sidebands are set up and that the carrier goes through several "zeros" and reversals in phase.

Frequency Multiplication

In amplitude modulation it is customary amateur practice to apply the modulation to the final r.f. stage of the transmitter. If a lower-level stage is modulated, a special type of operation is necessary in the following r.f. stages to pass the modulation envelope without distortion. These "linear" amplifiers are rather difficult to adjust properly and must be operated at low plate efficiency. Consequently, the simplest and most economical transmitter design results when the final stage is modulated.

In frequency or phase modulation there is no change in the amplitude of the signal with modulation. Consequently, an FM or PM signal can be amplified by an ordinary Class C amplifier without distortion. The modulation can take place in a very low-level stage and the signal can then be amplified by either frequency multipliers or straight amplifiers. In fact, this is the usual practice. The audio power required for modulating an FM or PM transmitter is negligible.

If the modulated signal is passed through one or more frequency multipliers, the modulation index is multiplied by the same factor that the carrier frequency is multiplied. For example, suppose that the controlling oscillator in the transmitter is on 3.5 Mc. and the final output is on 28 Mc. The total frequency multiplication is 8 times, and any FM or PM applied to the oscillator will likewise be multiplied by 8 in the 28-Mc. output. If the frequency deviation is 500 cycles at 3.5 Mc., it will be 4000 cycles at 28 Mc.

Frequency multiplication offers a means for obtaining practically any desired amount of frequency deviation, whether or not the modulator itself is capable of giving that much deviation without distortion. Also, if the same oscillator is modulated in a transmitter that operates on several bands, the frequency deviation is different on each band. The amount of frequency multiplication after modulation always must be taken into account in determining whether or not the final frequency deviation is the desired value on a given band.

● NARROW-BAND FM OR PM

Where FM or PM is used in crowded 'phone bands (particularly below 27 Mc.) it is of utmost importance that the transmissions should occupy no greater channel-width than would be occupied by an AM signal. It is evident from Fig. 9-51 that this requirement can be met only by using a relatively small modulation index. It must be realized that the higher-order sidebands always are present, even at very small indexes. It is therefore necessary to set an arbitrary level above which the extra sidebands should not go. If the modulation index (with single-tone modulation) does not exceed about 0.6 the most important extra sideband, the second, will be at least 20 db. below the unmodulated carrier level, and this should represent an effective channel-width about equivalent to that of an AM signal. In the case of speech, a somewhat higher modulation index can be used. This is because the energy distribution in a complex wave is such that the modulation index for any one frequency component is reduced, as compared to the index with a sine wave having the same peak amplitude as the voice wave.

The chief advantage of narrow-band FM or PM for frequencies below 30 Mc. is that it eliminates or reduces certain types of interference to broadcast reception. Also, the modulating equipment is relatively simple and inexpensive. However, assuming the same unmodulated carrier power in all cases, narrow-band FM or PM is not as effective as AM. As shown by Fig. 9-51, at an index of 0.6 the amplitude of the first sideband is about 25 per cent of the unmodulated-carrier amplitude; this compares with a sideband amplitude of 50 per cent in the case of a 100-per-cent modulated AM transmitter. In other words, so far as effectiveness is concerned a narrow-band FM or PM transmitter is about equivalent to a 100-per-cent modulated AM transmitter operating at one-fourth the power input. This assumes that the receiving system is equally efficient in all cases. This very often is not true on the low-frequency bands, since communications receivers are designed primarily for AM reception.

Methods of Frequency and Phase Modulation

FREQUENCY MODULATION

The simplest and most satisfactory device for amateur FM is the reactance modulator. This is a vacuum tube connected to the r.f. tank circuit of an oscillator in such a way as to act as a variable inductance or capacitance. Fig. 9-52 is a representative circuit. The control-grid circuit of the 6L7 tube is connected across the small capacitance, C_1, which is in series with the resistor, R_1, across the oscillator tank circuit. Any type of oscillator circuit may be used. The resistance of R_1 is made large compared to the reactance of C_1, so the r.f. current through R_1C_1 will be practically in phase with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_1 will lag the current by 90 degrees. The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage, and consequently is 90 degrees behind the current through C_1, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank. The frequency increases in proportion to the amplitude of the lagging plate current of the modulator. The value of plate current is determined by the voltage on the No. 3 grid of the 6L7; hence the oscillator frequency will vary when an audio signal voltage is applied to the No. 3 grid.

If, on the other hand, C_1 and R_1 are interchanged and the reactance of C_1 is made large compared to the resistance of R_1, the r.f. current in the 6L7 plate circuit will lead the oscillator-tank r.f. voltage, making the reactance capacitive rather than inductive.

A circuit using a receiving-type r.f. pentode of the high-transconductance type, such as the 6SG7, is shown in Fig. 9-53. In this case, both r.f. and audio are applied to the control grid. The audio voltage, introduced through a radio-frequency choke, *RFC*, varies the transconductance of the tube and thereby varies the r.f. plate current. The capacitance C_8 corresponds to C_1 in Fig. 9-52; it represents the input capacitance of the tube. (It is possible, also, to omit C_1 from Fig. 9-52 and depend upon the input capacitance of the 6L7 instead; the only disadvantage is that there is then no control over the modulator sensitivity. Likewise, a 3–30-$\mu\mu$fd. trimmer condenser can be connected at C_8 in Fig. 9-53 to permit controlling the sensitivity.) In Fig. 9-53 the r.f. circuit is

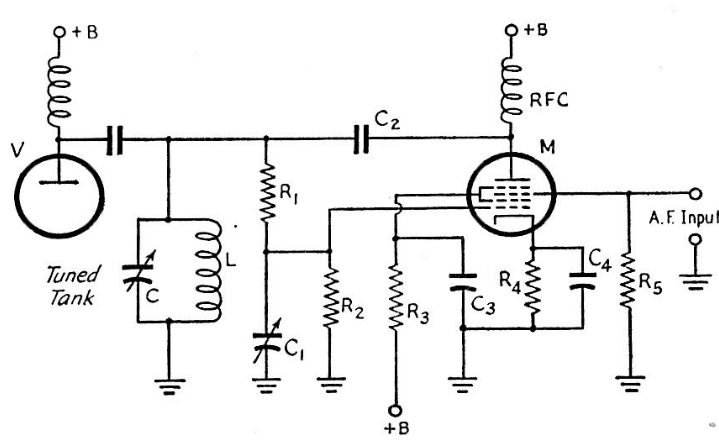


Fig. 9-52 — Reactance-modulator circuit using a 6L7 tube.
C — R.f. tank capacitance. C_1 — 3-30 $\mu\mu$fd. C_2 — 220 $\mu\mu$fd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 10-μfd. electrolytic in parallel with 0.01-μfd. paper.
L — R.f. tank inductance.
R_1 — 47,000 ohms.
R_3 — 33,000 ohms.
R_2, R_5 — 0.47 megohm.
R_4 — 330 ohms.
RFC — 2.5 mh.

series-fed, which is advantageous if the r.f. tube and the modulator can be operated at the same plate voltage. The use of different plate voltages on the two tubes calls for the parallel-feed arrangement shown in Fig. 9-52.

The modulated oscillator usually is operated on a relatively low frequency, so that a high order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired. The frequency deviation increases with the number of times the initial frequency is multiplied; for instance, if the oscillator is operated on 6.5 Mc. and the output frequency is to be 52 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 8000 cycles at the output frequency.

A reactance modulator can be connected to a crystal oscillator as well as to the self-controlled type. However, the resulting signal is more phase-modulated than it is frequency-modulated, for the reason that the frequency deviation that can be secured by varying the tuning of a crystal oscillator is quite small.

Design Considerations

The sensitivity of the modulator (frequency change per unit change in grid voltage) depends on the transconductance of the modulator tube. It increases when C_1 is made smaller, for a fixed value of R_1, and also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier stability of the oscillator depends on the L/C ratio, it is desirable to use the highest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 9-53 is used in connection with a 7-Mc. oscillator, a linear deviation of 1500 cycles above and below the carrier frequency can be secured when the oscillator tank capacitance is approximately 200 $\mu\mu$fd. A peak a.f. input of two volts is required for full deviation.

A change in *any* of the voltages on the modulator tube will cause a change in r.f. plate current, and consequently a frequency change. Therefore it is advisable to use a regulated plate power supply for both modulator and oscillator. At the low voltages used (250 volts), the required stabilization can be secured by means of gaseous regulator tubes.

Speech Amplification

The speech amplifier preceding the modulator follows ordinary design, except that no power is required from it and the a.f. voltage taken by the modulator grid usually is small — not more than 10 or 15 volts, even with large modulator tubes. Because of these modest requirements, only a few speech-amplifier stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance-coupled, will more than suffice for crystal microphones.

PHASE MODULATION

The same type of reactance-tube circuit that is used to vary the tuning of the oscillator tank in FM can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for PM. Hence the modulator circuits of Figs. 9-52 and 9-53 can be used for PM if the reactance tube works on an amplifier tank instead of directly on a self-controlled oscillator.

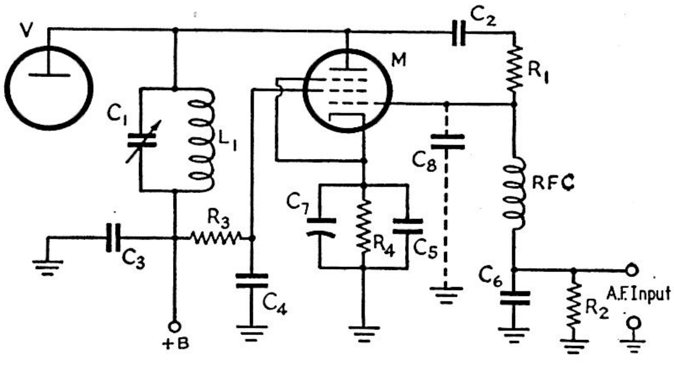


Fig. 9-53 — Reactance modulator using a high-transconductance pentode (6SG7, 6AG7, etc.).

C_1 — R.f. tank capacitance (see text).
C_2, C_3 — 0.001-μfd. mica.
C_4, C_5, C_6 — 0.0047-μfd. mica.
C_7 — 10-μfd. electrolytic.
C_8 — Tube input capacitance (see text).
R_1, R_2 — 0.47 megohm.
R_3 — Screen dropping resistor; select to give proper screen voltage on type of modulator tube used.
R_4 — Cathode bias resistor; select as in case of R_3.
L_1 — R.f. tank inductance.
RFC — 2.5-mh. r.f. choke.

The phase shift that occurs when a circuit is detuned from resonance depends on the amount of detuning and the Q of the circuit. The higher the Q, the smaller the amount of detuning needed to secure a given number of degrees of phase shift. If the Q is at least 10, the relationship between phase shift and detuning (in kilocycles either side of the resonant frequency) will be substantially linear over a range of about 25 degrees. From the standpoint of modulator sensitivity, the Q of the tuned circuit on which the modulator operates should be as high as possible. On the other hand, the effective Q of the circuit will not be very high if the amplifier is delivering power to a load, since the load resistance reduces the Q. There must therefore be a compromise between modulator sensitivity and r.f. power output from the modulated amplifier. An optimum figure for Q appears to be about 20; this allows reasonable loading of the modulated amplifier and the necessary tuning variation can be secured from a reactance modulator without difficulty.

It is advisable to modulate at a very low power level — preferably in a transmitter stage where receiving-type tubes are used. A practical phase-modulator unit is described later in this chapter.

A Frequency-Control and FM-Modulator Unit

The accompanying photographs show a complete VFO/reactance-modulator unit designed to work into a normally crystal-controlled transmitter using either 7- or 14-Mc. crystals. It has its own power supply, using a small "broadcast-replacement" power transformer. The VFO uses a 6SJ7 as the oscillator tube, and is followed by a 6SJ7 buffer that may be used as a straight amplifier for 7-Mc. output, or as a frequency doubler for 14-Mc. output. The reactance modulator is a 6L7. There are two stages of speech amplification, a 6SJ7 followed by a 6C5, a combination that provides ample gain for a crystal microphone. The r.f. output of the unit is intended to be fed through a link to a tuned circuit that substitutes for the crystal in the transmitter's regular crystal-oscillator circuit. This tuned circuit should be resonant at the same frequency as the output tank circuit in the control unit (L_2C_3 in Fig. 9-54) and can be identical in construction.

The constants of the oscillator tank circuit are chosen so that the frequency range 6000–7425 kc. can be covered. In the transmitter, the output can be multiplied in frequency to the 14-, 28-, 50- and 144-Mc. bands when the 6SJ7 oscillator is set to the appropriate frequency.

The sensitivity of the modulator is controlled by the setting of C_{11}. The higher the capacitance of this condenser the smaller the frequency deviation for a given audio input voltage to the modulator. At maximum sensitivity, with C_{11} at minimum capacitance, the linear deviation is approximately 1.5 kc.; this deviation requires a signal of 2 volts peak at the No. 3 grid of the 6L7. The actual deviation at the output frequency depends on the amount of frequency multiplication following the oscillator. The maximum linear deviation is 3 kc. at 14 Mc., 6 kc. at 28 Mc., 12 kc. at 50 Mc., and 36 kc. at 144 Mc. The unit is therefore suitable for narrow-band frequency modulation on 14 and 28 Mc., and for wide-band FM on 50 and 144 Mc. For narrow-band FM the speech-amplifier gain should be set so that the deviation does not exceed about 2 kc. at the *output* frequency.

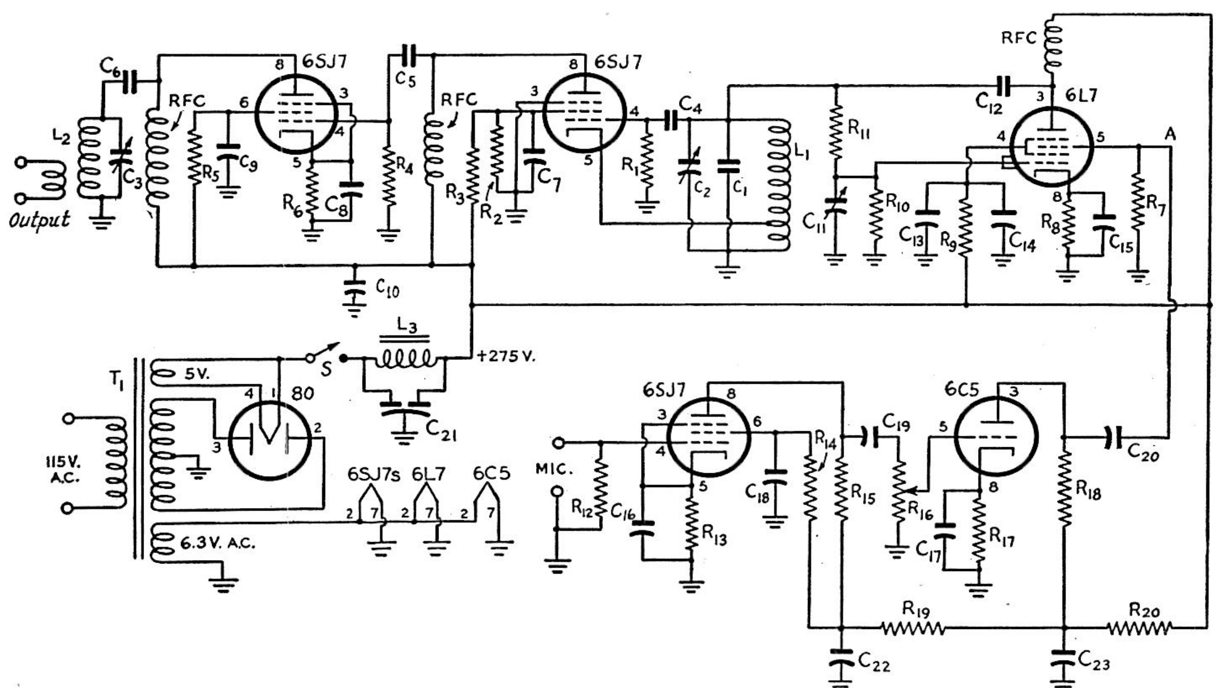


Fig. 9-54 — **Circuit diagram of the FM control unit for use with normally crystal-controlled v.h.f. transmitters.**

C_1 — 150-$\mu\mu$fd. silvered mica (for 7 Mc.)
C_2 — 100-$\mu\mu$fd. variable (National SE-100).
C_3 — 50-$\mu\mu$fd. variable (Hammarlund HF-50).
C_4 — 100-$\mu\mu$fd. mica.
C_5, C_{12} — 220-$\mu\mu$fd. mica.
C_6 — 0.001-μfd. mica.
C_7, C_8, C_9, C_{10}, C_{13}, C_{15}, C_{19}, C_{20} — 0.01-μfd. paper.
C_{11} — 3–30-$\mu\mu$fd. mica trimmer.
C_{14}, C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
C_{16}, C_{17} — 10-μfd. 25-volt electrolytic.
C_{18} — 0.1-μfd. 200-volt paper.
C_{21} — Dual 450-volt 8-μfd. electrolytic.
R_1 — 0.1 megohm, 1 watt.
R_2 — 22,000 ohms, 1 watt.
R_3, R_4, R_5, R_{11} — 47,000 ohms, 1 watt.
R_6, R_8 — 330 ohms, ½ watt.
R_7, R_{10} — 0.47 megohm, ½ watt.
R_9 — 33,000 ohms, 1 watt.
R_{12} — 4.7 megohms, ½ watt.
R_{13} — 1000 ohms, ½ watt.
R_{14} — 1 megohm, ½ watt.
R_{15}, R_{19} — 0.22 megohm, ½ watt.
R_{16} — 0.5-megohm volume control.
R_{17} — 2200 ohms, ½ watt.
R_{18} — 47,000 ohms, ½ watt.
R_{20} — 0.15 megohm, 1 watt.
L_1 — 7 Mc.: 11 turns No. 18 e., length ¾ inch, 1-inch diameter, tapped 3rd turn from ground.
L_2 — 3.5 Mc.: see text; 7 Mc.: 23 t. No. 24. e. close-wound on 1-inch diam. form; 14 Mc.: 11 t. No. 24 e. spaced wire diam. on 1-inch diam. form. Link: 3 to 5 turns (not critical).
L_3 — Filter choke, 10 hy., 40 ma.
RFC — 2.5-mh. r.f. choke.
S — S.p.s.t. toggle switch.
T_1 — 500 volts c.t., 40 ma.; 6.3 volts at 2 amp.; 5 volts at 2 amp. (Thordarson T-13R11).

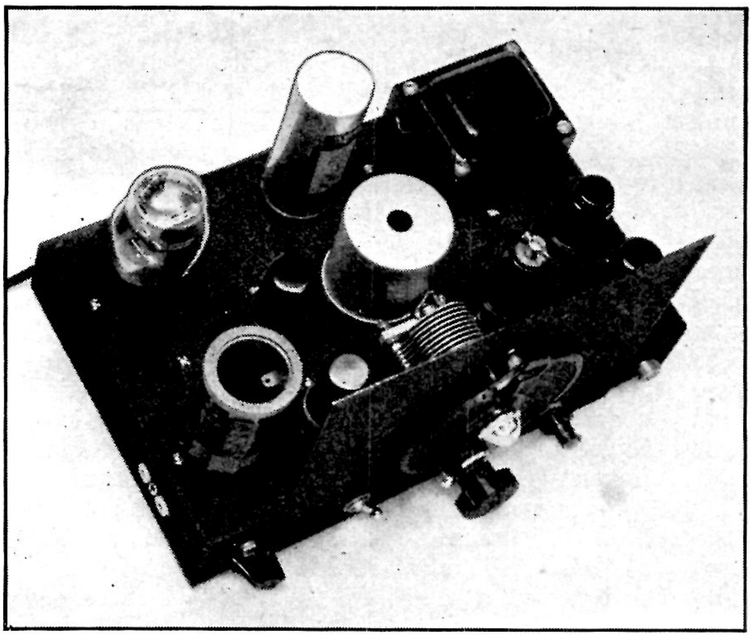


Fig. 9-55 — This modulator-oscillator unit is used with normally crystal-controlled v.h.f. transmitters for frequency-modulated output. It contains a speech amplifier and power supply, so that no additional equipment is needed. The oscillator coil is in the round shield can in the center. The coil in the left foreground is the buffer output circuit. The speech amplifier and modulator are at the right, with the power supply along the rear. A 7 × 11-inch chassis is used.

In the top view of the unit, the 6SJ7 oscillator tube is alongside the aluminum shield that covers the oscillator coil. The 6SJ7 r.f. amplifier is between the oscillator tuning condenser (at the center) and its output coil at the left. The tubes along the right-hand edge of the chassis are the 6SJ7 and 6C5 speech amplifiers, with the former in front. The 6L7 modulator is between the speech-amplifier tubes and the oscillator tuned circuit. The r.f. leads in the oscillator circuit, including the connections to R_{11}, C_{11}, and the No. 1 grid of the 6L7, should be kept short and rigid. The usual precautions as to wiring in r.f. and audio circuits should be observed.

For narrow-band FM in the portion of the 3.5-Mc. band in which such operation is permitted, it is advisable to operate the oscillator at half the desired frequency and use the r.f. output stage as a doubler to 3.5 Mc. For this purpose it is suggested that a 470-$\mu\mu$fd. silvered mica condenser be substituted at C_1, and that L_1 consist of 29 turns of No. 20 enameled wire close-wound on a 1-inch diameter form. The oscillator cathode tap should be at the 9th turn from ground. The r.f. output coil, L_2, should have 46 turns of No. 24 enameled wire close-wound on a 1-inch diameter form. The frequency range of the oscillator will be sufficient to cover the frequencies assigned for 'phone.

Somewhat more r.f. output can be obtained by substituting a 6AC7 for the 6SJ7 output amplifier. This substitution should be necessary only if the 6SJ7 is incapable of driving the crystal-oscillator tube in the transmitter. The considerations involved in coupling a VFO to various types of crystal-oscillator circuits are discussed in Chapter Six.

Before putting the unit on the air the carrier frequency stability and frequency deviation should be checked by the methods outlined later.

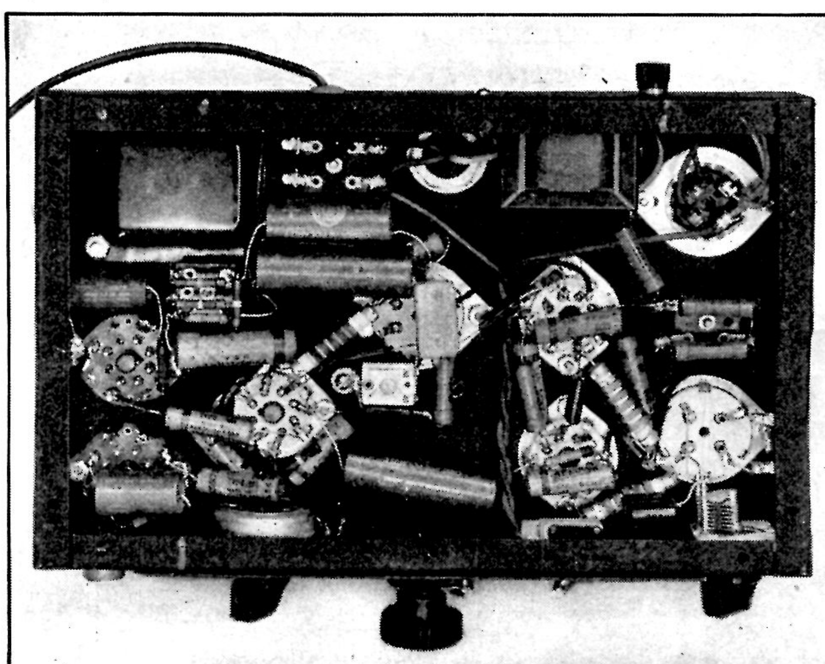


Fig. 9-56 — In this bottom view of the FM modulator unit, the r.f. section is at the right and the audio at the left. The oscillator socket is to the right of the coil socket in the center.

A Narrow-Band PM Exciter Unit

The unit shown in Figs. 9-57, 9-58 and 9-59 will deliver from 10 to 20 watts of phase-modulated output, depending upon the plate voltage used, and it can be used to replace the crystal oscillator of that power level in any existing transmitter. It can also be used with an existing VFO to obtain a phase-modulated signal. A low-pass filter is incorporated to limit the modulation frequencies to those below 3000 cycles, thus making it easier to comply with the regulations on narrow-band FM and PM.

As can be seen in the wiring diagram, Fig. 9-57, a 6J5 Pierce-type oscillator circuit is used to excite a 6SK7 r.f. amplifier. If VFO control is used, the 6J5 can be removed from its socket and the VFO output introduced at J_1. To accommodate various output levels from the VFO, a gain control, R_3, is included in the cathode circuit of the 6SK7 amplifier. The plate circuit of the 6SK7 amplifier is reactance-modulated (to give the PM signal) by a 6SG7 reactance modulator, and the output of the 6SK7 drives a 2E26 r.f. amplifier. With 500 volts on the plate of the 2E26, 15 watts output can be obtained, and the plate voltage can be raised to 600 if more output is required. The microphone input at J_4 is amplified through a 6SJ7 and a 6J5, and the low-pass filter is connected between the 6J5 and the 6SG7 modulator tube. The degree of modulation is controlled by the setting of R_{15}, the audio gain control.

Construction

The unit is built on a 7 × 12 × 3-inch chassis, and the location of the components can be seen from Figs. 9-58 and 9-59. A shield can (Millen 80016) is used over L_1, to avoid regeneration, and a small shield extends up 1 inch around the 2E26 for the same reason. No special care is necessary in wiring the unit, except that C_5 or C_6 should be mounted across the 6SK7 socket, to shield the grid pin from the plate pin, and the audio-circuit wiring should be kept away from the r.f. circuits. The r.f. input and output, from J_1 and J_3, is most conveniently run in short pieces of RG-58/U

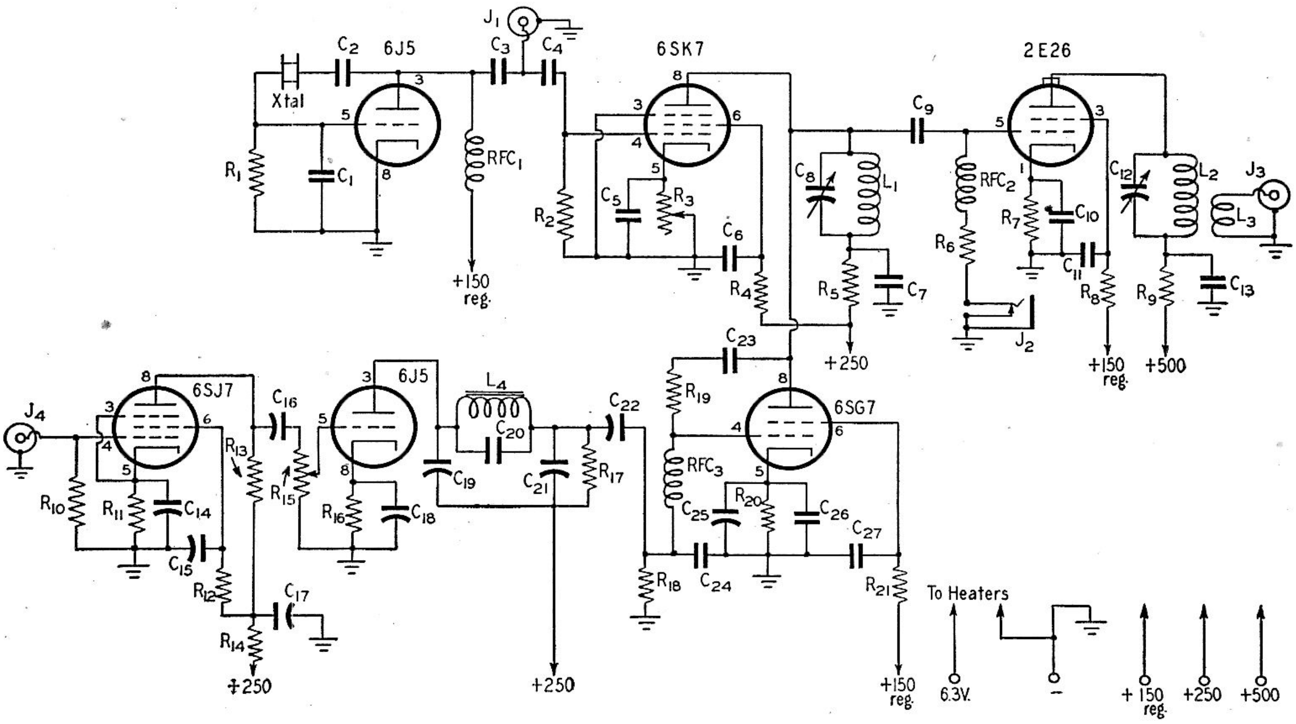


Fig. 9-57 — Wiring diagram of the narrow-band phase-modulation unit.

C_1 — 22-μμfd. mica.
C_2, C_5, C_6, C_7, C_{10}, C_{11}, C_{13}, C_{23}, C_{24}, C_{26}, C_{27} — 0.001-μfd. mica.
C_3, C_4, C_9 — 100-μμfd. mica.
C_8 — 50-μμfd. midget variable (Millen 21050).
C_{12} — 50-μμfd. variable (Millen 22050).
C_{14}, C_{18}, C_{25} — 10-μfd. electrolytic, 25 volts.
C_{15} — 0.1-μfd. paper, 200 volts.
C_{16} — 0.05-μfd. 400-volt paper.
C_{17} — 8-μfd. electrolytic, 450 volts.
C_{19}, C_{21}, C_{22} — 0.01-μfd. 400-volt paper.
C_{20} — 0.006-μfd. 200-volt paper.
R_1, R_2, R_4, R_{14} — 47,000 ohms.
R_3 — 5000-ohm potentiometer, wire-wound.
R_5, R_8, R_{21} — 470 ohms.
R_6 — 12,000 ohms.
R_7 — 330 ohms, 1 watt.
R_9 — 100 ohms.
R_{10}, R_{12} — 1.0 megohm.
R_{11} — 820 ohms.
R_{13} — 0.22 megohm.
R_{15} — 0.25-megohm volume control.
R_{16} — 1000 ohms.
R_{17} — 3900 ohms.
R_{18} — 0.1 megohm.
R_{19} — 22,000 ohms.
R_{20} — 220 ohms.
Resistors ½ watt unless otherwise specified.
L_1, L_2 — 3.5 Mc.: 40 turns No. 26 e., 1″ d., close-wound.
L_3 — 9 turns No. 22 enam., close-wound next to cold end of L_2.
L_4 — 0.25 henry (Millen 34400-250).
J_1, J_3 — Cable connector (Jones S-201).
J_2 — Closed-circuit midget jack.
J_4 — Microphone-cable connector (Amphenol PC1M).
RFC_1, RFC_2 — 2.5 mh.

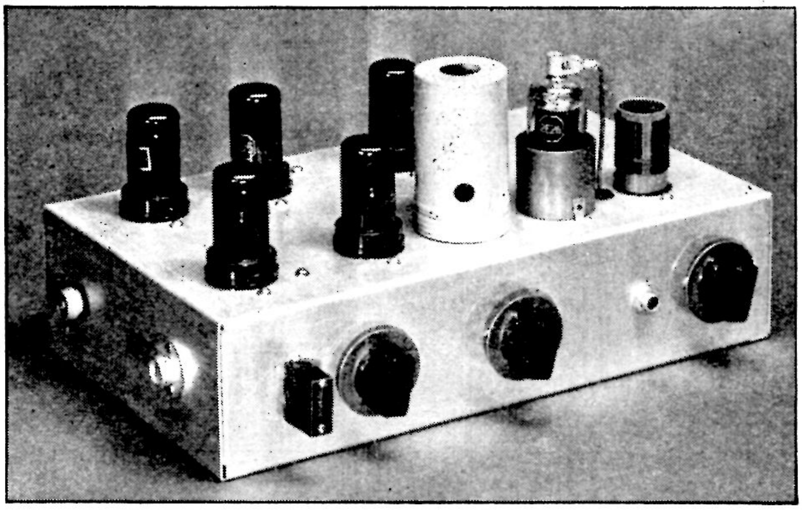


Fig. 9-58 — The narrow-band PM exciter is built in simple style. The r.f. stages are mounted along the front, and from left to right the tubes are 6J5, 6SK7 and 2E26. Note the shield can for the 6SK7 plate coil. The tubes at the rear, from left to right, are 6SJ7, 6J5 and 6SG7.

cable, but no other shielding should be necessary. The excitation control, R_3, was made to be adjusted by a screwdriver, since ordinarily there is little need for changing the setting once it has been established.

The power-supply requirements are 500 volts at 60 ma. (or whatever is required for input to the 2E26 plate), 250 volts at 25 ma., 150 volts regulated (by a VR-150) at about 12 ma., and 6.3 volts a.c. at 2.3 amperes. If the 150-volt supply is not regulated, C_{27} should be shunted with an 8-μfd. electrolytic condenser, to avoid audio degeneration.

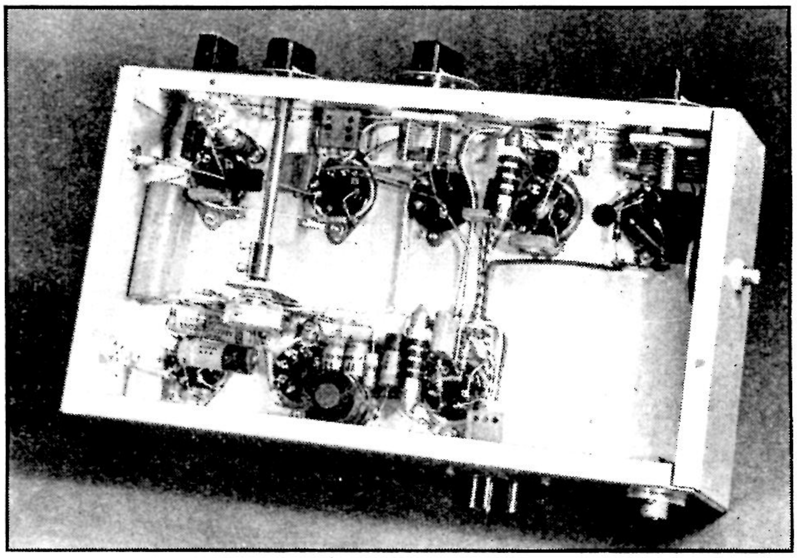


Fig. 9-59 — The r.f. and audio wiring of the PM exciter are kept separated as much as possible. To carry out this scheme, the audio gain control is mounted on a small bracket, and a long shaft is brought out to the panel.

Tuning

To set the unit in operation after the power supply has been connected, allow the heaters to warm up and apply plate power. If the crystal oscillator is being used, the circuit will work without adjustment and, with R_3 set at minimum resistance, a meter plugged in at J_2 should read 3 or 4 ma. when C_8 is tuned to resonate the 6SK7 plate circuit. With VFO input, the excitation may be more than this. The 2E26 plate circuit can be tuned using a milliammeter in the 500-volt lead or by reading grid current on the following stage. When the two r.f. circuits have been resonated, the gain of the 6SK7 stage should be reduced, by increasing the resistance at R_3 until the grid current is between 2.5 and 3 ma. Monitoring the signal in a receiver (at reduced gain and with no antenna connected), the proper setting of R_{15} for the microphone in use can be found. The audio gain control, R_{15}, will normally be set near maximum for work on 3.9 Mc., but it will be necessary to reduce the setting for 14- and 29-Mc. operation. Working with a 3.9-Mc. crystal or VFO, best results on 75-meter 'phone will be obtained when the receiving operator uses his crystal filter for pure PM reception.

Checking FM and PM Transmitters

Accurate checking of the operation of an FM or PM transmitter is considerably more difficult than the corresponding checks on an AM set. This is because the common forms of measuring devices either indicate amplitude variations only (a d.c. milliammeter, for example), or because their indications are most easily interpreted in terms of amplitude. There is no simple instrument that indicates frequency deviation in a modulated signal directly, in the same fashion that an oscilloscope will indicate the instantaneous modulation percentage of an

AM signal. To check an FM signal we must first establish the relationship between frequency deviation and the amplitude of the speech signal that is causing the deviation. Then the amplitude of the speech signal becomes a measure of the frequency deviation.

There is one very favorable feature in FM or PM checking. The modulation takes place at a very low level and the stages following the one that is modulated do not affect the frequency deviation so long as they are properly tuned. Therefore the modulation may be checked *without putting the transmitter on the air*, or even on a dummy antenna. The power is simply cut off the amplifiers following the modulated stage. This not only avoids unnecessary interference to other stations during testing periods, but also keeps the signal at such a low level that it may be observed quite easily on the station receiver. A good receiver with a crystal filter is an essential part of the checking equipment of an FM or PM transmitter, particularly for narrow-band FM or PM.

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

Reactance-Tube FM

It was explained earlier that in FM the frequency deviation is the same at any audio modulation frequency if the audio signal amplitude does not vary. Since this is true at *any* audio frequency it is true at zero frequency. Consequently it is possible to calibrate a reactance modulator by applying an adjustable d.c. voltage to the modulator grid and noting the change in oscillator frequency as the voltage is varied. A suitable circuit for applying the adjustable voltage is shown in Fig. 9-60. The battery, *B*, should have a voltage of 3 to 6 volts (two or more dry cells in series). The arrows indicate clip connections so that the battery polarity can be reversed.

The oscillator frequency deviation should be measured by using a receiver in conjunction with an accurately-calibrated frequency meter,

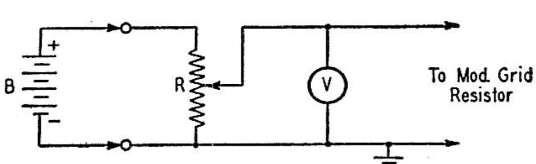


Fig. 9-60 — D.c. method of checking frequency deviation of a reactance-tube modulated oscillator. A 500- or 1000-ohm potentiometer may be used at *R*.

or by any means that will permit accurate measurement of frequency differences of a few hundred cycles. One simple method is to tune in the oscillator on the receiver (disconnect the receiving antenna, if necessary, to keep the signal strength well below the overload point) and then set the receiver b.f.o. to zero beat. Then increase the d.c. voltage applied to the modulator grid from zero in steps of about ½ volt and note the beat frequency at each change. Then reverse the battery terminals and repeat. The frequency of the beat-note may be measured by comparison with a calibrated audio-frequency oscillator, or by comparison with a piano or other musical instrument (see Chapter Twenty-Four for frequen-

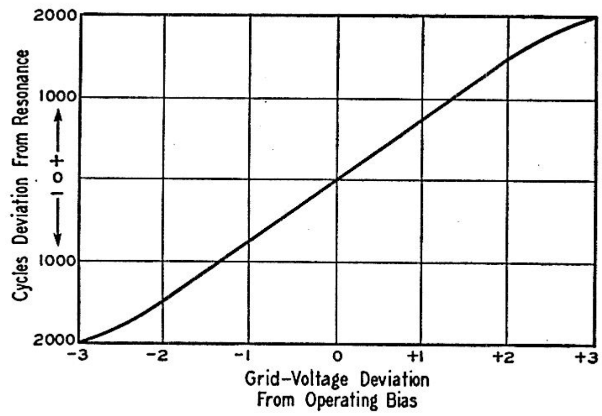


Fig. 9-61 — A typical curve of frequency deviation *vs.* modulator grid voltage. This curve was taken on the FM modulator unit described in this chapter (6L7 modulator and oscillator operating on 7 Mc.).

cies of musical tones). Note that with the battery polarity positive with respect to ground the radio frequency will move in one direction when the voltage is increased, and in the other direction when the battery terminals are reversed. When a number of readings has been taken a curve may be plotted to show the relationship between grid voltage and frequency deviation.

A sample curve is shown in Fig. 9-61. The usable portion of the curve is the center part which is essentially a straight line. The bending at the ends indicates that the modulator is no longer linear; this departure from linearity will cause harmonic distortion and will broaden the channel occupied by the signal. In the example, the characteristic is linear 1.5 kc. on either side of the center or carrier frequency. This is the maximum deviation permissible at the frequency at which the measurement is made. At the final output frequency the deviation will be multiplied by the same number of times that the measurement frequency is multiplied. This must be kept in mind when the check is made at a frequency that differs from the output frequency.

A good modulation indicator is a "magic-eye" tube such as the 6E5. This should be connected across the grid resistor of the reactance modulator as shown in Fig. 9-62. Note its deflection (using the d.c. voltage method as in Fig. 9-60) at the maximum deviation to be used. This deflection represents "100-per-cent modulation" and with speech input the gain should be kept at the point where it is just reached on voice peaks. If the transmitter is used on more than one band, the gain control should be marked at the proper setting for each band, because the signal amplitude that gives the correct deviation on one band will be either too great or too small on another. For narrow-band FM the proper deviation is approximately 2000 cycles (based on an upper a.f. limit of 3000 cycles and a deviation ratio of 0.7) at the final *output* frequency. If the output frequency is in the 29-Mc. band and the oscillator is on 7 Mc., the deviation at the *oscillator* frequency should not exceed 2000/4, or 500 cycles.

Checking with a Crystal-Filter Receiver

With PM the d.c. method of checking just described cannot be used, because the frequency deviation at zero frequency also is zero. For narrow-band PM it is necessary to check the actual channel-width occupied by the transmission. (The same method also can be used to check FM.) For this purpose it is necessary to have a crystal-filter receiver and an a.f. oscillator that generates a 3000-cycle sine wave.

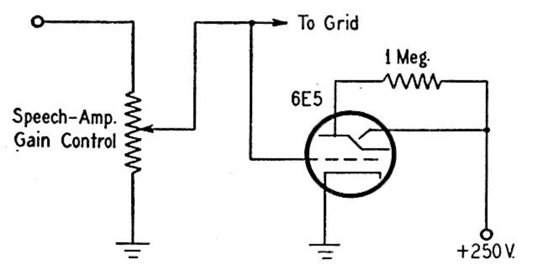


Fig. 9-62 — 6E5 modulation indicator for FM or PM modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.

Keeping the signal intensity in the receiver at a medium level, tune in the carrier at the *output* frequency. Do not use the a.v.c. Switch on the beat oscillator, and set the crystal filter at its sharpest position. Peak the signal on the crystal and adjust the b.f.o. for any convenient beat-note. Then apply the 3000-cycle tone to the speech amplifier (use the connections shown in Fig. 9-43 to avoid overloading) and increase the audio gain until there is a small amount of modulation. Tuning the receiver on either side of the carrier frequency will show the presence of sidebands 3 kc. from the carrier on both sides. With low audio input, these two should be the *only* sidebands detectable.

Now increase the audio gain and tune the receiver over a range of about 10 kc. on both sides of the carrier. When the gain becomes high enough, a second set of sidebands spaced 6 kc. on either side of the carrier will be detected. The signal amplitude at which these sidebands become detectable is the maximum speech amplitude that should be used. If the 6E5 modulation indicator is incorporated in the modulator, its deflection with the 3000-cycle tone will be the "100-per-cent modulation" deflection for speech.

When this method of checking is used with a reactance-tube modulated FM (not PM) transmitter, the linearity of the system can be checked by observing the *carrier* as the a.f. gain is slowly increased. The beat-note frequency will stay constant so long as the modulator is linear, but nonlinearity will be accompanied by a shift in the average carrier frequency that will cause the beat-note to change in frequency. If such a shift occurs at the same time that the 6-kc. sidebands appear, the extra sidebands may be caused by modulator distortion rather than by an excessive modulation index. This means that the modulator is not able to shift the frequency over a wide-enough range — a situation comparable to an AM transmitter that is not capable of 100-per-cent modulation. The 6-kc. sidebands should appear *before* there is any shift in the carrier frequency.

R.F. Amplifiers

The r.f. stages in the transmitter that follow the modulated stage may be designed and adjusted as in ordinary operation. In fact, there are no special requirements to be met except that all tank circuits should be carefully tuned to resonance (to prevent unwanted r.f. phase shifts that might interact with the modulation and thereby introduce hum, noise and distortion). In neutralized stages, the neutralization should be as exact as possible, also to minimize unwanted phase shifts. With FM and PM, all r.f. stages in the transmitter can be operated at the manufacturer's maximum c.w.-telegraphy ratings, since the average power input does not vary with modulation as it does in AM 'phone operation.

The output of the transmitter should be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated-carrier value when the transmitter is modulated. If there is no antenna ammeter in the transmitter, a flashlight lamp and loop can be coupled to the final tank coil to serve as a current indicator. If the carrier amplitude is constant, the lamp brilliance will not change with modulation.

Amplitude modulation accompanying FM or PM is just as much to be avoided as frequency or phase modulation that accompanies AM. A mixture of AM with either of the other two systems results in the generation of spurious sidebands and consequent widening of the channel. If the presence of AM is indicated by variation of antenna current with modulation, the cause is almost certain to be nonlinearity in the modulator. In very wide-band FM, it is possible for the selectivity of the tank circuits in the transmitter to cause the amplitude to decrease at high deviations, but this is not likely to occur on the amateur frequencies at which wide-band FM would be used. It is a practical certainty that it cannot occur with narrow-band FM or PM, except when an amplifier stage is on the verge of self-oscillation.

CHAPTER 10

Antennas and Transmission Lines

The radio-frequency power that is generated by a transmitter serves a useful purpose only when it is radiated out into space in the form of electromagnetic waves. It is the antenna's job to convert the power into radio waves as efficiently as possible, and to direct the waves where they will do the most good in communication. To do so, the antenna usually must be located well above the ground and kept as far as possible from buildings, trees, and other objects that might absorb energy. This raises a problem, because by some means or another the power that is generated inside the station, in the transmitter, must be conveyed to the antenna. The usual means is a **transmission line.**

There is thus a natural association between antennas and transmission lines — an association that has frequently led to the quite mistaken belief that an antenna fed by a particular type of transmission line is a better (or worse) radiator than exactly the same type of antenna fed by a different type of transmission line. The fact is that a transmission line can be used to carry power to any sort of device — not just an antenna — capable of receiving it. Nor does the antenna care by what means it gets the power; the amount it receives will be radiated just as well no matter by what system it was conveyed to the antenna.

While it would be dangerous to carry the comparison very far, there are nevertheless some similarities between transmission lines at radio frequencies and the lines used for carrying 60-cycle power from the generating station to the consumer. Some lines are best adapted to carrying power at high voltage and relatively low current; the reverse is true of others. Like a lot of other electrical devices, some antennas want a relatively large current at low voltage, while others want high voltage at low current. We connect a 115-volt lamp, for example, to a 115-volt power line, but if we have a 6-volt lamp and want to run it from the 115-volt line we have to use a transformer to reduce the voltage. Similarly, if an antenna wants high current at low voltage (low impedance) and the transmission line is of a type that is best adapted to carrying power at low current and high voltage (high impedance), we need a transforming device comparable to the transformer used with the 6-volt lamp.

Fig. 10-1 — **The principal elements in the system connecting the transmitter and antenna.**

The power company's generators usually do not generate the voltage that is wanted on the power line, so an appropriate transformer is connected between the generator and the line. Similarly, the voltage generated in the tank circuit of the final amplifier in the transmitter usually has to be transformed to a value that "fits" the transmission line used. At radio frequencies, it is more convenient to talk in terms of impedance rather than voltage, so we speak of "impedance transformations" rather than "voltage transformations." In general, we have a complete system like that shown in block form in Fig. 10-1. Perhaps this looks complicated, but the power-line analogy should help make it understandable. The equipment itself is not particularly complex, and seems even less so when the underlying necessity for it is appreciated.

Transmission Lines

At power-line frequencies — and even at rather high radio frequencies when we are dealing with tuned circuits that are physically rather small — it is habitual to think of current as flowing "around" the circuit. In a series circuit, for example, it is assumed that the current has the same value at every point in the circuit. Indeed, all the explanations of circuit action in Chapter Two are based on this assumption.

The assumption can be true only if electrical and magnetic effects take place instantaneously all around the circuit. The fact is, though, that the action is *not* instantaneous. The fastest that an electromagnetic field can travel is 300,000,000 meters, or 186,000 miles, per second. This is a tremendous speed, and is so great that in many circuits the action *appears* to be instantaneous. When that is so we can ignore the fact that it takes a certain amount of time for an electrical effect that occurs at one point to be felt at another a short distance away.

But there are other circuits in which time becomes an all-important factor. The transmission lines used to carry radio-frequency power are typical circuits in which time cannot be neglected.

CURRENT FLOW IN LONG LINES

Suppose we have a battery connected to a pair of parallel wires that extends to a very great distance, as in Fig. 10-2. At the moment

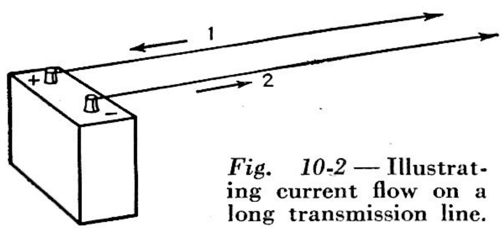


Fig. 10-2 — Illustrating current flow on a long transmission line.

the battery is connected to the wires, electrons in wire No. 1 near the positive terminal of the battery will be attracted to the battery, and the same number of electrons in wire No. 2, near the battery terminal, will be repelled outward along the wire. The directions are shown by the arrows. Thus a current flows in both wires at the instant the battery is connected. These currents do *not* flow throughout the entire length of both wires simultaneously. They start instantaneously *in both wires* at the battery terminals, but a definite time interval will elapse before they are evident at a distance from the battery.

The time interval may be very small. For example, one-millionth of a second (one microsecond) after the connection is made the currents in the wires will have traveled 300 meters, or nearly 1000 feet, from the battery terminals. Note that they flow in both wires simultaneously, even though there may be no connection between the two wires at the end (which is infinitely far away) to form what we ordinarily think of as a closed circuit.

The current is in the nature of a charging current, flowing to charge the capacitance between the two wires. But unlike an ordinary condenser, the conductors of this "linear" condenser have appreciable inductance. In fact, we may think of the line as being composed of a whole series of small inductances and capacitances connected as shown in Fig. 10-3, where each coil is the inductance of a very short section of one wire and each condenser is the capacitance between two such short sections.

Characteristic Impedance

An indefinitely-long chain of coils and condensers connected as in Fig. 10-3, where each L is the same as all others and all the Cs have the same value, has an interesting and important peculiarity. To an electrical impulse applied to one end, the combination (or transmission line) appears to have an impedance that is approximately equal to $\sqrt{L/C}$, where L and C are the inductance and capacitance per unit length. Furthermore, this impedance is purely resistive. The line will "look like" such an impedance only when it is infinitely long, but even a short line can be made to "think" it is infinitely long by means to be described a little later.

This inherent line impedance is called the **characteristic impedance** or **surge impedance** of the line. Its value is determined by the inductance and capacitance per unit length. These quantities in turn depend upon the size of the line conductors and the spacing between them. The closer the two conductors of the line and the greater their diameter, the higher the capacitance and the lower the inductance. A line with large conductors closely spaced will have low impedance, while one with small conductors widely spaced will have relatively high impedance.

The characteristic impedance of the line is a very important property. For one thing, it determines the amount of current that can flow when a voltage is applied to the line. When a line is infinitely long, the current is simply equal to E/Z_o, where E is the voltage applied to the line and Z_o is the characteristic impedance. This has nothing to do with the *resistance* of the conductors; in fact, in this simplified picture of a transmission line we

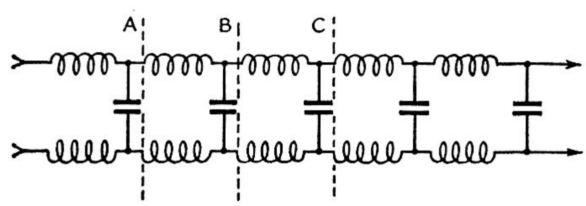


Fig. 10-3 — Equivalent of a transmission line in lumped circuit constants.

have tacitly assumed that the conductors do not have any resistance. The line is an impedance (like any circuit composed of L and C, without any R) that does not consume power. Actually, of course, the conductors do have resistance, so power cannot be transmitted along the line without some loss. But if the line is properly constructed and operated, this loss will be small compared with the amount of power carried to the load to do useful work.

R.F. on Lines

Bearing in mind that *time* must elapse before the currents initiated at the "input" end of the line — that is, the end to which the source of power is connected — can appear some distance away, consider now what happens when a radio-frequency voltage is applied to a transmission line. Suppose an r.f. generator is connected to a long line as shown in Fig. 10-4. To make the figures easy, assume that the frequency is 10 Mc., or 10,000,000 cycles per second. Then each cycle will occupy 0.1 microsecond, as shown by the drawing of the applied voltage. Suppose that the points *B* and *D* along the line are 30 meters away from *A* and *C*, respectively. If the current travels with the velocity of light, in 0.1 microsecond (one cycle) it will move 30 meters (300,000,000 meters divided by 10,000,000 cycles) along the line. This is a distance of one wavelength. Thus any currents observed at *B* and *D* occur just one cycle later in time than the currents at *A* and *C*. To put it another way, the currents initiated at *A* and *C* do not appear at *B* and *D*, one wavelength away, until the applied voltage has had time to go through a complete cycle.

Since the applied voltage is always changing, the currents at *A* and *C* are changing in proportion. The current a short distance away from *A* and *C* — for instance, at *X* and *Y* — is not the same as the current at *A* and *C* because the current at *X* and *Y* was caused by a value of voltage that occurred slightly earlier in the cycle. This is true all along the line; at any instant the current anywhere along the line from *A* to *B* and *C* to *D* is different from the current at every other point in that same distance. The series of drawings shows how the instantaneous currents might be distributed if we could snapshot them at intervals of one-quarter cycle. The current travels out from the input end of the line in waves.

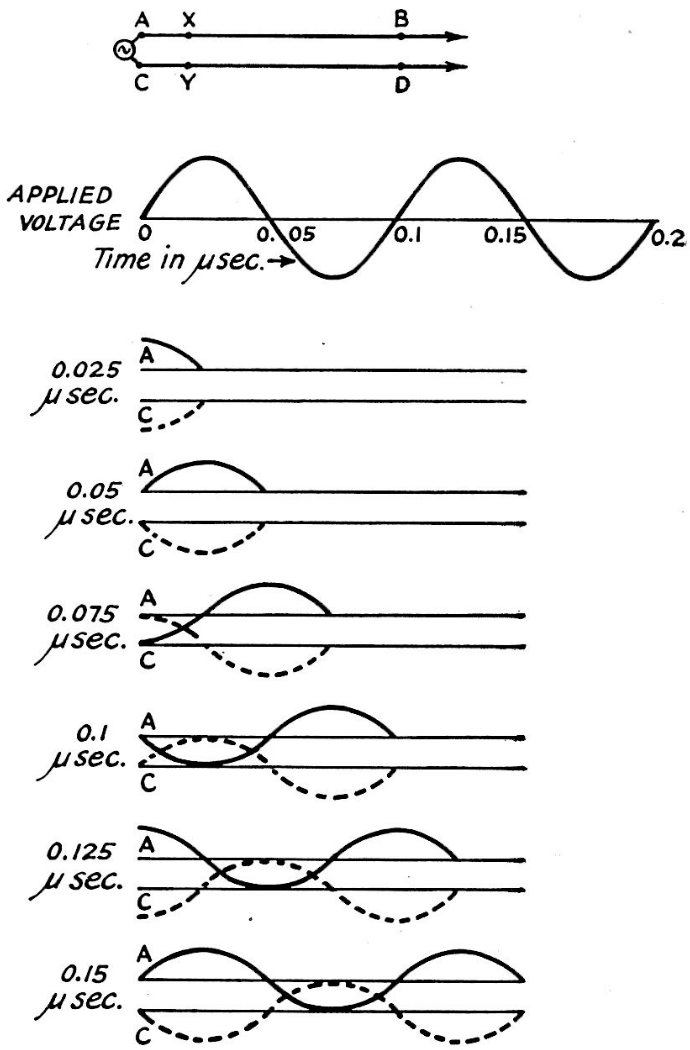


Fig. 10-4 — Progression of radio-frequency current flow in a transmission line.

At any selected point on the line the current goes through its complete range of a.c. values in the time of one cycle just as it does at the input end. Therefore (if there are no losses) an ammeter inserted in either conductor would read exactly the same current at any point along the line, because the ammeter averages the current over a whole cycle. The *phases* of the currents at any two separated points would be different, but an ammeter would not show this.

"Matched" Lines

In this picture of current traveling along a transmission line we have assumed that the line was infinitely long. Lines have a definite length, of course, and they are connected to or **terminated** in a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, the current traveling along the line to the load does not find conditions changed in the least when it meets the load; in fact, the load just looks like still more transmission line of the same characteristic impedance. Consequently, connecting such a load to a short transmission line allows the current to travel in exactly the same fashion as it would on an infinitely-long line.

In other words, a short line terminated in a purely-resistive load equal to the characteristic impedance of the line acts just as though it were infinitely long. Such a line is said to be **matched.** In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

STANDING WAVES

Now suppose that the line is terminated in a load that is not equal to the line's characteristic impedance. To take an extreme case, suppose that the output end of the line is short-circuited, as in Fig. 10-5.

With the infinitely-long line (or its matched counterpart) the impedance was the same at any point on the line and therefore the ratio of voltage to current was the same at any point on the line. However, the impedance at the end of the line in Fig. 10-5 is zero — or at least extremely small. A given amount of power in a very low impedance will result in a very large current and a very small voltage, as compared with the current-voltage ratio that exists in a few hundred ohms — which is a typical impedance value for some types of transmission lines. Something has to happen, therefore, when the power traveling along the transmission line meets the short-circuit at the end.

What happens is that the outgoing power, on meeting the short-circuit, simply reverses its direction of flow and goes back along the transmission line toward the input end. It has nowhere else to go. There is a very large current in the short-circuit, but substantially no voltage across the line at this point. We now have a voltage and current representing the power going outward toward the short-circuit, and a second voltage and current representing the **reflected** power traveling back toward the source.

Consider only the two current components for the moment. The reflected current travels at the same speed as the outgoing current, so its instantaneous value will be different at every point along the line, in the distance represented by the time of one cycle. At some points along the line the outgoing and reflected currents will be in phase while at other points they will be completely out of phase. At the out-of-phase points the currents cancel each other (if the outgoing and reflected currents have the same value, as they will if all the power is reflected) and so at those points the resultant current is zero. At the in-phase points the two currents add numerically. At in-between points the two currents are neither completely in nor completely out of phase and so their sum is not equal to their numerical sum, but is something less.

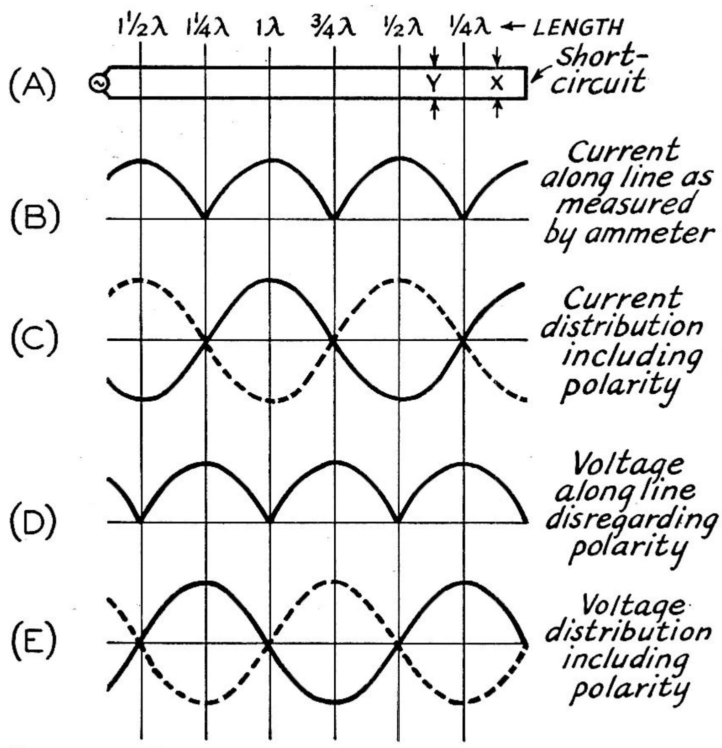


Fig. 10-5 — Standing waves of voltage and current along a short-circuited transmission line.

The points at which the currents are in and out of phase depend only on the *time* required for them to travel and so depend only on the *distance* along the line from the point of reflection. The phase is completely reversed when the current travels for one-half cycle — that is, a distance of one-half wavelength — and is back in the in-phase condition when the current has traveled for one whole cycle, or one wavelength.

In the short-circuit at the end of the line the total current is high and the two current components are in phase. Therefore at a distance of *one-half* wavelength back along the line from the short-circuit the outgoing and reflected components will again be in phase and the current will have its maximum value. This is also true at any point that is a multiple of a half-wavelength from the short-circuited end of the line. The distance along the line is one-half wavelength because the current has to travel the distance twice in order to "meet itself coming back."

Since a total distance of one-half wavelength gives a complete reversal of phase, the outgoing and reflected currents will cancel at a point *one-quarter* wavelength, along the line, from the short-circuit. At this point, then, the current will be zero. It will also be zero at all points that are an *odd* multiple of one-quarter wavelength from the short-circuit.

If the current along the line is measured at successive points with an ammeter, it will be found to vary about as shown in Fig. 10-5B. The same result would be obtained by measuring the current in either wire, since the ammeter cannot measure phase. However, if the phase could be checked, it would be found that in each successive half-wavelength section of the line the currents at any given instant are flowing in opposite directions, as indicated by the solid line in Fig. 10-5C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first wire, as a result of the electron movement discussed in connection with Fig. 10-2. This is indicated by the broken curve in Fig. 10-5C. The variations in current intensity along the transmission line are referred to as **standing waves.** The point of maximum line current is called a **current loop** and the point of minimum line current a **current node.**

Voltage Relationships

Since the end of the line is short-circuited, the voltage at that point has to be zero. This can only be so if the voltage in the outgoing wave is met, at the end of the line, by an equal voltage of opposite polarity. In other words,

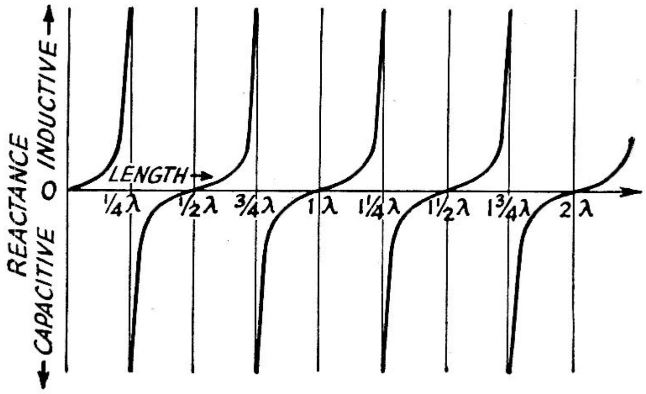


Fig. 10-6 — **Input reactance *vs.* length of a short-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.**

the phase of the voltage wave is *reversed* when reflection takes place from the short-circuit. This reversal is equivalent to an extra half-cycle or half-wavelength of travel. As a result, the outgoing and returning voltages are in phase a quarter wavelength from the end of the line, and again out of phase a half-wavelength from the end. The standing waves of voltage, shown at D in Fig. 10-5, are therefore displaced by one-quarter wavelength from the standing waves of current. The drawing at E shows the voltages on both wires when phase is taken into account. The polarity of the voltage on each wire reverses in each half-wavelength section of transmission line. A voltage maximum on the line is called a **voltage loop** and a voltage minimum is called a **voltage node.**

Input Impedance

It is apparent, from examination of B and D in Fig. 10-5, that at points that are a multiple of a half-wavelength — i.e., ½, 1, 1½ wavelengths, etc. — from the short-circuited end of the line the current and voltage have the same values that they do at the short-circuit. In other words, if the line were an exact multiple of a half-wavelength long the generator or source of power would "look into" a short-circuit. On the other hand, at points that are an odd multiple of a quarter wavelength — i.e., ¼, ¾, 1¼, etc. — from the short-circuit the voltage is maximum and the current is zero. Since $Z = E/I$, the impedance at these points is theoretically infinite. (Actually it is very high, but not infinite. This is because the current does not actually go to zero when there are losses in the line. Losses are always present, but usually are small enough so that the impedance is of the order of tens or hundreds of thousands of ohms.)

At either the odd or even multiples of a quarter wavelength the impedance is a pure resistance, because at these points the current and voltage in the transmission line are exactly in phase.

A detailed study of the outgoing and reflected components of voltage and current will show that at a point such as X in Fig. 10-5, lying anywhere in the section of line between the short-circuit and the first quarter-wavelength point, the current lags behind the voltage. This is exactly what happens in an inductance, so it can be said that a section of short-circuited transmission line less than a quarter wavelength long has inductive reactance. The value of reactance is determined by the ratio of voltage to current at the input end of such a line. It is evident from B and D in Fig. 10-5 that the reactance is low when the line is quite short, and highest when the line is nearly a quarter wavelength long. The line also has inductive reactance when its length is between one-half and three-quarter wavelength, between one and one-and-one-quarter wavelengths, and so on.

On the other hand, in the section of line between one-quarter and one-half wavelength from the short-circuit the current leads the voltage, so a short-circuited line having a length between these two limits "looks like" a capacitive reactance to the generator to which it is connected. The reactance is highest when the line is just over one-quarter wavelength long, and lowest when the line is just less than one-half wavelength long. Fig. 10-6 shows the general way in which the reactance varies with different line lengths.

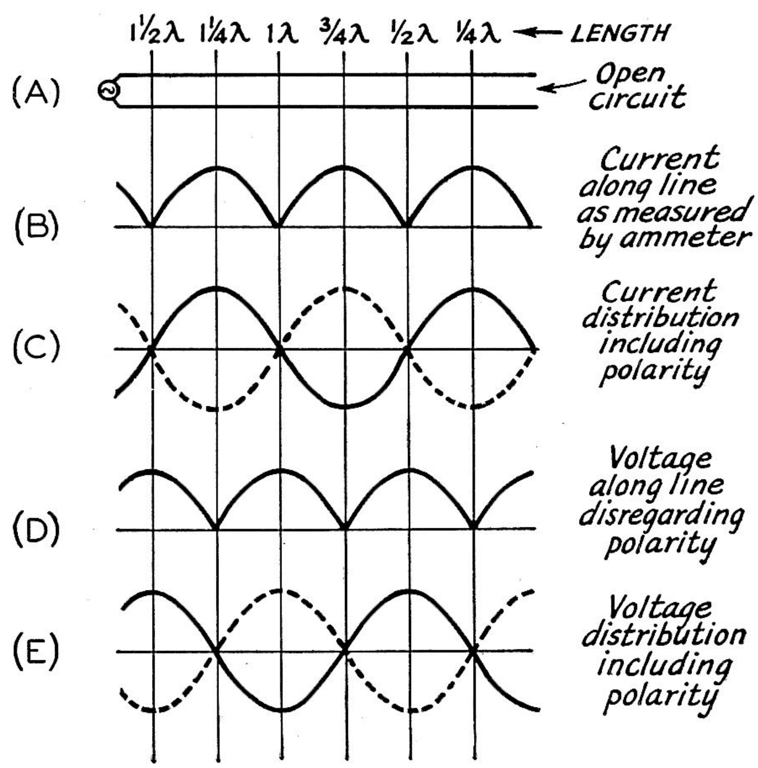


Fig. 10-7 — **Standing waves of current and voltage along an open-circuited transmission line.**

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source because it has nowhere else to go. In this case, the outgoing and reflected components of *current* must be equal and opposite in phase in order for the total current at the end of the line to be zero. The outgoing and reflected components of voltage are in phase, however, and add together. The result is that we again have standing waves, but the conditions are reversed. Fig. 10-7 shows the open-circuited line case. It may be compared directly with Fig. 10-5. The impedance looking into the line toward the open end is purely resistive at each multiple of one-quarter wavelength. It is very low at odd multiples of one-quarter wavelength, and very high at even multiples. In fact, an open-circuited line and short-circuited line behave just alike *if* the length of one differs by one-quarter wavelength from the length of the other.

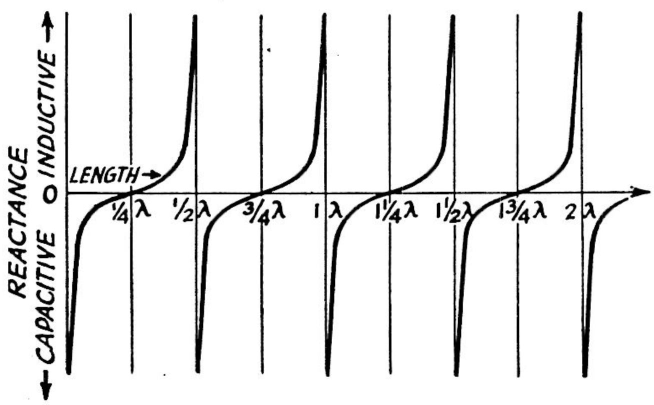


Fig. 10-8 — **Input reactance *vs.* length of an open-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.**

Fig. 10-8 shows how the reactance varies with line length for the open-circuited line. Comparing this with Fig. 10-6 shows that the reactance of any given length of line is of the opposite type to that obtained with a short-circuited line of the same length.

Lines Terminated in Resistive Load

An open- or short-circuited line does not deliver any power to a load, and for that reason is not, strictly speaking, a "transmission" line. However, the fact that a line of the proper length has an extremely high resistive input impedance at a given frequency or wavelength makes such lines useful as substitutes for the more common coil-and-condenser resonant circuits. With proper design, the effective Q of such a "linear" resonant circuit is much higher than is obtainable with coils and condensers. Linear circuits are particularly useful at v.h.f., and their application in that field is discussed in later chapters. In this chapter we are concerned with lines delivering power to a load such as an antenna.

Fig. 10-9 shows a line terminated in a resistive load. In such a case at least part of the outgoing power is absorbed in the load, and so is not available to be reflected back toward the source. Because only part of the power is reflected, the reflected components of voltage and current do not have the same magnitude as the outgoing components. Therefore there is no such thing as complete cancellation of either voltage or current at any point along the line. However, the *speed* at which the outgoing and reflected components travel is not affected by their amplitude, so the phase relationships are similar to those in open- or short-circuited lines.

It was pointed out earlier that if the load resistance, which we will call Z_r, is equal to the characteristic impedance, Z_o, of the line all the power is absorbed in the load. In such a case there is no reflected power and therefore no standing waves of current and voltage. This is a special case that represents the changeover point between "short-circuited" and "open-circuited" lines. If Z_r is less than Z_o, the current is largest at the load and the reflected component of voltage is out of phase with the outgoing component at the load. If Z_r is greater than Z_o, the voltage is largest at the load and the reflected component of current is out of phase with the outgoing component. Thus, if Z_r is less than Z_o the current will be minimum at a point one-quarter wavelength from the load and at every point an odd number of quarter wavelengths away, while the voltage will be maximum at these same points. The current will be maximum and the voltage minimum at points that are multiples of one-half wavelength from the load. If Z_r is greater

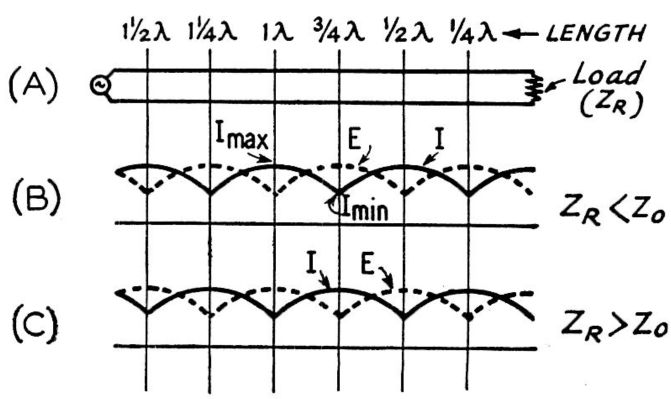


Fig. 10-9 — **Standing waves on a transmission line terminated in a resistive load.**

than Z_o, the opposite is true. The two conditions are shown at B and C, respectively, in Fig. 10-9.

The impedance looking into the line toward the load will be purely resistive (if Z_r is a pure resistance) when the line length is a multiple of a quarter wavelength, just as in the open- and short-circuited cases. The input impedance is *equal* to Z_r when the line is an *even* multiple of a quarter wavelength long. If Z_r is less than Z_o the impedance is maximum when the line is an odd multiple of a quarter wavelength long. In such a case the impedance looking into the line is

$$Z_s = \frac{Z_o^2}{Z_r} \qquad \textbf{(10-A)}$$

where Z_s = Impedance looking into line (line length an odd multiple of one-quarter wavelength)
Z_r = Impedance of load (must be pure resistance)
Z_o = Characteristic impedance of line

Example: A quarter-wavelength line having a characteristic impedance of 500 ohms is terminated in a resistive load of 75 ohms. The impedance looking into the input or sending end of the line is

$$Z_s = \frac{Z_o^2}{Z_r} = \frac{(500)^2}{75} = \frac{250{,}000}{75} = 3333 \text{ ohms}$$

When Z_r is greater than Z_o, the input impedance reaches its minimum value when the line is an odd multiple of a quarter wavelength long The value of input impedance in this case also is given by the equation above.

Example: A quarter-wave line is terminated in a resistive load of 1200 ohms. The characteristic impedance of the line is 600 ohms. Then the input impedance of the line is

$$Z_s = \frac{Z_o^2}{Z_r} = \frac{(600)^2}{1200} = \frac{360{,}000}{1200} = 300 \text{ ohms}$$

Impedance Transformation

If the formula in the preceding discussion is rearranged, we have

$$Z_o = \sqrt{Z_s Z_r} \qquad \textbf{(10-B)}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product. A quarter-wave line, in other words, has the characteristics of a transformer. This is a very useful attribute of transmission lines.

Example: A 600-ohm transmission line is to be used to feed an antenna that has a resistive impedance of 75 ohms. It is desired that the line operate without standing waves, and it must therefore be terminated in a resistive load equal to its characteristic impedance; i.e., 600 ohms. A quarter-wave line or "linear transformer" is to be used to match the 75-ohm load to the line impedance. To do this, the characteristic impedance of the quarter-wave transformer must be

$$Z_o = \sqrt{Z_s Z_r} = \sqrt{75 \times 600} = \sqrt{45{,}000} = 223 \text{ ohms}$$

Reactance of Terminated Lines

We have seen that a short-circuited line less than one-quarter wavelength long exhibits inductive reactance. Also, a line of any length terminated in a resistive load equal to its characteristic impedance always looks like a pure resistance to the source of power. When the load is purely resistive and has any value between zero and Z_o, a line less than a quarter wave long will show inductive reactance, but the reactive effects decrease the closer the value of Z_r approaches Z_o.

On the other hand, an open-circuited line less than one-quarter wavelength long exhibits capacitive reactance. If the line is terminated in a resistive load larger than Z_o, it continues to show capacitive reactance but the reactive effects are less the closer Z_r approaches Z_o in value.

In general, then, a line terminated in a resistive impedance less than Z_o will show reactance variations with length similar to those of a short-circuited line as given in Fig. 10-6. A line terminated in a resistive impedance greater than Z_o will show reactance variations with length similar to those of an open-circuited line as given in Fig. 10-8. The magnitudes of the reactances will be smaller the closer Z_r approaches Z_o in value.

Loads That Are Not Pure Resistance

In most amateur applications of transmission lines the load is — or should be — a pure resistance. At least, every attempt is made to make it so. However, there are cases where the load has reactance as well as resistance, and recognizing the symptoms of reactance in the load is of value in indicating what steps should be taken to convert the load to a pure resistance.

The situation is easier to visualize if a line terminated in a pure reactance is considered first. For example, suppose the line is terminated in a capacitive reactance as shown in Fig. 10-10A. It does not matter to the line what physical form the reactance takes; the important thing is that in it the current leads the voltage. The reactance might be a condenser, for example — or it might simply be an additional section of transmission line that exhibits capacitive reactance at its input end, as indicated in Fig. 10-10B.

From Fig. 10-8, we can see that a section of open-circuited transmission line less than one-quarter wavelength long will have capacitive reactance. By proper choice of line length, any desired value of reactance can be obtained. Conversely, any "lumped" reactance, such as a condenser, connected to the end of the transmission line can be replaced by a section of open-circuited line of appropriate length. If the condenser capacitance is small and its reactance therefore is high, only a short length of line is required. If the condenser capacitance is large and its reactance consequently is low, the additional line section must be nearly a quarter wavelength long. In other words, connecting a condenser across the end of the transmission line is equivalent to lengthening the line, electrically. The amount of effective lengthening depends on the capacitance of the condenser.

Once the equivalent lengthening is determined, we can simply look upon the line as one having the new length and apply all that has been said previously. In the case just considered, this would mean that the point of maximum current, instead of appearing exactly a quarter wavelength from the end of the open-circuited line, would appear at something less than a quarter wavelength from the end. This is shown in Fig. 10-10C. The larger the capacitance of the terminating condenser the closer the current loop comes to the physical end of the line. All the other loops and nodes of both current and voltage would be changed accordingly.

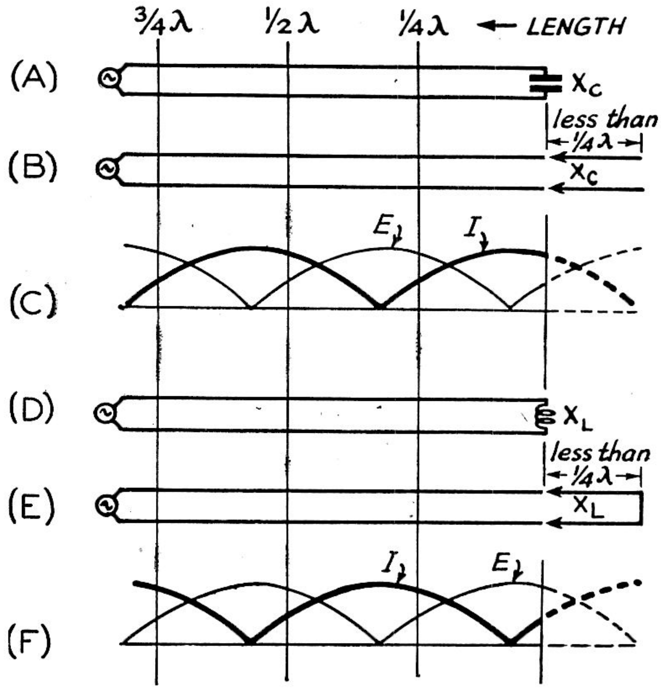


Fig. 10-10 — **Lines terminated in reactance. A reactive load is equivalent to a change in the length of the line.**

If the line is terminated in an inductance, we can substitute a short-circuited section of line less than one-quarter wavelength long for the lumped inductance. Thus, terminating a line in an inductance is equivalent to extending its length by something less than one-quarter wavelength and short-circuiting it. This is shown at D, E and F in Fig. 10-10. The larger the inductance, the greater the length of line, up to one-quarter wavelength, that must be added to obtain the electrical equivalent. When the equivalent section of line is substituted for the inductance, all that has been said about shorted lines applies, based on the new equivalent length.

When the load has both resistance and reactance the apparent length of the line is again increased. However, the amount of the apparent increase is affected by the resistance component of the load together with the reactive component. Inductive reactance will cause the first voltage maximum to appear less than one-quarter wavelength from the load, just as in Fig. 10-10F. Capacitive reactance will cause the first current maximum to appear less than one-quarter wavelength from the load, as in Fig. 10-10C. If the positions of either the voltage or current loops can be determined, it is always possible to tell whether the reactive component of the load is inductive or capacitive. When the load has both resistance and reactance, the voltage and current nodes do not reach zero because not all the outgoing power is reflected. The actual standing waves would be more like those shown in 10-9, but with the positions of the nodes and loops shifted as indicated in Fig. 10-10.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, as indicated in Fig. 10-11, is called the standing-wave ratio. It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is perfectly matched. (In that case the "maximum" and "minimum" current are the same, since the current does not vary along the line.) When the line is terminated in a purely-resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_r}{Z_o} \text{ or } \frac{Z_o}{Z_r} \qquad \textbf{(10-C)}$$

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

Z_r = Impedance of load (must be pure resistance)

Z_o = Characteristic impedance of line

Example: A line having a characteristic impedance of 300 ohms is terminated in a resistive load of 25 ohms. The s.w.r. is

$$S.W.R. = \frac{Z_o}{Z_r} = \frac{300}{25} = 12 \text{ to } 1$$

It is customary to put the larger of the two quantities, Z_r or Z_o, in the numerator of the fraction so that the s.w.r. will be expressed by a number larger than **1**.

It is easier to measure the standing-wave ratio than some of the other quantities (such as the impedance of an antenna) that enter into transmission-line computations. Consequently, the s.w.r. is a convenient basis for work with lines. The higher the s.w.r., the greater the mismatch between line and load. Also, the higher the s.w.r. the more marked are the reactive effects when the line length is not an exact multiple of a quarter-wavelength. *In practical* lines, the loss in the line itself increases with the s.w.r.

Resonant and Nonresonant Lines

A transmission line terminated in a resistive load equal to its characteristic impedance is commonly called a **flat, nonresonant** or **untuned** line. The line is "flat" because there are no standing waves, hence a graph of the current along the line is a straight line. It is "nonresonant" because the input impedance of such a line is pure resistance and does not change when the line length is changed.

When there are standing waves the line is said to be **resonant** or **tuned.** In this case the input impedance depends critically on the length of the line and the characteristics of the load. The input impedance is a pure resistance only when the line length is such that a current or voltage loop appears at the input end. The previous discussion has shown that the positions of these loops depends upon the characteristics of the load. At all other lengths the input impedance consists of both reactance and resistance. Under these conditions the line acts something like a circuit that is not tuned to resonance; it is difficult to make it "take power" until something is done to "tune" it — that is, to eliminate the reactance. When this is done the input impedance of the line is purely resistive and its resistance may be matched to the transmitter for optimum power transfer.

It should be noted that if there are standing waves on the line the input impedance, even when the reactance is tuned out, is never equal to the characteristic impedance of the line. Depending on the length of the line, the characteristics of the load, and the s.w.r., the input resistance may be considerably higher or considerably lower than the line's characteristic impedance. This introduces an element of uncertainty in coupling to the transmitter. In one special case, when the load is a pure resistance and the line is exactly one-half wavelength long, the input impedance of the line is a pure resistance equal to the load impedance.

The reactive or resonance effects increase with the s.w.r., as previously pointed out. Generally speaking, a line is satisfactorily flat if the s.w.r. does not exceed about 1.5 to 1, but if the s.w.r. is much larger it becomes necessary to tune out the input reactance.

Radiation

Whenever a wire carries alternating current the electromagnetic fields travel away into space with the velocity of light. At power-line frequencies the field that "grows" when the current is increasing has plenty of time to return or "collapse" about the conductor when the current is decreasing, because the alternations are so slow. But at radio frequencies fields that travel only a relatively short distance do not have time to get back to the conductor before the next cycle commences. The consequence is that some of the electromagnetic energy is prevented from being restored to the conductor; in other words, energy is radiated into space in the form of electromagnetic waves.

The amount of energy radiated depends, among other things, on the length of the conductor in relation to the frequency or wavelength of the r.f. current. If the conductor is very short compared to the wavelength the energy radiated will be small. However, a transmission line used to feed power to an antenna is not short in this sense; in fact, it is almost always an appreciable fraction of a wavelength long and may have a length of several wavelengths.

The lines previously considered have consisted of two parallel conductors of the same diameter. Provided there is nothing in the system to destroy symmetry, at every point along the line the current in one conductor has the same intensity as the current in the other conductor at that point, but the currents flow in opposite directions. This was shown in Figs. 10-5C and 10-7C. This means that the fields set up about the two wires have the same intensity, but *opposite directions.* The consequence is that the total field set up about such a transmission line is zero; the two fields "cancel out." Hence no energy is radiated.

Actually, the fields do not completely cancel out because for them to do so the two conductors would have to occupy the same space, whereas they are slightly separated. However, the cancellation is substantially complete if the distance between the conductors is very small compared to the wavelength. Radiation will be negligible if the distance between the conductors is 0.01 wavelength or less, provided the currents in the two actually are balanced as described.

The amount of radiation also is proportional to the current flowing in the line. Because of the way in which the current varies along the line when there are standing waves, the effective current, for purposes of radiation, becomes greater as the s.w.r. is increased. For this reason the radiation is least when the line is flat. However, if the conductor spacing is small and the currents are balanced, the radiation from a line with even a high s.w.r. is inconsequential. A small unbalance in the line currents is far more serious.

There is no factual basis for the common belief that the presence of standing waves on a transmission line always means that the line is radiating a great deal of r.f. energy. Tuned lines are perhaps more subject to the stray coupling effects described later in this chapter, simply because they are frequently cut to resonant lengths while any random length can be used for a flat line. It is the stray coupling that gives rise to excessive line radiation, not the presence of the normal type of standing wave on the transmission line.

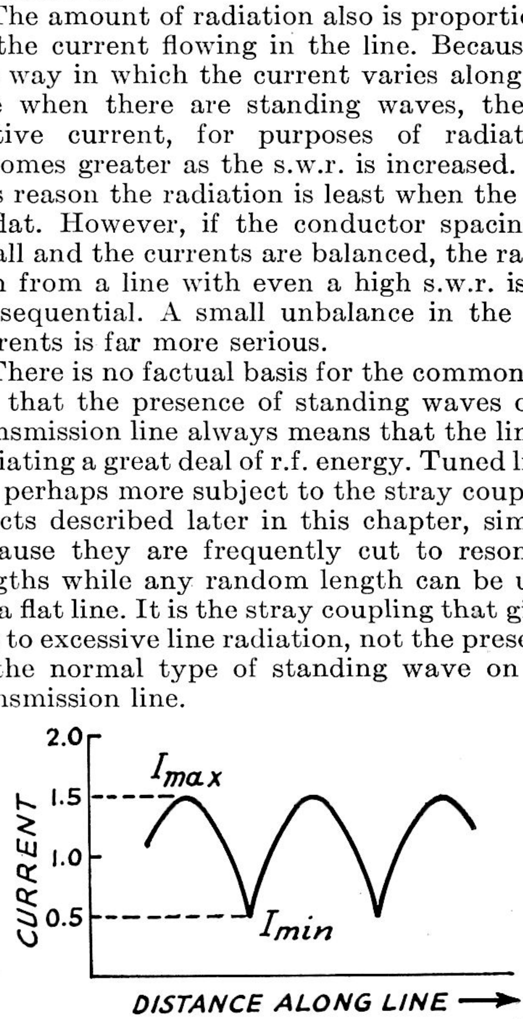


Fig. 10-11 — Measurement of standing-wave ratio. In this drawing, I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. $= I_{max}/I_{min} = 1.5/0.5 = 3$.

Practical Line Characteristics

The foregoing discussion of transmission lines has been based on a line consisting of two parallel conductors. Actually, the **parallel-conductor** line is but one of two general types. The other is the **coaxial** or **concentric** line. The coaxial line consists of a round conductor placed in the center of a circular tube. The inside surface of the tube and the outside surface of the smaller inner conductor form the two conducting surfaces of the line.

In the coaxial line the fields are entirely inside the tube, because the tube acts as a shield to prevent them from appearing outside. This reduces radiation to the vanishing point. So far as the electrical behavior of coaxial lines is concerned, all that has previously been said about the operation of parallel-conductor lines applies. There are, however, practical differences in their construction and use.

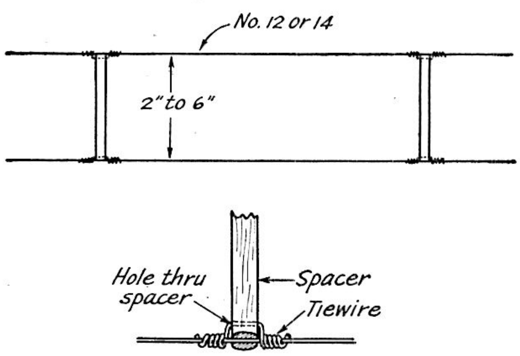


Fig. 10-12 — Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

Types of Construction

There are several constructional variations in both the basic types of transmission lines mentioned in the preceding section. Probably the most common type of transmission line used in amateur installations is a parallel-conductor line in which two wires (ordinarily No. 12 or No. 14) are supported a fixed distance apart by means of insulating rods called "spacers." The spacings used vary from two to six inches, the smaller spacings being necessary at frequencies of the order of 28 Mc. and higher so that radiation will be minimized. The construction is shown in Fig. 10-12. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 10-13. The characteristic impedance of such "open-wire" lines runs between about 400 and 600 ohms, depending on the wire size and spacing.

Parallel-conductor lines also are sometimes constructed of metal tubing of a diameter of ¼ to ½ inch. This reduces the characteristic impedance of the line. Such lines are mostly used as quarter-wave transformers, when different values of impedance are to be matched.

Two forms of "Twin-Lead" or "ribbon" transmission line are shown in Fig. 10-13. This is a parallel-conductor line with stranded conductors imbedded in low-loss insulating material (polyethylene). It has the advantages of light weight, compactness and neat appearance, together with close and uniform spacing. However, losses are higher in the solid dielectric than in air, and dirt or moisture on the line tends to change the characteristic impedance. Twin-Lead line is available in characteristic impedances of 75, 150 and 300 ohms.

The most common form of coaxial line consists of either a solid or stranded-wire inner conductor surrounded by polyethylene dielectric. Copper braid is woven over the dielectric to form the outer conductor, and a waterproof vinyl covering is placed on top of the braid. This cable is made in a number of different diameters. It is moderately flexible, and so is convenient to install. Some different types are shown in Fig. 10-13. This solid coaxial cable is commonly available in impedances approximating 50 and 70 ohms.

Air-insulated coaxial lines have lower losses than the solid-dielectric type, but are less used in amateur work because they are expensive and difficult to install as compared with the flexible cable. The common type of air-insulated coaxial line uses a solid-wire conductor inside a copper tube, with the wire held in the center of the tube by means of insulating "beads" at regular intervals.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_o = 276 \log \frac{b}{a} \qquad \textbf{(10-D)}$$

where Z_o = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

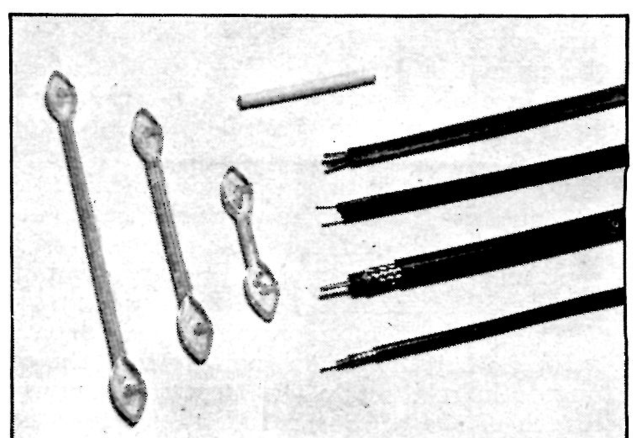


Fig. 10-13 — Typical manufactured transmission lines and spacers.

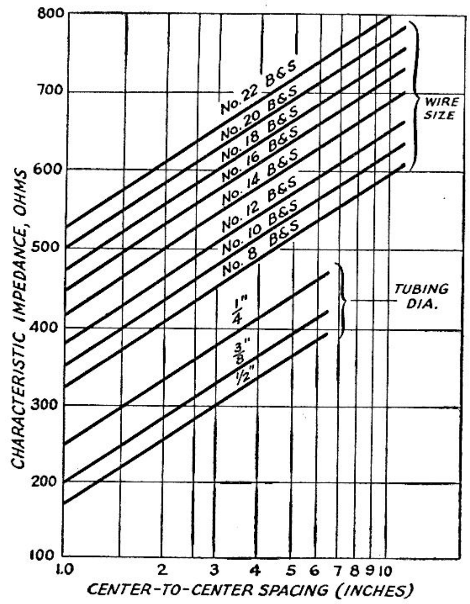


Fig. 10-14 — Chart showing the characteristic impedance of typical spaced-conductor parallel transmission lines. Tubing sizes given are for outside diameters.

It does not matter what units are used for *a* and *b*, so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 10-14 for a number of common conductor sizes.

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_o = 138 \log \frac{b}{a} \qquad \textbf{(10-E)}$$

where Z_o = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Again it does not matter what units are used for b and a, so long as they are the same. Curves for typical conductor sizes are given in Fig. 10-15.

The formula for coaxial lines is approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced. When the line is filled with a solid dielectric, the characteristic impedance as given by the chart should be multiplied by $1/\sqrt{K}$, where K is the dielectric constant of the material. In solid-dielectric parallel-conductor lines such as Twin-Lead the characteristic impedance cannot be calculated readily, because part of the electric field is in air as well as in the solid dielectric.

Electrical Length

In the discussion of line operation earlier in this chapter it was assumed that currents traveled along the conductors at the speed of light. Actually, the velocity is somewhat less, the reason being that electromagnetic fields travel more slowly in dielectric materials than they do in free space. In air the velocity is practically the same as in empty space, but a practical line always has to be supported in some fashion by solid insulating materials. The result is that the fields are slowed down; the currents travel a shorter distance in the time of one cycle than they do in space, and so the wavelength along the line is less than the wavelength would be in free space at the same frequency. (Wavelength is equal to velocity divided by frequency.)

Whenever reference is made to a line as being so many wavelengths (such as a "half-wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. Its actual physical length as measured by a tape always will be somewhat less. The physical length corresponding to an electrical wavelength is given by

$$\textit{Length in feet} = \frac{984}{f} \cdot V \qquad \textbf{(10-F)}$$

where f = Frequency in megacycles
V = Velocity factor

The velocity factor is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 10-I.

Example: A 75-foot length of 300-ohm Twin-Lead is used to carry power to an antenna at a frequency of 7150 kc. From Table 10-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is

$$\textit{Length}\text{ (feet)} = \frac{984}{f} \cdot V = \frac{984}{7.15} \times \mathbf{0.82}$$
$$= 137.6 \times 0.82 = 112.8 \text{ ft.}$$

The line length is therefore 75/112.8 = 0.665 wavelength.

Because a quarter-wavelength line is frequently used as a linear transformer, it is convenient to calculate the length of a quarter-wave line directly. The formula is

$$\textit{Length}\text{ (feet)} = \frac{246}{f} \cdot V \qquad \textbf{(10-G)}$$

where the symbols have the same meaning as above.

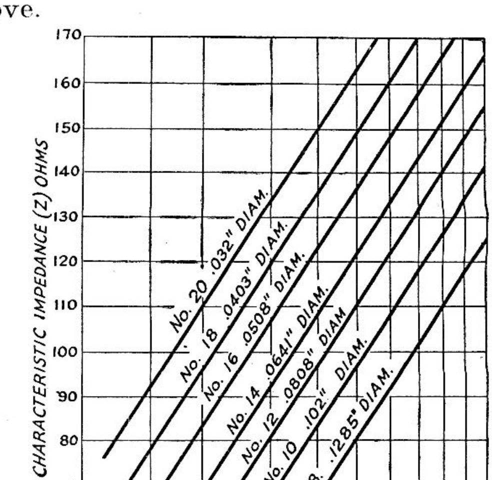


Fig. 10-15 — Chart showing characteristic impedance obtained with various air-insulated concentric lines.

Losses in Transmission Lines

There are three ways by which power may be lost in a transmission line: by radiation, by heating of the conductors (I^2R loss), and by heating of the dielectric, if any. Loss by radiation will occur if the line is unbalanced and, particularly with open-wire lines, may greatly exceed the heat losses. It can be reduced to a minimum by properly terminating the line in a balanced load and by symmetrical, uniform construction.

Heat losses in both the conductor and the dielectric increase with frequency. Conductor losses also are greater the lower the characteristic impedance of the line, because a higher current flows in a low-impedance line for a given power input. The converse is true of dielectric losses because these increase with the voltage, which is greater on high-impedance lines. The dielectric loss in air-insulated lines is negligible (the only loss is in the insulating spacers) and such lines operate at high efficiency when radiation losses are low. In solid-dielectric lines most of the loss is in the dielectric, the conductor losses being small.

It is convenient to express the loss in a transmission line in decibels per unit length, since the loss in db. is directly proportional to the line length. Losses in various types of lines operated without standing waves (that is, terminated in a resistive load equal to the characteristic impedance of the line) are given in Table 10-I. In these figures the radiation loss is assumed to be negligible.

When there are standing waves on the line the power loss increases as shown in Fig. 10-16. Whether or not the increase in loss is serious depends on what the original loss in watts would have been if the line were perfectly matched. If the line loss with perfect matching is very low, a large standing-wave ratio will not greatly affect the *efficiency* of the line — that is, 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 Table 10-I would be 1.5 × 0.41 = 0.615 db. Under these conditions the power delivered to the load would be 86.8% of the power input to the line. If the power input is 100 watts, the line loss is 100 − 86.8 = 13.2 watts. From Fig. 10-16, the loss is increased by a factor of 2.6 when the s.w.r. is 5 to 1, so the loss at this s.w.r. is 2.6 × 13.2 = 34.3 watts. Under these conditions the power delivered to the load is 100 − 34.3 = 65.7 watts. Therefore, 65.7/86.8 = 0.757, or approximately 76% as much power is delivered to the load with an s.w.r. of 5 as compared with perfect matching. The standing waves therefore cause the output power to be reduced by 1.2 db. (See discussion of the decibel in Chapter Twenty-Four.) With an open-wire line the loss caused by such an s.w.r. would be negligible, provided the line is well balanced to prevent radiation.

An appreciable s.w.r. on a solid-dielectric line may result in excessive loss of power at the higher frequencies. Such lines, whether of the parallel-conductor or coaxial type, should be operated as nearly flat as possible, particularly when the line length is more than 50 feet or so. As shown by Fig. 10-16, the increase in line loss is not too serious so long as the s.w.r. is below 2 to 1, but increases rapidly when the s.w.r. rises above 2.5 or 3 to 1. Tuned transmission lines such as are used with multiband antennas always should be air-insulated, in the interests of highest efficiency.

TABLE 10-I

Transmission-Line Velocity Factors and Attenuation

Type of Line	*Velocity Factor V*	** *Attenuation, db./100 ft.; Mc.* 3.5	7	14	28	50	144	*Capacitance per foot μμfd.*
Open-wire, 400 to 600 ohms	0.975*	0.03	0.05	0.07	0.1	0.13	0.25	
Parallel-tubing	0.95*	***						
Coaxial, air-insulated	0.85*	0.2	0.28	0.42	0.55	0.7	1.4	
RG-8/U (53 ohms)	0.66	0.28	0.42	0.64	1.0	1.4	2.6	29.5
RG-58/U (53 ohms)	0.66	0.53	0.8	1.2	1.9	2.7	5.1	28.5
RG-11/U (75 ohms)	0.66	0.27	0.41	0.61	0.92	1.3	2.4	20.5
Twin-Lead, 300 ohms	0.82	0.18	0.3	0.5	0.84	1.3	2.8	5.8
Twin-Lead, 150 ohms	0.77	0.2	0.35	0.6	1.0	1.6	3.5	10
Twin-Lead, 75 ohms	0.68	0.37	0.64	1.1	1.9	3.0	6.8	19
Transmitting Twin-Lead, 75 ohms	0.71	0.29	0.49	0.82	1.4	2.1	4.8	
Rubber-insulated twisted-pair or coaxial ****	0.56 to 0.65	0.96	1.6	2.5	4.2	6.2	13	

* Average figures for air-insulated lines taking into account effect of insulating spacers.
** For lines terminated in characteristic impedance.
*** Losses between open-wire line and air-insulated coaxial cable. Actual loss with both open-wire and parallel-tubing lines is higher than listed because of radiation, especially at higher frequencies.
**** Approximate figures for good-quality rubber insulation.

Unbalance in Parallel-Conductor Lines

When installing parallel-conductor lines care should be taken to avoid introducing electrical unbalance into the system. If for some reason the current in one conductor is higher than in the other, or if the currents in the two wires are not exactly out of phase with each other, the electromagnetic fields will not cancel completely and a considerable amount of power may be radiated by the line.

Maintaining good line balance requires, first of all, a balanced load at its end. For this reason the antenna should be fed, whenever possible, at a point where each conductor "sees" exactly the same thing. Usually this means that the antenna system should be fed at its electrical center. Even though the antenna appears to be symmetrical, physically, it can be unbalanced electrically if the part con-

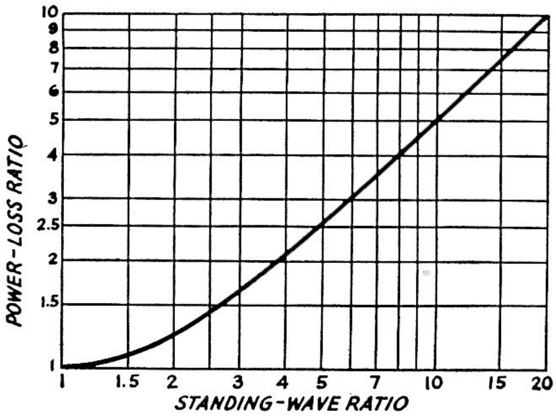


Fig. 10-16 — Effect of standing-wave ratio on line loss. The power-loss ratio given by the curve, multiplied by the power that would be lost in the same line if perfectly matched, gives the actual power lost in the line when standing waves are present.

nected to one of the line conductors is inadvertently coupled to something (such as house wiring or a metal pole or roof) that is not duplicated on the other part of the antenna. Every effort should be made to keep the antenna as far as possible from other wiring or sizeable metallic objects. The transmission line itself will cause some unbalance if it is not brought away from the antenna at right angles to it for a distance of at least a quarter wavelength.

In installing the line conductors take care to see that they are kept away from metal. The minimum separation between either conductor and all other wiring should be at least four or five times the conductor spacing. The shunt capacitance introduced by close proximity to metallic objects can drain off enough current (to ground) to unbalance the line currents, resulting in increased radiation. A shunt capacitance of this sort also constitutes a reactive load on the line, causing an impedance "bump" that will prevent making the line actually flat.

Coupling the Transmitter to the Line

In very general terms, the problem of coupling the transmission line and transmitter together is one of transforming the input impedance of the line into a value of impedance that will "load" the transmitter properly — that is, cause it to deliver the desired power output at as high efficiency as the transmitter design will permit. This is a question of impedance matching, and the impedance that must be matched is the value of resistance into which the tubes in the final stage of the transmitter should work. The value of this resistance is determined by the choice of tube operating conditions. The tubes are working into the proper resistance when the final tank circuit is tuned to resonance and the loading is such that the tubes are drawing rated plate current, as described in Chapter Six. The proper value of load resistance is thus reached automatically when the coupling is adjusted to bring the plate current up to the normal operating value. It is therefore not at all necessary to know what value of resistance is required. It is sufficient to note that, in general, it is in the neighborhood of a few thousand ohms, and is higher the higher the plate-voltage/plate-current ratio of the final stage.

The input impedance of the line can assume a wide range of values. As described earlier, it may be very much higher or very much lower than the impedance of the load at the end of the line, unless the line is matched to the load. Furthermore, it may or may not be a pure resistance, depending on the s.w.r., the line length, and the characteristics of the load.

Transforming Impedances

It was explained in Chapter Two that a resistive load tapped across part of a tuned circuit is equivalent to a higher value of resistance connected in parallel with the whole circuit. In other words, there is a transformer action in such an arrangement that enables us to change the value of a given resistance, such as *R* in Fig. 10-17A, into a new and higher value when the source of power looks into the terminals *AB*. Given reasonable values for *L* and *C*, the resistance looking into *AB* is determined practically wholly by the value of *R* and the position of the tap, so long as *LC* is tuned to resonance with the applied frequency. This is because the resonant impedance of *LC* alone (with *R* disconnected) is usually very high

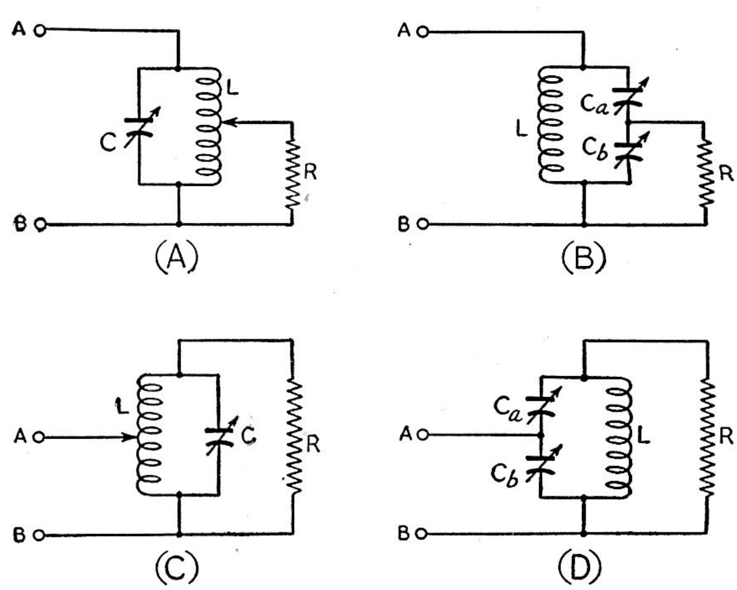


Fig. 10-17 — Using a resonant circuit for matching impedances.

compared with the resistance, R, of any practical load likely to be used, and also compared with any resistance that might be required between the terminals AB.

Fig. 10-17B shows a circuit that also provides a method for impedance transformation, using a capacitance voltage divider instead of tapping on the inductance. In this case, decreasing the capacitance of C_b (while increasing the capacitance of C_a correspondingly to maintain resonance) has the same effect as moving the tap toward the top of the coil in Fig. 10-17A. This type of circuit gives very smooth control. However, variable condensers of impracticable size would be necessary, to give as wide a range of impedance transformation as the circuit at A.

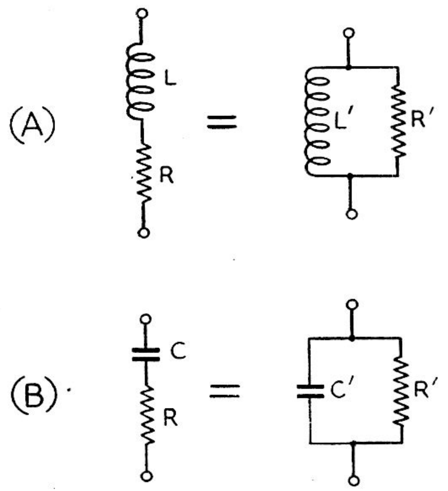


Fig. 10-18 — Series and parallel equivalents of a line whose input impedance has both reactive and resistive components.

When an r.f. amplifier is coupled to a transmission line the line impedance very seldom is larger than the load impedance required by the amplifier. However, should such a case arise the same circuits can be used by reversing the terminals. This is shown at C and D in Fig. 10-17. With R connected across the whole circuit, its resistance can be transformed to a lower value when the input terminals are tapped across part of the coil, as at A, or across C_b in Fig. 10-17B. The nearer the tap is to the bottom end of the coil, or the larger the capacitance of C_b compared with C_a, the smaller the resistance between terminals AB.

Complex Loads

In the foregoing it was assumed that the load, R, was a pure resistance. However, the input impedance of a line is more likely than not to have a reactive as well as a resistive component. This means, basically, that the current flowing into the line is not in phase with the voltage applied to the line. To represent such a condition by circuit symbols we can assume the input impedance of the line to consist either of a reactance (coil or condenser) in series with a resistance, or a reactance in parallel with a resistance. It does not matter which we choose, so long as the values assigned to the resistance and reactance are such that if the voltage were applied to the circuit instead of to the line, the current that flows would have exactly the same amplitude and phase angle as it actually does at the input terminals of the line.

These equivalent circuits are shown in Fig. 10-18. In practical work with lines it is not necessary to know the values of R, L or C. It is sufficient to know that they *symbolize* a condition that exists at the input end of the line — and then to know what to do about them. A few general points are worth noting: Given a fixed value of voltage, if the current at the input end of the line is high, then the impedance is relatively low; if the current is low, the impedance is relatively high. If the current is very nearly in phase with the voltage the reactance in the *series* equivalent circuit is small, but the reactance in the *parallel* equivalent circuit is large. On the other hand, if there is a considerable phase difference between current and voltage the reactance is large in the equivalent series circuit and is low in the equivalent parallel circuit. (In visualizing these reactances as coils and condensers it must be remembered that "large" and "small" are relative terms; for example, a "large" inductance at 28 Mc. would be a "small" inductance at 3.5 Mc. Also, the larger the capacitance of a condenser the smaller its reactance.)

Now suppose that a reactive line is to be connected to our impedance-transforming resonant circuit. Let us choose the parallel equivalent circuit, since it is somewhat easier to picture what happens. Fig. 10-19A shows a load with inductive reactance tapped across part of the resonant circuit (corresponding to Fig. 10-17A), and a load with capacitive reactance is shown in Fig. 10-19B. Imagine for the moment that the load has only reactance; the resistive component, R, is disconnected. Then, just as in the pure-resistance case previously discussed, a small reactance tapped across the coil L will appear as a larger reactance across the whole circuit, or between the input terminals AB. Thus, connecting a coil, L_1, across part of L is equivalent to connecting a larger coil across the whole circuit. Connecting a condenser, C_1, across part of L is equivalent to connecting a *smaller* condenser (larger reactance) across the whole circuit.

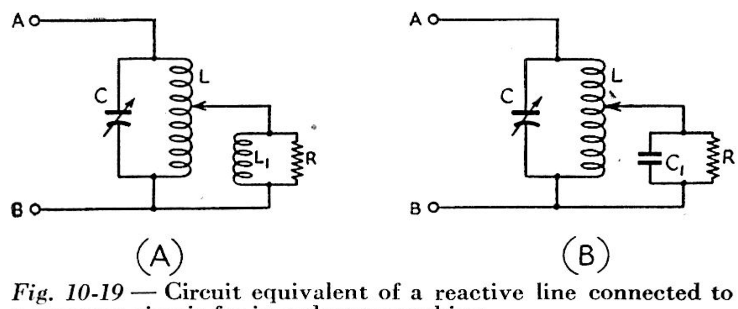


Fig. 10-19 — Circuit equivalent of a reactive line connected to a resonant circuit for impedance matching.

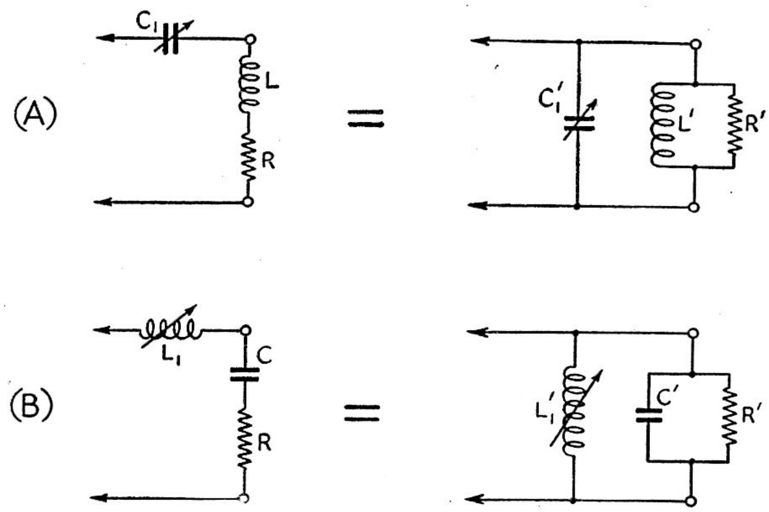


Fig. 10-20 — Methods for canceling the reactive component of the input impedance of a transmission line. In A the line input impedance is represented by L and R in series, or by L' and R' in parallel, and in B by C and R in series, or by C' and R' in parallel.

In either case this equivalent shunting reactance detunes the LC circuit from resonance, and C must be readjusted to bring it back. In the case of Fig. 10-19A, the capacitance of C must be increased because the "reflected" reactance in parallel with L decreases the total inductance (inductances in parallel) and so tunes the circuit to a higher frequency. The opposite is the case in Fig. 10-19B; the shunting reactance is capacitive and increases the total capacitance. Consequently the capacitance of C must be decreased to bring the circuit back to resonance.

The over-all effect, then, of coupling a reactive load to the circuit is to cause detuning as well as to cause the desired resistance loading. If the reflected reactance is large, corresponding to connecting a very large coil or a very small condenser across the whole LC circuit, it is readily possible to retune the circuit to resonance by adjusting C. The nearer the tap to the top end of L, the greater the change required in the tuning. But this simple method of compensating for the reactive component of the load is not always sufficient. In some cases the tap has to be moved so far up the coil, in order to obtain the right value of resistance loading, that the tuning condenser, C, no longer has sufficient range to compensate for the reflected reactance. When such a condition exists it is difficult, and sometimes impossible, to couple the desired amount of power to the transmission line.

Canceling Line Reactance

The remedy for this condition is to make the input end of the line look like a pure resistance *before* it is tapped on the impedance-transforming circuit. This can be done by "tuning out" the reactance of the line, by inserting a reactance of the same value but of the opposite kind. Again we have our choice between considering the line to be represented by reactance and resistance in series, or by reactance and resistance in parallel. The circuits are shown in Fig. 10-20. In A, a condenser, C_1, is used to cancel out the inductive reactance of the line, and in B an inductance, L_1, is used to cancel capacitive reactance. The same value of capacitance cannot be used for C_1 and C_1' under a given set of conditions because, as explained earlier, L and L' do not have the same values. For example, if L is small its parallel equivalent, L', is large, so a large capacitance would be required at C_1 and a small capacitance at C_1'. Because of limitations in practicable components (particularly in the capacitance range of variable condensers), there are conditions where the series circuit is the easiest to set up, from a practical standpoint. In others, the parallel circuit is easier to get working. For the large majority of cases either circuit will work equally well; from the standpoint of convenience, the parallel circuit is probably better.

To summarize, then, we have three general cases as shown in Fig. 10-21. If the line is purely resistive, or so nearly so that such reactance as is reflected across the LC circuit can be tuned out by readjusting C, the circuit at A may be used. Where the line shows more pronounced reactive effects, the line reactance can be tuned out, as indicated at B and C, so that the load tapped on L is purely resistive. It is easy to tell which should be used, inductance or capacitance, to compensate for the line reactance. If the line only (Fig. 10-21A)

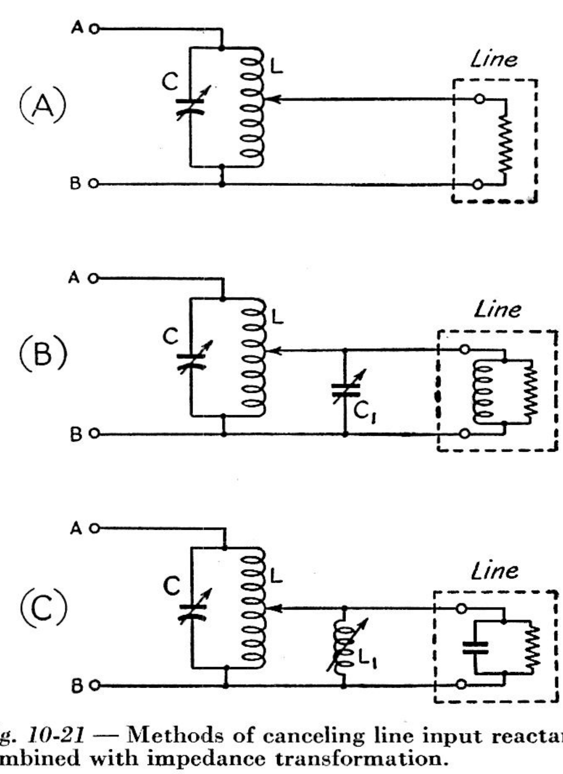


Fig. 10-21 — Methods of canceling line input reactance combined with impedance transformation.

is tapped across a very small portion of L, C will have to be readjusted slightly to bring the LC circuit back to resonance. If the capacitance of C has to be increased, a condenser, C_1, should be connected across the input terminals of the line. If the capacitance of C has to be decreased, an inductance, L_1, should be connected across the line. In either case the compensating reactance, C_1 or L_1, should be adjusted in value until the setting of C, for resonance with the applied frequency, is the same whether or not the line is tapped on L. When this condition is reached the loading may be adjusted by changing the tap position until the amplifier takes the desired plate current.

PRACTICAL COUPLING SYSTEMS

In practical work the two primary functions that a coupling system must perform — tuning out the line reactance, if any, and providing a method for control of loading on the transmitter — are not always enough. For one thing, it is desirable that the coupling system be such that the transmission line will operate only in the way it is intended that it should. For another, the coupling system should prevent transfer of any of the harmonic energy that always is present in the output of a transmitting amplifier. Both these points will be considered later in this section. For the moment, let us take a look at some of the simpler coupling systems.

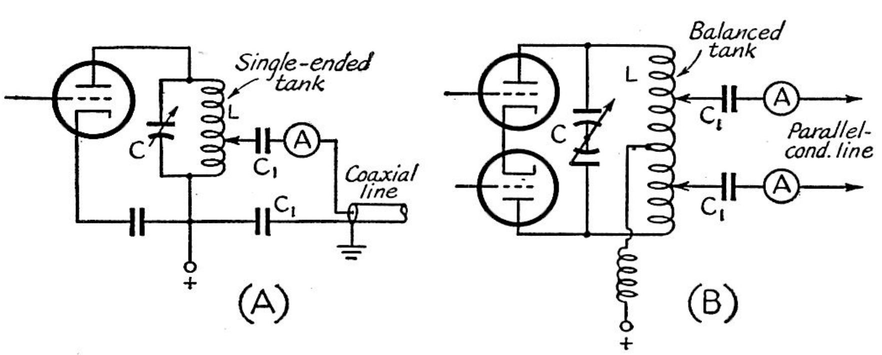


Fig. 10-22 — **Simple methods of coupling to a transmission line. The blocking condensers, C_1, should be 0.001-μfd. (or larger) mica condensers having a voltage rating in excess of the maximum d.c. voltage applied to the final amplifier (including the voltage applied on modulation up-peaks). The coaxial line can be coupled to a balanced tank circuit by connecting the grounded shield to the center of the coil (through a blocking condenser) and tapping the inner conductor on one side of the center. The parallel-conductor line requires a balanced tank circuit.**

The possibility of tapping the input end of the transmission line directly on the final-amplifier tank suggests itself from the discussion earlier. This method will work when the input impedance of the line is purely resistive, or nearly so. It can therefore be used with nonresonant or untuned lines, or with a resonant line when the line has the right length. As explained earlier, the input impedance of the line will be resistive when its length is a multiple of a quarter wavelength, provided the load at the output end of the line is a pure resistance. This will be so if the antenna itself is resonant, but will not be true if the antenna length is not correct for the operating frequency. The circuits are shown in Fig. 10-22. If the final amplifier is series-fed so that the tank circuit is "hot" with the plate voltage, it is necessary to connect a blocking condenser between the tank and the line. These circuits, although simple, are not recommended except perhaps in emergencies; there is little or no discrimination against harmonic frequencies.

Adjustment of this type of coupling is simple. First, resonate the amplifier tank circuit, with the line disconnected, by setting the tank condenser, C, to the minimum plate current point. Then tap the line across a turn or two of the tank coil, and readjust C for minimum plate current. The new minimum will be higher than with no load on the tank. Continue increasing the number of turns between the line taps, readjusting C each time, until the minimum plate current is the desired full-load value.

R.F. Ammeters

The r.f. ammeters shown in Fig. 10-22 and subsequent coupling circuits are useful accessories. The input impedance of the line is unaffected by any adjustments made in the coupling system (except for the effects of stray capacitance, as discussed later) so the greater the current flowing into the line the larger the amount of power delivered to the load. Measurement of r.f. current thus gives a check on the adjustment procedure and indicates when the largest power output is being obtained. Obviously, an adjustment that increases the input to the final stage of the transmitter without causing the line current to increase has simply increased the losses without increasing the output.

In the case of parallel-conductor lines two ammeters are shown, one in each conductor. This gives a check on line balance, since the two currents should be the same. It is not actually necessary to use two instruments; one ammeter can be switched from one side of the line to the other for comparative measurements. Also, it is to be understood that any current-indicating device (such as a flashlight lamp) that will work at r.f. may be used as a substitute for an actual ammeter.

The scale range required depends on the input impedance of the line and the power. The current to be expected can readily be calculated from Ohm's Law when the line is flat. In other cases the s.w.r. and the length of the line must be considered. The maximum current will occur when there is a current loop at the input end of the line, and if the load impedance and line impedance are known the input impedance at a current loop can be calculated from the formulas given earlier.

The ammeters are less useful when the input impedance of the line is high, because in that case the input current is quite small. It is to be noted that the value of current does not indicate, in any absolute sense, how well the system as a whole is working unless the actual value of the resistance component of the line input impedance is known. Current measurements taken on different lines, or on the same line if its length in wavelengths is changed, are not directly comparable.

Inductive Coupling

The circuits shown in Fig. 10-23, like those in Fig. 10-22, are useful only with lines having purely-resistive input impedance. The pick-up coil, which is inductively-coupled to the tank coil, is in fact simply a substitute for the tapped portion of the tank coil in Fig. 10-22. The number of turns required in the pick-up coil depends upon the resistance represented by the input end of the line. For flat lines, the number is governed by the characteristic impedance of the line. For 50- or 70-ohm lines it may range from one or two turns, at frequencies of the order of 14 to 28 Mc., to several turns at 3.5 Mc. For higher-impedance lines it may take half as many turns as there are in the tank coil, to get adequate coupling. In both cases the coupling between the coils will have to be very tight. The link windings provided on commercial coils are not usually adequate for this type of coupling except for low-impedance lines at the higher frequencies. When the number of turns on the pick-up coil is fixed, the loading on the final amplifier can be varied by varying the coupling between the two coils. Inductive coupling of this type is somewhat better than direct coupling from the standpoint of harmonic transfer.

Pick-up coil coupling introduces some reactance into the tank circuit, because of the leakage reactance of the coupling coil. This must be compensated for by retuning the final tank circuit when the desired degree of coupling is reached. If very much retuning is required, or if the amplifier loads with loose coupling between the two coils, it is an excellent indication that the line is not actually flat.

When a "swinging-link" assembly is used to obtain this type of coupling, the loading on the final amplifier can be adjusted to the desired value by varying the coupling between the two coils. The tank condenser, C, should be readjusted to minimum plate current each time the coupling is changed. If the desired loading cannot be obtained there is no alternative but to use a different coupling system.

The pick-up coil may be wound directly over the final tank coil, in which case the correct number of turns may be determined by trial. The insulation between the coils must be adequate for the plate voltage used, if the amplifier is series-fed.

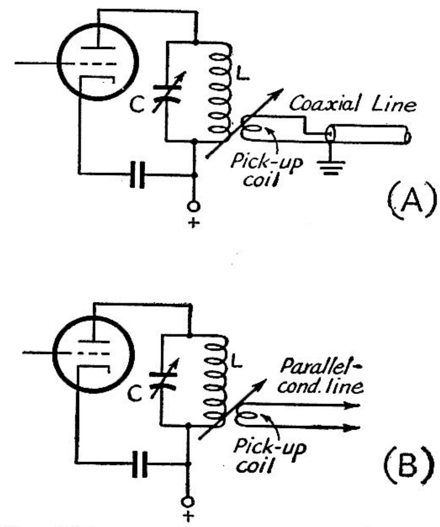


Fig. 10-23 — Using an untuned pick-up coil to couple to a transmission line. The method of adjustment is discussed in the text.

Series and Parallel Tuning

The circuits shown in Fig. 10-24 are useful with parallel-conductor lines operating at a relatively-high standing-wave ratio, particularly when the line length is such as to make the input impedance substantially a pure resistance. Assuming that the antenna is resonant, the optimum line lengths will be multiples of a quarter wavelength at the operating frequency. When the s.w.r. is high, the impedance at such points is considerably higher or considerably lower than the characteristic impedance of the line.

In these circuits the secondary, consisting of L_1, C_1 (and C_2, in the series circuit) and the input impedance of the line, is tuned to the operating frequency. As explained in Chapter Two, the degree of coupling between two resonant circuits is determined by their Qs, and it is necessary to keep the Qs fairly high (of the order of 10 or so). Assuming that the input impedance of the line is purely resistive, it can be inserted in series with the circuit (as in A) if its value is below about 100 ohms. The Q of the secondary circuit then can be brought to the proper value by making the reactance of L_1 of the order of 500 to 1000 ohms and setting the total capacitance of C_1 and C_2 to tune the circuit to resonance. With this type of tuning the current flowing into the line is rather large; in other words, the system is suitable for coupling into the line at a current loop.

On the other hand, if the line impedance is of the order of a few thousand ohms or more — which it will be at a voltage loop when the s.w.r. is high — the secondary circuit cannot be made to take power from the transmitter if the line resistance is inserted in series. The Q of the secondary circuit would be far too low to give adequate coupling. In such a case the parallel-tuned circuit at B may be used. As explained in Chapter Two, the Q of a circuit loaded by a parallel resistance will be equal to the resistance divided by the reactance of one of the tuned-circuit elements. Thus the reactance of L_1, in the parallel-tuned circuit, should not exceed a few hundred ohms at the operating frequency, to ensure a high-enough Q for good coupling.

If the input impedance of the line has reactance along with resistance, the reactance can be tuned out (within reasonable limits) by adjusting C_1 to compensate for it. This can easily be understood by reference to the section on canceling line reactance earlier in this chapter. The line will show reactance when its length is such that neither a voltage loop nor a current loop appears at the input end. There is a limit to how far the compensation can be carried, because at some line lengths the resistance component of the impedance has a value that is neither low enough to be inserted in series with the tuned secondary circuit, nor high enough to be placed in parallel with it. In such cases it is impossible to get adequate coupling between the final tank circuit and the line. In general, the value of the resistance component changes more slowly in the vicinity of a current loop than it does in the vicinity of a voltage loop. For this reason the series circuit is usable over a fair range of line lengths either longer or shorter than the length giving a current loop. There is less tolerance in the case of a voltage loop, so parallel tuning cannot be expected to give good results if the line length departs too far from the length giving a voltage loop at the input end.

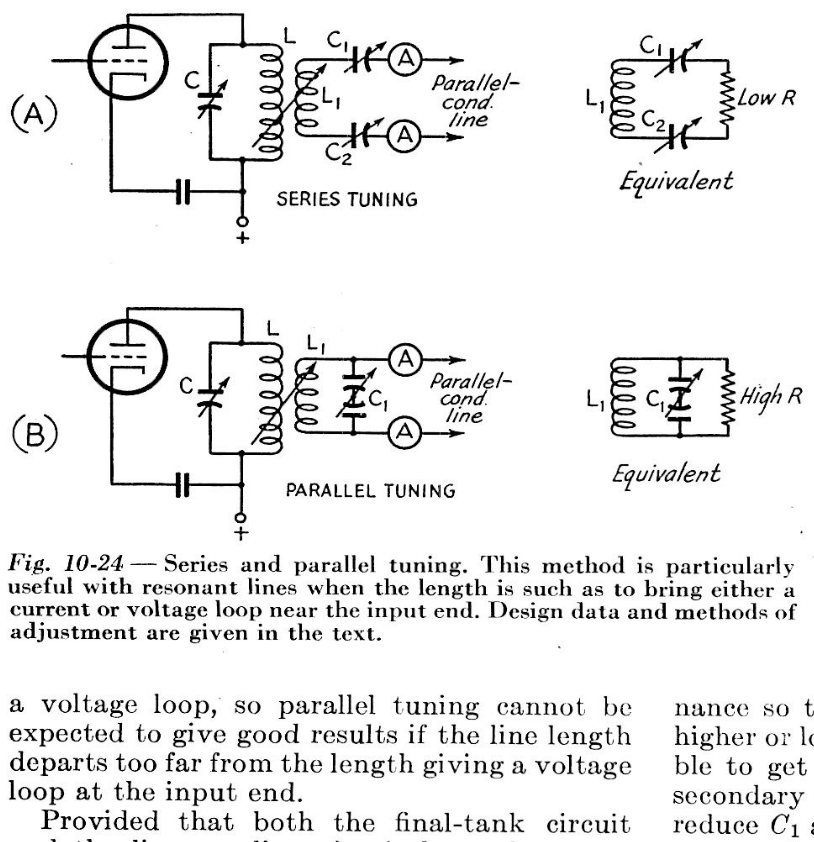


Fig. 10-24 — Series and parallel tuning. This method is particularly useful with resonant lines when the length is such as to bring either a current or voltage loop near the input end. Design data and methods of adjustment are given in the text.

Provided that both the final-tank circuit and the line-coupling circuit have Qs of the order of 10 or more, adequate power transfer can be obtained with fairly loose coupling between L and L_1. This is desirable, in that it aids in maintaining a balanced line and helps reduce the amount of harmonic energy transferred. The fact that the secondary circuit is tuned to the operating frequency is also greatly helpful in preventing harmonics from getting into the line.

Adjustment of Series Tuning

The tuning procedure with series tuning is as follows: With C_1 and C_2 at minimum capacitance, couple the antenna coil, L_1, loosely to the transmitter output tank coil, and observe the plate current. Then increase C_1 and C_2 simultaneously until a setting is reached that gives maximum plate current, indicating that the antenna system is in resonance with the transmitting frequency. Readjust the plate tank condenser to minimum plate current. This is necessary because tuning the antenna circuit will have some effect on the tuning of the plate tank. The new minimum plate current will be higher than with the antenna system detuned, but should still be well below the rated value for the tube or tubes. Increase the coupling between L_1 and L_2 by a small amount, readjust C_1 and C_2 for maximum plate current, and again set the plate tank condenser to minimum. Continue this process until the minimum plate current is equal to the rated plate current for the amplifier. Always use the degree of coupling between L_1 and L_2 that just brings the amplifier plate current to the rated value when C_1 and C_2 pass through resonance.

The values of inductance and capacitance to be used in the secondary circuit will depend largely on the frequency of operation. A coil of 3 or 4 turns (diameter 2 to 3 inches) will usually be adequate at 28 Mc., but 15 or 20 turns may be needed at 3.5 Mc. It is best to be able to adjust the number of turns on L_1 (by a tap), and there must of course be some means for varying the coupling between L and L_1. If the transmitter does not load properly, the secondary circuit may not be tuned to resonance; if it is found that maximum loading is secured with C_1 and C_2 at maximum capacitance, L_1 is too small and more turns must be used. If the maximum loading occurs with C_1 and C_2 at minimum, reduce the number of turns in L_1. In either case, make sure that C_1 and C_2 can be tuned *through* resonance so that the loading drops off at either higher or lower settings. Should it not be possible to get adequate loading even though the secondary circuit is resonated, increase L_1 and reduce C_1 and C_2 correspondingly, to raise the Q of the secondary circuit. If the practical limit of this process is reached and the transmitter output stage still does not load properly, the transmission-line length is not suitable for series tuning.

Two condensers are used in the series-tuned circuit in order to keep the line balanced to ground. This is because two identical condensers, both connected with either their stators or rotors to the line, will have the same capacitance to ground. A single condenser will slightly unbalance the circuit, since the frame has more capacitance to ground than the stator, but the unbalance is not serious unless the condenser is mounted near a large mass of metal, such as a chassis.

Adjustment of Parallel Tuning

Coupling and tuning adjustments with parallel tuning are carried out in much the same way as with series tuning. There is only one condenser to adjust, of course. Start with very loose coupling between L and L_1, resonate the secondary circuit by adjusting C_1 to make the final-amplifier plate current rise, then readjust C for minimum plate current. Increase the coupling in small steps, reresonating C_1 and C each time, until the desired loading is obtained.

Just as in the case of series tuning, it should be possible to tune *through* resonance with C_1. If the resonant point is at either maximum or minimum capacitance on C_1, change the number of turns on L_1 to bring the resonant point well on the condenser scale. In general, L_1 and C_1 will have about the same values as L and C, respectively, when the input impedance of the line is purely resistive. If the line shows reactance, the reactance can be tuned out, within limits, by adjustment of C_1 and, if necessary, by changing the number of turns on L_1 to achieve a combination that will permit the secondary circuit to resonate at the operating frequency.

If the input resistance of the line is very high, the secondary circuit will tune quite sharply. On the other hand, if the input resistance is relatively low the tuning will be broad and the resonance point will not be well marked. In such a case the number of turns in L_1 should be reduced and the capacitance of C_1 increased, to increase the Q of the circuit. This will permit power transfer with relatively loose coupling between L and L_1. Should it not be possible to load the transmitter properly with any combination of L_1 and C_1, the input resistance of the line is too low for parallel tuning.

In the parallel-tuned circuit C_1 is shown as a balanced or split-stator condenser. This type of condenser is used so that the system will be balanced to ground for stray capacitances. This is particularly desirable in the case of parallel tuning, because the voltage at the input end of the line is high, causing a relatively large current to flow through a small stray capacitance. An alternative scheme to maintain balance is to use two single-ended condensers in parallel, but with the frame of one connected to one side of the line and the frame of the other connected to the other side of the line. The same two condensers may be switched in series, as in Fig. 10-24A, when series tuning is to be used.

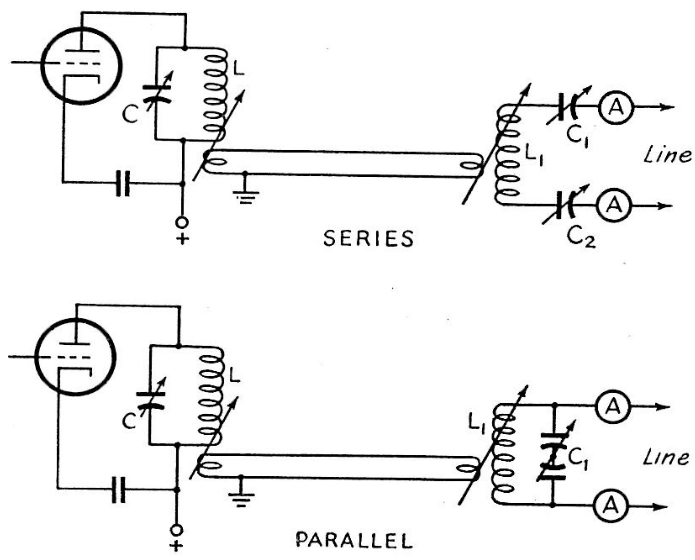


Fig. 10-25 — **Link-coupled series and parallel tuning.**

Link Coupling

The circuits shown in Fig. 10-24 require a means for varying the coupling between two sizable coils, a thing that is somewhat inconvenient constructionally. It is easier to use separate fixed mountings for the final tank and antenna coils and couple them by means of a link. As explained in Chapter Two, a short length of link line is equivalent to providing mutual inductance between two tuned circuits. Typical arrangements for series and parallel tuning are shown in Fig. 10-25. Although these drawings show variable coupling at both ends of the link circuit, a fixed link can be used at either end so long as a variable link is used at the other.

There is no essential difference between the tuning procedures with these circuits and those of Fig. 10-24. The only change is that the coupling is adjusted by means of a link instead of by varying the spacing between L and L_1.

"Universal" Antenna Couplers

An antenna-coupling system that is adaptable to a wide range of line input impedances can be constructed on the basis of the coupling principles described earlier. Combined with link coupling to the final tank circuit and provision for tuning out the input reactance of the line, such a system is suitable for working into either resonant or nonresonant lines, and introduces additional selectivity into the coupling system that helps discriminate against harmonics.

The circuit diagram is given in Fig. 10-26. The final tank is coupled to a second tuned circuit, L_1C_1, through a link. Taps are provided on L_1 so that the resistive component of the line impedance can be matched to the transmitter. To take care of cases where the input impedance of the line has a considerable reactive component, provision is made for switching in either a shunt capacitance or inductance, both of which are variable (see earlier discussion). The coupling should be variable at least at one end of the link circuit.

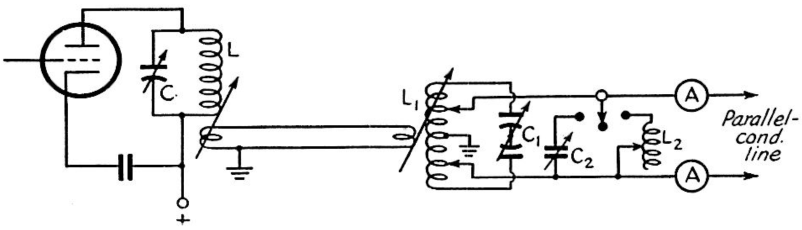


Fig. 10-26 — "Universal" antenna-coupling system. This circuit can be used with both resonant and nonresonant parallel-conductor lines.

In general, it is advisable to make the inductance of L_1 about the same as that of L, and to use for C_1 a condenser of the same capacitance as that used for C. The voltage rating for C_1 also should be the same as that of C. In other words, L_1C_1 may be a duplicate of LC for the operating frequency in use. The link coils can consist of two or three turns at each end. Provision should be made for tapping L_1 at frequent intervals — every turn, if possible. C_2 should have as large a maximum capacitance as is convenient — 250 to 500 $\mu\mu$fd. — but its voltage rating need not be high in the average case. For most installations where the power output does not exceed a few hundred watts a plate spacing of the order of 0.025 to 0.05 inch is sufficient. The inductance L_2 can consist of 20 or 25 turns approximately 2 inches in diameter and spaced 8 to 10 turns to the inch. The coil should be tapped every few turns.

The tuning procedure is as follows: First, disconnect the feeder taps on L_1 and use the loosest possible coupling, through the variable link coupling, to the final tank circuit. Tune C_1 until the plate current rises to a peak, indicating that L_1C_1 is resonated, and note the setting of C_1. Cut C_2 and L_2 out of the circuit and then connect the line taps across a turn or two at the center of L_1. Readjust C_1 to resonance, as indicated by a rise in plate current. It should be necessary to use closer coupling to get an observable change in plate current with the line connected. Note the new setting of C_1. If the capacitance is lower, switch in L_2 and find the tap that permits returning C_1 as nearly as possible to its original setting; if the capacitance is higher, switch in C_2 and adjust it to bring C_1 back to the original setting. Then increase the coupling, keeping C_1 at resonance as indicated by maximum plate current, and keeping C at resonance as indicated by minimum plate current. Continue until the minimum plate current reaches the desired load value. If C_1 flashes over as the coupling is increased, or if tuning C_1 back and forth a small amount either side of resonance makes it necessary to change the setting of C appreciably to maintain the final tank in resonance, the taps on L_1 are too close together. Move each tap one turn toward the ends of L_1, and again try increasing the coupling for rated load on the amplifier. When the proper loading is obtained, the tuning of L_1C_1 will be reasonably sharp, and changing the coupling will not necessitate more than "touching up" C to maintain resonance. If the taps on L_1 are too far apart the antenna tank circuit, L_1C_1, will be loaded heavily and its tuning will be broad. Under these conditions it may also be impossible to load the amplifier to rated plate current, even with the tightest available coupling. On the other hand, if the taps on L_1 are too close together the antenna tank will be too lightly loaded; its tuning will be critical and will affect the tuning of the plate tank circuit to a marked degree, and L_1 may overheat when the coupling is adjusted to make the amplifier take normal input.

When the reactive effects at the input end of the line are small, neither C_2 nor L_2 will be required. When this is the case, the setting of C_1 for resonance will not change much when the line is tapped on L_1. The greater the number of turns between the taps, the greater the detuning of the antenna tank by a given amount of reactance in the transmission-line input impedance.

This coupling system is equally effective with flat lines or those operating at a high s.w.r. If the line is actually flat, C_2 and L_2 will not be needed and the resonance setting of C_1 will not be affected by connecting the line. Regardless of the s.w.r., the positions of the line taps will depend on the resistive component of the line input impedance. If the resistance is low, the taps will be close together; if it is very high, the taps may have to be set right at the ends of L_1.

Coupling to Coaxial Lines

The principles of coupling to coaxial lines are just the same as for coupling to parallel-conductor lines. However, this type of line is unbalanced to ground, has inherently low impedance, and always should be operated with a low standing-wave ratio. The input impedance of a properly-operated coaxial transmission line therefore will be principally resistive, and of a value varying between perhaps 30 to 100 ohms, depending on the type of line and the s.w.r.

It is possible to couple such a line by means of a small coil inductively coupled to the final tank coil, as shown in Fig. 10-23A. The small amount of reactance introduced by the pick-up coil — and by the line, if the s.w.r. is slightly greater than 1 — can readily be tuned out by adjustment of the final tank condenser. However, additional selectivity is desirable for the purpose of reducing harmonic transfer from the final tank. Circuits are shown in Fig. 10-27. Except that it is adapted for single-ended rather than balanced operation, the circuit at A operates in much the same way as the circuit in Fig. 10-26. Also, because the load is known to be in the region of 100 ohms or less, it is possible to tap it across a capacitance voltage divider (see earlier discussion) for impedance matching. This avoids the necessity for tapping L_1.

The circuit of Fig. 10-27B is similar in operation to that at A, but dispenses with the link circuit. For convenience, it uses a link coil on the final tank for inductive transfer of energy, the rest of the inductance in the antenna tank circuit being made up by L_1.

In the circuit at A, L_1 may be the same as L; in B, L_1 plus the pick-up coil should have about the same inductance as L. Except at perhaps 28 Mc., it is satisfactory, practically, to make L_1 the same as L in this circuit also, since the pick-up coil will not ordinarily have much inductance itself. In both circuits C_2 should have about the same capacitance as C, and C_1 should have approximately the value suggested in Fig. 10-27.

To adjust the circuit, set C_1 at maximum, loosen the coupling between L and the link or pick-up coil, and tune C_2 to resonance. This will be indicated, as usual, by a rise in the amplifier plate current. Adjust C to minimum plate current and increase the coupling in small steps, reresonating C_2 and C each time, until the amplifier plate current is normal. The loading on the antenna tank circuit is least when C_1 is at maximum capacitance, and increases when the capacitance of C_1 is decreased (with C_2 increased correspondingly to maintain resonance). The symptoms of under- and over-loading of the antenna tank are the

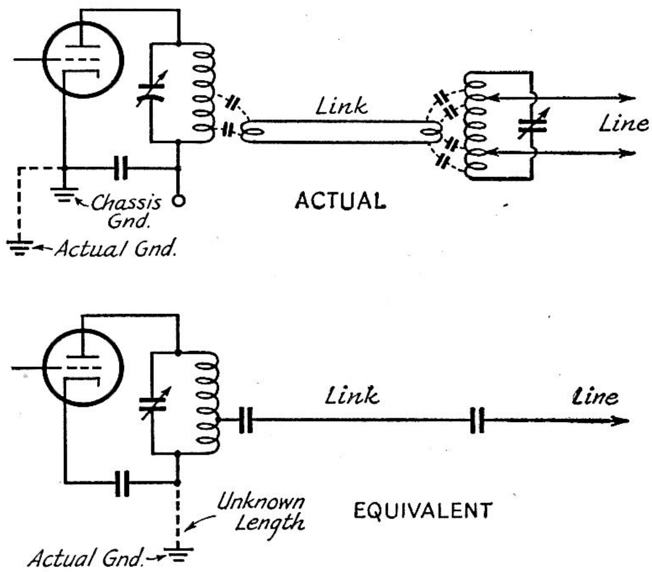


Fig. 10-28 — The stray capacitive coupling between coils in the upper circuit leads to the equivalent circuit shown below. The effect on the performance of the antenna system is discussed in the text.

same as described in connection with the universal antenna coupler. Adjust the loading by means of C_1, so that at normal plate input the antenna tank tuning is reasonably sharp and the setting of C is not greatly affected when C_2 is tuned a small amount either side of resonance.

Stray Coupling

In most of the circuits in Figs. 10-22 to 10-27, inclusive, a single-ended tank circuit has been indicated for the final amplifier. The amplifier itself has been shown only sketchily. The fact is that any type of antenna coupling circuit can be used with any type of amplifier — screen grid or neutralized triode, single-ended or push-pull. However, the actual arrangement, physically, of the circuit elements usually has an important bearing on the performance of the system. As it happens, a coupling system that is poorly designed, constructionally speaking, usually will do what it is supposed to do. But, equally important, it may do a lot of things it is *not* supposed to do.

Most of the unwanted effects that occur on transmission lines can be traced to stray capacitances in the system. Fig. 10-28 is an illustration. The upper drawing shows the ordinary link-coupled system as it might be used to couple into a parallel-conductor line. Inasmuch as a coil is a sizeable metallic object, it will have capacitance to any other metallic objects in its vicinity, including other coils. Consequently there is capacitance between the final tank coil and its associated link coil, and between the antenna tank coil and its link. These capacitances

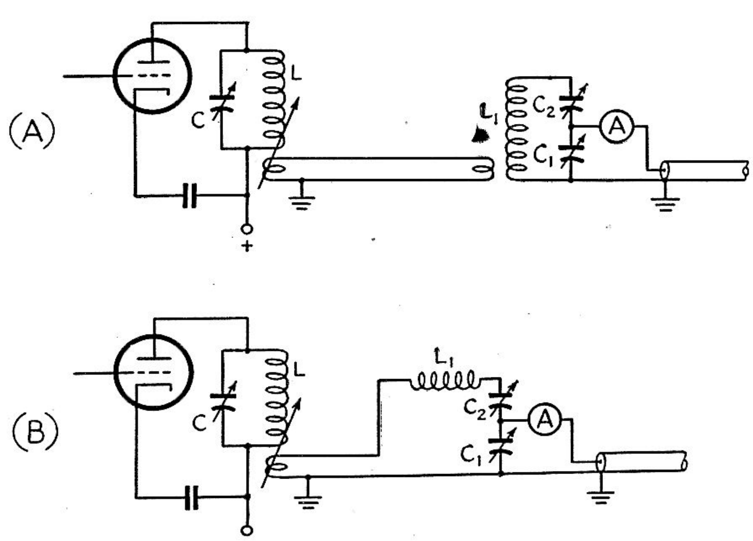


Fig. 10-27 — Coupling to coaxial lines. These circuits are used for harmonic suppression when working into a nonresonant coaxial line. Recommended capacitance values for C_1 are as follows: 28 Mc., 100 $\mu\mu$fd.; 14 Mc., 200 $\mu\mu$fd.; 7 Mc., 400 $\mu\mu$fd.; 3.5 Mc., 800 $\mu\mu$fd.

are small, but not negligible. In addition, the transmitter, particularly with metal-chassis construction, has appreciable capacitance to ground. Even if it did not, there is always a path from the transmitter to ground through the power wiring and the many stray capacitances associated with it.

There is a fundamental difference between the inductive coupling between coils and the capacitive coupling between them. Inductive coupling induces a voltage in the secondary coil that causes a current to flow, in common terminology, "around" the circuit. In Fig. 10-28, this means that the same current flows in both conductors of the link but, if the wires are parallel, the current flows in opposite directions in the two as it completes its travel around the loop. The same is true of the currents in the two conductors of the line. But with stray capacitive coupling the voltages at all points on the secondary coil are essentially in phase; for this type of coupling the secondary coil is just a mass of metal. Consequently, whatever current flows in the link (or in the line) flows in the *same* direction in both wires. Although both the link and line have two conductors and apparently form an ordinary go-and-return circuit, to the currents that flow as a result of capacitive coupling they simply look like a pair of conductors in parallel — in effect, that is, like a single conductor. The equivalent circuit is shown in the lower drawing in Fig. 10-28.

This single-wire circuit is an antenna system in itself, working in conjunction with a ground lead of unknown composition and length. It includes the regular antenna as well as the entire transmission line. If the various lengths happen to be just right, a fairly-large "parallel" current of this type can flow in it. This means that a considerable proportion of the total power output of the transmitter can be wasted in losses and radiation from a very undesirable sort of antenna system. Furthermore, despite the tuned tank circuits in the amplifier and antenna coupler, harmonic currents will flow in such an "antenna" even more readily than the fundamental current.

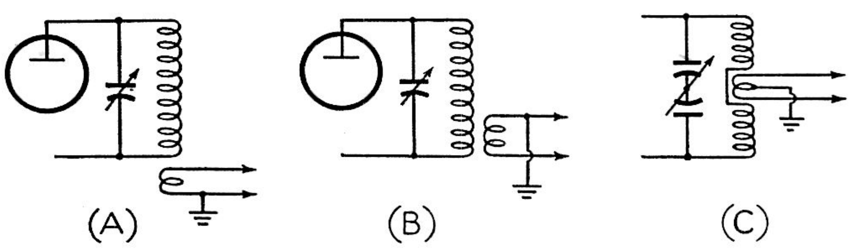


Fig. 10-29 — Methods of coupling and grounding link circuits to reduce energy transfer through stray capacitance.

There are other undesirable results, too. The fact that the power wiring becomes part of an "antenna" system means that the transmitter itself may perforce be at a considerable r.f. potential above ground. The chassis becomes "hot" with r.f., r.f. feed-back is prone to occur in speech equipment, and a considerable amount of r.f. power may be pumped into receiving and other equipment connected to the same a.c. power outlet. (A similar type of coupling in the input circuits of a receiver leads to stray pick-up of signals that may partially or completely mask the directive effects of the proper antenna.) On top of all this, it is impossible to tell much about the operation of the transmission line because the parallel current is more or less in phase with the regular line current in one wire and out of phase with it in the other. Thus the resultant currents in the two wires are unbalanced, and there is no way to separate the "parallel" and "line" currents in measurement.

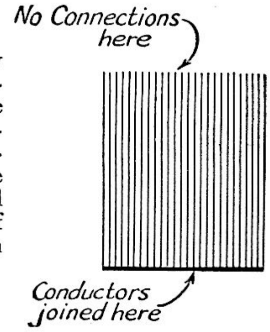


Fig. 10-30 — The Faraday screen. Stiff wire or small-diameter rod may be used for the conductors, separated by a distance about equal to the diameter of the wire or rod. The dimensions of the screen should be greater than the diameters of the coils to be shielded from each other.

These effects can only be eliminated if the stray capacitances are eliminated. However, they can be reduced by arranging the coils so the amount of energy coupled from the primary to the secondary is small, even though the capacitance itself still exists. This can be done by using a link coil that is physically small — that is, has few turns — and coupling it to the "cold" point on the tank coil. The cold point will be at the end of the coil that is grounded for r.f., either directly or through a by-pass condenser, in the case of single-ended tanks. In balanced tank circuits, the cold point is at the center. The coupling is further reduced if one side of the link circuit is grounded to the transmitter chassis as close as possible to the point where the tank itself is grounded. If the link is at the end of the tank coil the side farthest from the tank should be grounded, as indicated in Fig. 10-29A. If the link is wound *over* one end of the tank coil, ground the side toward the hot end of the tank, as indicated in Fig. 10-29B. With a balanced tank circuit the link should be at the center of the coil. In this case the best point to ground is the center of the link coil, but if this is impracticable good results will be secured by grounding either end of the coil. Ground directly to the chassis and keep the lead as short as possible.

This treatment of link circuits does not eliminate capacitive coupling. It simply makes it less troublesome, by making certain that the coupling occurs between parts of circuits that

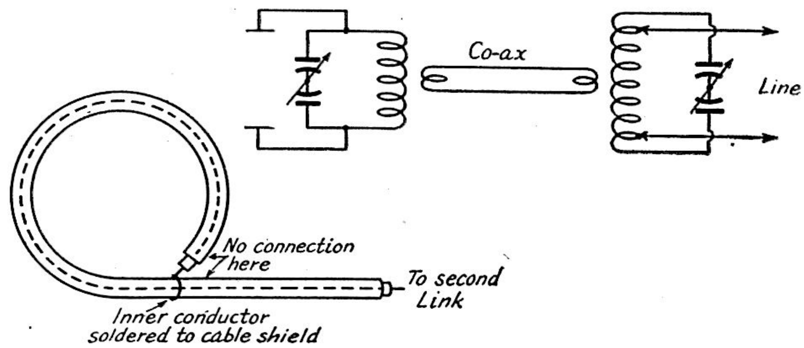


Fig. 10-32 — A shielded link coil constructed from coaxial cable. The smaller sizes of cable such as RG-59/U are most convenient, except when the coils have a diameter of 3 inches or more. For larger coils, RG-8/U or RG-11/U can be used.

are not at high r.f. voltage. However, there are cases, particularly with balanced tank circuits, where the point on the tank coil that is cold for the fundamental frequency is hot at the even harmonics. This means that even though the transmitter and line behave properly on the fundamental frequency, harmonics still can be radiated at considerable intensity. The only way to be sure that these effects do not exist is to eliminate the stray capacitance entirely.

Capacitive coupling between coils can be eliminated by means of a **Faraday screen.** This is a shield that prevents the electric field from one coil from reaching the other, but which has no effect on the magnetic field. As shown in Fig. 10-30, it consists of a group of parallel conductors, insulated from each other, and connected together at one end only. This forms an effective shield for the electric field, but since the conductors are open-circuited the voltages induced in them by the magnetic field cannot cause any current to flow. (Such current flow is essential to magnetic shielding with nonmagnetic materials, as explained in Chapter Two.)

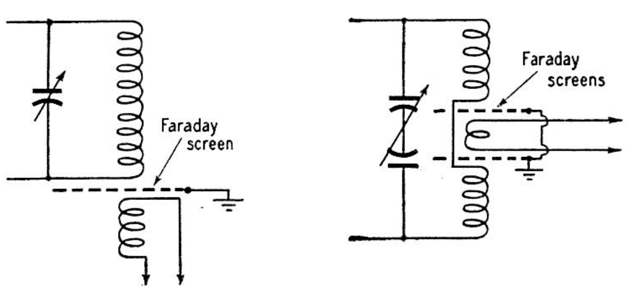


Fig. 10-31 — Installation of Faraday screens to eliminate capacitive coupling between coils.

The Faraday screen should be somewhat larger than the diameter of the coils with which it is used. It is simply mounted between the two coils that are to be shielded from each other, and then grounded to the chassis through a short lead, as indicated in Fig. 10-31. In the case of a balanced tank circuit with a swinging link, two shields must be used, one on each side of the link coil. In the case of fixed links wound over the tank coil, a satisfactory screen can be made by using several turns of the same type of coil, cutting them parallel to the axis to open-circuit the conductors, and then soldering them together at one end only. This shield can then be inserted between the tank coil and link, making sure that it is adequately insulated from both.

An alternative, and perhaps simpler, type of screening is shown in Fig. 10-32. In this case the inner conductor of a piece of coaxial cable is used to form a one-turn link. The outer conductor serves as an open-circuited shield around the turn, this shield being grounded to the chassis. The circuit to the link line is made by connecting the inner conductor to the outer conductor at the finish of the turn, as shown, and from there on the coaxial line is used to transfer the power to a second, and similar, link coil at the antenna tuner. This type of shielded link is simpler to make than the regular Faraday screen.

Aside from the adverse effects on the performance of the antenna system, stray capacitive coupling frequently is responsible for interference to near-by broadcast receivers. It is not difficult to appreciate that radiation taking place from transmission lines and power wiring is, in general, more likely to get into a broadcast receiver than radiation from an antenna that is intentionally kept away from other antennas — particularly when the receivers are connected to that same power wiring.

Antenna-Coupler Construction

The apparatus used to cancel line reactance and match the line resistance to the transmitter is commonly called an "antenna coupler" or "antenna tuner." (The name is really a misnomer, because the coupling and tuning equipment at the input end of the line does not have any effect on the antenna itself; if there is any antenna tuning to be done it must be done at the antenna, independently of the line.) The design principles and the important constructional points have been covered earlier in this chapter; in this section we show a few examples of typical construction.

Bearing in mind the precautions mentioned earlier as to maintaining balance in parallel-conductor transmission lines, it is usually good practice to install the coupling equipment close to the point where the line enters the station. This is a simple matter when the tuning equipment is link-coupled to the trans-

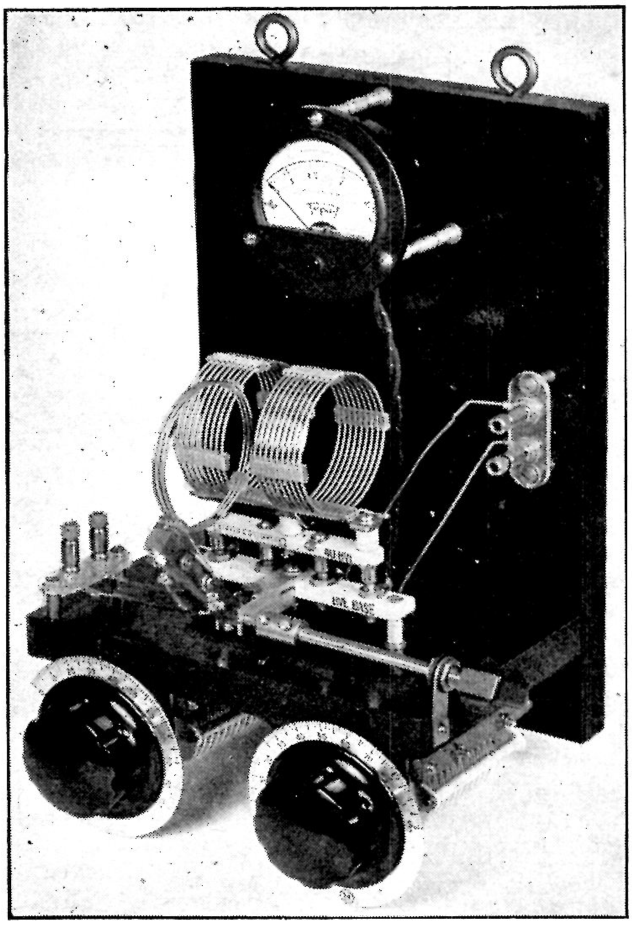


Fig. 10-33 — A wall-mounting antenna coupler for medium-power transmitters. This unit provides a choice of either series or parallel tuning for resonant feeders. Standard transmitting coils of the variable-link type are used.

mitter, since there are no particular restrictions on the length of the link that can be used. However, if the link line is fairly long it should be treated as a transmission line rather than merely as a means of providing mutual inductance between two separated coils. In such a case it is advisable to have variable coupling at both ends of the link. This permits matching the link line to the line tank circuit, and once the match is obtained the power output of the transmitter can be varied by changing the coupling at the transmitter tank. If the link line is not properly matched its current may be excessive, leading to unnecessary power loss.

The most desirable form of link line is coaxial cable. Properly handled, its losses are low; and since it is shielded it can be on or near metal objects with impunity.

SERIES-PARALLEL COUPLER FOR WALL MOUNTING

Fig. 10-33 shows a link-coupled coupler designed for series or parallel tuning of a resonant line. It is suitable for transmitters having a power output in the neighborhood of 250 watts. A higher-power version easily could be made using a similar layout, but substituting heavier coils and condensers with greater plate spacing.

As shown in Fig. 10-34, the change from series to parallel tuning is made by means of jumpers and extra pins on the coil plug-bar. A separate coil is used for each band, and after determining which should be used, series or parallel tuning, on a particular band, the jumpers may be installed permanently or left off as required. The tuning condensers specified, together with a set of standard plug-in transmitting coils, should provide adequate coupling if the transmission-line length is such as to bring a voltage or current loop near the input end.

The unit is mounted on an 8 × 12 × ⅞-inch board for hanging on the wall in any convenient location near the entrance point of the feeders. The 2.5-ampere r.f. ammeter is mounted centrally by long wood screws through spacers at the top of the unit. A short length of twisted pair connects it to the thermocouple, secured in a horizontal position at the bottom of the backboard. The tuning condensers are mounted on the underside of a 4-inch shelf extending the width of the unit. Atop the shelf, the jack-bar for the coil is supported on pillars by wood screws. An extension shaft to vary the degree of coupling is supported by a bushing fastened to a short strip of brass at the right of the shelf. A short length of 300-ohm ribbon (coaxial cable can be used instead) connects the input terminals to the movable link, while the output terminals are located at the middle right of the backboard. Two screw eyes at the top permit the unit to be hung from screws or nails in the wall.

RACK-MOUNTING SERIES-PARALLEL COUPLER

The rack-mounting coupling unit shown in Fig. 10-35 is suitable for power outputs of 25 to 50 watts, and provides either series or parallel tuning for resonant lines. Separate condensers are used for this purpose, and while

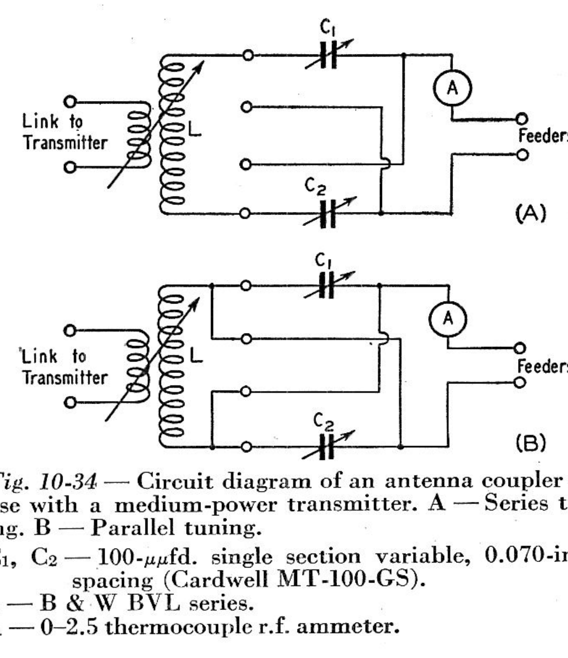


Fig. 10-34 — Circuit diagram of an antenna coupler for use with a medium-power transmitter. A — Series tuning. B — Parallel tuning.

C_1, C_2 — 100-μμfd. single section variable, 0.070-inch spacing (Cardwell MT-100-GS).

L — B & W BVL series.

A — 0–2.5 thermocouple r.f. ammeter.

three are required, this system has the advantage that no switching is necessary when changing from series to parallel tuning. It is also possible to cover a somewhat wider range of line input impedances with parallel tuning because the series condensers can be used to help cancel out inductive reactance that cannot be handled by the parallel circuit alone.

The coupler is mounted on a 5¼ × 19-inch panel. The parallel condenser, C_1, is in the center, with C_2 and C_3 on either side. The variable condensers are mounted on National GS-1 stand-off insulators which are fastened to the condenser tie-rods by means of machine screws with the heads cut off. Small ceramic shaft couplings are used to insulate the control knobs from the condenser shafts.

Clips with flexible leads attached are provided for the parallel condenser, C_1, so that the sections may be used either in series or parallel to form either a high-C or low-C tank circuit, as required. When the high-C tank is necessary the two stators are connected together by means of the clips, as indicated by the dotted lines in the circuit diagram, Fig. 10-36. When the two sections are connected in series for low-C operation the breakdown voltage is increased.

Two sets of variable condensers are suggested in the list of parts. The smaller receiving-type condensers with 0.03-inch air gap are satisfactory for transmitter power outputs up to 50 watts. The larger condensers, with 0.045-inch spacing, are required for transmitter outputs of the order of 100 watts.

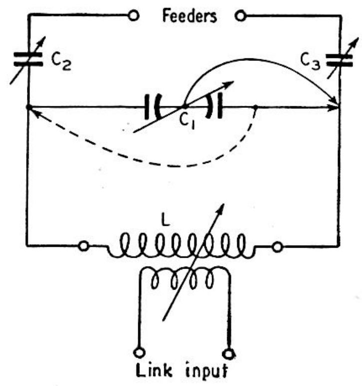


Fig. 10-36 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers that are operated at less than 1000 volts on the plate.

All coils are 1⅞ inches in diameter and 2¼ inches long, with the variable link located at the center. For series tuning, use the coil specified for the next-higher frequency band, which will be approximately correct.

C_1 — 100 μμfd. per section, 0.045-inch spacing (National TMK-100-D) for high voltages; receiving type for low voltages (Hammarlund MCD-100).

C_2, C_3 — 250 μμfd., 0.026-inch spacing (National TMS-250) for high voltages; receiving type for low voltages (Hammarlund MC-250).

L — B & W JVL-series coils. Approximate dimensions for parallel tuning for each band are as follows:
3.5-Mc. band — 40 turns No. 20.
7-Mc. band — 24 turns No. 16.
14-Mc. band — 14 turns No. 16.
28-Mc. band — 8 turns No. 16.

BANDSWITCHING UNIVERSAL COUPLER

The coupling unit shown in Figs. 10-37 and 10-39 is of the "universal" type discussed earlier. It is a bandswitching unit using commercially-available coils. Provision is made for switching either capacitance or inductance across the transmission line to compensate for its input reactance. Impedance matching is achieved by tapping the tank coils at the proper points.

In the circuit diagram, Fig. 10-38, only one set of coils is shown. For other bands the connections shown for L_1 and L_2 would be duplicated. Bandswitching is accomplished by a five-gang switch, S_1. Compensating reactances can be switched in or out of the circuit by S_2. The coupling links, L_2, are the shielded type using coaxial cable described earlier in this chapter (Fig. 10-32).

The coupler is wholly supported by a 7 × 19-inch relay-rack panel. The variable condensers are mounted from the panel by small stand-off insulators, and insulated couplings are used between the condenser shafts and the National Type AM dials. The tank condenser, C_1, is mounted at the right-hand end of the panel with the bandswitch, S_1, to its left. The four coils are grouped around the bandswitch, with the 28-Mc. coil placed so that the leads to it are the shortest. The coils are Millen 44000 series with the plug bases removed from the

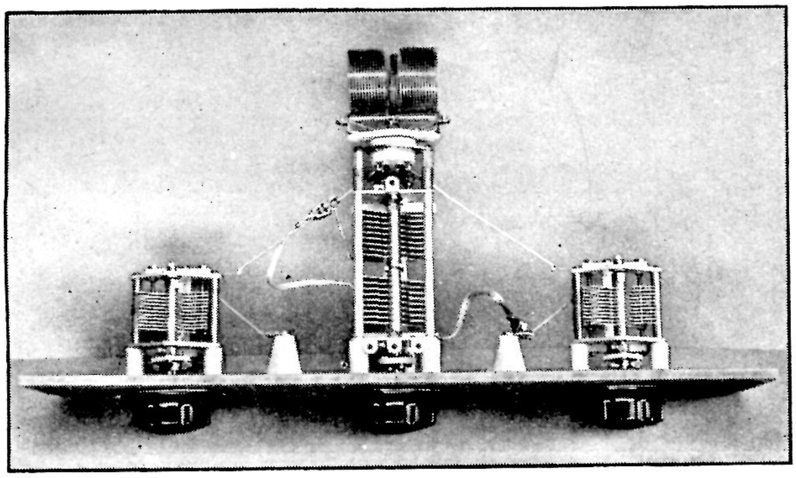


Fig. 10-35 — Rack-mounted coupler for low-power transmitters. This unit uses three variable condensers to provide either series or parallel tuning without condenser switching.

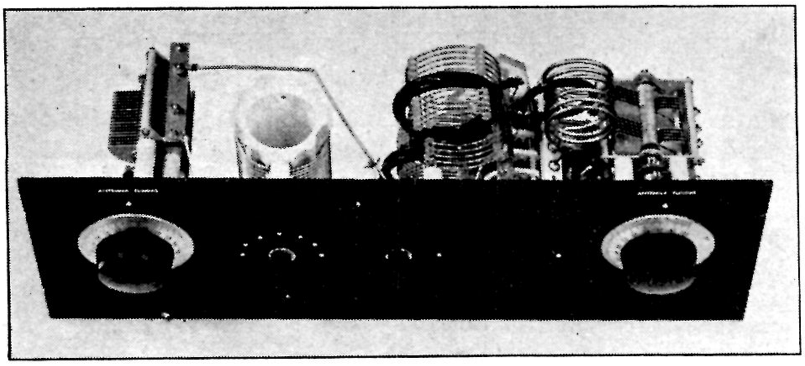


Fig. 10-37 — Bandswitching universal-type coupler for parallel-conductor lines. This unit can be used with transmitters having power outputs of the order of 100 watts.

3.5-, 7- and 14-Mc. coils. It is not practicable to remove the base from the 28-Mc. coil because it does not have the polystyrene supporting strip that is part of the lower-frequency coil

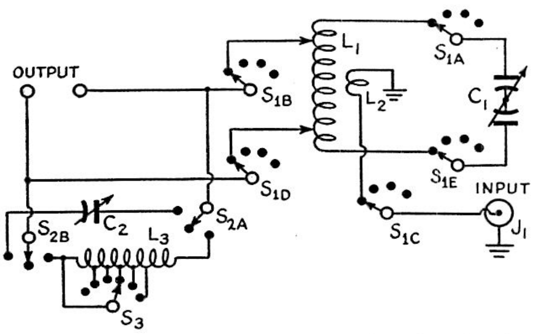


Fig. 10-38 — Circuit diagram of the bandswitching coupler. In this diagram the ground symbol indicates points that are connected together. Wiring to coils is shown for one band only, to avoid complicating the diagram; the wiring for other coils is identical.

C_1 — 100-$\mu\mu$fd.-per-section variable (Cardwell MR-100-BD).
C_2 — 335-$\mu\mu$fd. variable (Cardwell MR-335-BS).
L_1 — Millen 44000-series coils (see text).
L_2 — Shielded link; one turn for 28 and 14 Mc.; 2 turns for 7 and 3.5 Mc.
L_3 — 26 turns No. 12 on 2½-inch diameter form (National XR-10A), 7 turns per inch. Tapped 8, 14, 18, 22 and 24 turns from end to which arm of S_3 is connected.
J_1 — Coaxial-cable connector (Amphenol).
S_1 — 5-section 4-position ceramic wafer switch (Centralab 2546).
S_2 — 2-section 4-position ceramic wafer switch (Centralab 2543).
S_3 — 1-section 6-position ceramic wafer switch (Centralab 2501).

assemblies. The coils are partly supported by the wiring to the switch and partly by the polystyrene plate mounted on the back of the switch. The ends of the coil mounting strips are cemented into holes cut in the plate.

The compensating condenser, C_2, is mounted at the left-hand end of the panel. L_3 is mounted vertically to its right, with S_3 directly in front of it on the panel. S_2 is mounted centrally on the panel. The output terminals to the line are mounted above S_3. The link input terminal is a coaxial cable socket mounted on a small bracket in the lower right-hand corner.

The link coils, L_2, are supported by the wiring, and the coupling is changed by bending the link into or out of its associated tank coil. Since the links fit rather tightly in the tank coils, the pressure helps hold them in place once the proper coupling is determined. The link shields are all connected together and to the input connector; the inner conductors go to the switch contacts. The link coils are made from RG-59/U cable.

With the coils and condensers specified, this coupler can handle power outputs of the order of 100 to 150 watts. The method of adjustment is covered earlier in this chapter.

A WIDE-RANGE ANTENNA COUPLER

The photograph of Fig. 10-40 shows the constructional details of a wide-range antenna coupler suitable for use with high-power transmitters. Various combinations of parallel and series tuning, with high- and low-*C* tanks and high- and low-impedance outputs, are available. Diagrams of the various circuit combinations possible with this arrangement are given in Fig. 10-41.

A separate coil is used for each band, and the desired connections for series or parallel tuning with high or low *C*, or for low-impedance output

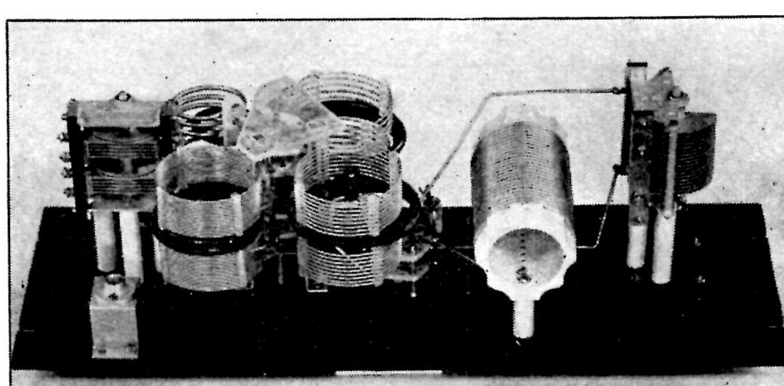


Fig. 10-39 — Rear view of the bandswitching coupler. Details of coil mountings are shown in this view.

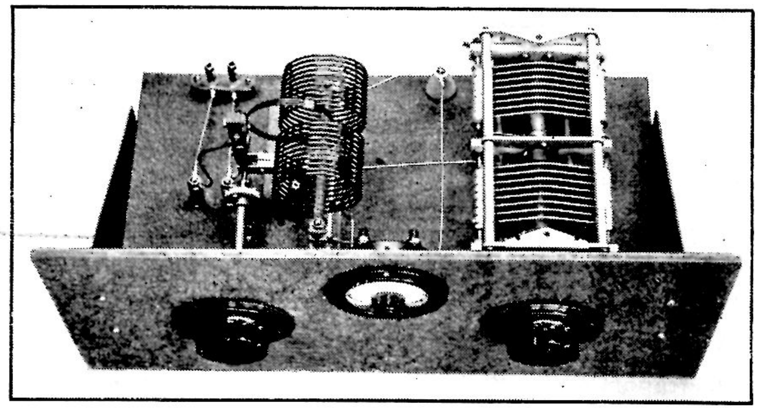


Fig. 10-40 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel 8¾ × 19 inches in size. The variable condenser is a split-stator unit with a capacitance of 200 μμfd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B & W TVL series. The r.f. ammeter has a 4-ampere scale.

with high or low *C*, are automatically made when the coil is plugged in. Coil connections to the pins for various circuit arrangements are shown in Fig. 10-41.

The tuning condenser specified, together with a set of standard plug-in transmitting coils, should cover nearly all coupling conditions likely to be encountered.

Because the switching connections require the use of a central pin, a slight alteration in the B & W coil-mounting unit is required. The central link-mounting unit should be removed from the jack-bar and an extra jack placed in the central hole thus made available. The link assembly should then be mounted on a 2-inch cone insulator to one side of the jack bar.

Correspondingly, the central nut on each coil plug base must be removed and a Johnson tapped plug, similar to those furnished with the coils, substituted. An extension shaft may then be fitted on the link shaft and a control brought out to a knob on the panel.

The split-stator tank condenser is mounted by means of angle brackets on four 1-inch cone-type ceramic insulators, and an insulated flexible coupling is provided for the shaft.

If desired, the coils may be wound with fixed links on ceramic transmitting coil forms. The links should be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for seven connections.

The unit as described should be satisfactory for transmitters having an output of 500 watts with plate modulation and somewhat more on c.w. For higher-power 'phone, a tank condenser with larger plate spacing should be used.

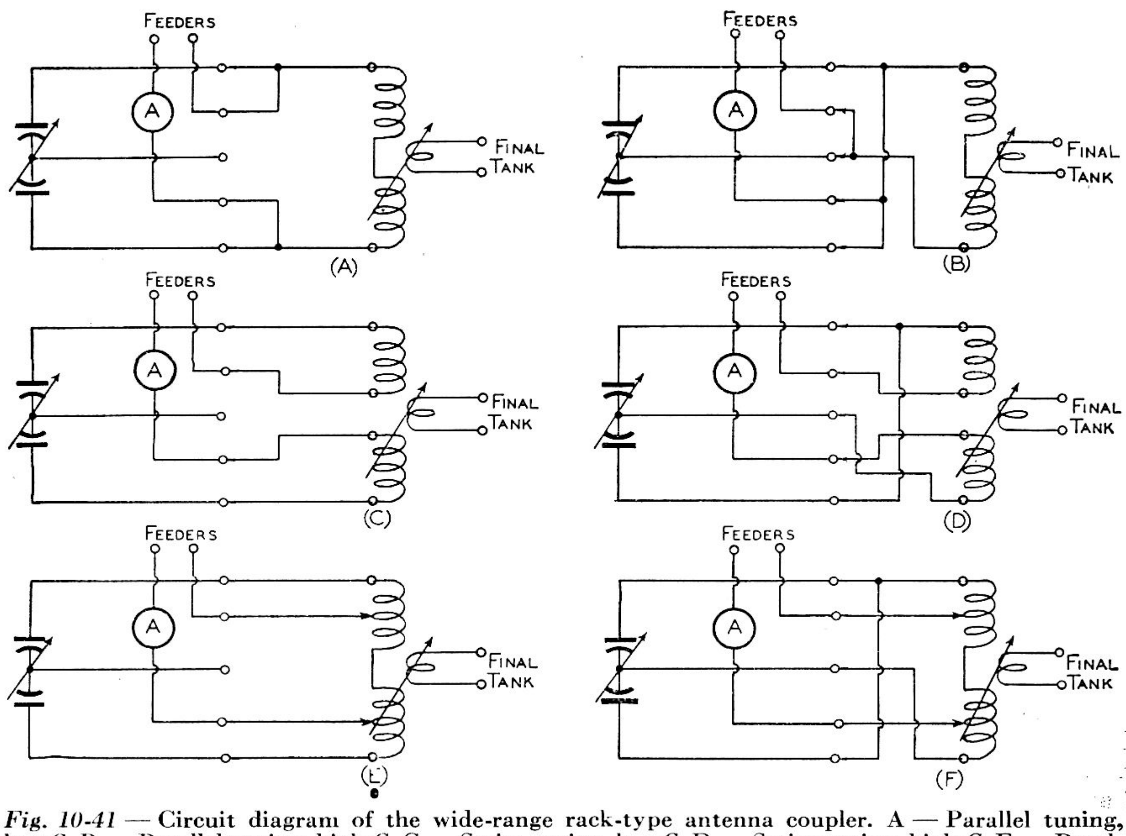


Fig. 10-41 — Circuit diagram of the wide-range rack-type antenna coupler. A — Parallel tuning, low *C*. B — Parallel tuning, high *C*. C — Series tuning, low *C*. D — Series tuning, high *C*. E — Parallel tank, low-impedance output, low *C*. F — Parallel tank, low-impedance output, high *C*. After the inductance required for each of the various bands has been determined experimentally, the connections to the coils can be made permanent. Then it will be necessary only to plug in the right coil for each band, tune the condenser for resonance, and adjust the link loading.

Antennas

In selecting the type of antenna to use, the propagation characteristics of the frequency band or bands to be used should be given due consideration. These are outlined in Chapter Four. In general, antenna construction and location become more critical and important on the higher frequencies. On the lower frequencies (3.5 and 7 Mc.) the angle of radiation and plane of polarization may be of relatively little importance; at 28 Mc. and higher they may be all-important. On a given frequency, the particular type of antenna best suited for long-distance transmission may not be as good for shorter-range work as would a different type. The important properties of an antenna or antenna system are its polarization, angle of radiation, impedance, directivity and gain.

Polarization

The polarization of a straight-wire antenna is its position with respect to the earth. That is, a vertical wire transmits vertically-polarized waves and a horizontal antenna generates horizontally-polarized waves in its direction of maximum radiation (broadside). The wave from an antenna in a slanting position contains both horizontal and vertical components.

Angle of Radiation

The wave angle (or vertical angle) at which an antenna radiates best is determined by its polarization, height above ground, and the nature of the ground. Radiation is not all at one well-defined angle, but rather is generally dispersed over a more or less large angular region, depending upon the type of antenna. The angle is measured in a vertical plane with respect to a tangent to the earth at that point.

Impedance

The impedance of the antenna at any point is the ratio of the voltage to the current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load represented by the antenna. It is a pure resistance only at current loops (maxima) and nodes (minima) on resonant antennas. The antenna impedance is high at the current node and low at the current loop.

Directivity

All antennas radiate more power in certain directions than in others. This characteristic, called *directivity,* must be considered in three dimensions, since directivity exists in the vertical plane as well as in the horizontal plane. Thus the directivity of the antenna will affect the wave angle as well as the actual compass directions in which maximum transmission takes place.

Current

The **field strength** produced by an antenna is proportional to the current flowing in it. When there are standing waves on an antenna, the parts of the wire carrying the higher current have the greater radiating effect. All resonant antennas have standing waves — only terminated types, like the terminated rhombic and terminated "V," have substantially uniform current along their lengths.

Power Gain

The ratio of power required to produce a given field strength, with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the **power gain** of the latter antenna. The field is measured in the optimum direction of the antenna under test. In amateur work, the comparison antenna is generally a half-wave antenna at the same height and having the same polarization as the antenna under consideration. Power gain usually is expressed in decibels.

Front-to-Back Ratio

In unidirectional beams (antenna systems with maximum radiation in only one direction) the front-to-back ratio is the ratio of power radiated in the maximum direction to power radiated in the opposite direction. It is also a measure of the reduction in received signal when the beam direction is changed from that for maximum response to the opposite direction. Front-to-back ratio is usually expressed in decibels.

Ground Effects

The radiation pattern of any antenna that is many wavelengths distant from the ground and all other objects is called the **free-space pattern** of that antenna. The free-space pattern of an antenna is almost impossible to obtain in practice, except in the v.h.f. and u.h.f. ranges. Below 30 Mc., the location of the antenna with respect to ground plays an important part in determining the actual radiation pattern of the antenna.

When any antenna is near the ground the free-space pattern is modified by reflection of radiated waves from the ground, so that the actual pattern is the resultant of the free-space pattern and ground reflections. This resultant is dependent upon the height of the antenna, its position or orientation with respect to the surface of the ground, and the electrical characteristics of the ground. The effect of a perfectly-reflecting ground is such that the

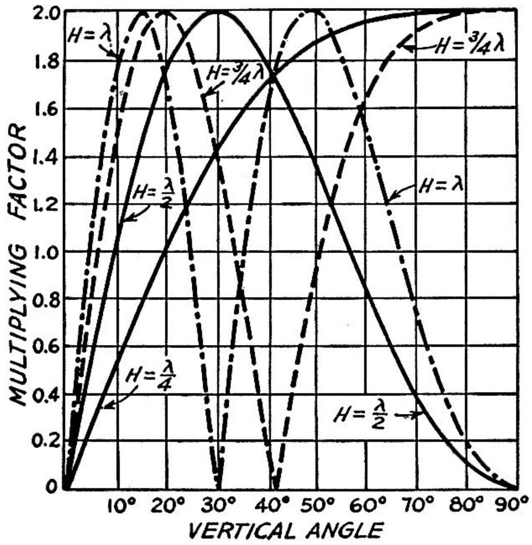


Fig. 10-42 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.

original free-space field strength may be multiplied by a factor which has a maximum value of 2, for complete reinforcement, and having all intermediate values to zero, for complete cancellation. These reflections only affect the radiation pattern in the vertical plane — that is, in directions upward from the earth's surface — and not in the horizontal plane, or the usual geographical directions.

Fig. 10-42 shows how the multiplying factor varies with the vertical angle for several representative heights for horizontal antennas. As the height is increased the angle at which complete reinforcement takes place is lowered, until for a height equal to one wavelength it occurs at a vertical angle of 15 degrees. At still greater heights, not shown on the chart, the first maximum will occur at still smaller angles.

Radiation Angle

The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies. It is advantageous, therefore, to erect the antenna at a height that will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low radiation angles usually are desirable, this generally means that the antenna should be high — at least one-half wavelength at 14 Mc., and preferably three-quarter or one wavelength; at least one wavelength, and preferably higher, at 28 Mc. and the very-high frequencies. The physical height required for a given height in wavelengths decreases as the frequency is increased, so that good heights are not impracticable; a half-wavelength at 14 Mc. is only 35 feet, approximately, while the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower frequencies the higher radiation angles are effective, so that again a reasonable antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures being preferable.

Imperfect Ground

Fig. 10-42 is based on ground having perfect conductivity, whereas the actual earth is not a perfect conductor. The principal effect of actual ground is to make the curves inaccurate at the lowest angles; appreciable high-frequency radiation at angles smaller than a few degrees is practically impossible to obtain over horizontal ground. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the sort of result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance

Waves that are reflected directly upward from the ground induce a current in the antenna in passing, and, depending on the antenna height, the phase relationship of this induced current to the original current may be such as either to increase or decrease the total current in the antenna. For the same power input to the antenna, an increase in current is equivalent to a decrease in impedance, and vice versa. Hence, the impedance of the antenna varies with height. The theoretical curve of variation of radiation resistance for an antenna above perfectly-reflecting ground is shown in Fig. 10-43. The impedance approaches the free-space value as the height becomes large, but at low heights may differ considerably from it.

Choice of Polarization

Polarization of the transmitting antenna is generally unimportant on frequencies between 3.5 and 30 Mc. However, the question of whether the antenna should be installed in a horizontal or vertical position deserves consideration for other reasons. A vertical half-wave or quarter-wave antenna will radiate

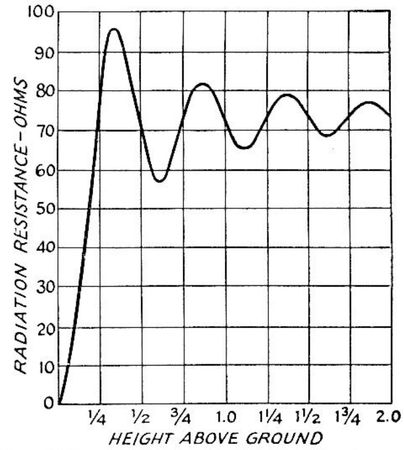


Fig. 10-43 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

equally well in all *horizontal* directions, so that it is substantially nondirectional, in the usual sense of the word. If installed horizontally, however, the antenna will tend to show directional effects, and will radiate best in the direction at right angles, or broadside, to the wire. The radiation in such a case will be least in the direction toward which the wire points.

The vertical angle of radiation also will be affected by the position of the antenna. If it were not for ground losses at high frequencies, the vertical half-wave antenna would be preferred because it would concentrate the radiation horizontally.

The Half-Wave Antenna

The fundamental form of antenna is a single wire whose length is approximately equal to half the transmitting wavelength. It is the unit from which many more-complex forms of antennas are constructed. It is variously known as a **half-wave dipole, half-wave doublet,** or **Hertz** antenna.

The length of a half-wavelength in space is:

$$\textit{Length}\ \text{(feet)} = \frac{492}{\textit{Freq.}\ \text{(Mc.)}} \qquad \textbf{(10-H)}$$

The actual length of a half-wave antenna will not be exactly equal to the half-wave in space, but depends upon the thickness of the conductor in relation to the wavelength as shown in Fig. 10-44, where K is a factor that must be multiplied by the half-wavelength in free space to obtain the resonant antenna

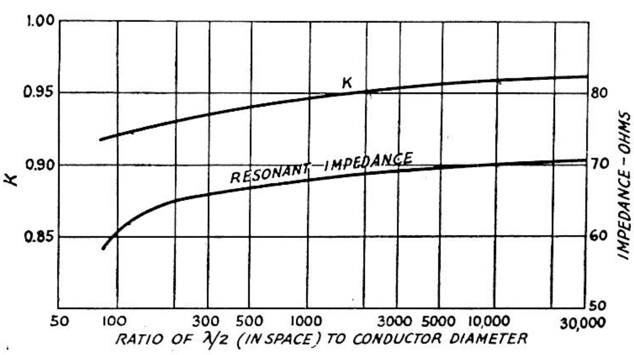


Fig. 10-44 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half-wavelength (Equation 10-H). The effect of conductor diameter on the impedance measured at the center also is shown.

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).** Under average conditions the following formula is sufficiently accurate for wire antennas at frequencies up to 30 Mc.:

$$\textit{Length of half-wave antenna}\ \text{(feet)} = \frac{492 \times 0.95}{\textit{Freq.}\ \text{(Mc.)}} = \frac{468}{\textit{Freq.}\ \text{(Mc.)}} \qquad \textbf{(10-I)}$$

Example: A half-wave antenna for 7150 kc. (7.15 Mc.) is $\frac{468}{7.15}$ = 65.45 feet, or 65 feet 5 inches.

Above 30 Mc. the following formulas should be used, particularly for antennas constructed from rod or tubing. The factor K is taken from Fig. 10-44.

$$\textit{Length of half-wave antenna}\ \text{(feet)} = \frac{492 \times K}{\textit{Freq.}\ \text{(Mc.)}} \qquad \textbf{(10-J)}$$

$$\textit{or length}\ \text{(inches)} = \frac{5905 \times K}{\textit{Freq.}\ \text{(Mc.)}} \qquad \textbf{(10-K)}$$

Example: Find the length of a half-wavelength antenna at 29 Mc., if the antenna is made of 2-inch diameter tubing. At 29 Mc., a half-wavelength in space is $\frac{492}{29}$ = 16.97 feet, from Eq. 10-H. Ratio of half-wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2}$ = 101.8. From Fig. 10-44, K = 0.92 for this ratio. The length of the antenna, from Eq. 10-J, is $\frac{492 \times 0.92}{29}$ = 15.6 feet, or 15 feet 7 inches. The answer is obtained directly in inches by substitution in Eq. 10-K: $\frac{5905 \times 0.92}{29}$ = 187 inches.

Current and Voltage Distribution

When power is fed to such an antenna, the current and voltage vary along its length. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage. The current does not actually reach zero at the current nodes, because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance of the wire (*ohmic resistance*) and the **radiation resistance.** The radiation resistance is an *equivalent* resistance, a convenient conception to indicate the radiation properties of an antenna. The radiation resistance is the equivalent resistance that would dissipate the power the antenna radiates, with a current flowing in it equal to the antenna current at a current loop (maximum). The ohmic resistance of a half-wavelength antenna is usually small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

Impedance

The radiation resistance of an infinitely-thin half-wave antenna in free space — that is, sufficiently removed from surrounding objects so that they do not affect the antenna's characteristics — is 73 ohms, approximately. The value under practical conditions is commonly taken to be in the neighborhood of 70 ohms. It is pure resistance, and is measured at the center of the antenna. The impedance

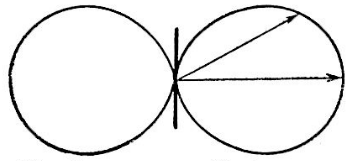


Fig. 10-45 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining the drawing glued to a piece of cardboard, with a short length of wire fastened on it to represent the antenna. Twirling the wire will give a visual representation of the solid radiation pattern.

is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends. The actual value at the ends will depend on a number of factors, such as the height, the physical construction, the insulators at the ends, and the position with respect to ground.

Conductor Size

The impedance of the antenna also depends upon the diameter of the conductor in relation to the wavelength, as shown in Fig. 10-44. If the diameter of the conductor is made large, the capacitance per unit length increases and the inductance per unit length decreases. Since the radiation resistance is affected relatively little, the decreased L/C ratio causes the Q of the antenna to decrease, so that the resonance curve becomes less sharp. Hence, the antenna is capable of working over a wide frequency range. This effect is greater as the diameter is increased, and is a property of some importance at the very-high frequencies where the wavelength is small.

Radiation Characteristics

The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions perpendicular to the wire and zero along the direction of the wire, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 10-45, which represents the radiation pattern in free space. The relative intensity of radiation is proportional to the length of a line drawn from the center of the figure to the perimeter. If the antenna is vertical, as shown in the figure, then the field strength will be uniform in all horizontal directions; if the antenna is horizontal, the relative field strength will depend upon the direction of the receiving point with respect to the direction of the antenna wire. The variation in radiation at various vertical angles from a half-wavelength horizontal antenna is indicated in Figs. 10-46 and 10-47.

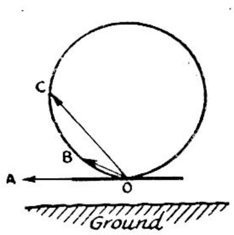


Fig. 10-46 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing of the free-space field pattern of a horizontal antenna.

FEEDING THE HALF-WAVE ANTENNA

Direct Feed

If possible, it is advisable to locate the antenna at least a half-wavelength from the transmitter and use a transmission line to carry the power from the transmitter to the antenna. However, in many cases this is impossible, particularly on the lower frequencies, and direct feed must be used. Three examples of direct feed are shown in Fig. 10-48. In the method shown at A, C_1 and C_2 should be about 150 $\mu\mu$fd. each for the 3.5-Mc. band, 75 $\mu\mu$fd. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil

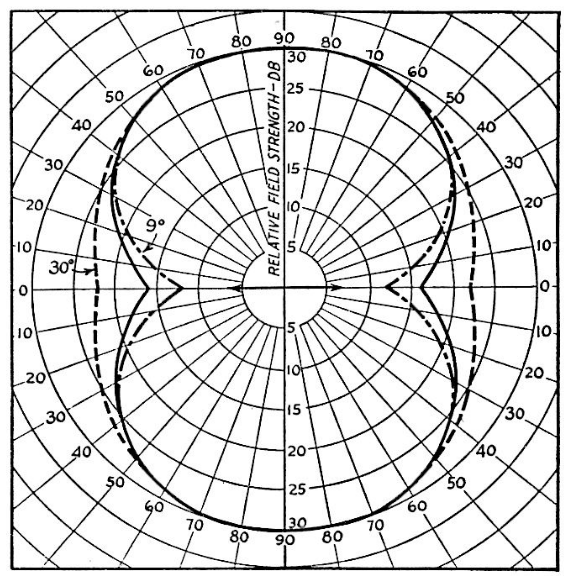


Fig. 10-47 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will depend upon the height of the antenna above ground and the vertical angle considered. The patterns for all three angles have been proportioned to the same scale, but this does not mean that the maximum amplitudes necessarily will be the same. The arrow indicates the direction of the horizontal antenna wire.

connected between them should resonate to 3.5 Mc. with about 60 or 70 $\mu\mu$fd., for the 80-meter band, for 40 meters it should resonate with 30 or 35 $\mu\mu$fd., and so on. The circuit is adjusted by using loose coupling between the antenna coil and the transmitter tank coil and adjusting C_1 and C_2 until resonance is indicated by an increase in plate current. The coupling between the coils should then be increased until proper plate current is drawn. It may be necessary to reresonate the transmitter tank circuit as the coupling is increased, but the change should be small.

The circuits in Fig. 10-48B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the

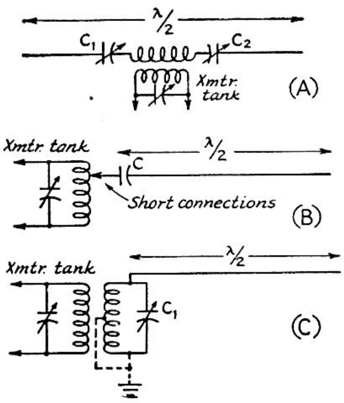


Fig. 10-48 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitance coupling; C, voltage feed, with inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

tap toward the "hot" or plate end of the tank coil — the condenser C may be of any convenient value that will stand the voltage, and it doesn't have to be variable. In the circuit at C, the antenna tuned circuit (C_1 and the antenna coil) should be similar to the transmitter tank circuit. The antenna tuned circuit is adjusted to resonance with the antenna connected but with loose coupling to the transmitter. Heavier loading of the tube is then obtained by tightening the coupling between the antenna coil and the transmitter tank coil.

Of the three systems, that at A is preferable because it is a symmetrical system and generally results in less r.f. power "floating" around the shack. The system of B is undesirable because it provides practically no protection against the radiation of harmonics, and it should only be used in emergencies.

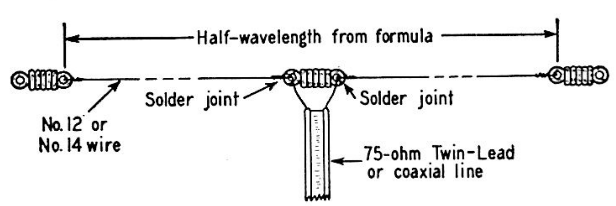


Fig. 10-49 — Construction of a half-wave doublet fed with 75-ohm line. The length of the antenna is calculated from Equation 10-I.

Transmission-Line Feed for Half-Wave Antennas

Since the impedance at the center of a half-wavelength antenna is in the vicinity of 75 ohms, it offers a good match for 75-ohm two-wire transmission lines. Several types are available on the market, with different power-handling capabilities. They can be connected in the center of the antenna, across a small strain insulator to provide a convenient connection point. Coaxial line of 75-ohms impedance can also be used, but it is heavier and thus not as convenient. In either case, the transmission line should be run away at right angles to the antenna for at least one-quarter wavelength, if possible, to avoid current unbalance in the line caused by pick-up from the antenna. The antenna length is calculated from Equation **10-I**, for a *half*-wavelength antenna. When No. 12 or No. 14 enameled wire is used for the antenna, as is generally the case, the length of the wire is the over-all length measured from the loop through the insulator at each end. This is illustrated in Fig. 10-49.

The use of 75-ohm line results in a "flat" line over most of any amateur band. However, by making the half-wave antenna in a special manner, called the **two-wire** or **folded doublet**, a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 10-50, with another version in Fig. 10-79B. The two differ only in the construction of the antenna proper. The open-wire line shown in Fig. 10-50 is made of No. 12 or No. 14 enameled wire, separated by lightweight spacers of Lucite or other material (it doesn't have to be a *low-loss* insulating material), and the spacing can be on the order of from 4 to 8 inches, depending upon what is convenient and what the operating frequency is. At 14 Mc., 4-inch separation is satisfactory, and 8-inch or even greater spacing can be used at 3.5 Mc.

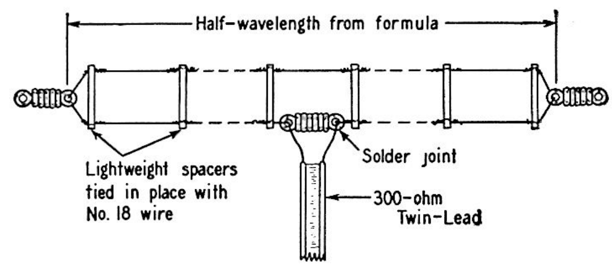


Fig. 10-50 — The construction of an open-wire folded doublet fed with 300-ohm line. The length of the antenna is calculated from Equation 10-I.

If a half-wavelength antenna is fed at the center with other than 75-ohm line, or if a folded doublet is fed with other than 300-ohm line, standing waves will appear on the line and coupling to the transmitter may become awkward for some line lengths, as described earlier in this chapter. However, in many cases it is not convenient to feed the half-wave antenna with the correct line (as is the case where multiband operation of the same antenna is desired), and sometimes it is not convenient to feed the antenna at the center. Where multiband operation is desired (to be discussed later) or when the antenna must be

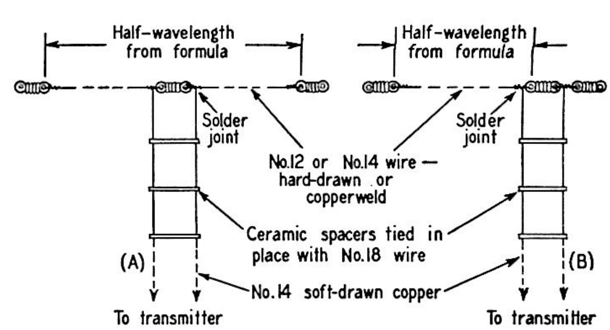


Fig. 10-51 — The antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 10-I.

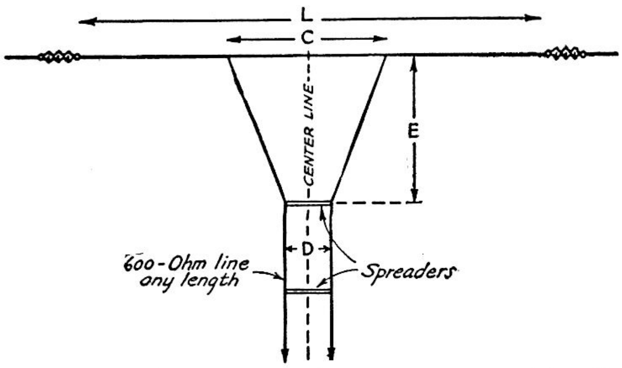


Fig. 10-52 — **Delta-matched antenna system. The dimensions *C*, *D*, and *E* are found by formulas given in the text. It is important that the matching section, *E*, come straight away from the antenna without any bends.**

fed at one end by a transmission line, an open-wire line of from 450 to 600 ohms impedance is generally used. The impedance at the end of a half-wavelength antenna is in the vicinity of several thousand ohms, and hence a standing-wave ratio of 4 or 5 is not unusual when the line is connected to the end of the antenna. It is advisable, therefore, to keep the losses in the line as low as possible. This requires the use of ceramic or Micalex feeder spacers, if any appreciable power is used. For low-power installations in dry climates, dry wood spacers that have been boiled in paraffin are satisfactory. Mechanical details of half-wavelength antennas fed with open-wire lines are given in Fig. 10-51. If the power level is low, below 100 watts or so, 300-ohm Twin-Lead can be used in place of the open line.

One method for offering a match to a 600-ohm open-wire line with a half-wavelength antenna is shown in Fig. 10-52. The system is called a **delta match.** The line is "fanned" as it approaches the antenna, to have a gradually-increasing impedance that equals the antenna impedance at the point of connection. The dimensions are fairly critical, but careful measurement before installing the antenna and matching section is generally all that is necessary. The length of the antenna, L, is calculated from Equation 10-I. The length of section C is computed from:

$$C \text{ (feet)} = \frac{118}{Freq. \text{ (Mc.)}} \qquad \textbf{(10-L)}$$

The feeder clearance, E, is found from

$$E \text{ (feet)} = \frac{148}{Freq. \text{ (Mc.)}} \qquad \textbf{(10-M)}$$

Example: For a frequency of 7.1 Mc., the length

$$L = \frac{468}{7.1} = 65.91 \text{ feet, or 65 feet 11 inches.}$$

$$C = \frac{118}{7.1} = 16.62 \text{ feet, or 16 feet 7 inches.}$$

$$E = \frac{148}{7.1} = 20.84 \text{ feet, or 20 feet 10 inches.}$$

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires 4¾-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or 3¾-inch spaced No. 16 wire.

Long-Wire Antennas

An antenna will be resonant so long as an integral number of standing waves of current and voltage can exist along its length; in other words, so long as its length is some integral multiple of a half-wavelength. When the antenna is more than a half-wave long it usually is called a long-wire antenna, or a harmonic antenna.

Current and Voltage Distribution

Fig. 10-53 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is equal to a half-wavelength) and at its second, third and fourth harmonics. For example, if the fundamental frequency of the antenna is 7 Mc., the current and voltage distribution will be as shown at A. The same antenna excited at 14 Mc. would have current and voltage distribution as shown at B. At 21 Mc., the third harmonic of 7 Mc., the current and voltage distribution would be as in C; and at 28 Mc., the fourth harmonic, as in D. The number of the harmonic is the number of half-waves contained in the antenna at the particular operating frequency.

The polarity of current or voltage in each standing wave is opposite to that in the ad-

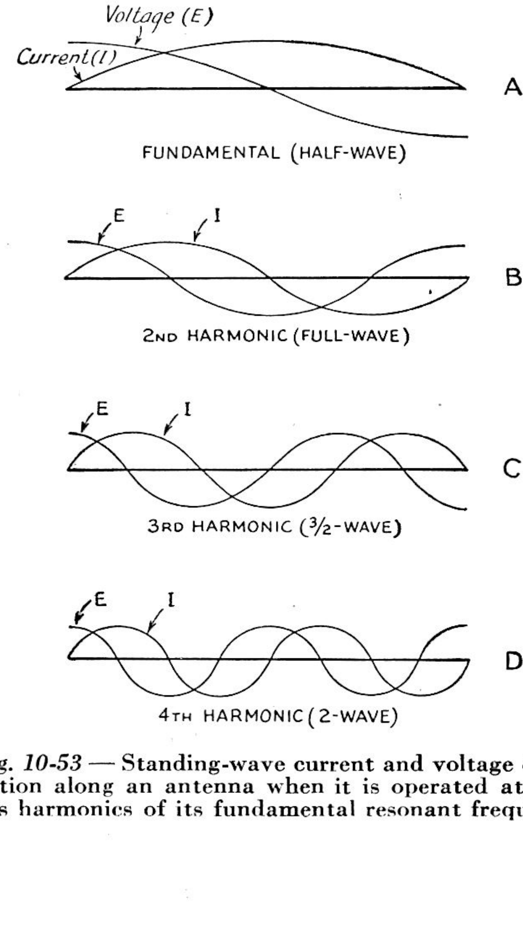


Fig. 10-53 — **Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.**

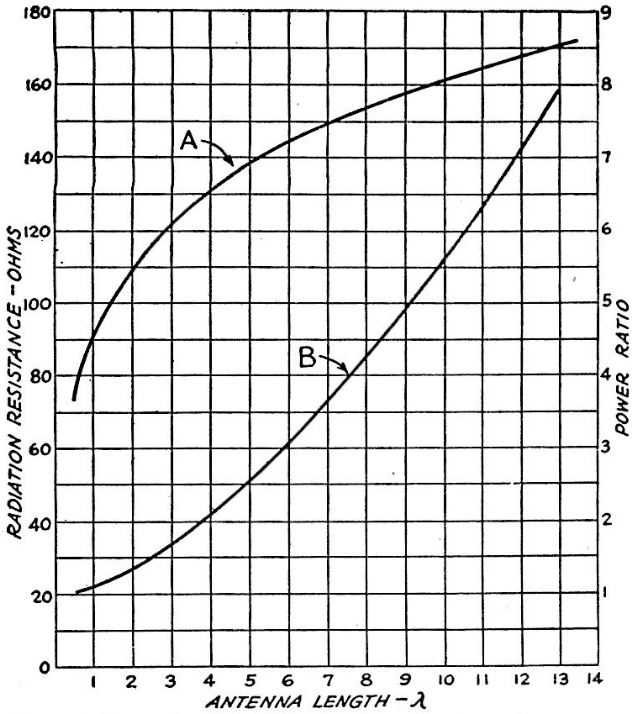


Fig. 10-54 — **Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.**

jacent standing waves. This is shown in the figure by drawing the current and voltage curves successively above and below the antenna (taken as a zero reference line), to indicate that the polarity reverses when the current or voltage goes through zero. Currents flowing in the same direction are *in phase;* in opposite directions, *out of phase.*

It is evident that one antenna may be used for harmonically-related frequencies, such as the various amateur bands. The long-wire or harmonic antenna is the basis of multiband operation with one antenna.

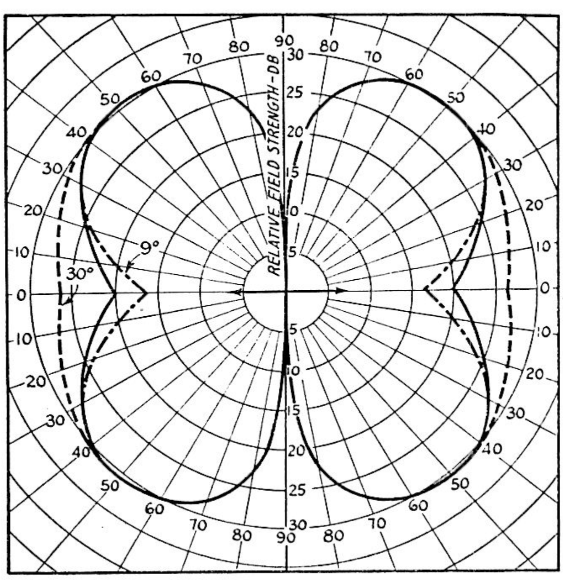


Fig. 10-55 — **Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale; actual amplitudes will depend upon the height of the antenna.**

Physical Lengths

The length of a long-wire antenna is not an exact multiple of that of a half-wave antenna because the end effects operate only on the end sections of the antenna; in other parts of the wire these effects are absent, and the wire length is approximately that of an equivalent portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$\textit{Length}\text{ (feet)} = \frac{492\ (N-0.05)}{\textit{Freq.}\ (\text{Mc.})} \quad \textbf{(10-N)}$$

where N is the number of *half*-waves on the antenna.

Example: An antenna 4 half-waves long at 14.2 Mc. would be $\frac{492\ (4\ -\ 0.05)}{14.2} = \frac{492 \times 3.95}{14.2}$ = 136.7 feet, or 136 feet 8 inches.

It is apparent that an antenna cut as a half-wave for a given frequency will be slightly off

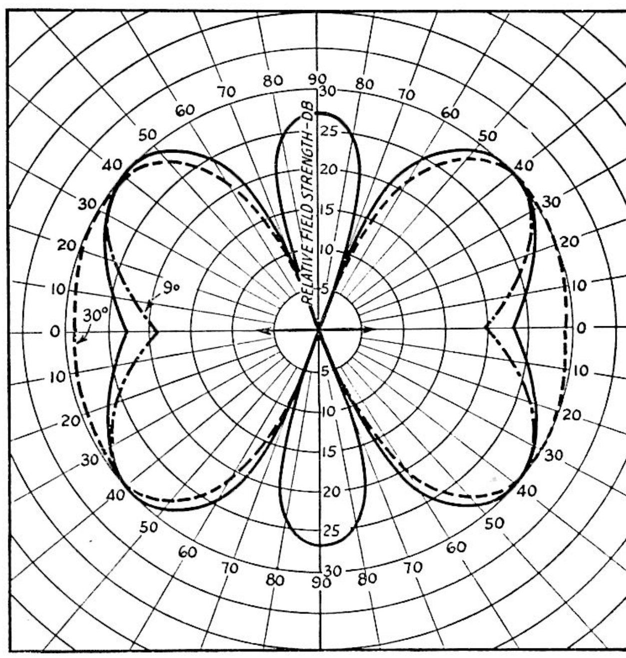


Fig. 10-56 — **Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.**

resonance at exactly twice that frequency (the second harmonic), because of the decreased influence of the end effects when the antenna is more than one-half wavelength long. The effect is not very important, except for a possible unbalance in the feeder system and consequent radiation from the feedline. If the antenna is fed in the exact center, no unbalance will occur at any frequency, but end-fed systems will show an unbalance in all but one frequency band, the band for which the antenna is cut.

Impedance and Power Gain

The radiation resistance as measured at a current loop becomes larger as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable 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. 10-54 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

Directional Characteristics

As the wire is made longer in terms of the number of half-wavelengths, the directional effects change. Instead of the "doughnut" pattern of the half-wave antenna, the directional characteristic splits up into "lobes" which make various angles with the wire. In general, as the length of the wire is increased the direction in which maximum radiation occurs tends to approach the line of the antenna itself.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 10-55, 10-56 and 10-57, for three vertical angles of radiation. Note that, as the wire length increases, the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to have practically "end-on" directional characteristics, even at the lower radiation angles.

Methods of Feeding

In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 10-53. The feeder system must not upset this phase relationship. This requirement is met by feeding the antenna at either end or at any current *loop*. A two-wire feeder cannot be inserted at a current *node*, however, because this invariably brings the currents in two adjacent half-wave sections in phase; if the phase in one section could be reversed, then the currents in the feeders necessarily would have to be in phase and the feeder radiation would not be canceled out.

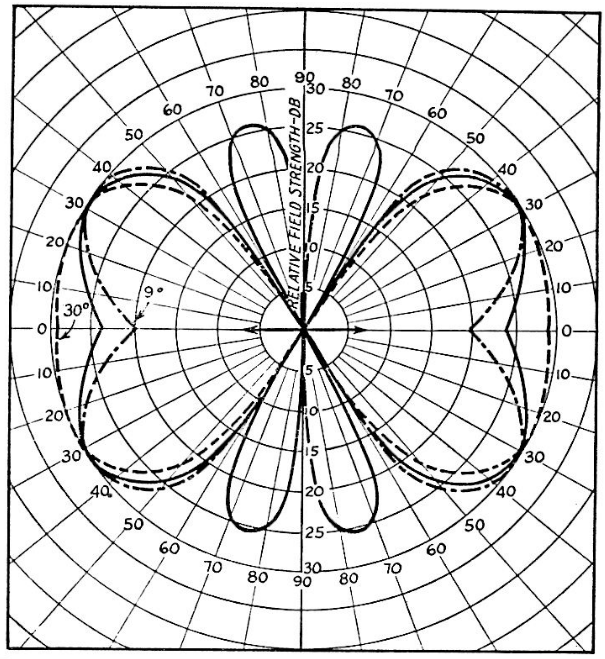


Fig. 10-57 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

No point on a long-wire antenna offers a reasonable impedance for a direct match to any of the common types of transmission lines. The most common practice is to feed the antenna at one end or at a current loop with a low-loss open-wire line and accept the resulting standing-wave ratio of 4 or 5. When a better match is required, "stubs" are generally used (described later in this chapter).

Multiband Antennas

As suggested in the preceding section, the same antenna may be used for several bands by operating it on harmonics. When this is done it is necessary to use resonant feeders, since the impedance matching for nonresonant feeder operation can be accomplished only at one frequency unless means are provided for changing the length of a matching section and shifting the point at which the feeder is attached to it.

Furthermore, the current loops shift to a new position on the antenna when it is operated on harmonics, further complicating the feed situation. It is for this reason that a half-wave antenna which is center-fed by a rubber-insulated line is practically useless for harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses arise in the rubber dielectric. It is also wise not to attempt to use a half-wave antenna center-fed with coaxial cable on its harmonics. Higher-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

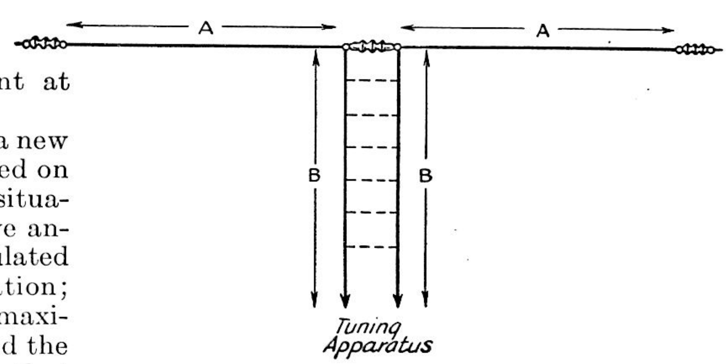


Fig. 10-58 — Practical arrangement of a shortened antenna. The total length, $A + B + B + A$, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table 10-III for lengths and tuning data.

TABLE 10-II
Multiband Resonant-Line Fed Antennas

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed: 120	60	4-Mc. 'phone	series
136	67	3.5-Mc. c.w. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
134	67	3.5-Mc. c.w. 7 Mc.	series parallel
67	33	7 Mc. 14 Mc. 28 Mc.	series parallel parallel
With center feed: 137	67	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel parallel
67.5	34	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc., but will work well in the region that quadruples into the 14-Mc. band (3500–3600 kc.). Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will vary with the band in use.

Simple Systems

The most practical simple multiband antenna is one that is a half-wavelength long at the lowest frequency and is fed either at the center or one end with an open-wire line. Although the standing-wave ratio on the feedline will not approach 1.0 on any band, if the losses in the line are low the system will be efficient. From the standpoint of reduced feedline radiation, a center-fed system is superior to one that is end-fed, but the end-fed arrangement is often more convenient and should not be ignored as a possibility. The center-fed antenna will not have the same radiation pattern as an end-fed one of the same length, except on frequencies where the over-all length of the antenna is a half-wavelength or less. The end-fed antenna acts like a long-wire antenna on all bands (for which it is longer than a half-wavelength), but the center-fed one acts like two antennas of half that length fed in phase. For example, if a full-wavelength antenna is fed at one end, it will have a radiation pattern as shown in Fig. 10-55, but if it is fed in the center the pattern will be somewhat similar to Fig. 10-47, with the maximum radiation broadside to the wire. Either antenna is a good radiator, but if the radiation pattern is a factor, the point of feed must be considered.

Since multiband operation of an antenna does not permit matching of the feedline, some attention must be paid to the length of the feedline if convenient transmitter-coupling arrangements are to be obtained. Table 10-II gives some suggested antenna and feeder lengths for multiband operation. In general, the length of the feedline should be some integral multiple of a quarter wavelength at the lowest frequency.

Antennas for Restricted Space

If the space available for the antenna is not large enough to accommodate the length necessary for a half-wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half-wave long. Nevertheless, such a system is useful where operation on the desired band otherwise would be impossible.

Resonant feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance. With end feed the feeder currents become badly unbalanced.

With center feed practically any convenient length of antenna can be used, if the feeder length is adjusted to accommodate at least

TABLE 10-III
Antenna and Feeder Lengths for Short Multiband Antennas, Center-Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	38	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel series series series or parallel
67.5	34	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
50	43	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	51	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	31	7 Mc. 14 Mc. 28 Mc.	parallel series parallel

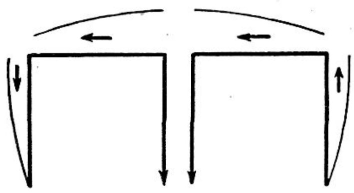


Fig. 10-59 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

one half-wave around the whole system.

A practical antenna of this type can be made as shown in Fig. 10-58. Table 10-III gives a few recommended lengths. However, the antenna can be made any convenient length, provided the total length of wire is a half-wavelength at the lowest frequency, or an integral multiple of a half-wavelength.

Bent Antennas

Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a half-wave antenna (the center quarter wave, approximately) does most of the radiating. Advantage can be taken of this fact when the space available does not permit erecting an antenna a half-wave long. In this case the ends may be bent, either horizontally or vertically, so that the total length equals a half-wave, even though the straightaway horizontal length may be as short as a quarter wave. The operation is illustrated in Fig. 10-59. Such an antenna will be a somewhat better radiator than a quarter-wavelength antenna on the lowest frequency, but is not so desirable for multiband operation because the ends play an increasingly important part as the frequency is raised. The performance of the system in such a case is difficult to predict, especially if the ends are vertical (the most convenient arrangement) because of the complex combination of horizontal and vertical polarization which results as well as the dissimilar directional characteristics. However, the fact that the radiation pattern is incapable of prediction does not detract from the general usefulness of the antenna.

Long-Wire Directive Arrays

THE "V" ANTENNA

It has been emphasized that, as the antenna length is increased, the lobe of maximum radiation makes a more acute angle with the wire. Two such wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bidirectional) along a line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate. It can also be used on harmonics, so that it is suitable for multiband work. The "V" antenna is shown in Fig. 10-60.

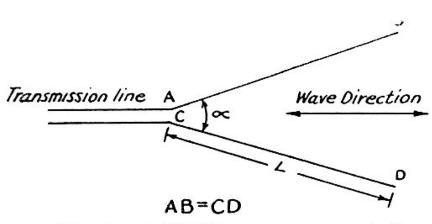


Fig. 10-60 — The basic "V" antenna, made by combining two long wires.

Fig. 10-61 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multiband operation. Angle ∝ is approximately equal to twice the angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 10-61 is the vertical angle of maximum radiation. Tilting the whole horizontal plane of the "V" will tend to increase the low-angle radiation off the low end and decrease it off the high end.

The gain increases with the length of the

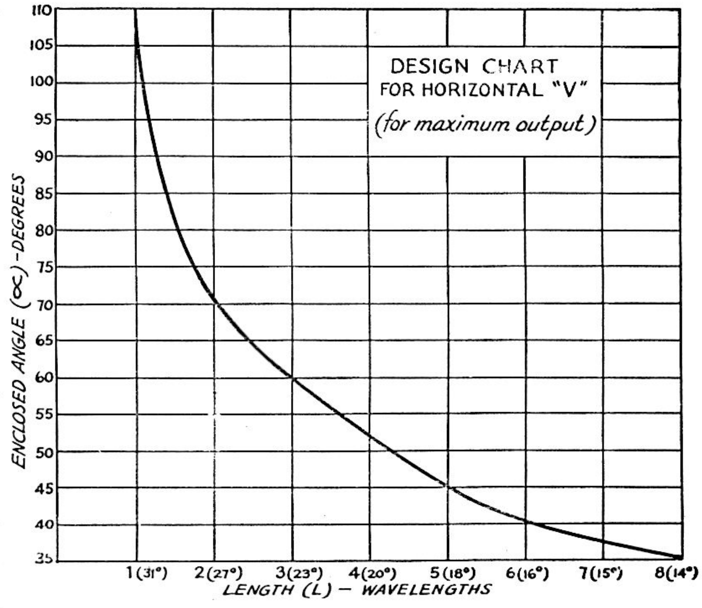


Fig. 10-61 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires. Values in parentheses represent approximate wave angle for height of one-half wavelength.

wires, but is not exactly twice the gain for a single long wire as given in Fig. 10-54. In the longer lengths the gain will be somewhat increased, because of mutual coupling between the wires. A "V" eight wavelengths on a leg, for instance, will have a gain of about 12 db. over a half-wave antenna, whereas twice the gain of a single eight-wavelength wire would be only approximately 9 db.

The two wires of the "V" must be fed out of phase, for correct operation. A resonant line may simply be attached to the ends, as shown in Fig. 10-60. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a nonresonant line. If the antenna wires are made multiples of a half-wave in length (use Equation 10-N for computing the length), the matching section will be closed at the free end. A stub can be connected across the resonant line to provide a match, as described later.

THE RHOMBIC ANTENNA

The horizontal rhombic or "diamond" antenna is shown in Fig. 10-62. Like the "V," it requires a great deal of space for erection, but it is capable of giving excellent gain and directivity. It also can be used for multiband operation. In the terminated form shown in Fig. 10-62, it operates like a nonresonant transmission line, without standing waves, and is unidirectional. It may also be used without the terminating resistor, in which case there are standing waves on the wires and the antenna is bidirectional.

The important quantities influencing the design of the rhombic antenna are shown in Fig. 10-62. While several design methods may be used, the one most applicable to the conditions existing in amateur work is the so-called "compromise" method. The chart of Fig. 10-63 gives design information based on a given length and wave angle to determine the remaining optimum dimensions for best operation. Curves for values of length of two, three

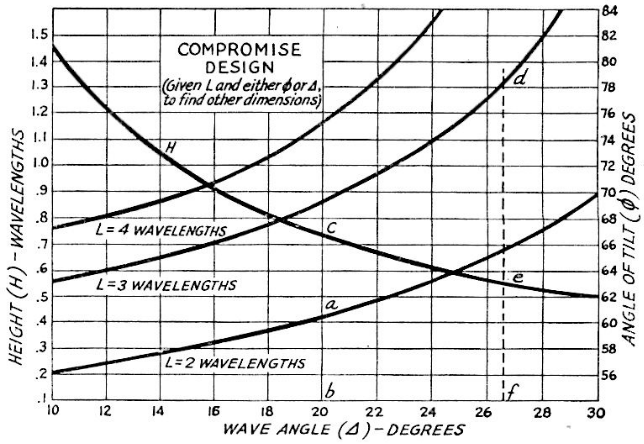


Fig. 10-63 — Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:

(1) Given:

Length (L) = 2 wavelengths.
Desired wave angle (Δ) = 20°.

To Find: H, Φ.

Method:

Draw vertical line through point a (L = 2 wavelengths) and point b on abscissa (Δ = 20°). Read angle of tilt (Φ) for point a and height (H) from intersection of line ab at point c on curve H.

Result:

Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:

Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.

To Find: H, Δ.

Method:

Draw a vertical line from point d on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point e) for height, and intersection at point f on the abscissa for Δ.

Result:

H = 0.56 wavelength.
Δ = 26.6°.

and four wavelengths are shown, and any intermediate values may be interpolated.

With all other dimensions correct, an increase in length causes an increase in power gain and a slight reduction in wave angle. An increase in height also causes a reduction in wave angle and an increase in power gain, but not to the same extent as a proportionate increase in length. For multiband work, it is satisfactory to design the rhombic antenna on the basis of 14-Mc. operation, which will permit work from the 7- to 28-Mc. bands as well.

A value of 800 ohms is correct for the terminating resistor for any properly-constructed rhombic, and the system behaves as a pure resistive load under this condition. The terminating resistor must be capable of safely dissipating one-half the power output (to eliminate the rear pattern), and should be noninductive. Such a resistor may be made up from a carbon or graphite rod or from a long 800-ohm transmission line using

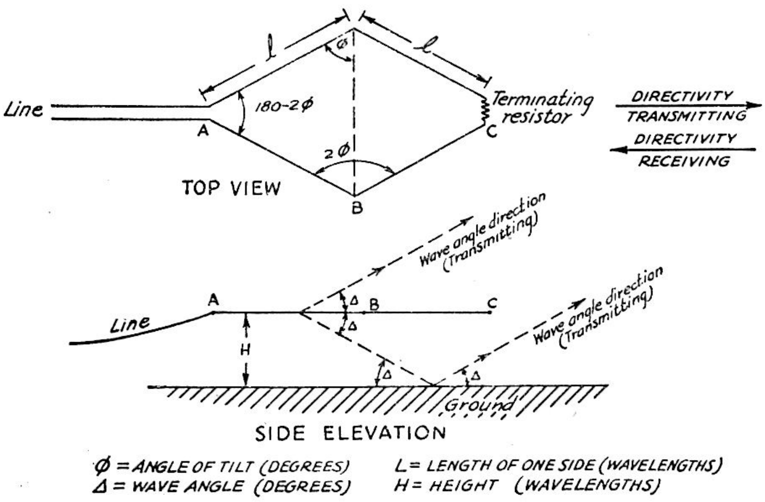


Fig. 10-62 — The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

resistance wire. If the carbon rod or a similar form of lumped resistance is used, the device should be suitably protected from weather effects, i.e., it should be covered with a good asphaltic compound and sealed in a small lightweight box or fiber tube. Suitable nonreactive terminating resistors are also available commercially.

For feeding the antenna, the antenna impedance will be matched by an 800-ohm line, which may be constructed from No. 16 wire spaced 20 inches or from No. 18 wire spaced 16 inches. The 800-ohm line is somewhat ungainly to install, however, and may be replaced by an ordinary 600-ohm line with only a negligible mismatch. Alternatively, a matching section may be installed between the antenna terminals and a low-impedance line. However, when such an arrangement is used, it will be necessary to change the matching-section constants for each different band on which operation is contemplated.

The same design details apply to the unterminated rhombic as to the terminated type. When used without a terminating resistor, the system is bidirectional. Resonant feeders are preferable for the unterminated rhombic. A nonresonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to satisfactory multiband work.

Rhombic antennas will give a power gain of 8 to 12 db. or more for leg lengths of two to four wavelengths, when constructed according to the charts given. In general, the larger the antenna, the greater the power gain.

Directive Arrays with Driven Elements

By combining individual half-wave antennas into an **array** with suitable spacing between the antennas (called **elements**) and feeding power to them simultaneously, it is possible to make the radiated fields from the individual elements add in a favored direction, thus increasing the field strength in that direction as compared to that produced by one antenna element alone. In other directions the fields will more or less oppose each other, giving a reduction in field strength. Thus a power gain in the desired direction is secured at the expense of a power reduction in other directions.

Besides the spacing between elements, the instantaneous direction of current flow (*phase*)

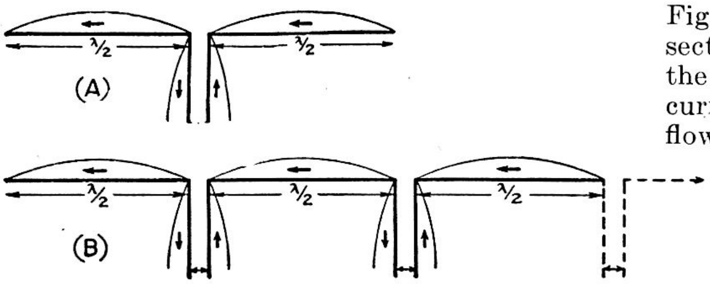


Fig. 10-64 — Collinear half-wave antennas in phase. The system at A is generally known as "two half-waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

in individual elements determines the directivity and power gain. There are several methods of arranging the elements. If they are strung end to end, so that all lie on the same straight line, the elements are said to be **collinear.** If they are parallel and all lying in the same plane, the elements are said to be **broadside** when the phase of the current is the same in all, and **end-fire** when the currents are not in phase. Elements that receive power from the transmitter through the transmission line are called **driven elements.**

The power gain of a directive system increases with the number of elements. The proportionality between gain and number of elements is not simple, however. The gain depends upon the effect that the spacing and phasing has upon the radiation resistance of the elements, as well as upon their number.

Collinear Arrays

Simple forms of collinear arrays, with the current distribution, are shown in Fig. 10-64. The two-element array at A is popularly known as "two half-waves in phase." It will be recognized as simply a center-fed antenna operated at its second harmonic. The way in which the number of elements may be extended for increased directivity and gain is shown in Fig. 10-64B. Note that quarter-wave phasing sections are used between elements; these give the reversal in phase necessary to make the currents in individual antenna elements all flow in the same direction at the same instant.

Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder, or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end-fed by any of the systems previously described, or any element may be center-fed. It is best to feed at the center of the array, so that the energy will be distributed as uniformly as possible among the elements.

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table 10-IV. Although three-quarter wave spacing gives greater gain, it is difficult to construct a suitable phase-reversing system when the ends of the antenna elements are widely separated. For this reason, the half-wave spacing is most generally used in actual practice.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mount-

TABLE 10-IV
Theoretical Gain of Collinear Half-Wave Antennas

Spacing between centers of adjacent half-waves	*Number of half-waves in array vs. gain in db.*				
	2	3	4	5	6
½ wave	1.8	3.3	4.5	5.3	6.2
¾ wave	3.2	4.8	6.0	7.0	7.8

ing gives increased horizontal directivity, while the vertical directivity remains the same as for a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles. It is seldom practicable to use more than two elements vertically at frequencies below 14 Mc. because of the excessive height required.

Broadside Arrays

Parallel antenna elements with currents in phase may be combined as shown in Fig. 10-65 to form a **broadside** array, so named because

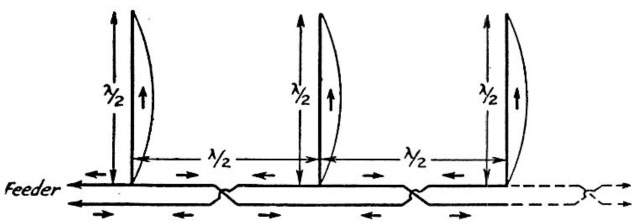


Fig. 10-65 — **Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).**

the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the number of elements and the spacing, the gain for different spacings being shown in Fig. 10-66. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table 10-V gives theoretical gain as a function of the number of elements with half-wave spacing.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other **(stacked).** In the former case the horizontal pattern becomes quite sharp, while the vertical pattern is the same as that of one element alone. If the array is suspended horizontally, the horizontal pattern is equivalent to that of one element while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines or through quarter-wave matching sections and nonresonant lines. In Fig. 10-65, note the "crossing over" of the feeders, which is necessary to bring the elements into proper phase relationship.

Combined Broadside and Collinear Arrays

Broadside and collinear arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. The general plan of constructing such antennas is shown in Fig. 10-67. The lower angle of radiation resulting from stacking elements in the vertical plane is desirable at the higher frequencies. In general, doubling the number of elements in an array by stacking will raise the gain from 2 to 4 db., depending upon whether vertical or horizontal elements are used — that is, whether the stacked elements are of the broadside or collinear type.

The arrays in Fig. 10-67 are shown fed from one end, but this is not especially desirable in the case of large arrays. Better distribution of energy between elements, and hence better over-all performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the eight-element array at A, the feeders could be introduced at the middle of the transmission line between the second and third set of elements, in which case the connecting line would not be transposed between the second and third set of elements. Alternatively, the antenna could be constructed with the transpositions as shown and the feeder connected between the adjacent ends of either the second or third pair of collinear elements.

A four-element array of the general type shown in Fig. 10-67B, known as the "lazy-H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 10-68.

End-Fire Arrays

Fig. 10-69 shows a pair of parallel half-wave elements with currents out of phase. This is known as an **end-fire** array, because it radiates best along the line of the antennas, as shown.

The end-fire array may be used either ver-

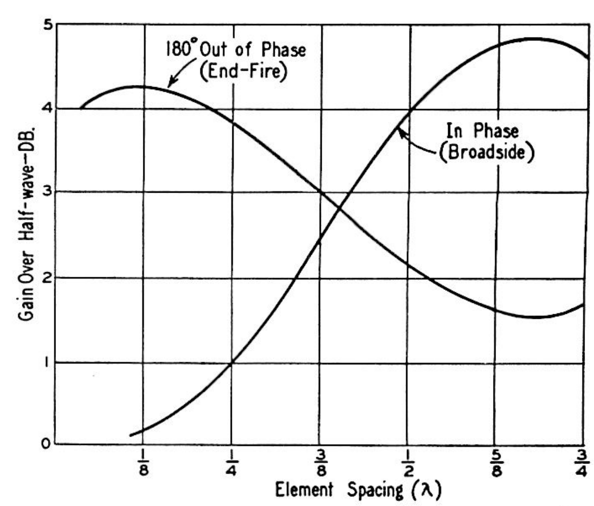


Fig. 10-66 — **Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.**

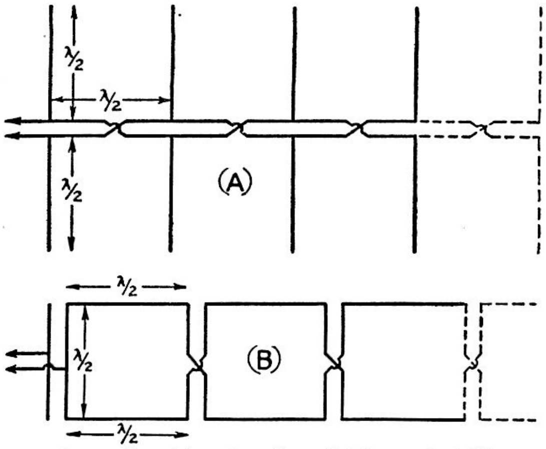


Fig. 10-67 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table 10-V) plus the gain of one set of collinear elements (Table 10-IV). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.), giving a total gain of 8.8 db. In B, each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.), making the total gain 7.3 db. The result is not strictly accurate, because of mutual coupling between the elements, but is good enough for practical purposes.

tically or horizontally (elements at the same height), and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 10-66 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements to give a further increase in gain and directivity.

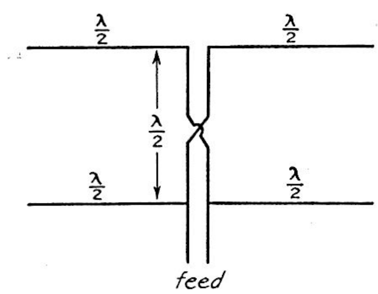


Fig. 10-68 — A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

Either resonant or nonresonant lines may be used with this type of array. Nonresonant lines preferably are matched to the antenna through a quarter-wave matching section or phasing stub.

Phasing

Figs. 10-67 and 10-69 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 10-69, when the transmission line is connected as at A there is no crossover in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the crossover becomes necessary (B). This is because in B the two halves of the connecting line are simply *branches* of the same line. In other words, even though the connecting line in B is a half-wave in length, it is not actually a half-wave line but *two quarter-wave lines in parallel.* The same thing is true of the untransposed line of Fig. 10-67. Note that, under these conditions, the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two antenna elements and does not have the feedline connected to its center, as in Fig. 10-65.

Adjustment of Arrays

With arrays of the types just described, using half-wave spacing between elements, it

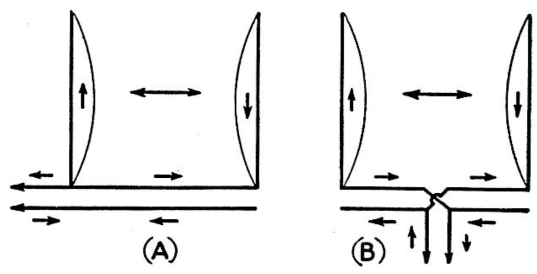


Fig. 10-69 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 10-66. Direction of maximum radiation is shown by the large arrows.

will usually suffice to make the length of each element that given by Equations 10-I or 10-J. The half-wave phasing lines between the parallel elements should be of open-wire construction, and their length can be calculated from:

$$\textit{Length of half-wave line (feet)} = \frac{480}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-O)}$$

Example: A half-wavelength phasing line for 28.8 Mc. would be $\frac{480}{28.8} = 16.66$ feet = 16 feet 8 inches.

The spacing between elements can be made equal to the length of the phasing line. No special adjustments of line or element length or spacing are needed, provided the formulas are followed closely.

TABLE 10-V
Theoretical Gain *vs.* Number of Broadside Elements (Half-Wave Spacing)

No. of elements	*Gain*
2	4 db.
3	5.5
4	7
5	8
6	9

With collinear arrays of the type shown in Fig. 10-64B, the same formula may be used for the element length, while the length of the quarter-wave phasing section can be found from the following formula:

$$\textit{Length of quarter-wave line}\ \text{(feet)} = \frac{240}{\textit{Freq.}\ \text{(Mc.)}} \qquad \textbf{(10-P)}$$

Example: A quarter-wavelength phasing line for 14.25 Mc. would be $\frac{240}{14.25}$ = 16.84 feet = 16 feet 10 inches.

If the array is fed in the center it should not be necessary to make any particular adjustments, although, if desired, the whole system can be resonated by connecting an r.f. ammeter in the shorting link of each phasing section and moving the link back and forth to find the maximum-current position. This refinement is hardly necessary in practice, however, so long as all elements are the same length and the system is symmetrical.

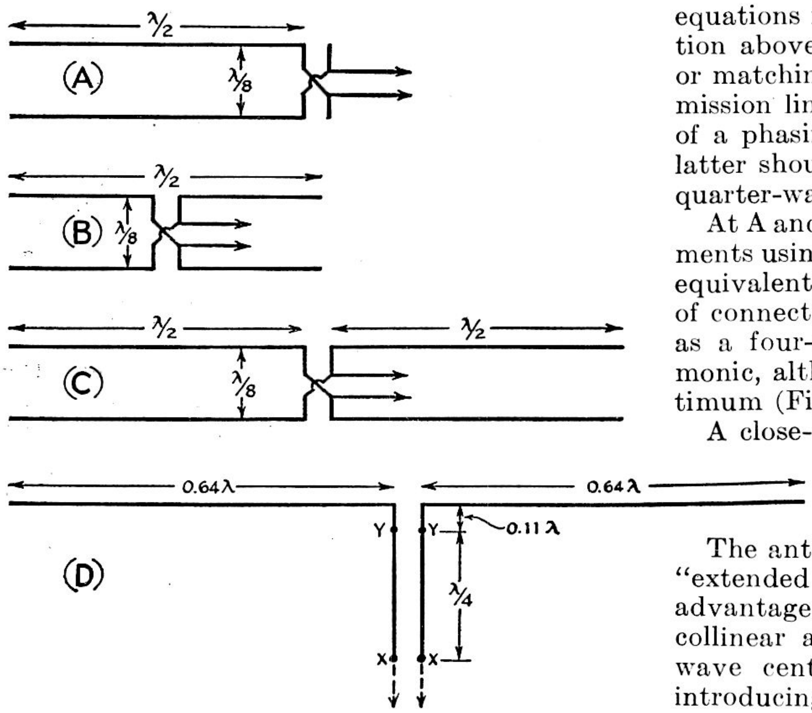


Fig. 10-70 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is a simple two-element broadside array using extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5-db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db. In A, B and C, the phasing line contributes about 1/16 wavelength to the transmission line; when B is used on the second harmonic, this contribution is ⅛ wavelength. Alternatively, the antenna ends may be bent to meet the transmission line, in which case each feeder is simply connected to one antenna. In D, points *Y-Y* indicate a quarter-wave point (high current) and *X-X* a half-wave point (high voltage). The line may be extended in multiples of quarter waves if resonant feeders are to be used. A, B and C may be suspended on wooden spreaders. The plane containing the wires should be parallel to the ground.

The phasing sections can be made of 300-ohm Twin-Lead, if low power is used. However, the lengths of the phasing sections must be only 84 per cent of the length obtained in the two formulas above.

Example: The half-wavelength line for 28.8 Mc. would become 0.84 × 16.66 = 13.99 feet = 14 feet 0 inches.

Using Twin-Lead for the phasing sections is most useful in arrays such as that of Fig. 10-64B, or any other system in which the element spacing is not controlled by the length of the phasing section.

Simple Arrays

Several simple directive-antenna systems using driven elements have achieved rather wide use among amateurs. Four of these systems are shown in Fig. 10-70. Tuned feeders are assumed in all cases; however, a matching section readily can be substituted if a nonresonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths can be calculated from the equations for the antenna and from the equation above for the resonant transmission line or matching section. In cases where the transmission line proper connects to the midpoint of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 10-66) for such operation.

A close-spaced four-element array is shown at C. It will give about 2 db. more gain than the two-element array.

The antenna at D, commonly known as the "extended double-Zepp," is designed to take advantage of the greater gain possible with collinear antennas having greater than half-wave center-to-center spacing, but without introducing feed complications. The elements are made longer than a half-wave in order to bring this about. The gain is 3 db. over a single half-wave antenna, and the broadside directivity is fairly sharp.

The antennas of A and B may be mounted either horizontally or vertically; *horizontal* suspension (with the elements in a plane parallel to the ground) is recommended, since this tends to give low-angle radiation without an unduly sharp horizontal pattern. Thus these systems are useful for coverage over a wide horizontal angle. The system at C, when mounted horizontally, will have a sharper horizontal pattern than the two-element arrays because of the effect of the collinear arrangement. The vertical pattern, however, will be the same as that of the antennas in A and B.

Matching the Antenna to the Line

Except in the several cases of half-wave antennas mentioned earlier, most antenna systems do not have center impedances that readily match open-wire lines or available solid-dielectric ones. However, any antenna can be matched to practically any line by any of the several means to be described. The matching is accomplished by first resonating the antenna to the proper frequency and then introducing either a matching transformer between the antenna and the line or by applying corrective stubs to the line.

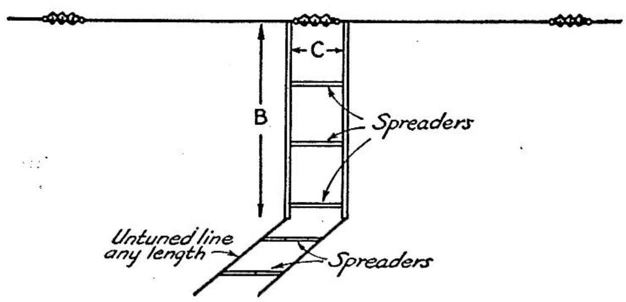


Fig. 10-71 — **The "Q" antenna, using a quarter-wave impedance-matching section with close-spaced conductors.**

An impedance mismatch of 10 or 20 per cent is of little consequence so far as power transfer to the antenna is concerned. It is relatively easy to get the standing-wave ratio down to 1.5- or 2-to-1, a perfectly satisfactory condition in practice. Of considerably greater importance is the necessity for getting the currents in the two wires balanced, both as to amplitude and phase. If the currents are not the same at corresponding points on adjacent wires and the loops and nodes do not also occur at corresponding points, there will be considerable radiation loss. Perfect balance can be brought about only by perfect symmetry in the line, particularly with respect to ground. This symmetry should extend to the coupling apparatus at the transmitter.

In the following discussion of ways in which different types of lines may be matched to the antenna, a half-wave antenna is used as an example. Other types of antennas may be treated by the same methods, making due allowance for the order of impedance that appears at the end of the line when more elaborate systems are used.

"Q"-Section Transformer

The impedance of a two-wire line of ordinary construction (400 to 600 ohms) can be matched to the impedance of the center of a half-wave antenna by utilizing the impedance-transforming properties of a quarter-wave line, Equation 10-B. The matching section must have low surge impedance and therefore is commonly constructed of large-diameter conductors such as aluminum or copper tubing, with fairly-close spacing. This system is known as the "Q" antenna. It is shown in Fig. 10-71. Important dimensions are the length of the antenna itself, the length of the matching section, B, the spacing between the two conductors of the matching section, C, and the impedance of the untuned transmission line connected to the lower end of the matching section.

The required characteristic impedance for the matching section is

$$Z_m = \sqrt{Z_1 Z_2} \qquad \textbf{(10-B)}$$

where Z_1 and Z_2 are the antenna and feedline impedances.

Example: To match a 600-ohm line to an antenna presenting a 72-ohm load, the quarter-wave matching section would require a characteristic impedance of $\sqrt{72 \times 600} = \sqrt{43{,}200} = 208$ ohms.

The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 10-14. With ½-inch tubing, the spacing should be 1.5 inches for an impedance of 208 ohms.

The length of the matching section, B, should be equal to a quarter wavelength, and is given by Equation 10-G. The length of the antenna can be calculated from Equations 10-I or 10-J.

This system has the advantage of the simplicity of adjustment of the 75-ohm feeder system and at the same time the superior insulation of an open-wire system.

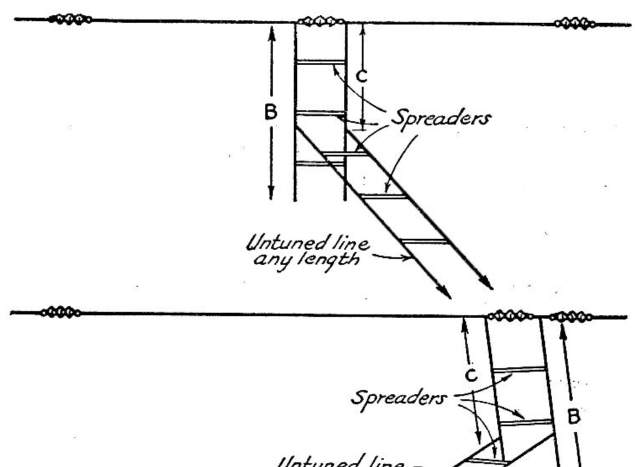


Fig. 10-72 — **Antenna systems with quarter-wave open-wire linear impedance-matching transformers.**

Linear Transformers

Fig. 10-72 shows two methods of coupling a nonresonant line to an antenna through a quarter-wave linear transformer or matching section. In the case of the center-fed antenna, the free end of the matching section, B, is open (high impedance) if the other end is connected to a low-impedance point (current loop) on the antenna. With the end-fed antenna, the free end of the matching section is closed through a shorting bar or link; this end of the section has low impedance, since the other end is connected to a high-impedance point on the antenna.

When the connection between the matching section and the antenna is unbalanced, as in the end-fed system, it is important that the antenna be the right length for the operating frequency if a good match is to be obtained. The balanced center-fed system is less critical in this respect. The shorting-bar method of tuning the center-fed system to resonance may be used if the matching section is extended to a half-wavelength, bringing a current loop at the free end.

In the center-fed system, the antenna and matching section should be cut to lengths found from Equations 10-I, 10-N and 10-P. Any necessary on-the-ground adjustment can be made by adding to or clipping off the open ends of the matching section. In the end-fed system the matching section can be adjusted

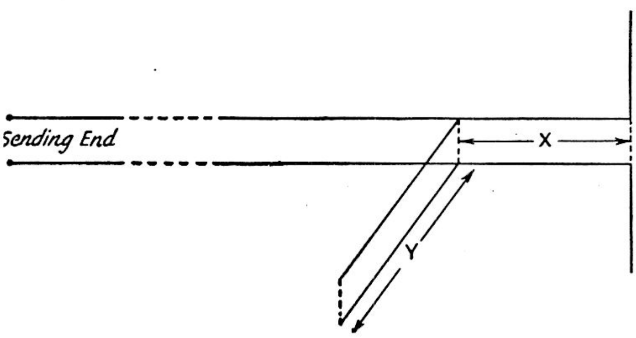


Fig. 10-73 — **When antenna and transmission line differ in impedance, they may be matched by a short length of transmission line, *Y*, called a stub. Determination of the critical dimensions, *X* and *Y*, for proper matching depends on whether the stub is open or closed at the end.**

by making the line a little longer than necessary and adjusting the system to resonance by moving the shorting link up and down. Resonance can be determined by exciting the antenna at the proper frequency from a temporary antenna near by and measuring the current in the shorting bar by a low-range r.f. ammeter or galvanometer using one of the devices of this type described in the chapter on measurements. The position of the bar should be adjusted for maximum current reading. This should be done before the transmission line is attached to the matching section.

The position of the line taps will depend upon the impedance of the line as well as on the antenna impedance at the point of connection. The procedure is to take a trial point, apply power to the transmitter, and then check the transmission line for standing waves. This can be done by measuring the current in, or voltage along, the wires. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a

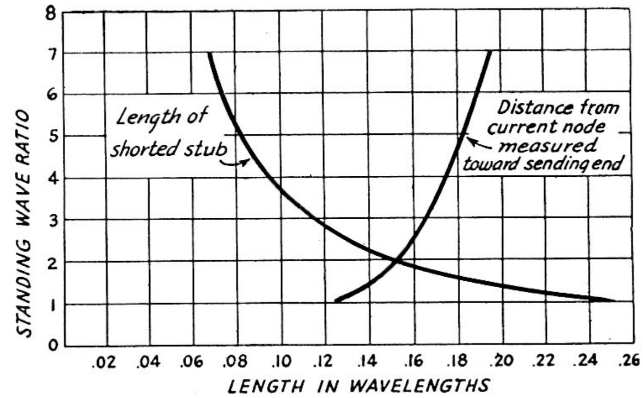


Fig. 10-74 — **Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.**

quarter wavelength will indicate whether or not standing waves are present.

It will not usually be possible to obtain complete elimination of standing waves when the matching stub is exactly resonant, but the line taps should be adjusted for the smallest obtainable standing-wave ratio. Then a further "touching up" of the matching-stub tuning will eliminate the remaining standing waves, provided the adjustments are carefully made. The stub must be readjusted, because when resonant it exhibits some reactance as well as resistance at all points except at the ends, and a slight lengthening or shortening of the stub is necessary to tune out this reactance.

Matching Stubs

The operation of the quarter-wave matching transformer of Fig. 10-72 may be considered from another — and more general — viewpoint. Suppose that section *C* is looked upon simply as a continuation of the transmission line. Then the "free" end of the transformer becomes a "stub" line, shunting a section of the main transmission line. From this viewpoint, matching the line to the antenna becomes a matter of selecting the right type and length of stub and attaching it to the proper spot along the line.

Referring to Fig. 10-73, at any distance (*X*) from the antenna, the line will have an imped-

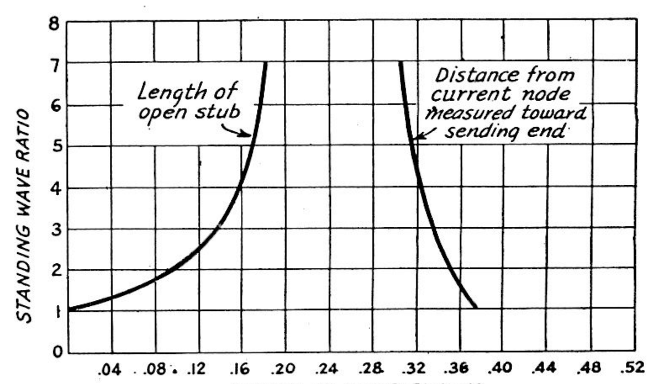


Fig. 10-75 — **Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.**

ance that may be considered to be made up of reactance (either inductive or capacitive) and resistance, in parallel. The reactive component can be eliminated by shunting the line at distance *X* from the antenna with another reactance equal in value but opposite in sign to the reactance presented by the line at that point. If distance *X* is such that the line presents an inductive reactance, a corresponding shunting capacitive reactance will be required.

The required compensating reactance may be supplied by shunting the line with a stub cut to proper length, *Y*. With the reactances canceled only a pure resistance remains as a termination for the remainder of the line between the sending end and the stub, and this resistance can be adjusted to match the characteristic impedance of the line by adjusting the distance *X*.

Distances *X* and *Y* may be determined experimentally, but since their values are interdependent the cut-and-try method is somewhat laborious. If the standing-wave ratio and the positions of the current loops and nodes can be measured, the length and position of the stub can be found from Figs. 10-74 and 10-75.

While it is relatively easy to locate the position of the current (or voltage) loops and nodes by examining the line with a neon bulb, r.f. galvanometer, or pick-up loop and crystal detector, other means are more direct for determining the standing-wave ratio. Several devices of this type are described in Chapter Sixteen, and the use of these also affords a simple method for determining the location of current loops (voltage nodes). With the meter or indicator in the line near the transmitter, points will be found on the transmission line where touching the line with a screwdriver will have a minimum effect on the meter indication. These points correspond to voltage nodes.

Once the standing-wave ratio is known, the length and position of the stub, in terms of wavelength, can be found directly from Figs. 10-74 and 10-75. The wavelength in feet for any frequency can be found from Equation 10-O.

Measuring Standing Waves

In adjusting a "Q-match" or linear transformer, or a delta or "T"-match to an antenna, one of the standing-wave indicators described in Chapter Sixteen should be used. If 300-ohm Twin-Lead is used, the simple "twin-lamp" indicator is the most convenient and the simplest to use. For lines of other impedance, or for coaxial line, the Micro-Match type or the bridge type should be used. In any event, the absolute value of standing-wave ratio is not as important as the proper adjustment for a minimum ratio, since ratios of 1.5-to-1 or less represent good amateur practice.

Where two-wire lines are used, the standing-wave-ratio indicator should give the same reading regardless of the polarity of the transmission line — any discrepancy indicates an unbalance in the line.

Directive Arrays with Parasitic Elements

Parasitic Excitation

The antenna arrays previously described are bidirectional; that is, they will radiate in directions both to the "front" and to the "back" of the antenna system. If radiation is wanted in only one direction, it is necessary to use different element arrangements. In most of these arrangements the additional elements receive power by induction or radiation from the driven element, generally called the "antenna," and reradiate it in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line. They are widely used to give additional gain and directivity to simple antennas.

The parasitic element is called a **director** when it reinforces radiation on a line pointing to it from the antenna, and a **reflector** when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic-element tuning (which usually is adjusted by changing its length) and, particularly when the element is self-resonant, upon the spacing between it and the antenna.

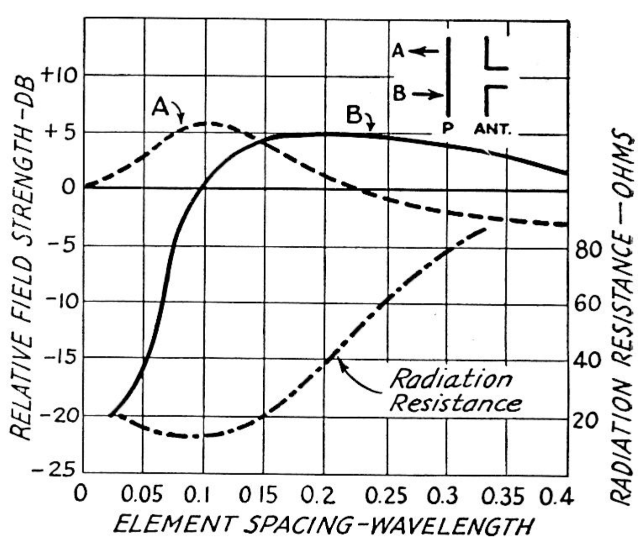


Fig. 10-76 — Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back ratio is the difference in db. between curves *A* and *B*. Variation in radiation resistance of the driven element also is shown. These curves are for a self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; the gain as a director can be increased by shortening. This also improves the front-to-back ratio.

Gain vs. Spacing

The gain of an antenna-reflector or an antenna-director combination varies chiefly with the spacing between the elements. The way in which gain varies with spacing is shown in Fig. 10-76, for the special case of self-resonant parasitic elements. This chart also shows how the attenuation to the "rear" varies with spacing. The same spacing does not necessarily give both maximum forward gain and maximum backward attenuation. Backward attenuation is desirable when the antenna is used for receiving, since it greatly reduces interference coming from the opposite direction to the desired signal.

Element Lengths

The antenna length is given by the formula for a half-wavelength antenna. The director and reflector lengths must be determined experimentally for maximum performance. The preferable method is to aim the antenna at a receiver a mile or more distant and have an observer check the signal strength (on the receiver S-meter) while the reflector or director is adjusted a few inches at a time, until the length which gives maximum signal is found. The attenuation may be similarly checked, the length being adjusted for minimum signal. In general, for best front-to-back ratio the length of a director will be about 4 per cent less than that of the antenna. The reflector will be about 5 per cent longer than the antenna.

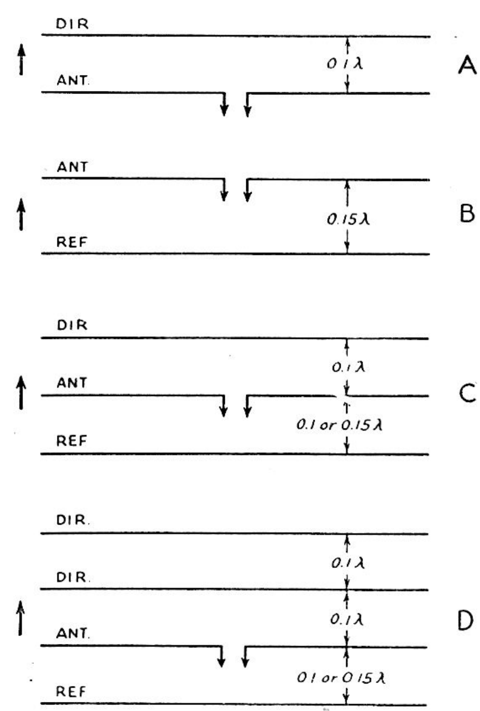


Fig. 10-77 — Half-wave antennas with parasitic elements. A, with director; B, with reflector; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 10-76, in the first two cases, and depends upon the spacing and length of the parasitic element. In the three- and four-element arrays a reflector spacing of 0.15 wavelength will give slightly more gain than 0.1-wavelength spacing. Arrows show the direction of maximum radiation.

Simple Systems: the Rotary Beam

Four practical combinations of antenna, reflector and director elements are shown in Fig. 10-77. Spacings which give maximum gain or maximum front-to-back ratio (ratio of power radiated in the desired direction to power radiated in the opposite direction) may be taken from Fig. 10-76. In the chart, the front-to-back ratio in db. will be the sum of gain and attenuation at the same spacing.

Systems of this type are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

Arrays using more than one parasitic element, such as those shown at C and D in Fig. 10-77, will give more gain and directivity than is indicated for a single reflector or director by the curves of Fig. 10-76. The gain with a properly-adjusted three-element array (antenna, director and reflector) will be 5 to 7 db. over a half-wave antenna. Somewhat higher gain still can be secured by adding a second director to the system, making a four-element array. The front-to-back ratio is correspondingly improved as the number of elements is increased.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic resistance but also has lower *Q;* both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a single half-wave dipole. With 3- and 4-element arrays the radiation resistance of the driven element may be as low as 6 or 8 ohms, so that ohmic losses in the conductor can consume an appreciable fraction of the power. Low radiation resistance means that the antenna will work over only a small frequency range without retuning unless large-diameter conductors are used. In addition, the antenna elements should be rigid because if they are free to move with respect to each other, the array will tend to show troublesome detuning effects under windy conditions.

Feeding Close-Spaced Arrays

While any of the usual methods of feed may be applied to the driven element of a parasitic array, the fact that, with close spacing, the radiation resistance as measured at the center of the driven element drops to a very low value makes some systems more desirable than others. The preferred methods are shown in Fig. 10-78. Resonant feeders are not recommended for lengths greater than a half-wavelength.

The quarter- or half-wave matching stubs shown at A and B in Fig. 10-78 preferably

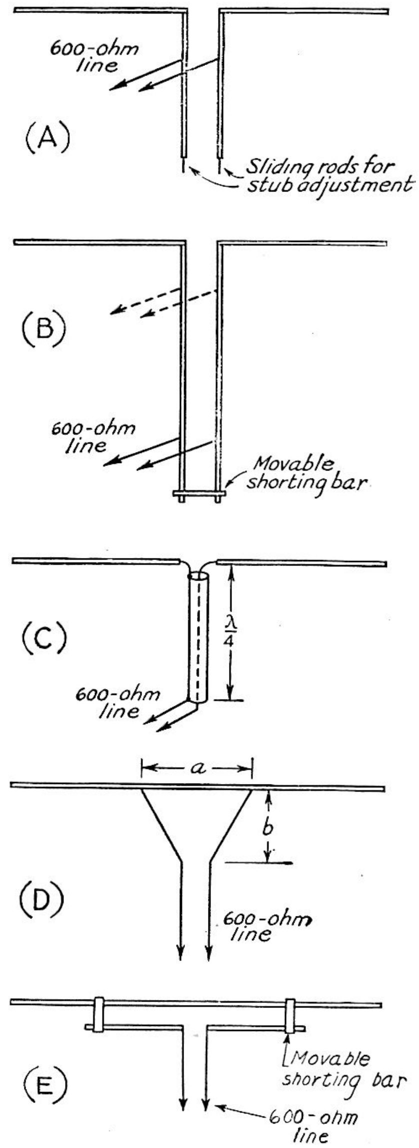


Fig. 10-78 — **Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer; E, "T" matching transformer. Adjustment details are discussed in the text.**

should be constructed of tubing with rather close spacing, in the manner of the "Q" section. This lowers the impedance of the matching section and makes the position of the line taps somewhat less difficult to determine accurately. The line adjustment should be made only with the parasitic elements in place, and after the correct element lengths have been determined it should be checked to compensate for changes likely to occur because of element tuning.

The concentric-line matching section at C will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as an impedance-inverting transformer, and, if its characteristic impedance is 70 ohms (RG-11/U), it will give a good match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2-to-1. The length of the quarter-wave section may be calculated from Equation 10-G.

The delta matching transformer shown at D is probably easier to install, mechanically, than any of the others. The positions of the taps (dimension a) must be determined experimentally, along with the length, b, by checking the standing-wave ratio on the line as adjustments are made. Dimension b should be about 15 per cent longer than a.

The system shown at E ("T"-match) resembles the delta match in principles of operation. It has the advantage that, with close spacing between the two parallel conductors, line radiation from the matching section is negligible whereas radiation from a delta may be considerable. It is adjusted by moving the shorting bars, keeping them equidistant from the center, until there are no standing waves on the line. The matching section may be made of the same type of conductor used for the driven element and spaced a few inches from it.

The "folded-dipole" type of antenna may be used as the driven element of a close-spaced parasitic array to secure an impedance step-up to the transmission line and also to broaden the resonance curve of the antenna. The folded dipole consists of two or more half-wave antennas connected together at the ends with the feeder connected to the center of only one of the antennas. The spacing between the parallel antennas should be small — of the order of the spacing used between wires of a transmission line. The current in the system divides in approximate proportion to the areas of the conductors, resulting in an impedance step-up at the input terminals. With two similar conductors (equal areas) the impedance step-up is 4-to-1; if there are three similar conductors (or if the one not connected to the transmission line has twice the diameter of the other) the step-up is 9-to-1; if the ratio of the areas is 3-to-1 the step-up is 16-to-1, and so on. Thus if a 3-conductor dipole (all conductors the same diameter) is used as the driven element of a four-element parasitic array the center impedance of approximately 8 ohms is multiplied by 9 and appears as approximately 72 ohms at the input terminals. Such a system therefore can be fed directly from a 70-ohm line with no additional means for matching.

Fig. 10-80 shows the impedance step-up obtained in a folded dipole when conductors of different sizes are used.

Sharpness of Resonance

Peak performance of a multielement parasitic array depends upon proper phasing or tuning of the elements, which can be exact for one frequency only. In the case of close-spaced arrays, which because of the low radiation resistance usually are quite sharp-tuning, the frequency range over which optimum results can be secured is only of the order of 1 or 2

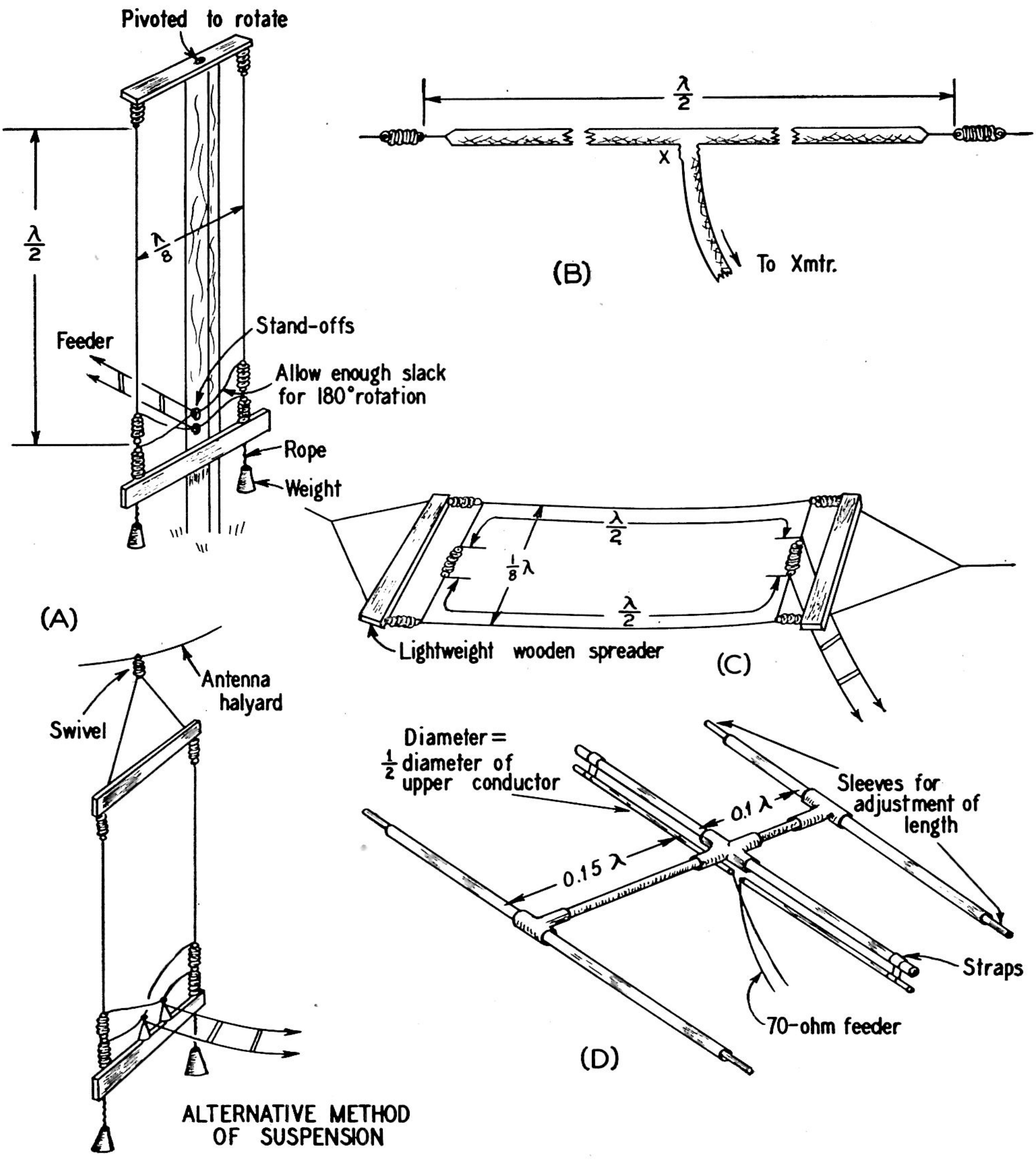
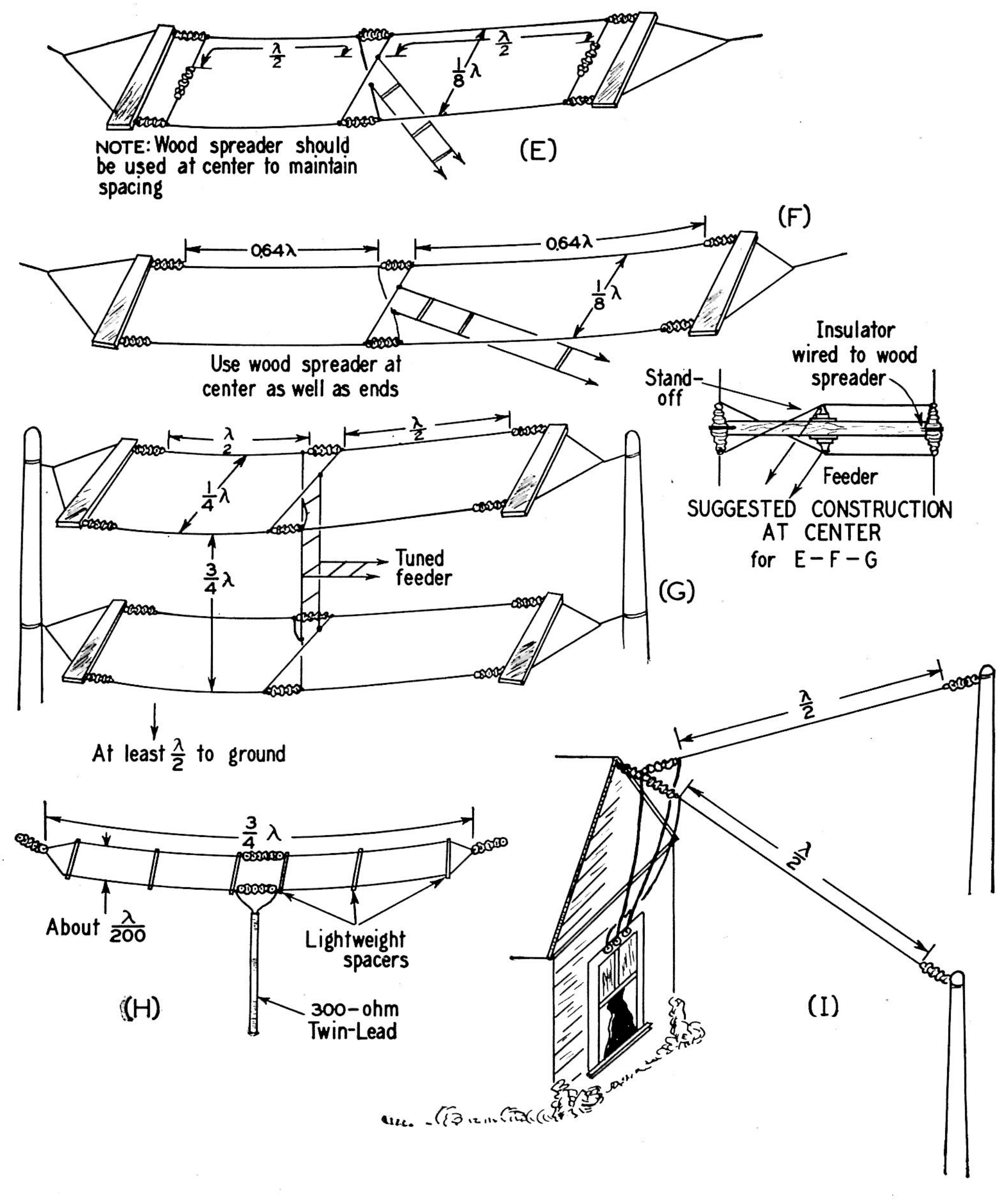


Fig. 10-79 — Some suggested antenna systems. A — Simple bidirectional rotatable end-fire array using ⅛-wave spacing between out-of-phase elements. It is suitable for either 14 or 28 Mc. and can be hand-rotated. It can also be suspended from the halyard holding another antenna, as suggested in the lower drawing. B — Folded dipole using 300-ohm Twin-Lead for both antenna and feeder. The junction *X* at the center is made by opening one conductor of the antenna section and soldering to the feeder leads. The joint may be made mechanically firm by heating the dielectric with a soldering iron, using extra bits of dielectric for a good bond. C — An end-fire array for use where space is limited. The ends of the two half-wave elements are folded to meet at an insulator in the center. The antenna may be made still shorter by increasing the spacing; spacings up to ¼ wavelength may be used. D — Pipe-assembly three-element beam ("plumber's delight") with folded-dipole driven element. Because all three elements are at the same r.f. potential at their centers it is possible to join them electrically as well as mechanically with no effect on the performance. Provision is made for adjusting the element lengths for optimum performance at a given frequency. E — An extension of the folding principle shown in C. The collinear in-phase elements give additional gain and directivity. F — End-



fire array with extended double-Zepps. This antenna should give a gain of about 7 db. in the direction perpendicular to the line of the antenna. G — An 8-element array combining broadside, end-fire and collinear elements. The gain of an antenna of this type is about 10 db. This antenna also can be used at half the frequency for which it is designed. H — A three-quarter wavelength folded antenna offers a fairly-close match for a 500- or 600-ohm open-wire line. Its pattern is quite similar to a half-wavelength antenna. Note that, unlike the half-wavelength folded dipole, the far side is *open* at the center. I — Using two half-wave antennas at right angles to change direction. With the three feeders indicated, either antenna alone can be fed as a Zepp and will radiate best perpendicular to its direction. By feeding the two together, leaving the third feeder wire idle, the optimum direction is the bisector of the angle between the wires. This system is most useful at high frequencies.

In these drawings, wavelength dimensions on conductors refer to lengths calculated for the conductor size as described in Equation 10-J. Dimensions between elements are free-space dimensions.

The feeders to the various directive systems in A, C, E, F and G must be tuned if used as shown. For one-band operation, matching stubs may be attached to the feeders if a matched line is desired.

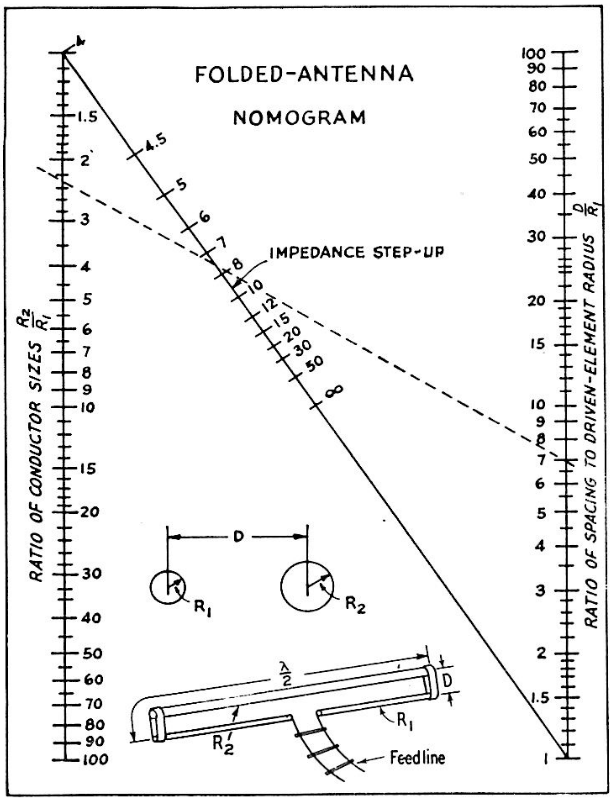


Fig. 10-80 — Nomogram for computing impedance step-up in a folded dipole with dissimilar conductors. The line at the left is the ratio of conductor diameters, and the line at the right is the ratio of conductor spacing (center-to-center) to the driven-element radius. The solid slanting line is the impedance step-up ratio. Laying a straightedge between any two known quantities will give the value of the third.

Example: Find the diameter of the large conductor when the driven-element diameter is 0.5 inch, line impedance 300 ohms, antenna impedance 40 ohms, and spacing 1.75 inches.

Impedance step-up required = 300/40 = 7.5

Spacing-to-element-radius ratio = 1.75/0.25 = 7

Laying a straightedge across the figure (dashed line), ratio of conductor diameters = 2.3

Diameter of large conductor = 2.3 × 0.5 = 1.15 inches

per cent of the resonant frequency, or up to about 500 kc. at 28 Mc. However, the antenna can be made to work satisfactorily over a wider frequency range by adjusting the director or directors to give maximum gain at the *highest* frequency to be covered, and by adjusting the reflector to give optimum gain at the *lowest* frequency. This sacrifices some gain at all frequencies, but maintains more uniform gain over a wider frequency range.

As mentioned in the preceding paragraphs, the use of large-diameter conductors will broaden the response curve of an array because the larger diameter lowers the Q. This causes the reactances of the elements to change rather slowly with frequency, with the result that the tuning stays near the optimum over a considerably-wider frequency range than is the case with wire conductors.

Combination Arrays

It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broadside-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array could be treated in the same fashion.

When combination arrays are built up, a rough approximation of the gain to be expected may be obtained by adding the gains for each type of combination. Thus the gain of two broadside sets of four collinear arrays with a set of reflectors, one behind each element, at quarter-wave spacing for the parasitic elements, would be estimated as follows: From Table 10-IV, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 10-66 or Table 10-V, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 10-76, the gain of a parasitic reflector at quarter-wave spacing is 4.5 db. The total gain is then the sum, or 13 db. for the sixteen elements. Note that using two *sets* of elements in broadside is equivalent to using two elements, so far as gain is concerned; similarly with sets of reflectors, as against one antenna and one reflector. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements and upon the effect which mutual coupling between elements has upon the radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive-antenna combinations can be worked out by combining elements according to these principles.

RECEIVING ANTENNAS

Nearly all of the properties possessed by an antenna as a radiator also apply when it is used for reception. Current and voltage distribution, impedance, resistance and directional characteristics are the same in a receiving antenna as if it were used as a transmitting antenna. This reciprocal behavior makes possible the design of a receiving antenna of optimum performance based on the same considerations that have been discussed for transmitting antennas.

The simplest receiving antenna is a wire of random length. The longer the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, a large antenna is not necessary for picking up signals at good strength. An indoor wire only 15 to 20 feet long will serve at frequencies below the v.h.f. range, although a longer wire outdoors is better.

The use of a tuned antenna improves the operation of the receiver, however, because the signal strength is raised more in proportion to the stray noises picked up than is the case with wires of random length. Since the transmitting antenna usually is given the best loca-

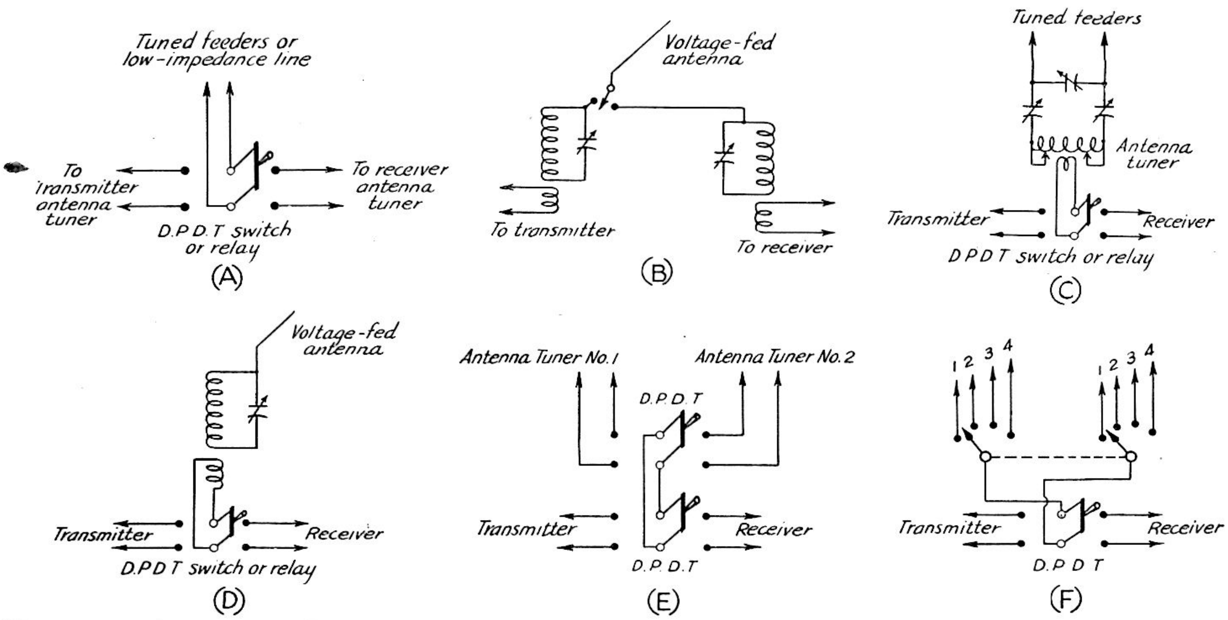


Fig. 10-81 — **Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. E — For two tuned-line antennas with a tuner for each antenna or for two low-impedance lines. F — For combinations of several two-wire lines.**

tion, it can also be expected to serve best for receiving. This is especially true when a directive antenna is used, since the directional effects and power gain of directive transmitting antennas are the same for receiving as for transmitting.

In selecting a directional receiving antenna it is preferable to choose a type that gives very little response in all but the desired direction (small minor lobes). This is even more important than high gain in the desired direction, because the cumulative response to noise and unwanted-signal interference in the smaller lobes may offset the advantage of increased desired-signal gain. The feedline from the antenna should be balanced so that it will not pick up signals and destroy the directivity.

Antenna Switching

Switching of the antenna from receiver to transmitter is commonly done with a change-over relay, connected in the antenna leads or the coupling link from the antenna tuner. If the relay is one with a 115-volt a.c. coil, the switch or relay that controls the transmitter plate power will also control the antenna relay. If the convenience of a relay is not desired, porcelain knife switches can be used and thrown by hand.

Typical arrangements are shown in Fig. 10-81. If coaxial line is used, the use of a coaxial relay is recommended, although on the lower-frequency bands a regular switch or change-over relay will work almost as well.

Antenna Construction

The use of good materials in the antenna system is important since the antenna is exposed to wind and weather. To keep electrical losses low, the wires in the antenna and feeder system must have good conductivity and the insulators must have low dielectric loss and surface leakage, particularly when wet.

For short antennas, No. 14 gauge hard-drawn enameled copper wire is a satisfactory conductor. For long antennas and directive arrays, No. 14 or No. 12 enameled copper-clad steel wire should be used. It is best to make feeders and matching stubs of ordinary soft-drawn No. 14 or No. 12 enameled copper wire, since hard-drawn or copper-clad steel wire is difficult to handle unless it is under considerable tension at all times. The wires should be all in one piece; where a joint cannot be avoided, it should be carefully soldered.

In building a resonant two-wire feeder, the spacer insulation should be of as good quality as in the antenna insulators proper. For this reason, good ceramic spacers are advisable. Wooden dowels boiled in paraffin may be used with untuned lines, but their use is not recommended for tuned lines. The wooden dowels can be attached to the feeder wires by drilling small holes and binding them to the feeders with wire.

At points of maximum voltage, insulation is most important, and Pyrex glass, Isolantite or steatite insulators with long leakage paths are recommended for the antenna. Glazed porcelain also is satisfactory. Insulators should be cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung

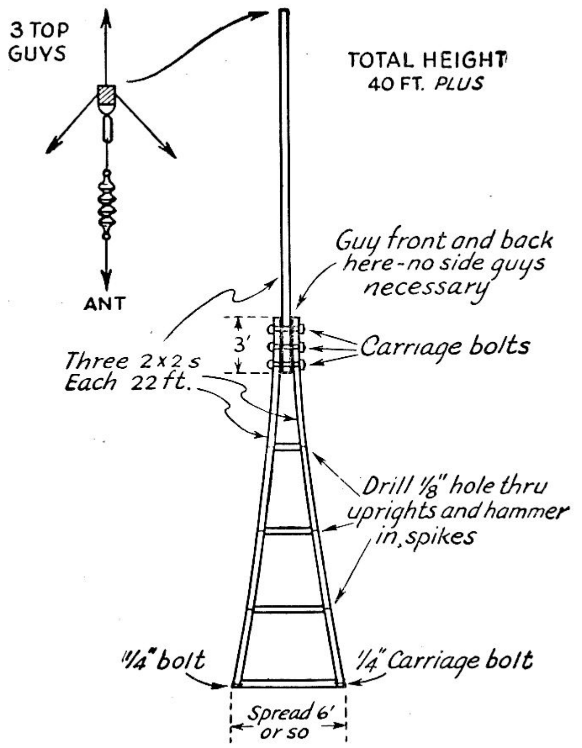


Fig. 10-82 — **Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.**

from one chimney to another or from a chimney to a tree. Small trees usually are not satisfactory as points of suspension for the antenna because of their movement in windy weather. If the antenna is strung from a point near the center of the trunk of a large tree, this difficulty is not so serious. Where the antenna wire must be strung from one of the smaller branches, it is best to tie a pulley firmly to the branch and run a rope through the pulley to the antenna, with the other end of the rope attached to a counterweight near the ground. The counterweight will keep the tension on the antenna wire reasonably constant even when the branches sway or the rope tightens and stretches with varying climatic conditions.

"A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 10-82 is satisfactory for heights up to 35 or 40 feet. Clear, sound lumber should be selected. The completed mast may be protected by two or three coats of house paint.

If the mast is to be erected on the ground, a couple of stakes should be driven to keep the bottom from slipping and it may then be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation — lifting the mast, carrying it to its permanent berth and fastening the guys — with the mast vertical all the while. It is entirely practicable, therefore, to erect this type of mast on any small, flat area of roof.

By using 2 × 3s or 2 × 4s, the height may be extended up to about 50 feet. The 2 × 2 is too flexible to be satisfactory at such heights.

SIMPLE 40-FOOT MAST

The mast shown in Fig. 10-83 is relatively strong, easy to construct, readily dismantled, and costs very little. Like the "A"-frame, it is suitable for heights of the order of 40 feet.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about two feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the bottom the two legs are bolted to a length of 2 × 4 which is set in the ground. A short length of 2 × 3 is placed between the two legs about halfway up the bottom section, to maintain the spacing.

The two back guys at the top pull against the antenna, while the three lower guys prevent buckling at the center of the pole.

The 2 × 4 section should be set in the ground so that it faces the proper direction, and then made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. With the lower section laid on the ground, bolt *A* should be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section may be bolted in place and the mast pushed up, using a ladder or another 20-foot 2 × 3 for the job. As the mast goes up, the slack in the guys can be taken up so that the whole structure is in some meas-

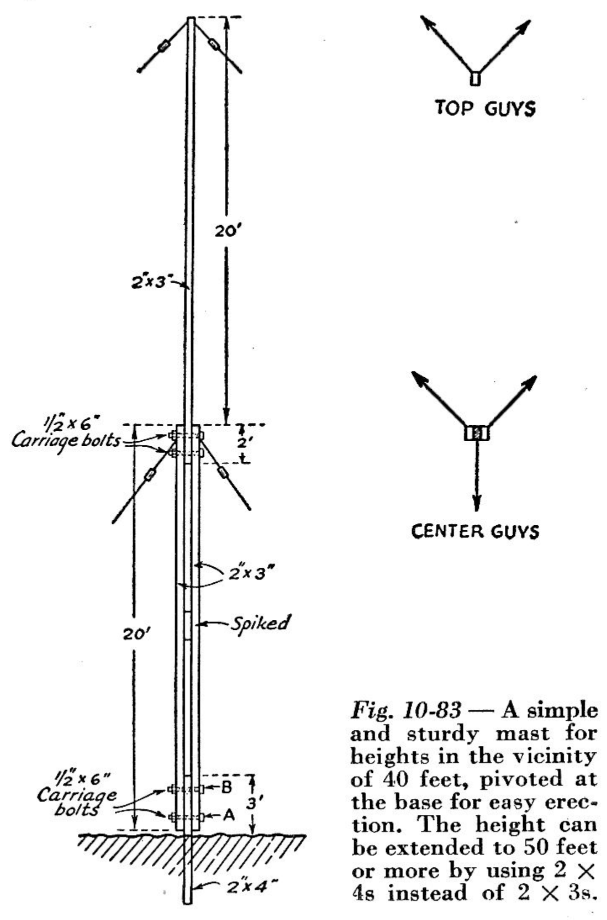


Fig. 10-83 — **A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.**

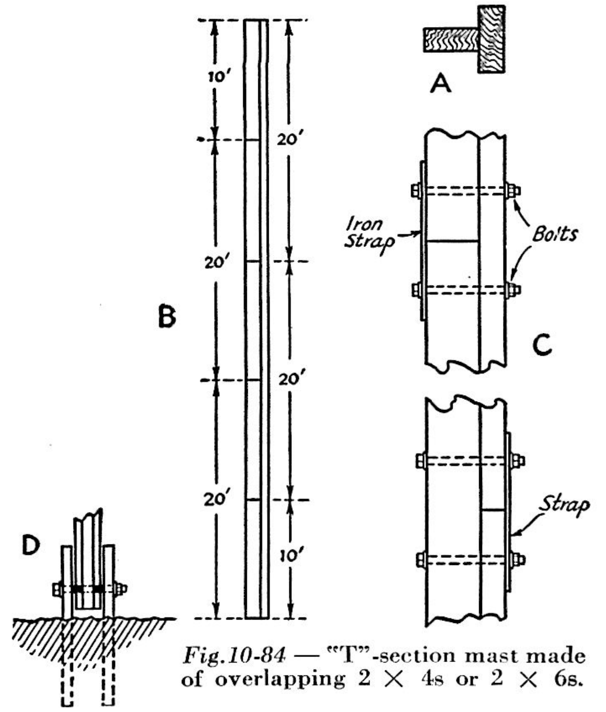


Fig. 10-84 — "T"-section mast made of overlapping 2 × 4s or 2 × 6s.

ure continually supported. When the mast is vertical, bolt *B* should be slipped in place and both *A* and *B* tightened. The lower guys can then be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they should be adjusted to pull the top section into line.

"T"-SECTION MAST

A type of mast suitable for heights up to about 80 feet is shown in Fig. 10-84. The mast is built up by butting 2 × 4 or 2 × 6 timbers flatwise against a second 2 × 4, as shown at A, with alternating joints in the edgewise and flatwise sections. The construction can be carried out to greater lengths simply by continuing the 20-foot sections. Longer or shorter sections may be used.

The method of making the joints is shown at C. Quarter-inch or 3/16-inch iron, 1½ to 2 inches wide, is recommended for the straps, with ½-inch bolts to hold the pieces together. One bolt should be run through the pieces midway between joints, to provide additional rigidity.

Although there are many ways in which such a mast can be secured at the base, the "cradle" illustrated at D has many advantages. Heavy timbers set firmly in the ground, spaced far enough apart so the base of the mast will pass between them, hold a large carriage bolt or steel bar which serves as a bearing. The bolt goes through a hole in the mast so that it is pivoted at the bottom.

Half of the guys can be put in place and tightened up before the mast leaves the ground. Four sets of guys should be used, *one in front*, one directly in the rear, and two on each side at right angles to the direction in which the mast will face. A set of guys should be used at each of the joints in the edgewise sections, the guy wires being wrapped around the pole for added strength.

For heights up to 50 feet, 2 × 4-inch members may be used throughout. For greater heights, use 2 × 4s for the edgewise sections; 2 × 6-inch pieces will do for the flat sections.

POLE AND TOWER SUPPORTS

Poles, which often may be purchased at a reasonable price from the local telephone or power company, have the advantage that they do not require guying unless they are called upon to carry a very heavy load. The life of a pole can be extended many years by proper precautions before erecting, and regular maintenance thereafter.

Before setting the pole, it should be given four or five coats of creosote, applying it liberally so it can soak into and preserve the wood. The bottom of the pole and the part that will be buried in the ground should have a generous coating of hot pitch, poured on while the pole is warm. This will keep termites out and prevent rotting.

The pole should be set in the ground four to eight feet depending upon the height. It is a good idea to pour concrete around the bottom three feet of the base, packing the rest of the excavation with soil. The concrete will help hold the pole against strong winds. After filling the hole with dirt, a stream from a hose should be played on the dirt slowly for several hours.

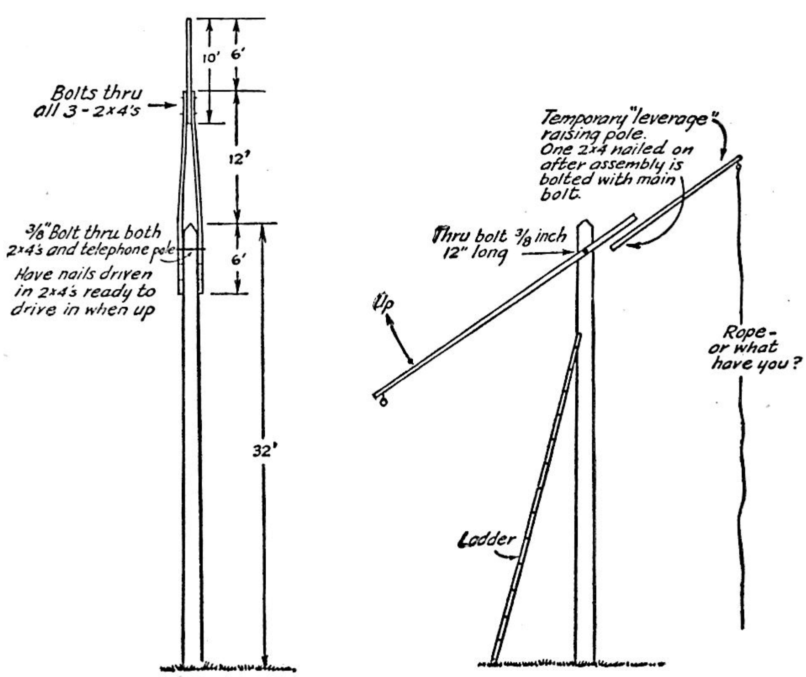


Fig. 10-85 — This type of mast may be carried to a height of fifty feet or more. No guy wires are required.

This will help to settle the soil quickly.

If desired, the pole may be extended by the arrangement shown in Fig. 10-85. Three 2 × 4s are required for the top section, two being 18 feet long and one 10 feet long. The 10-foot section is placed between the other two and bolted in place. A half-inch hole should be bored through the pole about 2 feet from its top and through both 18-foot 2 × 4s about 5 feet from their bottom ends, which are spread apart to fit the top of the pole. The bottom end of the extension is then hauled up to the top of the pole and bolted loosely so that the section can be swung up into place by the leverage of another 2 × 4 temporarily fastened to the section, as shown in Fig. 10-85.

Lattice towers built of wood should be assembled with brass screws and casein glue, rather than with nails which work loose in a short time. A tower constructed in this manner will give trouble-free service if treated with a coat of paint every year.

In painting outside structures, use pure white lead, thinned with three parts of pure linseed oil to one part of turpentine, for the first coat on new wood. The use of a drier is not recommended if the paint will possibly dry without it, since it may cause the paint to peel after a short time. For the second and third coats pure white lead thinned only with pure

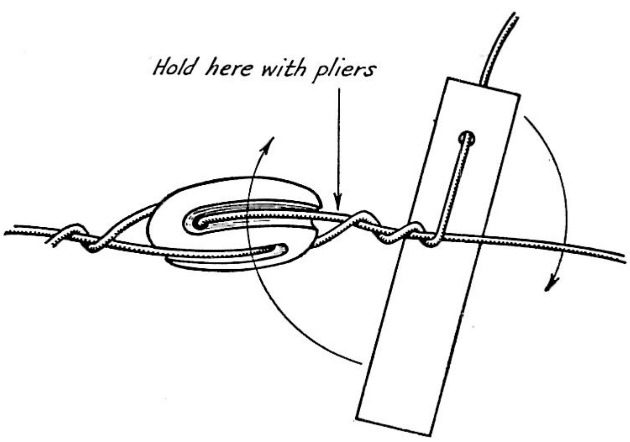


Fig. 10-86 — **Using a lever for twisting heavy guy wires.**

linseed oil is recommended. Plenty of time for drying should be allowed between coats. White paint will last fifty per cent longer than any colored paint.

GUYS AND GUY ANCHORS

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material. Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

More than three guy wires in any one set usually are unnecessary. If a horizontal antenna is to be supported, two guy wires in the top set will be sufficient in most cases. These should run to the rear of the mast about 100 degrees apart to offset the pull of the antenna.

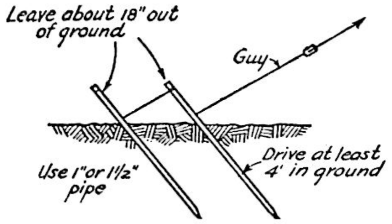


Fig. 10-87 — **Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.**

Intermediate guys should be used in sets of three, one running in a direction opposite to that of the antenna, while the other two are spaced 120 degrees either side. This leaves a clear space under the antenna. The guy wires should be adjusted to pull the pole slightly back from vertical before the antenna is hoisted so that when the antenna is pulled up tight the mast will be straight.

When raising a mast that is big enough to tax the facilities available, it is some advantage to know nearly exactly the length of the guys. Those on the side on which the pole is lying can then be fastened temporarily to the anchors beforehand, which assures that when the pole is raised, those holding opposite guys will be able to pull it into nearly-vertical position with no danger of its getting out of control. The guy lengths can be figured by the right-angled-triangle rule that "the sum of the squares of the two sides is equal to the square of the hypotenuse." In other words, the distance from the base of the pole to the anchor should be measured and squared. To this should be added the square of the pole length to the point where the guy is fastened. The square root of this sum will be the length of the guy.

Guy wires should be broken up by strain insulators, to avoid the possibility of resonance at the transmitting frequency. Common practice is to insert an insulator near the top of each guy, within a few feet of the pole, and then cut each section of wire between the insulators to a length which will not be resonant either on the fundamental or harmonics. An insulator every 25 feet will be satisfactory for frequencies up to 30 Mc. The insulators should be of the "egg" type with the insulating material under compression, so that the guy will not part if the insulator breaks.

Twisting guy wires onto "egg" insulators may be a tedious job if the guy wires are long and of large gauge. The simple time- and finger-saving device shown in Fig. 10-86 can be made from a piece of heavy iron or steel by drilling a hole about twice the diameter of the guy wire about a half inch from one end of the piece. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in the sketch. By passing the wire through the hole in the iron and rotating the iron as shown, the wire may be quickly and neatly twisted.

Guy wires may be anchored to a tree or building when they happen to be in convenient spots. For small poles, a 6-foot length of 1-inch pipe driven into the ground at an angle will suffice. Additional bracing will be provided by using two pipes, as shown in Fig. 10-87.

HALYARDS AND PULLEYS

Halyards or ropes and pulleys are important items in the antenna-supporting system. Particular attention should be directed toward the choice of a pulley and halyards for a high mast since replacement, once the mast is in position, may be a major undertaking if not entirely impossible.

Galvanized-iron pulleys will have a life of only a year or so. Especially for coastal-area installations, marine-type pulleys with hardwood blocks and bronze wheels and bearings should be used.

An arrangement that has certain advantages over a pulley when a mast is used is shown in Fig. 10-88. In case the rope breaks, it may be possible to replace it by heaving a line over the brass rod, making it unnecessary to climb or lower the pole.

For short antennas and temporary installations, heavy clothesline or window-sash cord may be used. However, for more permanent jobs, ⅜-inch or ½-inch waterproof hemp rope should be used. Even this should be replaced about once a year to insure against breakage.

Nylon rope, used during the war as glider tow rope, is, of course, one of the best materials for halyards, since it is weatherproof and has extremely long life.

It is advisable to carry the pulley rope back up to the top in "endless" fashion in the manner of a flag hoist so that if the antenna breaks close to the pole, there will be a means for pulling the hoisting rope back down.

BRINGING THE ANTENNA OR FEEDLINE INTO THE STATION

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 10-89, 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

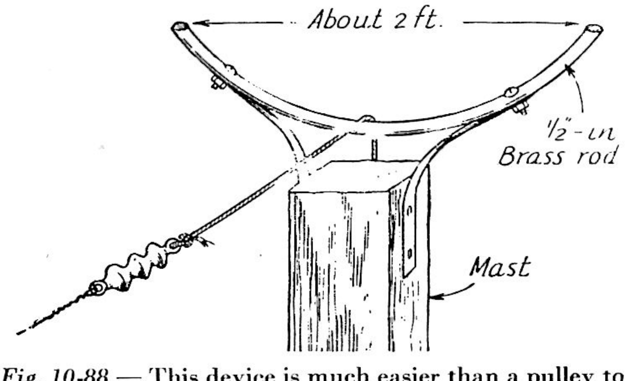


Fig. 10-88 — This device is much easier than a pulley to "rethread" when the rope breaks.

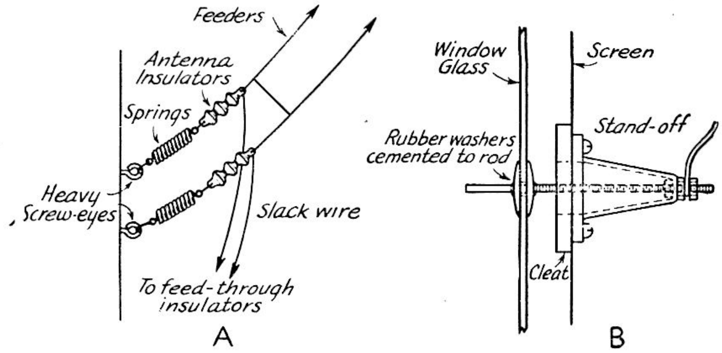


Fig. 10-89 — A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

bringing the line into the station. The holes should have plenty of air clearance about the conducting rod, especially when using tuned lines that develop high voltages. Probably the best place to go through the walls is the trimming board at the top or bottom of a window frame which provides flat surfaces for lead-in insulators. Either cement or rubber

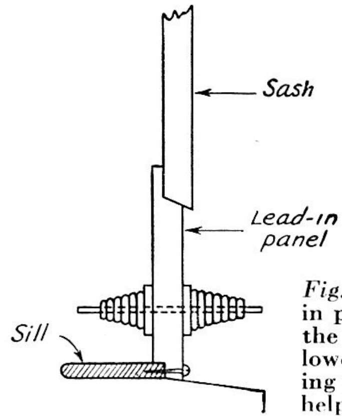


Fig. 10-90 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Sealing the overlapping joint will help make it weatherproof.

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. 10-89B may be used.

As a less permanent method, the window may be raised from the bottom or lowered from the top to permit insertion of a board which carries the feed-through insulators. This lead-in arrangement can be made weatherproof by making an overlapping joint between the board and window sash, as shown in Fig. 10-90, and

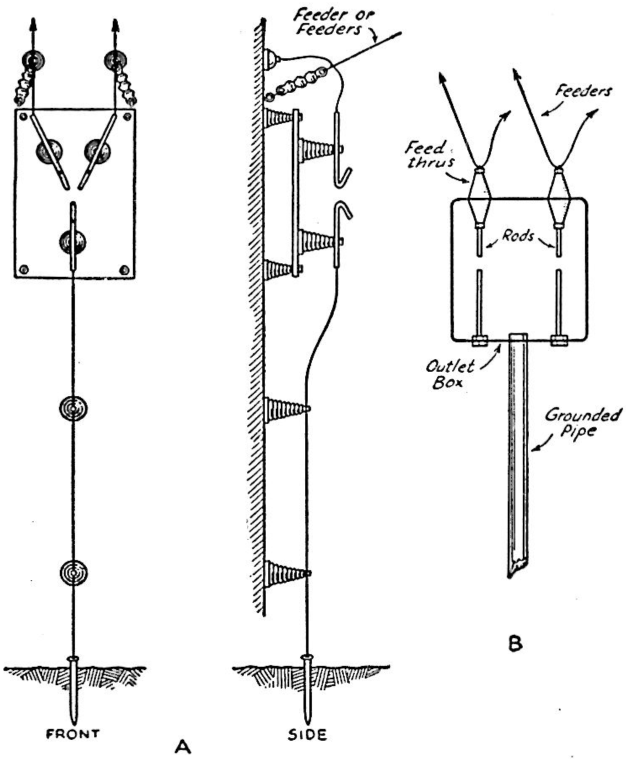


Fig. 10-91 — Low-loss lightning arresters for transmitting-antenna installations.

covering the opening between sashes with a sheet of soft rubber from a discarded inner tube.

LIGHTNING PROTECTION

An ungrounded radio antenna, particularly if large and well elevated, is a lightning hazard. When grounded, it provides a measure of protection. Therefore, grounding switches or lightning arresters should be provided. Examples of construction of low-loss arresters are shown in Fig. 10-91. At A, the arrester electrodes are mounted by means of stand-off insulators on a fireproof asbestos board. At B, the electrodes are enclosed in a standard steel outlet box. The gaps should be made as small as possible without danger of breakdown during operation. Lightning-arrester systems require the best ground connection obtainable.

The most positive protection is to ground the antenna system when it is not in use; grounded flexible wires provided with clips for connection to the feeder wires may be used. The ground lead should be short and run, if possible, directly to a driven pipe or water pipe where it enters the ground outside the building.

Rotary-Beam Construction

It is a distinct advantage to be able to shift the direction of a beam antenna at will, thus securing the benefits of power gain and directivity in any desired compass direction. A favorite method of doing this is to construct the antenna so that it can be rotated in the horizontal plane. Obviously, the use of such rotatable antennas is limited to the higher frequencies — 14 Mc. and above — and to the simpler-antenna element combinations if the structure size is to be kept within practicable bounds. For the 14- and 28-Mc. bands such antennas usually consist of two to four elements and are of the parasitic-array type described earlier in this chapter. At 50 Mc. and higher it becomes possible to use more elaborate arrays because of the shorter wavelength and thus obtain still higher gain. Antennas for these bands are described in Chapter Fourteen.

The problems in rotary-beam construction are those of providing a suitable mechanical support for the antenna elements, furnishing a means of rotation, and attaching the transmission line so that it does not interfere with the rotation of the system.

Elements

The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the supporting structure. The large diameter of the conductor is beneficial also in reducing resistance, which becomes an important consideration when close-spaced elements are used.

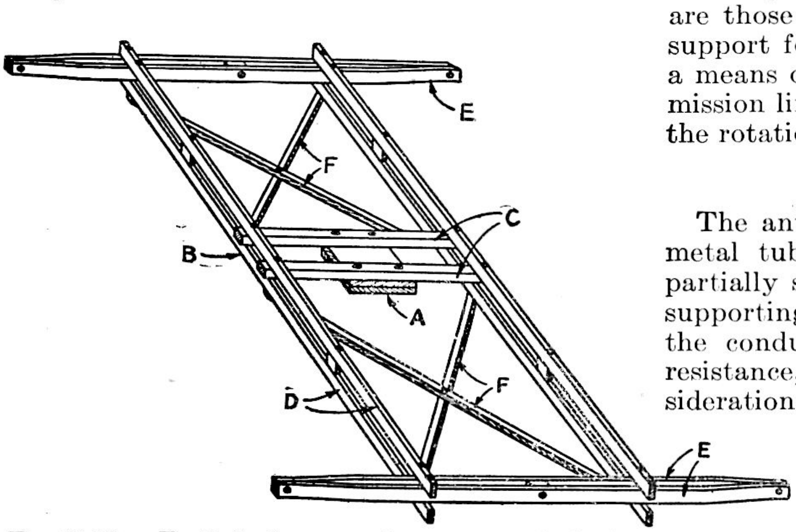


Fig. 10-92 — **Easily-built supporting structure for horizontal rotary beams. Made chiefly of 1 × 2″ wood strip, it is strong yet lightweight. Antenna elements are supported on stand-off insulators on the arms, *E*. The length of the *D* sections will depend upon the element spacing, while the length of the *E* sections and the spacing between the *D* sections should be ¼ to ½ the length of the antenna elements.**

Dural tubes often are used for the elements, and thin-walled corrugated steel tubes with copper coating also are available for this purpose. The elements frequently are constructed of sections of telescoping tubing, making length adjustments for tuning quite easy. Electricians' thin-walled conduit also is suitable for rotary-beam elements.

If steel elements are used, special precautions

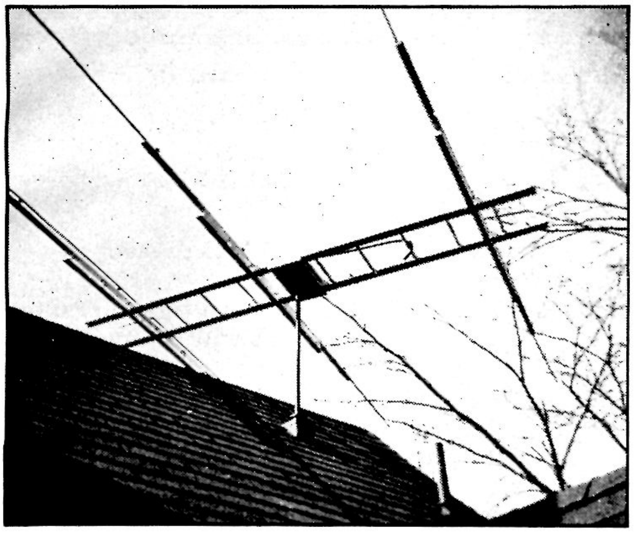


Fig. 10-93 — A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)

should be taken to prevent rusting. Even copper-coated steel does not stand up indefinitely, since the coating usually is too thin. The elements should be coated both inside and out with slow-drying aluminum paint. For coating the inside, a spray gun may be used, or the paint may be poured in one end while rotating the tubing. The excess paint may be caught as it comes out the bottom end and poured through again until it is certain that the entire inside wall has been covered. The ends should then be plugged up with corks sealed with glyptal varnish.

Supports

The supporting framework for a rotary beam usually is made of wood but sometimes of metal, using as lightweight construction as is consistent with the required strength. Generally, the frame is not required to hold much weight, but it must be extensive enough so that the antenna elements can be supported near enough to their ends to prevent excessive sag, and it must have sufficient strength to stand up under the maximum wind in the locality. The design of the frame will depend chiefly on the size of the antenna elements, whether they are mounted horizontally or vertically, and the method to be employed for rotating the antenna.

The general preference is for horizontal polarization, primarily because less height is required to clear surrounding obstructions when all the antenna elements are in the horizontal plane. This is important at 14 and 28 Mc. where the elements are fairly long.

An easily-constructed supporting frame for a horizontal array is shown in Fig. 10-92. It may be made of 1 × 2-inch lumber, preferably oak, for the center sections *B*, *C*, and *D*. The outer arms, *E*, and cross braces, *F*, may be of white pine or cypress. The square block, *A*, at the center supports the whole structure and may be coupled to the pole by any convenient means which permits rotation. Alternatively, the block may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

Another type of construction is shown in Fig. 10-93, with details in Figs. 10-94 and 10-95. This method, suitable for 28-Mc. beams, uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements. Fig. 10-94 also indicates a method of adjusting the lengths of the parasitic elements and bringing the transmission line down through the supporting pole from a delta match. The latter is especially adapted to construction in which the pole rather than the framework alone is rotated.

Metal Booms

Metal can be used to support the elements of the rotary beam. For 28 Mc., a piece of 2-inch diameter duraluminum tubing makes a good "boom" for supporting the elements. The elements can be made to slide through suitable holes in the boom, or special clamps and brackets can be fashioned to support the elements. The antenna of Fig. 10-79D shows one example of such construction.

Generally it is not practicable to support the elements of a 14-Mc. beam by a single-piece boom, because the size of the elements requires a stronger structure. However, by making use of tubing or duraluminum angle, a lightweight support for a 20-meter antenna can be built. The four-element beam shown in

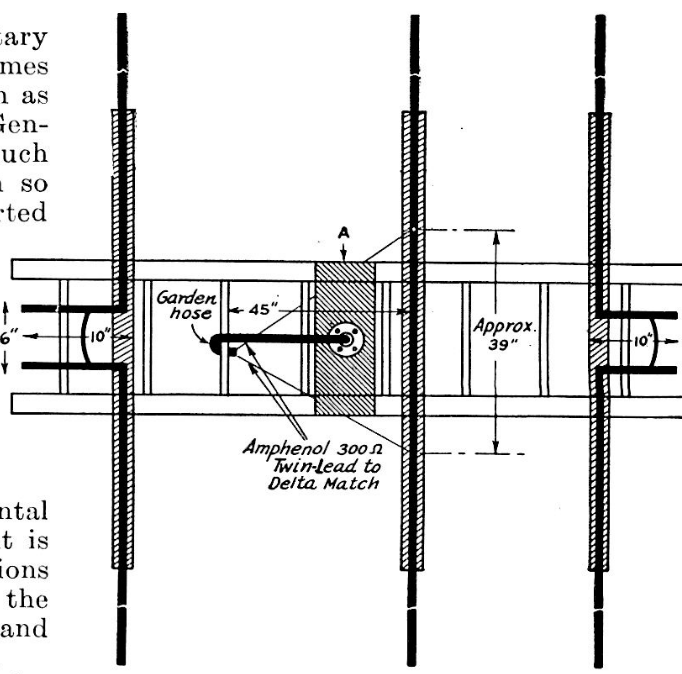


Fig. 10-94 — Top-view drawing of the ladder support and mounted elements. Lengths of director and reflector are adjusted by means of the shorting bars on the small stubs at the center. The drawing also shows a method for pulling off the wires of a delta match and feeding 300-ohm Twin-Lead transmission line through the pipe support.

Figs. 10-96, 10-97 and 10-98 is an example. It uses 1¾-inch angle for the main pieces and ¾-inch angle for the other members, and the entire framework plus elements weighs only forty pounds. This simplifies considerably the problem of supporting the beam.

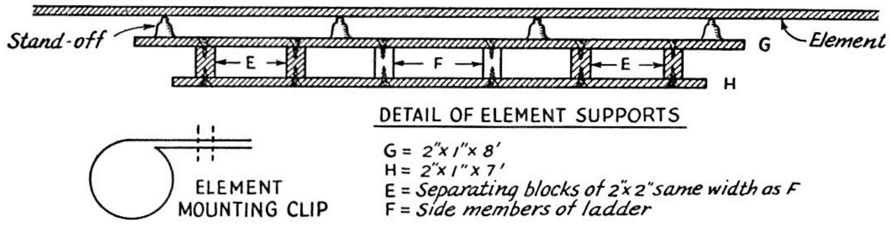


Fig. 10-95 — Detail of element supports for the ladder beam.

The following aluminum pieces are required:

4 — 1-inch diameter tubing, 12 feet long, 1/16-inch wall
8 — ⅞-inch diameter tubing, 12 feet long, 1/32-inch wall. Must fit snugly into 1-inch tubing.
2 — 1¾-inch angle, 21 feet long
2 — ¾-inch angle, 21 feet long
4 — ¾-inch angle, 1 foot long
2 — ½-inch diameter tubing, 6 feet long

Aluminum tubing and angle corresponding to the above sizes can possibly be bought from scrap dealers at reasonable prices, if not directly from the manufacturer. If the sections of the elements do not fit snugly, insert shims or make some other provision for a tight fit, since the appearance of the beam will be spoiled by sagging elements. Some amateurs reinforce their beam elements with copper-clad steel wire supported a foot above the elements at the boom and tied to the extreme ends of the elements.

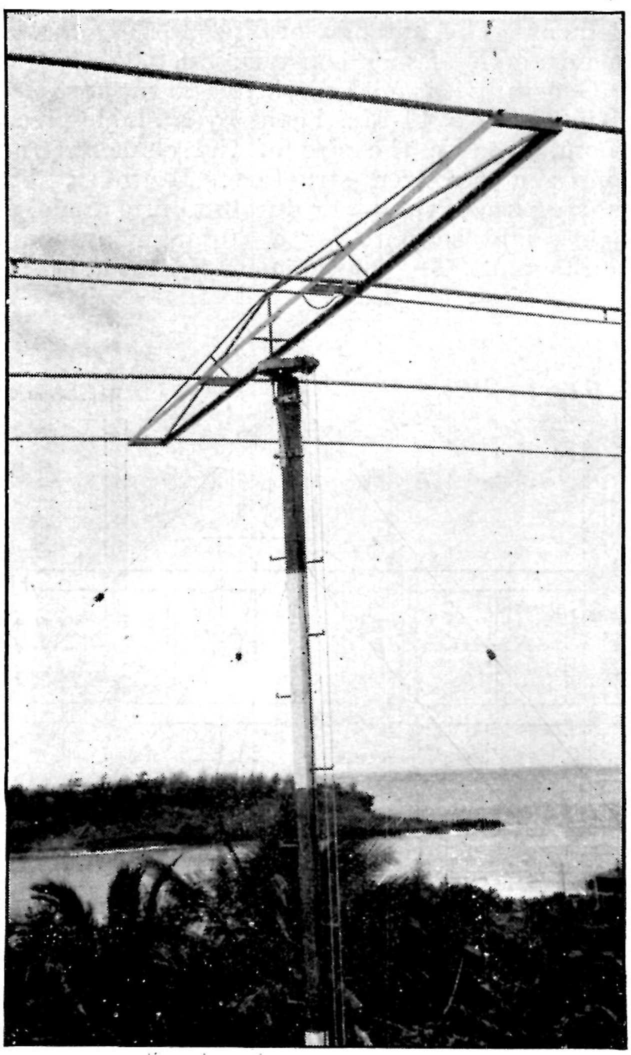


Fig. 10-96 — A four-element 14-Mc. beam of light-weight all-metal construction. Fed by coaxial cable and hand-rotated, the antenna and boom assembly weighs only 40 pounds (KH6IJ, Dec., 1947, *QST*.)

As shown in Fig. 10-97A, two 1¾-inch aluminum angles 21 feet long serve as the main members of the boom. They are spaced one foot apart. The elements are spaced 7 feet apart. Wooden spacers of 2 × 2 are placed at the end of the boom and screwed on with brass screws. These spacers are also placed under each element where it crosses the boom. These spacers may be unnecessary if the elements are bolted to the boom, but if the construction is as in Fig. 10-97B the spacers are recommended.

The cross braces shown in Fig. 10-98 are put into position at the very last, after the beam is hung in position on the central pivot, since they offer a means for truing up minor sag in the elements.

The central pivot consists of structure made from ¾-inch angle iron and ½-inch pipe, as shown in Fig. 10-97C. It has to be brazed. The crossbar rest is made separate from the boom and central pivot, and affords a means for tilting the beam when unbolted from these structures. The ½-inch pipe is drilled for the coaxial line that is fed through this pipe. The pinion gear on the ½-inch pipe should be brazed on.

A washing-machine gear train is well suited for this type of beam. Another possibility (used in this instance) is a discarded forge blower. It was fitted with a ½-inch pipe which serves as the central pivot. The gear train ends up in a "V"-pulley, and the beam is easily rotated by a system of ropes and pulleys that ends up in an automobile steering wheel at the operating position. A plumb bob attached to the shaft of the steering wheel serves as a direction indicator. A small cardboard scale mounted along the line of plumb-bob travel can be readily calibrated to show the direction of the beam.

The supporting structure for this beam consists of a 4 × 4 pole 30 feet long, with ten-foot extensions of 2 × 4 bolted to both sides of the bottom, making the total length about 36 feet. Two sets of guy wires should be used, approximately 2 feet and 15 feet from the top. As an alternative, the pole can be set against the side of the house, and only the top set of guys used to provide additional support.

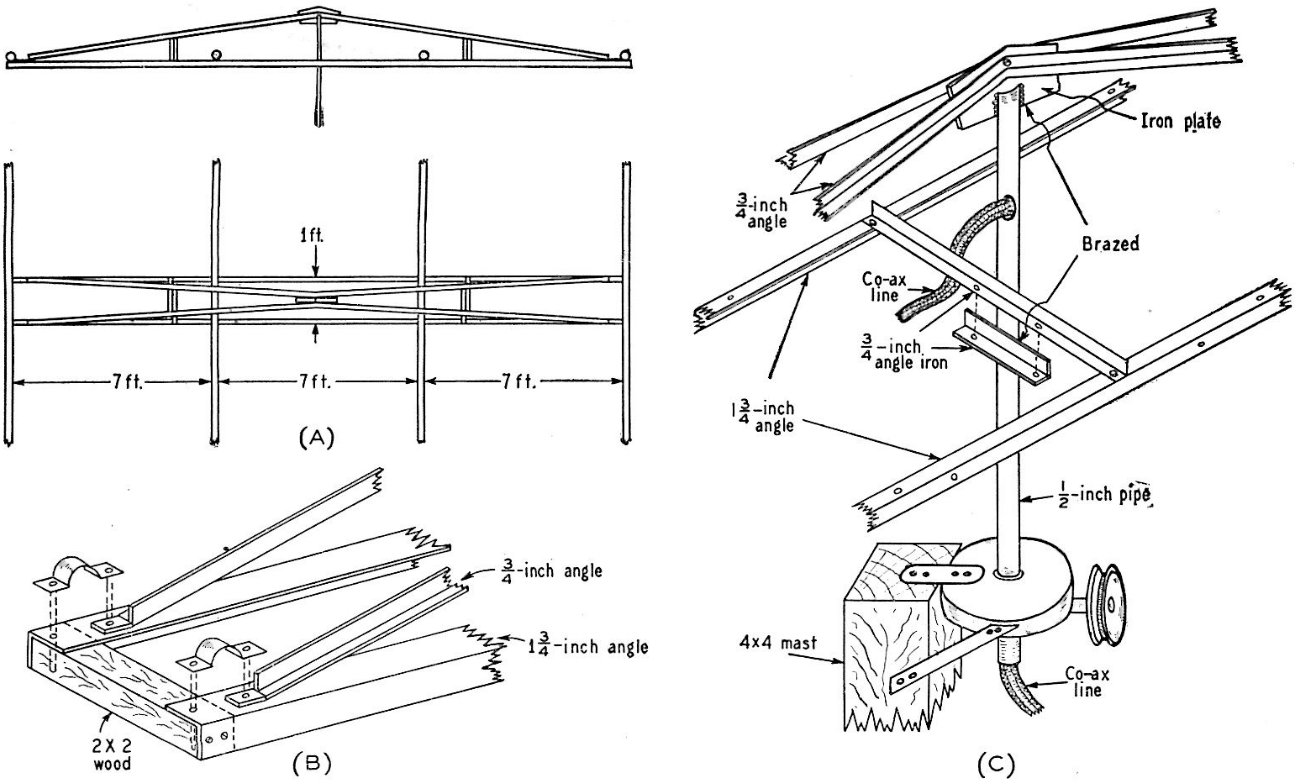


Fig. 10-97 — Details of the 4-element beam construction. The general dimensions and arrangement of the beam are given in A, the detail of the ends of the boom is shown at B, and C shows the construction of the central pivot. A discarded-forge blower gear train is used to drive the assembly.

With all-metal construction, delta match or "T"-match 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.

A Wooden Boom for 14 Mc.

Many amateurs prefer to build their beam booms from standard pieces of lumber, and the beam shown in Figs. 10-99 and 10-100 is an example of excellent design in wooden-boom construction. The boom members are two 20-foot 2 × 4s fastened to the 4 × 12 × 24-inch center block with six lag screws. The two center screws serve as the axis for tilting — the other four lock the boom in position after final assembly and adjustment have been completed. The blocks midway from each end are 2 × 4s spaced about six inches apart, with a long bolt between them. When this bolt is drawn tight, a very sturdy box brace is formed.

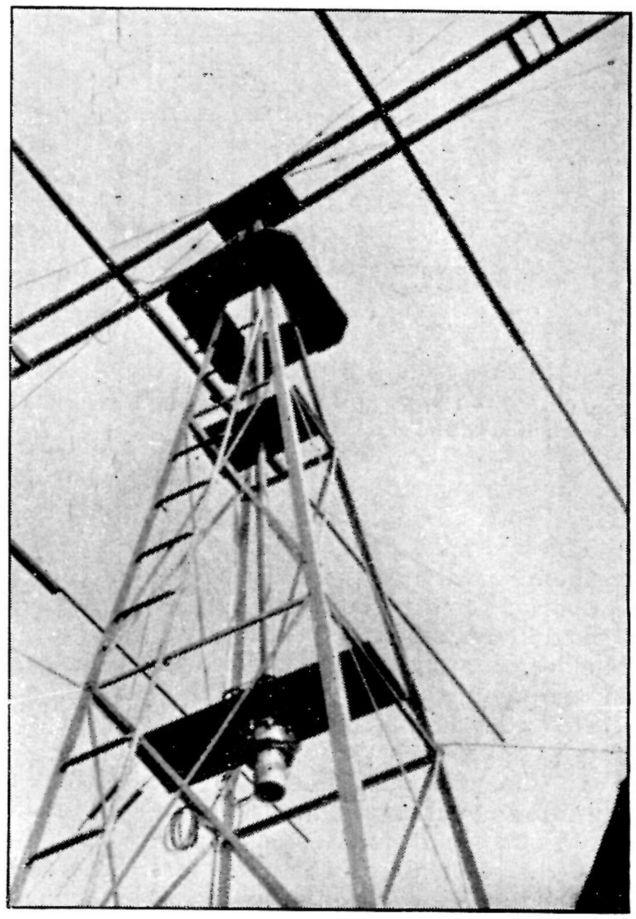


Fig. 10-99 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. This installation is made on a windmill tower, and the drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST*.)

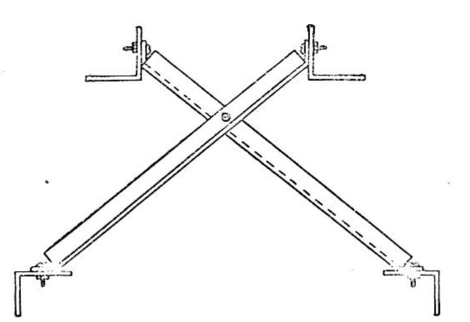


Fig. 10-98 — The boom for the 4-element beam is cross-braced at two points, about 6½ feet in from the ends.

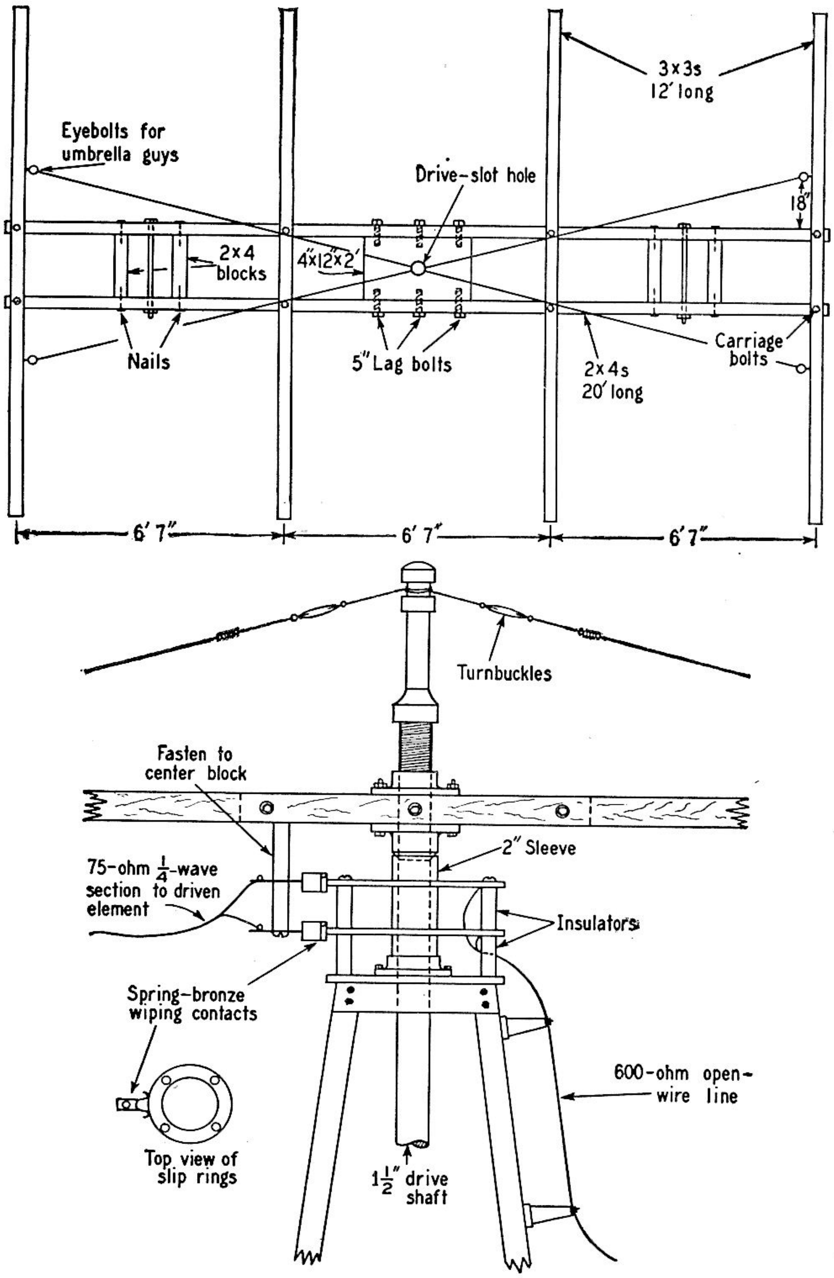


Fig. 10-100 — **Details of the wooden boom, its method of support and the construction of the slip rings.**

The crossarms are 3 × 3s twelve feet long, bolted to the boom with carriage bolts.

The umbrella guys should have turnbuckles in them, and the guys are fastened to the center support after the beam has been permanently locked in its horizontal position. With the turnbuckles properly adjusted, there will be no sag in the boom, the elements will be parallel and neat, and weaving in the wind will be eliminated.

The elements are 1⅜- and 1½-inch diameter duralumin tubing, supported by 1½-inch stand-off insulators. Hose clamps are used to hold the elements on the insulators. Final adjustment of element lengths is possible through "hairpin" loops. The tower, for the beam shown in Fig. 10-99, was a Sears-Roebuck windmill tower. The driving motor for the beam was located halfway down the tower, the torque being transmitted through a length of 1½-inch drive shaft. A pipe flange is welded to the drive shaft and bolted to the center block. A cone bearing is obtained by turning both the flange and a sleeve of 2-inch pipe to match, as shown in Fig. 10-100.

One method of matching the line to the antenna is to use a quarter wavelength of 75-ohm Twin-Lead between the radiator and the slip-ring contacts, to match a 600-ohm line from the slip rings to the transmitter.

A 600-ohm open-wire line is run to a point about halfway up on the tower, then up the side of the tower to the slip rings. The slip rings are mounted on the top of the tower, directly under the center block. A quarter-wavelength matching section of transmitting-type 75-ohm Amphenol Twin-Lead hangs in a loop between the driven element and the slip-ring contacts.

Feeder Connections

For beams that rotate only 180 degrees, it is relatively simple to bring off feeders by making a short section of the feeder, just where it leaves the rotating member, of flexible wire. Enough slack should be left so that there is no danger of breaking or twisting. Stops should be placed on the rotating shaft of the antenna so that the feeders cannot "wind up." This method also can be used with antennas that rotate the full 360 degrees, but again a stop is necessary to avoid jamming the feeders.

For continuous rotation, the sliding contact is simple and, when properly built, quite practicable. Fig. 10-101 shows two methods of making sliding contacts. The chief points to keep in mind are that the contact surfaces should be wide enough to take care of wobble in the rotating shaft, and that the contact surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition. Sliding contacts preferably should be used with nonresonant open lines where the impedance is of the order of 500 to 600 ohms, so that the current is low.

The possibility of poor connections in sliding contacts can be avoided by using inductive coupling at the antenna, with one coil rotating on the antenna and the other fixed in position, the two coils being arranged so that the coupling does not change when the antenna is rotated. Such an arrangement is shown in Fig. 10-102,

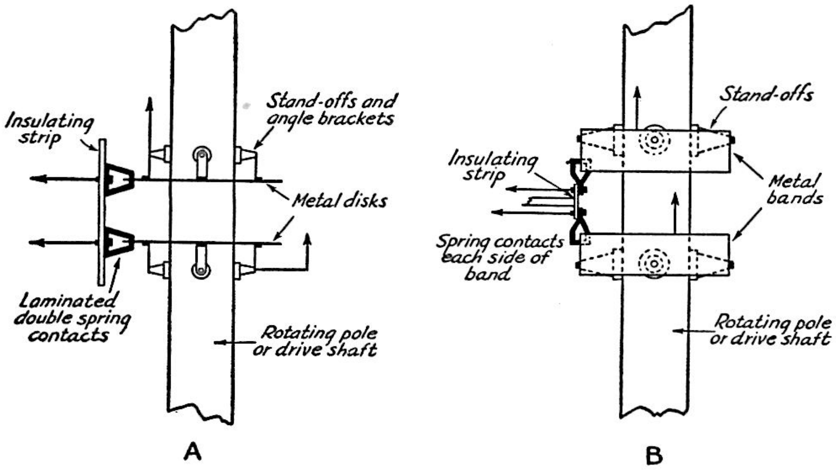


Fig. 10-101 — Ideas in sliding contacts for rotatable antenna feeder connection to permit continuous rotation. The broad bearing surfaces take care of any wobble in the rotating mast or driving shaft.

adapted to an antenna system in which the pole itself rotates. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. In the drawing, the link coil connects to a twisted-pair transmission line, but any type of line such as flexible coaxial cable can be used. The circuit would be adjusted in the same way as any link-coupled circuit, and the number of turns in the link should be varied to give proper loading on the transmitter. The rotating coupling circuit of course tunes to the transmitting frequency. The whole thing is equivalent to a link-coupled antenna tuner mounted on the pole, using a parallel-tuned tank at the end of a quarter-wave line to center feed the antenna. To maintain constant coupling, the two coils should be quite rigid and the pole should rotate without wobble. The two coils might be made a part of the upper bearing assembly holding the rotating pole in position.

Other variations of the inductive-coupled system can be worked out. The tuned circuit might, for instance, be placed at the end of a 600-ohm line, and a one-turn link used to couple directly to the center of the antenna, if the construction of the rotary member permits. In this case the coupling can be varied by changing the *L/C* ratio in the tuned circuit. For mechanical strength the coils preferably should be made of copper tubing, well braced with insulating strips to keep them rigid.

Rotation

It is convenient to use a motor to rotate the beam, but it is not always necessary, especially if a rope-and-pulley arrangement can be brought into the operating room. If the pole can be mounted near a window in the operating room, hand rotation of the beam will work out quite well, as has been proven by many amateur installations.

If the use of a rope and pulleys is impracticable, motor drive is about the only alternative. There are several complete motor-driven rotators on the market, and they are easy to mount, convenient to use, and require little or no maintenance. However, to many the cost of such units puts them out of reach, and a homemade unit must be considered. Generally speaking, lightweight units are better because they reduce the load on the mast or tower.

The speed of rotation should not be too great — one or two r.p.m. is about right. This requires a considerable gear reduction from the usual 1750-r.p.m. speed of small induction motors; a large reduction is advantageous because the gear train will prevent the beam from turning in weather-vane fashion in a wind. The ordinary structure does not require a great deal of power for rotation at slow speed, and a ⅛-hp. motor will be ample. Even small series motors of the sewing-machine type will develop enough power to turn a 28-Mc. beam at slow speed. If possible, a reversible motor should be used so that it will not be necessary to go through nearly 360 degrees to bring the beam back to a direction only slightly different, but in the opposite direction of rotation, to the direction to which it may be pointed at the moment. In cases where the pole is stationary and only the supporting framework rotates, it will be necessary to mount the motor and gear train in a housing on or near the top of the pole. If the pole rotates, the motor can be installed in a

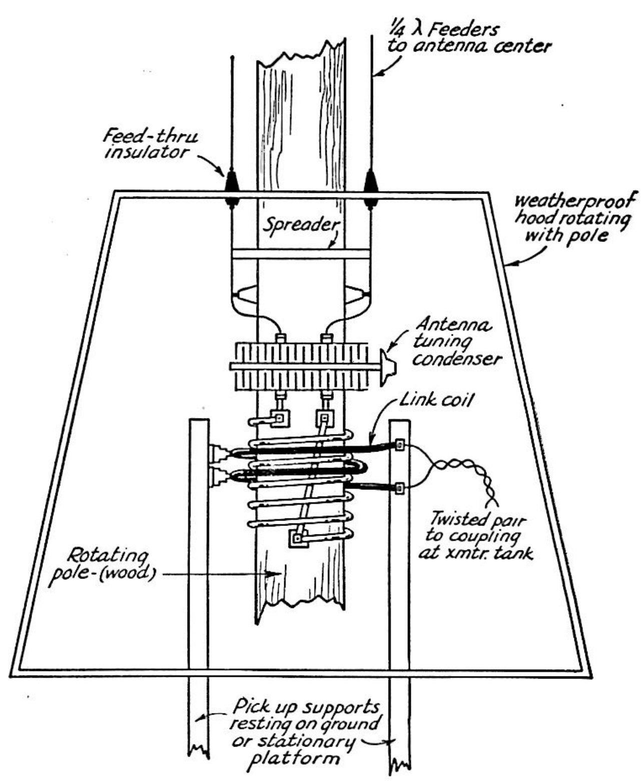


Fig. 10-102 — One method of transmission line–antenna system coupling which eliminates sliding contacts. The low-impedance line is link-coupled to a tuned line.

more accessible location (see Fig. 10-99).

Parts from junked automobiles often provide gear trains and bearings for rotating the antenna. Rear axles, in particular, can readily be adapted to the purpose. Driving motors and gear housings will stand the weather better if given a coat of aluminum paint followed by two coats of enamel and a coat of glyptal varnish. Even commercial units will last longer if treated with glyptal varnish. Be sure, of course, that the surfaces are clean and free from grease before painting them. Grease can be removed by brushing it with kerosene and then squirting the surface with a solid stream of water. The work can then be wiped dry with a rag.

If hand rotation of the beam is used, or if the rotating motor drives the beam through a pulley system, bronze cable or chain drive is preferable to rope. However, if you must use rope, be sure to soak it overnight in pure linseed oil and then let it dry for several days before installation.

The power and control leads to the rotator should be run in electrical conduit or in lead covering, and the metal should be grounded. Often r.f. appearing in power leads can be reduced by suitable filtering, but running wires in conduit is generally easier and more satisfactory. Any r.f. in the wiring can sometimes be responsible for feed-back in a 'phone transmitter. "Hash" from the motor is also reduced by shielding the wires, but it is often necessary to install a small filter at the motor to reduce this source of interference. Motor noise appearing in the receiver is a nuisance, since it is usual practice to determine the proper direction for the beam by rotating it while listening to the station it is desired to work and setting the antenna at the point that gives maximum signal strength.

The outside electrical connections should be soldered, bound with rubber tape followed by regular friction tape, and then given a coat of glyptal varnish.

About V.H.F.

In the days when DX activity first burgeoned on our lower frequencies the assignments above 30 Mc. were not too highly regarded. It was assumed that propagation on these frequencies was limited to distances only slightly beyond the visual horizon, and thus the bands allocated to amateurs in this region were used principally in areas where large concentrations of population brought hundreds of workers within local range of one another. In the early thirties activity boomed on 56 Mc. in the larger cities of the United States, but there were few stations elsewhere. Use of frequencies higher than 60 Mc. was confined to a few experimentally-inclined amateurs here and there.

In 1934, '35 and '36, new types of propagation were discovered by amateurs, and the opportunities for v.h.f. DX so brought to light caused a tremendous growth in activity, particularly in areas where it had not previously existed. Up to this time, practically all v.h.f. work had been done with the simplest sort of gear, mainly modulated-oscillator transmitters and superregenerative receivers; but when our available space began to fill with DX signals it became obvious that, if we were to realize anything like the possibilities inherent in this type of work, we must have improved techniques, whereby more stations could be accommodated in a given area. Crystal-controlled transmitters and superheterodyne receivers, permitting utilization of the 56-Mc. band on a scale comparable with that obtaining on lower frequencies, became the order of the day, and by the end of 1938 stabilization of transmitters used on all frequencies up to 60 Mc. became mandatory. Our 5-meter band had grown up!

With the impetus of improved techniques, operating ranges on 56 Mc. grew by leaps and bounds. Meanwhile the use of the simplest form of equipment was transferred to the next higher band, then 112 Mc; and this band, in turn, took over the burden of heavy urban occupancy formerly carried by the 5-meter band. Soon our principal cities were teeming with 112-Mc. activity, and before long it was found that this band, too, had much of interest to offer. Even more than had been the case on 56 Mc., it was found that weather conditions had a profound effect on 112-Mc. propagation, and before the close-down of amateur activity, at the entry of our country into the war, the record for 112-Mc. work had passed the 300-mile mark. There was a smattering of activity on the still higher frequencies of 224 and 400 Mc. as well.

During the war years the vast use of v.h.f., u.h.f. and s.h.f. equipment in countless war applications demonstrated that these frequencies, once thought to be almost useless, were of untold importance. In the postwar world the v.h.f. amateur need no longer apologize for his interests. His frequencies are among the most highly prized in the whole radio-frequency spectrum, and his is now regarded as one of the major fields of amateur endeavor.

Propagation Phenomena

A thorough understanding of the basic principles of wave propagation, outlined in Chapter Four, is a most useful tool for the v.h.f. worker. Much of the pleasure and satisfaction to be derived from v.h.f. endeavor lie in making the best possible use of propagation vagaries resulting from natural phenomena. Contrary to the impression of many newcomers to the field, a working knowledge of v.h.f. propagation is not difficult of attainment. Below are listed the principal ways by which v.h.f. waves may be propagated over abnormal distances.

F_2-Layer Reflection

The "normal" contacts made on 28 Mc. and lower frequencies are the result of reflection of the transmitted wave by the F_2 layer, the ionization density of which varies with solar activity, the highest frequencies being reflected at the peak of the 11-year solar cycle. The maximum usable frequency (m.u.f.) for F_2 reflection also rises and falls with other well-defined cycles, including daily, monthly, and seasonal variations, all related to conditions on the sun and its position with respect to the earth.

At the low point of the 11-year cycle, such as the period we were entering at the outbreak of war, the m.u.f. may reach 28 Mc. only during a short period each spring and fall, whereas it may go to 60 Mc. or higher at the peak of the cycle. The fall of 1946 saw the first authentic

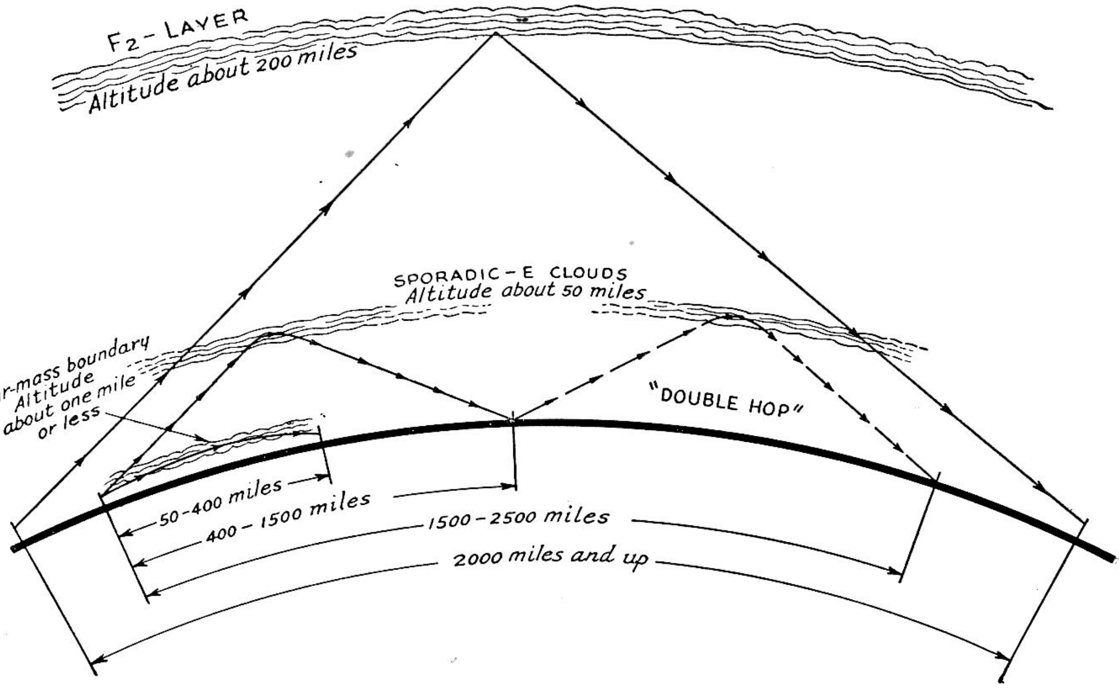


Fig. 11-1 — **The principal means by which v.h.f. waves may be returned to earth. The F_2 layer, highest of the known ionospheric layers, is capable of reflecting 50-Mc. signals during the period around the peak of the 11-year solar cycle, and may support communication over world-wide distances. Sporadic ionization of the *E* layer produces "short-skip" contacts at medium distances. It is a fairly frequent occurrence regardless of the solar cycle, but is most common in May through August. Refraction of v.h.f. waves also takes place at air-mass boundaries in the lower atmosphere, making possible reception of signals at distances up to 300 miles or more without a skip zone.**

instances of long-distance 50-Mc. work by this medium, and it is probable that F_2 DX will be workable on 50 Mc. until about 1950. In the northern latitudes there are peaks of m.u.f. each spring and fall, with a low period during the summer and a slight dropping off during the midwinter months. At or near the Equator conditions are more or less constant at all seasons.

Fortunately the F_2 m.u.f. is quite readily determined by observation, and means are available whereby it may be estimated quite accurately for any path at any time. It is predictable for months in advance,[1] enabling the v.h.f. worker to arrange test schedules with distant stations at propitious times. As there are numerous signals, both harmonics and fundamental transmissions, on the air in the range between 28 and 50 Mc., it is possible for the listener to determine the approximate m.u.f. by careful listening in this range. A series of daily observations will serve to show if the m.u.f. is rising or falling from day to day, and once the peak for a given month is determined it can be assumed that the peak for the following month will occur about 27 days later, this cycle coinciding with the turning of the sun on its axis. The working range, via F_2 skip, will be roughly comparable to that on 28 Mc., though the *minimum* distance is somewhat longer. Two-way work on 50 Mc. by means of reflection from the F_2 layer has been accomplished over distances ranging from 2200 to 10,500 miles. The maximum frequency for F_2 reflection is believed to be in the vicinity of 70 Mc.

[1] *Basic Radio Propagation Predictions*, issued monthly, three months in advance, by the Central Radio Propagation Laboratory of the National Bureau of Standards. Order from the Supt. of Documents, Washington 25, D. C.; $1.50 per year.

Sporadic-E Skip

Patchy concentrations of ionization in the *E*-layer region are often responsible for reflection of signals on 28 and 50 Mc. This is the popular "short skip" that provides fine contacts on both bands in the range between 400 and 1300 miles. It is most common in May, June and July, during the early evening hours, but it may occur at any time or season. Since it is largely unpredictable, at our present state of knowledge, sporadic-*E* skip is of high "surprise value." Multiple-hop effects may appear, when ionization develops simultaneously over large areas, making possible work over distances of more than 2500 miles. As far as is known, no 144-Mc. effects have yet been observed, the known limit for sporadic-*E* skip being in the vicinity of 100 Mc.

Aurora Effect

Low-frequency communication is occasionally wiped out by absorption of these frequencies in the ionosphere, when ionospheric storms, associated with variations in the earth's magnetic field, occur. During such disturbances, however, 50-Mc. signals may be reflected back to earth, making communication possible over distances not normally workable on this band. Magnetic storms may be accompanied by an aurora-borealis display, if the disturbance occurs at night and visibility is good. When the aurora is confined to the northern sky, aiming a directional array at the auroral curtain will bring in 50-Mc. signals strongest, regardless of the true direction to the transmitting station. When the display is widespread there may be only a slight improvement noted when the array is aimed north. The latter condition is often noticed during the period around the peak of the 11-year cycle, when solar activity is spread well over the sun's surface, instead of being concentrated extensively in the region near the solar equator.

Aurora-reflected signals are characterized by a rapid flutter, which lends a "dribbling" sound to 28-Mc. carriers and may render modulation on 50-Mc. signals completely unreadable. The only satisfactory means of communication then becomes straight c.w. The effect may be noticeable on signals from any distance other than purely local, and stations up to about 500 miles in any direction may be worked at the peak of the disturbance. Unlike the two methods of propagation previously described, aurora effect exhibits no skip zone. It has been observed, to date, only on the frequencies up to about 60 Mc.

Reflections from Meteor Trails

Probably the least-known means of v.h.f. wave propagation is that resulting from the passage of meteors across the signal path. Reflections from the ionized meteor trails may be noted as a Doppler-effect whistle on the carrier of a signal already being received, or they may cause bursts of reception from stations not normally receivable. Sudden large increases in strength of normally-weak signals are another manifestation of this effect. Ordinarily such reflections are of little value in extending communication ranges, since the increases in signal strength are of short duration, but meteor showers of considerable magnitude and duration may provide fluttery 50-Mc. signals from distances up to 1000 miles or more. Signals so reflected have a combination of the characteristics of aurora and sporadic-*E* skip.

Tropospheric Bending

Refraction of radio waves takes place whenever a change in refractive index is encountered. This may occur at one of the ionized layers of the ionosphere, as mentioned above, or it may exist at the boundary area between two different types of air masses, in the region close to the earth's surface. A warm, moist air mass from over the Gulf of Mexico, for instance, may overrun a cold, dry air mass which may have had its origin in northern Canada. Each tends to retain its original characteristics for considerable periods of time, and there may be a well-defined boundary between the two for as much as several days. When such an air-mass boundary exists near the midpoint between two v.h.f. stations separated by 50 to 300 miles or more, a considerable degree of refraction takes place, and signals run high above the average value. Under ideal conditions there may be almost no attenuation, and signals from far beyond the visual horizon will come through with strength comparable to that of local stations.

Many factors other than air-mass movement of a continental character may provide increased v.h.f. operating range. The convection that takes place along our coastal areas in warm weather is a good example. The rapid cooling of the earth after a hot day in summer, with the air aloft cooling more slowly, is another, producing a rise in signal strength in the period around sundown. The early-morning hours, when the sun heats the air aloft, before the temperature of the earth's surface begins its daily rise, may frequently be the best hours of the day for extended v.h.f. range, particularly in clear calm weather, when the barometer is high and the humidity low.

Any weather condition that produces a pronounced boundary between air masses of different temperature and humidity characteristics provides the medium by which v.h.f. signals cover abnormal distances. The ambi-

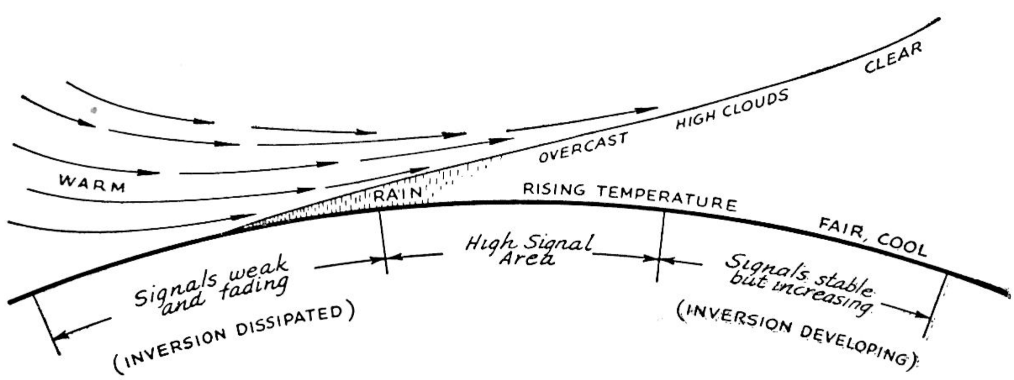


Fig. 11-2 — **Illustrating a typical weather sequence, with associated variations in v.h.f. propagation. At the right is a cold air mass (fair weather, high or rising barometer, moderate summer temperatures). Approaching this from the left is a warm moist air mass, which overruns the cold air at the point of contact, creating a temperature inversion and considerable bending of v.h.f. waves. At the left, in the storm area, the inversion is dissipated and signals are weak and subject to fading. Barometer is low or falling at this point.**

tious v.h.f. enthusiast soon learns to correlate various weather manifestations with radio-propagation phenomena. By watching temperature, barometric pressure, changing cloud formations, wind direction, visibility, and other easily-observed weather signs, he is able to tell with a reasonable degree of accuracy what is in prospect on the v.h.f. bands.

The responsiveness of radio waves to varying weather conditions increases with frequency. Our 50-Mc. band is considerably more sensitive to weather variations than is the 28-Mc. band, and the 144-Mc. band may show strong signals from far beyond visual distances when the lower frequencies are relatively inactive. It is probable that this tendency continues on up through the microwave range, and that our assignments in the u.h.f. and s.h.f. portions of the frequency spectrum may someday support communication over distances far in excess of the optical range. Already 144-Mc. communication by amateurs has passed the 600-mile mark, and even greater distances are believed possible on this and higher frequencies.

● STATION LOCATIONS

In line with our early notions of v.h.f. wave propagation, it was once thought that only highly-elevated v.h.f. stations had any chance of working beyond a few miles. Almost all the work was done by portable stations operating from mountain tops, and only hilltop home sites were considered suitable for fixed-station work. It is still true that the fortunate amateur who lives at the top of a hill enjoys a certain advantage over his fellows on the v.h.f. bands, but high elevation is not the all-important factor it was once thought to be.

Improvements in equipment, the wide use of high-gain antenna systems, and an awareness of the opportunities afforded by weather phenomena have enabled countless v.h.f. workers to achieve excellent results from seemingly poor locations. In 50-Mc. DX work particularly, elevation has ceased to be an important factor, though it may help in extending the range of operation somewhat under normal conditions. A high elevation is somewhat more helpful on 144 Mc. and higher frequencies, particularly when no unusual propagation factors are present, as during the winter months. Other factors, such as close proximity to large bodies of water, may more than compensate for lack of elevation during the other seasons of the year, however.

Stations situated in sea-level locations along our coasts have been consistent in their ability to set distance records on 144 Mc.; weather variations provide interesting propagation effects over our Middle Western plain areas; and even the worker situated in mountainous country need not necessarily feel that he is prevented by the nature of his horizon from doing interesting work. Contacts have been made on 50 and 144 Mc. over distances in excess of 100 miles in all kinds of terrain.

The consistently-reliable nature of 50 and 144 Mc. for work over such a radius and more, regardless of weather, time or season, and the occasional opportunities these frequencies afford for exciting DX, have caused an increasing number of amateurs to migrate to the v.h.f. bands for extended-local communication, once thought possible only on the lower frequencies.

V.H.F. Techniques

Recognition of the value of the very-high frequencies has resulted in the development of many tubes and other components especially suited for use at these frequencies. Where, not so many years ago, it was necessary to remove the bases from available tubes, and otherwise cut down components designed for use at lower frequencies, we now have tubes and circuit components specifically designed for high-efficiency operation, not only on the v.h.f. bands, but on up through the microwave range. Examples of transmitting tubes especially made for v.h.f. work are shown in Fig. 11-3.

The higher frequencies are rapidly becoming the primary field of interest for those amateurs who like to design and build their own equipment. While there is an increasing tendency to the purchase of commercially-built equipment, both transmitting and receiving, for low-frequency operation, most gear used for v.h.f. and higher bands is still a product of the amateur's own ingenuity.

In the field of antenna design, too, the v.h.f. bands offer much to the amateur who is interested in experimental work. With their smaller physical size making for greater ease of construction and adjustment, the development of high-gain directional antennas continues to occupy much of the time devoted by the v.h.f. enthusiast to experimental work.

● TRANSMITTER DESIGN

The use of crystal control, or its equivalent in stability, is standard for 50-Mc. work. The design of transmitters for this band differs hardly at all from that employed for lower amateur frequencies, except that much more care must be exercised in the selection of component parts and their placement in the equipment, in order to avoid more than the absolute minimum length in the connecting leads. Customary procedure is to start with a crystal or variable-frequency oscillator, operating at 6, 8, or 12 Mc., and follow with such frequency-multiplying stages as may be required to reach 50 Mc. The power level is not particularly important, as interference is not a critical factor in 50-Mc. communication. Much good work has, in fact, been accomplished with power inputs under 100 watts and even stations in the 10-watt class are quite capable of working out well, particularly if equipped with well-designed antenna systems.

At 144 Mc. crystal control is becoming more popular daily. It is somewhat more difficult of attainment than at 50 Mc., but the construction of a crystal-controlled transmitter for 144 Mc. is not beyond the capabilities of the average amateur. The number of usable tube types is limited, however, and only those specifically designed for v.h.f. applications can be used successfully. Even with such tubes, great care must be exercised to keep leads and circuit capacitances down. Conventional coil-and-condenser combinations designed for lower frequencies are generally unsatisfactory, and only well-designed tank circuits will operate efficiently at this frequency. High power is seldom employed in 144-Mc. operation, most workers preferring to use high-gain antenna systems rather than high-powered transmitters, at this and higher frequencies.

For 235 Mc. and higher, crystal control may be employed, and its use is desirable where possible, but the modulated-oscillator type of transmitter still bears the brunt of operation on 235 Mc., and is used almost exclusively for 420 Mc. and higher. Since occupancy is relatively low, the broader signals radiated by such equipment and the inefficiencies of the superregenerative receivers necessary to accommodate them, are not major problems.

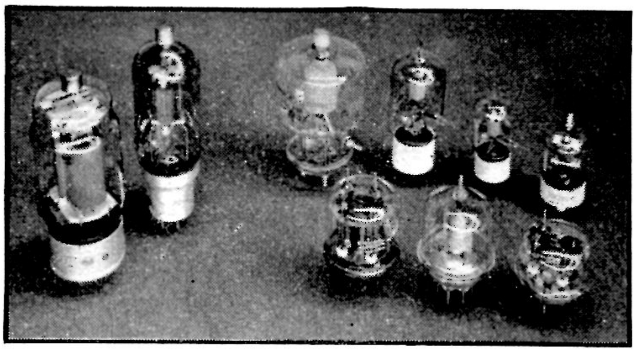


Fig. 11-3 — Vacuum tubes designed especially for high-efficiency operation at very-high frequencies are now available for amateur use. Several such tubes are shown above, in comparison to typical low-frequency tubes, the 813 and V-70-D at the left. The v.h.f. types are the GL-592, 35TG, 24G, HY-75-A, all triodes; and the 829-B, HK-57, and 832-A, all tetrodes.

● RECEIVER CONSIDERATIONS

Even more than in work on lower frequencies, a good receiver is all-important in the v.h.f. station. Though commercial receivers that cover the 50-Mc. band are slowly appearing on the amateur market, the most satisfactory and inexpensive solution to the receiver problem is still that of a converter that works into a communications receiver designed for the lower frequencies. Such a combination is almost certain to give better results on 50 Mc. than a complete receiver, unless the latter is designed especially for v.h.f. use.

Converters are replacing the once-popular superregenerative receivers for 144-Mc. use also, particularly for fixed-station work in localities where the use of stabilized transmitters has become more or less standard procedure. Many types of superhet receivers used for radar and aircraft service during the war are convertible to amateur use, and hundreds of such surplus units are now employed by amateurs working on 144 Mc.

For portable or emergency use, where small size and low battery drain are important, the simple superregenerative receiver is still popular. For the number of tubes and parts required, it is still an efficient receiving system, especially in areas where there is not extensive activity. For frequencies higher than 148 Mc. it is still the principal receiving system, though the converter approach is practical for any frequency. To accommodate the broader signals generally found on these frequencies, a converter may be used in conjunction with a wide-band i.f. system, such as a receiver designed for FM broadcast reception.

V.H.F. Receivers

In its essentials most modern receiving equipment for the 28- and 50-Mc. bands differs very little from that used on lower frequencies. The 28-Mc. band serves as the meeting ground between what are ordinarily termed "communications frequencies" and the very-highs, and it will be found that most of the receivers described in Chapter Five are capable of working on 28 Mc. In this chapter are described receivers and converters capable of good performance on 50 Mc. and higher.

Federal regulations impose identical requirements on all frequencies below 54 Mc. respecting stability of frequency and, when amplitude modulation is used, freedom from frequency modulation. Thus receivers for 50-Mc. AM reception may have the same selectivity as those designed for the lower frequencies. This order of selectivity is not only possible but desirable, since it permits a considerable increase in the number of transmitters that can work in the band without undue interference. High selectivity also aids greatly in improving the signal-to-noise ratio, both as concerns noise originating in the receiver itself and in its response to external noise. The effective sensitivity of such a receiver can be made considerably higher than is possible with nonselective receivers.

Superheterodynes for V.H.F.

The superheterodyne system of reception is used almost universally on 50 Mc., and to a considerable extent on 144 Mc., because it is the only type that fulfills the stability, selectivity and sensitivity requirements. AM superheterodynes and those for FM reception differ only in the i.f. amplifier and second detector, so that a single high-frequency converter may be used for either AM or FM.

Superheterodynes for 50 Mc. and higher should have fairly-high intermediate frequencies to reduce both image response and oscillator "pulling." For example, a difference between signal and image frequencies of 900 kc. (the difference when the i.f. is 450 kc.) is a very small percentage of the signal frequency; consequently, the response of the r.f. circuits to the image frequency is nearly as great as to the desired frequency. To obtain discrimination against the image equal to that obtainable at 3.5 Mc. would require an i.f. 16 times as high, or about 7 Mc. However, the Q of tuned circuits is less in the v.h.f. range than it is at lower frequencies, chiefly because the tube loading is considerably greater, and thus still higher intermediate frequencies are desirable. A practical compromise is reached at about 10 Mc., and the standard i.f. for converters and commercial v.h.f. receivers is 10.7 Mc.

To obtain high selectivity with a reasonable number of i.f. stages, the double-superheterodyne principle is often employed. A 10-Mc. intermediate frequency, for example, is changed to a second i.f. of perhaps 450 kc. by an additional oscillator-mixer combination.

Few amateurs build complete 50-Mc. superheterodyne receivers. General practice in this band has been to use a conventional communications receiver to handle the i.f. output of a simple 50-Mc. frequency converter. Even an all-wave broadcast receiver may be used with excellent results on 50 Mc. by the addition of a relatively simple converter.

The superheterodyne type of receiver is finding increased favor for 144-Mc. work also, as the occupancy of that band increases. Especially in heavily-populated areas, stabilization of transmitters and an improvement in the selectivity of receivers are becoming almost mandatory, particularly for those operators who are interested in exploiting the full possibilities of this band.

With a well-designed converter, a considerable improvement in signal-to-noise ratio can be achieved in 144-Mc. reception by using such a converter in conjunction with a communications receiver. Only crystal-controlled or other stable signals can be received in this manner, but the effective sensitivity will be better than is possible with a less critical broad-band receiver. An example of a simple but effective converter for 2-meter use is shown in Figs. 12-16 through 12-19.

In any superheterodyne for 144 Mc. a primary problem is that of oscillator stability. One satisfactory solution is the use of a crystal-controlled oscillator and frequency multiplier to supply the injection voltage, the method used in the converter shown in Figs. 12-13–12-15. All r.f. circuits are then fixed-tuned, and coverage of the band is attained by tuning the communications receiver (or i.f. system) over a range of 4 Mc. This can be 14 to 18 Mc. in the case of certain war-surplus receivers, or it may be any higher frequency within the range of the communications receiver.

A converter working into an FM receiver, or into a broad-band i.f. channel designed for either AM or FM reception, provides a quite satisfactory means of reception of signals, not only at 144 Mc., but on up through the microwave range. This approach has been used in most of the recent pioneering efforts by amateurs working in the microwave field.

The Superregenerative Receiver

The simplest type of v.h.f. receiver is the **superregenerator,** long favored in amateur work. It affords good sensitivity with few tubes and elementary circuits. Its disadvantages are lack of selectivity and, if the oscillating detector is coupled to an antenna, a tendency to radiate a signal which may cause severe interference to other receivers. To some extent the lack of selectivity is advantageous, since it makes for easy tuning, and permits reception of all signals within its tuning range, however unstable they may be. To reduce radiation, a superregenerative detector should be preceded by an r.f. stage, or, if the detector is coupled directly to the antenna, it should be operated at the lowest plate voltage that will permit superregeneration.

From a practical aspect, superregenerative receivers may be divided into two general types. In the first the quenching voltage is developed by the detector tube functioning as a "self-quenched" oscillator. In the second, a separate oscillator tube is used to generate the quench voltage. Self-quenched superregenerators have found wide favor in amateur work. The simpler types are particularly suited for portable equipment, which must be kept as simple as possible. Many amateurs have "pet" circuits claimed to be superior to all others, but the probability is that the arrangement of a particular circuit has led to correct operating conditions. Time spent in minor adjustments will result in a smooth-working receiver.

Superregeneration Principles

The limit to which ordinary regenerative amplification can be carried is the point at which oscillations commence, since at that point further amplification ceases. The *superregenerative* detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range (of the order of 20 to 200 kilocycles), in such a way as to vary the detector's operating point. As a consequence of the introduction of this **quench** or **interruption** frequency, the detector can oscillate only when the varying operating point is in a region suitable for the production of oscillations. Because the oscillations are constantly being interrupted, the regeneration can be greatly increased, and the amplified signal will build up to tremendous proportions. A one-tube superregenerative detector is capable of an inherent sensitivity approaching the thermal-agitation noise level of the tuned circuit, and may have an antenna input sensitivity of two microvolts or better.

Because of its inherent characteristics, the superregenerative circuit is suitable only for the reception of modulated signals, and operates best on the very-high frequencies. Typical superregenerative circuits for v.h.f. are shown in Fig. 12-1, but the basic circuit may be any of the various arrangements used for straight regenerative detectors.

In Fig. 12-1A the quench frequency is obtained from a separate oscillator and introduced into the plate circuit of the detector.

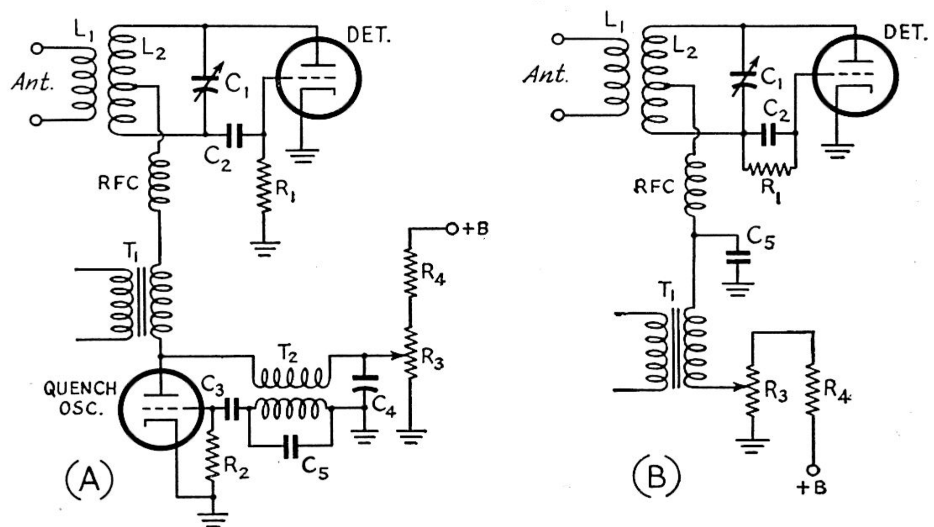


Fig. 12-1 — (A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:

C_2 — 47 $\mu\mu$fd.
C_3 — 470 $\mu\mu$fd.
C_4 — 0.1 μfd.
C_5 — 0.001–0.005 μfd.
R_1 — 2-10 megohms.
R_2 — 47,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 47,000 ohms.
RFC — R.f. choke, value depending upon frequency. Small low-capacitance chokes are required for v.h.f. operation.
T_1 — Audio transformer, plate-to-grid type.
T_2 — Quench-oscillator transformer.

The quench oscillator, operating at a low radio frequency, alternately allows oscillations to build up in the regenerative circuit and then causes them to die out. In the absence of a signal, the thermal-agitation noise in the input circuit produces the voltage that initiates the build-up process. However, when an incoming signal provides the initiating pulse, it has the effect of advancing the starting time of the oscillations. This causes the area within the envelope to increase, as indicated in Fig. 12-2C.

If regeneration in an ordinary regenerative circuit is carried sufficiently far, the circuit will

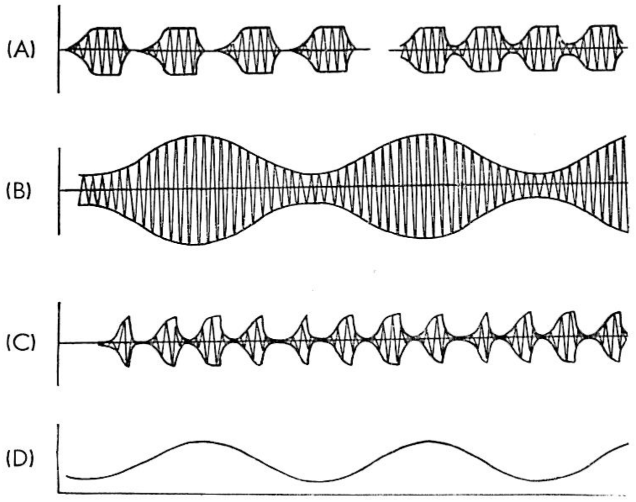


Fig. 12-2 — R.f. oscillation envelopes in a self-quenched superregenerative detector. Without signal (A at left) oscillations are completely quenched after each period, resuming in random phase depending on momentary noise voltages. At right, when the initiating pulses are supplied by a received signal the starting time of the oscillations is advanced causing the build-up period to begin before damping is complete. This advance is proportional to the carrier amplitude when modulated (B). Since the building-up period varies in accordance with modulation (C), when these wave trains are rectified the average rectified current is proportional to the amplitude of the signal. Amplitude modulation is therefore reproduced as an audio wave in the output circuit (D).

break into a low-frequency oscillation simultaneously with that at the operating radio frequency. This low-frequency oscillation has much the same quenching effect as that from a separate oscillator, hence a circuit so operated is called a *self-quenching* superregenerative detector. The frequency of the quench oscillation depends upon the feed-back and upon the time constant of the grid leak and condenser, the oscillation being a "blocking" or "squegging" in which the grid accumulates a strong negative charge which does not leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

The greater the difference between the quenching and signal frequencies the greater the amplification, because the signal then has a longer period in which to build up during the nonquenching half-cycle when the resistance of the circuit is negative. This ratio should not exceed a certain limit, however, for during the quenched or nonregenerative intervals the input selectivity is merely that of the Q of the tuned circuit alone.

Because of the greater amplification, the hiss noise when a superregenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in, the hiss will disappear to a degree that depends upon the signal strength.

Lack of hiss indicates insufficient feed-back at the signal frequency, or inadequate quench voltage. Antenna-loading effects will cause dead spots that are similar to those in regenerative detectors and can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid-leak and grid-condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

T.R.F. Superregenerative Receiver

The 144-Mc. receiver in Figs. 12-3–12-7 uses miniature tubes throughout and is intended for either home or portable/mobile use. The r.f. amplifier stage furnishes some additional gain over a straight superregenerative detector, affords freedom from antenna effects, and — most important of all — prevents radiation from the receiver. Although the r.f. and detector circuits are individually tuned, the broad tuning of the r.f. stage makes the receiver essentially a single-dial affair — important in mobile work — and the miniature tubes permit compact assembly and low current consumption. Total heater current is 625 ma. at 6.3 volts, and the total plate-current drain from 135 volts of "B" battery is less than 10 ma.

The tuned r.f. amplifier stage uses a 6AK5 pentode which is coupled through C_5 to the 6C4 superregenerative triode detector. This in turn is transformer-coupled to a 6C4 audio stage which drives the 6AK6 output stage. A plate coupling choke, L_4, and the coupling condenser C_{12} remove d.c. from the output jack, J_2, and eliminate the possibility of short-circuiting the plate supply at this point.

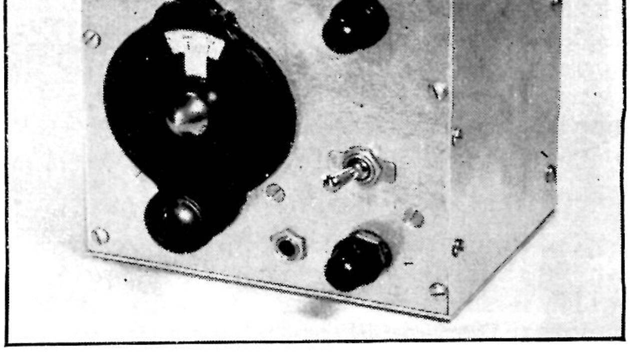


Fig. 12-3 — Front view of the 144-Mc. t.r.f. receiver. The pointer knob above the vernier dial tunes the r.f. stage. The small round knobs are for audio volume (lower right) and detector plate-voltage variation. Dimensions of the handmade case are 7 × 5½ × 4 inches.

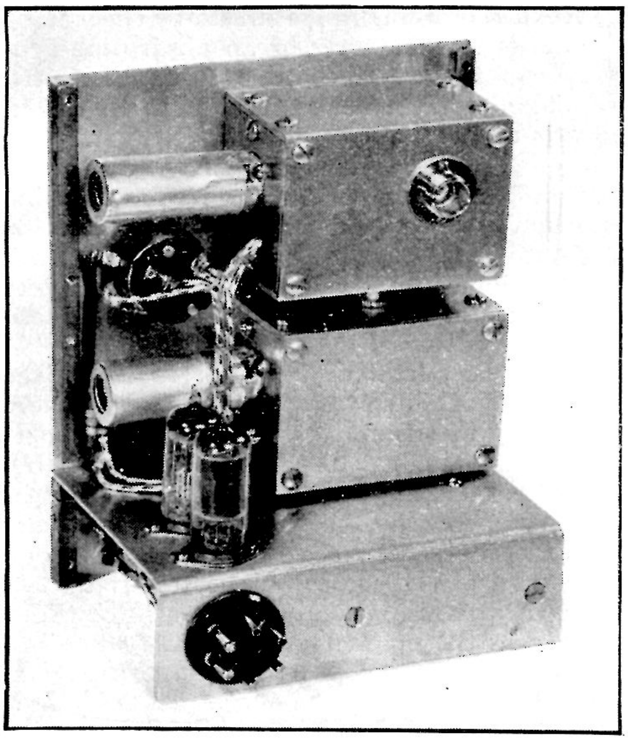


Fig. 12-4 — Rear view of the complete receiver. Note that the r.f. stage and superregenerative-detector circuit components are in separate completely-enclosed compartments, for elimination of radiation. Miniature tubes are used throughout, for compactness and low current consumption.

boxes are completely assembled. The wire between C_5 and L_3 runs through two Millen 32150 bushings in the walls of the two shield compartments. This interconnection, the only one except for the power circuits, is made by running separate leads from the condenser and coil through the bushings and then soldering the two ends together after the two units are mounted on the front panel.

The detector tuning condenser, C_8, is a regular Cardwell ZV-5-TS modified by adding a single circular plate to the regular one-plate rotor. This additional constant capacitance across the circuit increases the bandspread and, because it decreases the L-to-C ratio, smooths out the regeneration so that the regeneration control, R_{10}, does not have to be readjusted within the 144-Mc. band.

Mechanical Details

The receiver chassis and partitions are built from pieces of 1/16-inch aluminum held together at the corners with machine screws and strips of 1/4-inch-square brass rod. The over-all dimensions are 7 × 5½ × 4 inches — the chassis that mounts the audio components is 4 × 5 inches with a 1¾-inch folded lip. To eliminate oscillation in the r.f. stage and radiation from the detector, completely-separate compartments are used for the r.f. and detector stages. These compartments consist of identical boxes that measure 1⅞ inches square and 3 inches long. The tube sockets are mounted on the end plates, and all of the connections to the sockets are made before the

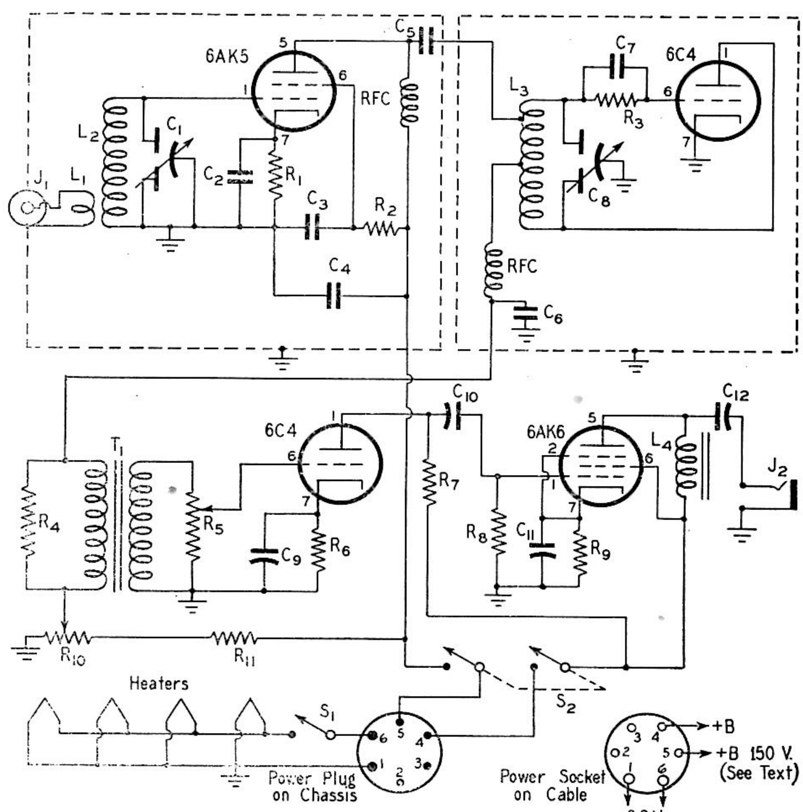


Fig. 12-5 — Wiring diagram of the 4-tube t.r.f. superregenerative receiver. Shield compartments housing r.f. and detector stages are shown by dotted lines.

C_1, C_8 — Split-stator condenser (Cardwell ZV-5-TS). See text.
C_2, C_3, C_4 — 470-μμfd. midget mica.
C_5, C_7 — 47-μμfd. midget mica.
C_6 — 0.0022-μfd. midget mica.
C_9, C_{11} — 10-μfd. 25-volt midget electrolytic.
C_{10}, C_{12} — 0.1-μfd. paper.
R_1 — 1500 ohms, ½ watt.
R_2, R_7, R_8 — 0.1 megohm, ½ watt.
R_3 — 3.3 megohms, ½ watt.
R_4 — 39,000 ohms, ½ watt. See text.
R_5 — 0.5-megohm potentiometer.
R_6 — 2200 ohms, ½ watt.
R_9 — 680 ohms, ½ watt.
R_{10} — 50,000-ohm potentiometer.
R_{11} — 22,000 ohms, 1 watt.
L_1 — 2 t. ⅜-inch i.d. No. 18 enam. inserted between turns of L_2 at cold end.
L_2 — 4 t. ⅜-inch i.d., ¾ inch long, No. 18 tinned.
L_3 — 5 t. center-tapped, ½ inch long, No. 18 tinned. R.f. coupling tap 1 t. from grid end.
L_4 — Midget audio or filter choke (Inca D-92).
J_1 — Coaxial socket (Jones S-201). Matching plug for antenna is P-101 or P-201.
J_2 — Headphone or 'speaker jack.
RFC — See text.
S_1 — S.p.s.t. switch on R_{10}.
S_2 — D.p.s.t. toggle switch.
T_1 — Midget audio transformer (Thordarson T-13A34).

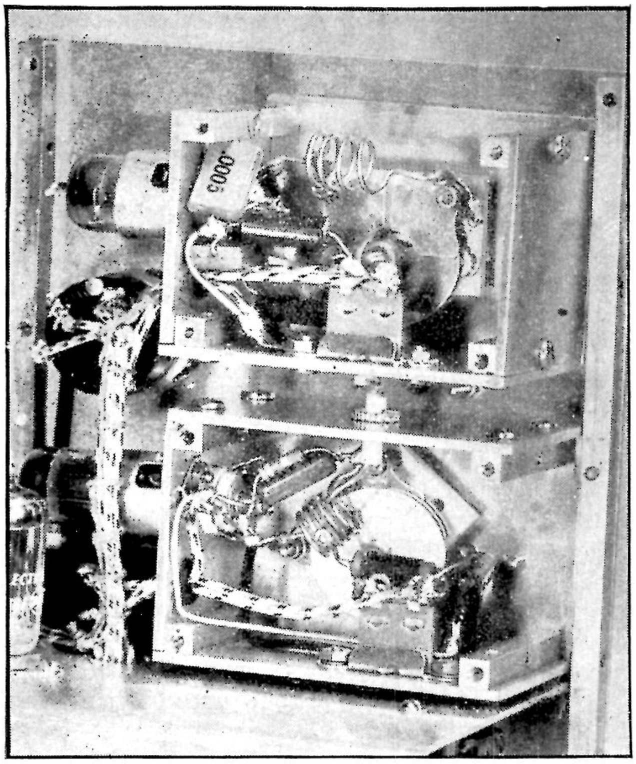


Fig. 12-6 — Close-up view of the r.f. and superregenerative-detector compartments, with back plates removed to show details. Top, back, and right side may be removed from either assembly, providing accessibility despite compact design.

The two r.f. chokes, *RFC*, are homemade affairs wound on 1-watt IRC composition resistors — 0.22 megohm or higher — of the insulated type that is 1/4 inch in diameter and 21/32 inch long. The ends are notched with a small file or saw, to prevent the ends of the coil wire from slipping after they have been soldered to the pigtail leads of the resistor, and then a single layer of No. 30 d.s.c. is wound on for a length of 17/32 inch. No lacquer or dope should be used on the winding because of the increased distributed capacitance that would result.

Adjustments

When the receiver is completely wired the first move should be to check detector operation. With the 6AK5 in its socket, but with no plate or screen voltage applied to it, apply the plate voltage to the detector and check for the customary hiss. Try the regeneration control, R_{10}, to determine whether the detector goes in and out of superregeneration smoothly. Some variation in values of R_3, R_4 and C_6 may be necessary to attain this end, and some 6C4s work better than others in this respect.

Next, the tuning range should be checked by means of Lecher wires or an absorption-type wavemeter. With the values given, 144 Mc. should fall at about 80 on the dial, with 148 Mc. at around 60. The position of the r.f. coupling tap on L_3 will have considerable effect on the resonant frequency of the combination. Its position is not critical, except for its effect on the tuning range of the detector circuit, but the spacing of the turns in the coil will have to be changed if the position of the tap is materially different from that given.

When the detector is found to be in the band, the r.f. stage may be put into operation. With any of the shields removed, or with no antenna connected, the 6AK5 will probably oscillate, blocking the detector, but this effect will disappear when the two compartments are completely assembled and an antenna attached by means of the coaxial connector. If the r.f. stage is operating properly there will be slight change in the character of the hiss when the stage is tuned through resonance. Using a signal generator (the harmonic of any oscillator which falls in the 144-Mc. band will do) or the signal of a 144-Mc. station, there will be a pronounced drop in background noise and a slight change in dial setting of the detector when the r.f. stage is tuned "on the nose." Once the r.f. tuning is adjusted for maximum response, preferably on a weak signal near the middle of the band, it may be left at that setting for all except the very weakest signals at either end.

Power Supply

Power-supply filtering and regulation are important factors in attaining smooth and efficient performance with superregenerative detectors. The power plug mounted on the back of the chassis provides a separate connection (Pin 5) for the detector and r.f. +B, in order that this may be drawn from a regulated source, such as a VR-150. The other pin marked "+B" (Pin 4) supplies the audio tubes, and the voltage used here need not be regulated. If "B" batteries are used — and they are highly recommended for mobile oper-

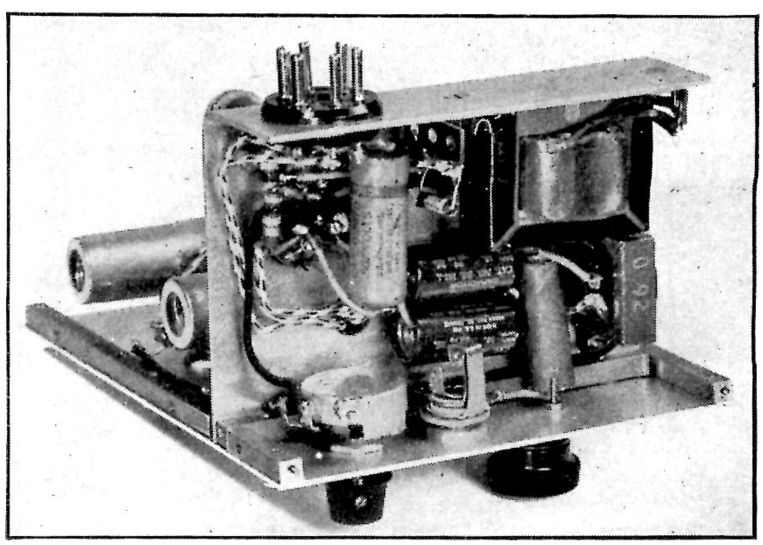


Fig. 12-7 — Bottom view, showing audio-component arrangement.

ation — Pins 4 and 5 may be connected together in the power socket on the cable. The use of "B" batteries in mobile work will result in better sensitivity and more quiet operation than will be available with any sort of mobile power supply, vibrator or dynamotor, and the drain from the car battery will be negligible during receiving periods. Medium-size "B" batteries will last through a year or more of normal operation. When batteries are used, the on-off switch, S_2, should be thrown to the "off" position when the receiver is not in use, otherwise there will be a small continuous drain on the batteries through the R_{10}-R_{11} bleeder.

Simple Two-Tube Converter for 50 Mc.

When a high intermediate frequency is used, image rejection is not a problem, and r.f. selectivity in the converter is not particularly important, especially when the converter is used in conjunction with a highly-selective communications receiver. Thus quite satisfactory performance can be obtained without the use of an r.f. amplifier stage. The new high-transconductance miniature pentodes, such as the 6AK5, are excellent as mixers, and a two-tube converter incorporating the 6AK5 in an appropriate circuit will give a degree of performance formerly obtainable only with more complex designs. Such a converter is shown in Figs. 12-8–12-12. It was designed by Richard W. Houghton, W1NKE, and was described in detail in *QST* for June, 1946. Though it was laid out particularly for use with an HRO it may be used effectively with any communications receiver capable of tuning to 10.5 Mc.

As shown in the schematic diagram, Fig. 12-10, the oscillator voltage is injected at the screen grid of the mixer tube. The coupling condenser, C_9, has sufficient capacitance to act as the 6AK5 screen by-pass condenser as well. The grid tank circuit, comprised of L_2 in parallel with C_1, C_2 and C_3, resonates over the operating frequency range, 49.5 to 54.8 megacycles. Capacitor C_3 is ganged with the oscillator tuning condenser, C_6.

The oscillator operates over a range 10.5 Mc. higher than that of the mixer, and the mixer plate circuit is tuned to this intermediate frequency. With this i.f., the fifth harmonic of the receiver's local oscillator (10.955 × 5 = 54.775 Mc.) appears just outside the high end

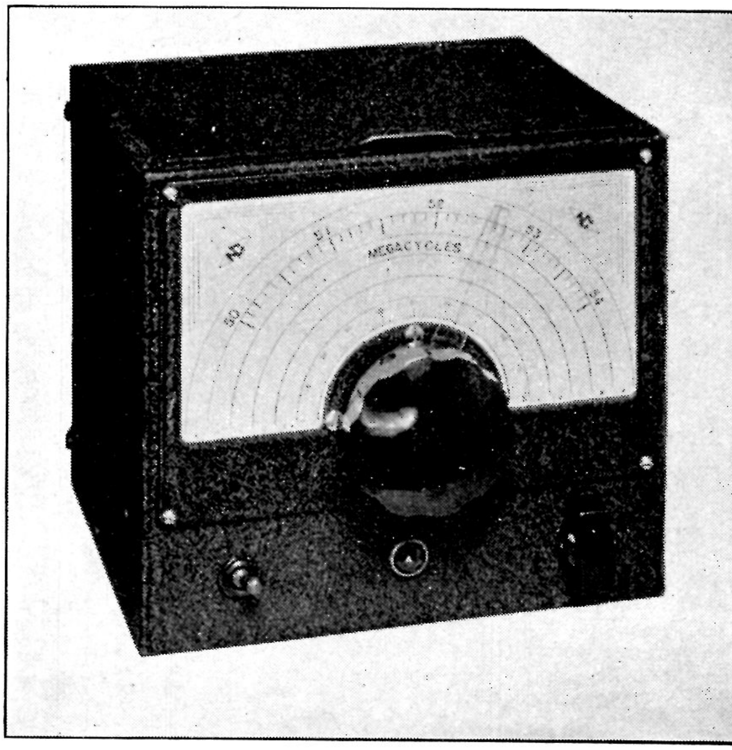


Fig. 12-8 — This two-tube 50-Mc. converter incorporates miniature tubes and obtains its power from the communications receiver with which it is used. The toggle switch at the left cuts the filament circuit when the unit is not in use. The control at the lower right transfers the antenna from the converter to the receiver for normal reception.

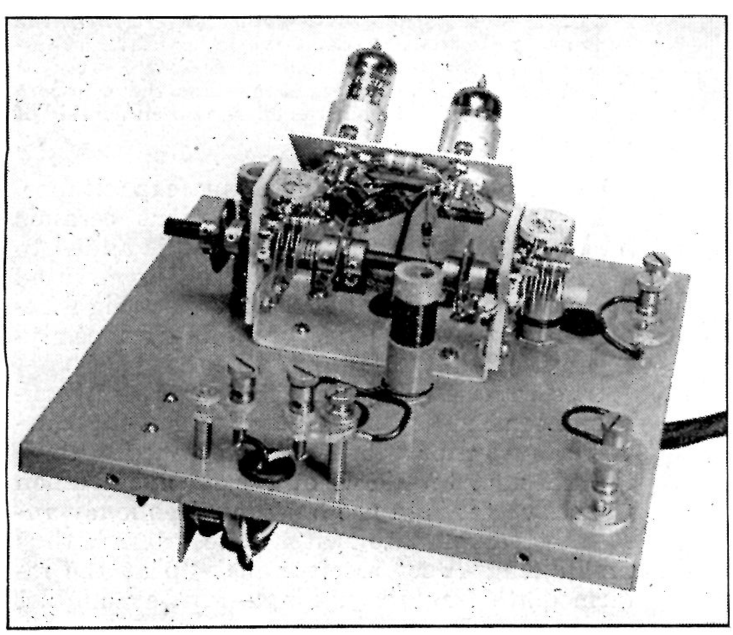


Fig. 12-9 — The r.f. construction of the 50-Mc. converter is shown in this above-chassis view. The 6C4 oscillator is at the left and 6AK5 mixer at the right on the subchassis. The 10.5-Mc. i.f. output coil is in the foreground. Flexible ground leads are shown connected to their binding posts in the position normally used for grounded antenna systems.

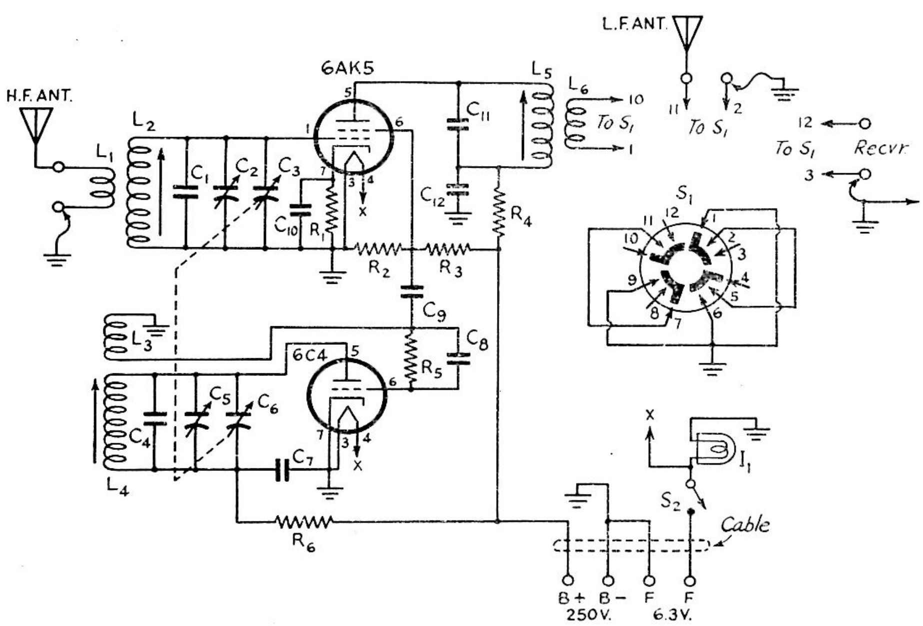


Fig. 12-10 — Circuit diagram of the 50-Mc. converter.

C_1 — 15-$\mu\mu$fd. fixed ceramic, zero temp. coef. (Erie NPOA).
C_2, C_5 — 2–6-$\mu\mu$fd. ceramic trimmer (Centralab 820-A).
C_3 — 11-$\mu\mu$fd. variable (National UMA-10 with 1 stator plate removed).
C_4 — 12-$\mu\mu$fd. fixed ceramic, zero temp. coef. (Erie NPOA).
C_6 — 9-$\mu\mu$fd. variable (National UMA-10 with 1 stator and 1 rotor plate removed).
C_7, C_8, C_9 — 100-$\mu\mu$fd. mica or ceramic.
C_{10}, C_{12} — 47-$\mu\mu$fd. mica or ceramic.
C_{11} — 35-$\mu\mu$fd. fixed ceramic, zero temp. coef. (Erie NPOA).
R_1 — 6800 ohms, ½ watt.
R_2 — 1.5 megohms, ½ watt.
R_3 — 0.47 megohm, ½ watt.
R_4 — 0.1 megohm, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 10,000 ohms, 1 watt.
L_1 to L_6, inc. — See Fig. 12-11.
I_1 — 6.3-volt pilot lamp.
S_1 — 4-pole double-throw switch, preferably with ceramic wafers (Oak Type HC).
S_2 — S.p.s.t. toggle.

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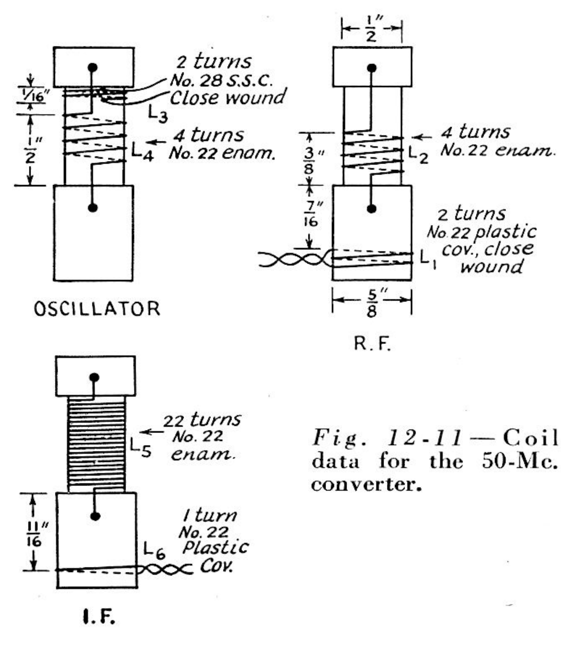


Fig. 12-11 — Coil data for the 50-Mc. converter.

of the tuning range, sufficiently far from the calibrated band so that it does not interfere with normal operation. The i.f. may be shifted slightly from the 10.5-Mc. figure, if necessary, to avoid strong signals at that frequency.

Tracking is easily accomplished over the frequency range under consideration because the percentage of frequency change is small. Starting with two identical tuning condensers (National Type UMA-10), two plates are removed from the one used in the oscillator and one plate from the one in the mixer. Sufficient fixed padding capacitance, using a zero-temperature-coefficient ceramic for low over-all temperature drift, is added to give the required range. The coil forms used are provided with adjustable cores of high-frequency powdered iron, providing an easily-accessible inductance adjustment.

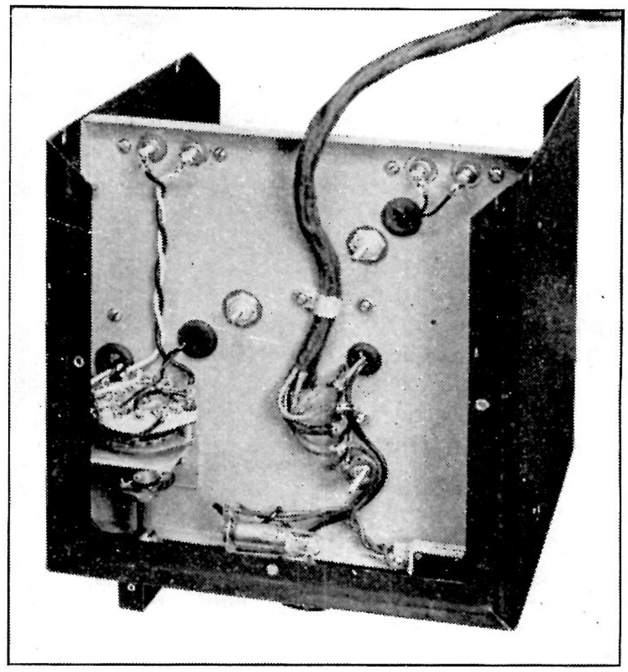


Fig. 12-12 — A bottom view of the converter. S_1, the antenna-transfer switch, is at the lower left. Low-impedance antenna leads should be twisted loosely as shown. The three adjusting screws for the iron-core inductances protrude from the chassis on either side of the power cord.

The wafer-type switch, S_1, provides a convenient means of channeling either the converter output or a low-frequency antenna into the antenna terminals of the receiver. When the converter is in use both low-frequency antenna terminals are switched to ground, thus minimizing direct receiver pick-up at the intermediate frequency. Single-wire or doublet antennas may be used at either high- or low-frequency inputs.

When operating the receiver over its normal frequency range, the converter filaments may be turned off by means of switch S_2. This function also could be accomplished by means of an additional wafer on S_1.

A four-prong-to-four-prong adapter, of the sort used for making tube substitutions, is used on the power cord to enable both it and the receiver cord to be plugged into the HRO power pack simultaneously. With receivers having integral power packs a different arrangement would be required, one possibility being to use a similar plug adapter under one of the power tubes in the receiver, picking up the "B" voltage at the screen-grid pin. A separate supply may be used if desired.

Crystal-Controlled Converter for 144 Mc.

While most converters are used in the manner described above (by leaving the communications receiver set at a given intermediate frequency and tuning the converter over the desired frequency range), it is quite possible to reverse the procedure, using a fixed-frequency oscillator in the converter and tuning the receiver. This approach is particularly advantageous at 144 Mc. and higher, where the selectivity of the tuned circuits is such that no adjustment of the converter circuits is required when the i.f. (in this case usually a broad-band receiver) is varied over a four-megacycle range.

Several converters employing this principle were described by Calvin F. Hadlock, W1CTW, in the May 1946 issue of *QST*. The simplest is shown in Figs. 12-13, 12-14 and 12-15. It uses a 6J6 oscillator-doubler, operating with a 28-Mc. crystal, followed by a 6C4 doubler and a 6AK5 mixer, the grid circuit of which is tuned to 146 Mc. and coupled to the antenna. The plate circuit of the mixer is the input circuit of a receiver (see Fig. 12-14) that tunes the range between 30 and 34 Mc. The converter was designed for use with the National One-Ten, a superregenerative receiver, but it should provide excellent results when used with any

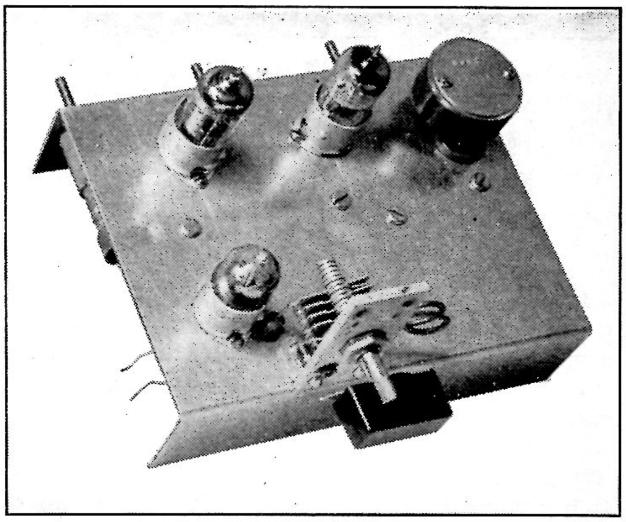


Fig. 12-13 — Top view of the three-tube 144-Mc. converter using a 10-meter crystal. Space is provided at the right of the mixer for addition of an r.f. stage.

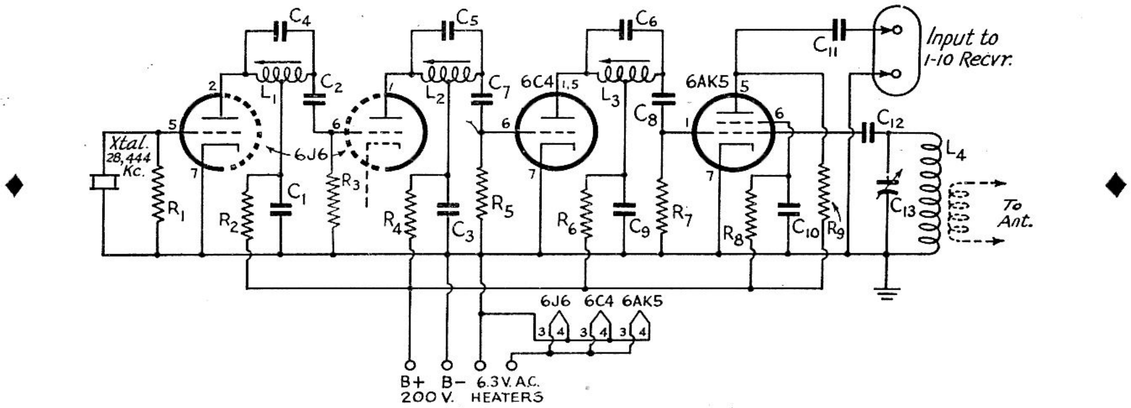


Fig. 12-14 — Schematic of the 3-tube 2-meter converter, using a 28-Mc. crystal.

C_1, C_3, C_9 — 470-μμfd. mica.
C_2, C_7 — 100-μμfd. mica.
C_4, C_6 — 15-μμfd. (10 to 20) ceramic or mica.*
C_5 — 22-μμfd. (15 to 25) ceramic or mica.*
C_8 — 2.2-μμfd. ceramic or mica.
C_{10}, C_{12} — 47-μμfd. mica.
C_{11} — 100-μμfd. mica.
C_{13} — 15-μμfd. variable, National UMA-15.
R_1 — 22,000 ohms, ½ watt.
R_2 — 4700 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 4700 ohms, ½ watt.
R_5 — 0.1 megohm, ½ watt.
R_6 — 4700 ohms, ½ watt.
R_7 — 0.25 megohm, ½ watt.
R_8 — 0.75 megohm, ½ watt.
R_9 — 4700 ohms, ½ watt.
L_1 — XR-50 coil form, ungrooved, 11 turns No. 22 enam., close-wound, center-tapped.
L_2 — XR-50 coil form, ungrooved, 5 turns No. 16 enam., spaced ½ dia. of wire, center-tapped.
L_3 — XR-50 coil form, ungrooved, 3 turns copper strip, 3/32 inch wide, spaced 3/32 inch, center-tapped.
L_4 — 1½ turns of No. 14 copper wire, ½ inch in diameter.
* C_4, C_5 and C_6 should be selected in value so that plugs extend fairly well out from center of coil at resonance.

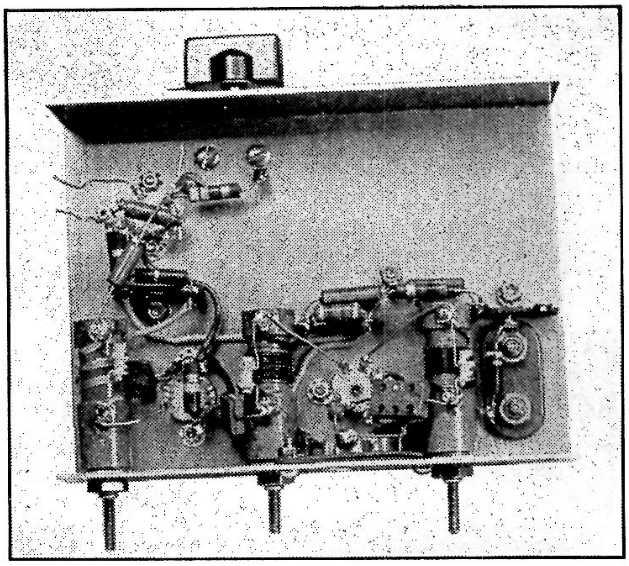


Fig. 12-15 — **Bottom view of the three-tube 2-meter converter. Note the fixed-tuned tank circuits mounted along the back edge of the chassis. The two short leads at the upper left connect to the antenna terminals of a One-Ten receiver.**

of several A-M-FM receivers that are capable of tuning this range.

It is built on a chassis of folded aluminum 6 × 4½ × 1½ inches in size. Space is left on the chassis for addition of an r.f. stage, if desired. The first half of the 6J6 is a conventional triode crystal oscillator, the second half acting as a doubler, driving a 6C4 doubler. With the values shown, the second 6J6 grid will have about 20 volts of excitation, as measured with a high-resistance voltmeter across R_3. The voltage developed across R_5 will be about 25 to 30 volts. The 6C4 doubler provides about 10 volts on the mixer grid before the r.f. input circuit is connected. With the input circuit connected and adjusted to approximately the middle of the 2-meter band, the excitation voltage drops to about 1 volt, which is sufficient for good conversion with the grid-leak injection shown. A very high-resistance voltmeter should be used for these measurements. A 100-microampere meter with a 0.5-megohm resistor in series is suitable.

One-Tube Converter for 144 Mc.

A simple converter employing a single 7F8 tube is shown in Figs. 12-16–12-19. It is designed to work into a communications receiver on either 10.7 or 27.9 Mc., the latter frequency being provided so that the converter may be used with v.h.f. superheterodynes such as the Five-Ten, NHU, S-27, S-36, and others which do not tune to the lower frequency. While it was designed for maximum simplicity, it is capable of outperforming the best superregenerative receivers in weak-signal work. If greater sensitivity is desired, one or two stages of r.f. amplification (Figs. 12-20–12-22) may be added.

From the schematic diagram, Fig. 12-18, it may be seen that one section of the 7F8 dual triode is used as a mixer and the other as a Colpitts oscillator. Stability, an important factor in v.h.f.-converter design, is assured as the result of several precautions. The tuned circuit has a high C/L ratio, the coil is mounted rigidly on the tuning condenser, and the tube is mounted below the chassis to minimize heating effects. The Colpitts oscillator circuit permits grounding the cathode, preventing a.c. hum modulation, a common trouble when the cathode is operated above ground in v.h.f. circuits.

Mechanical Details

No attempt was made to gang the oscillator and mixer tuning controls, as the mixer setting is sufficiently broad so that it may be peaked at 146 Mc. and left in that position for the whole

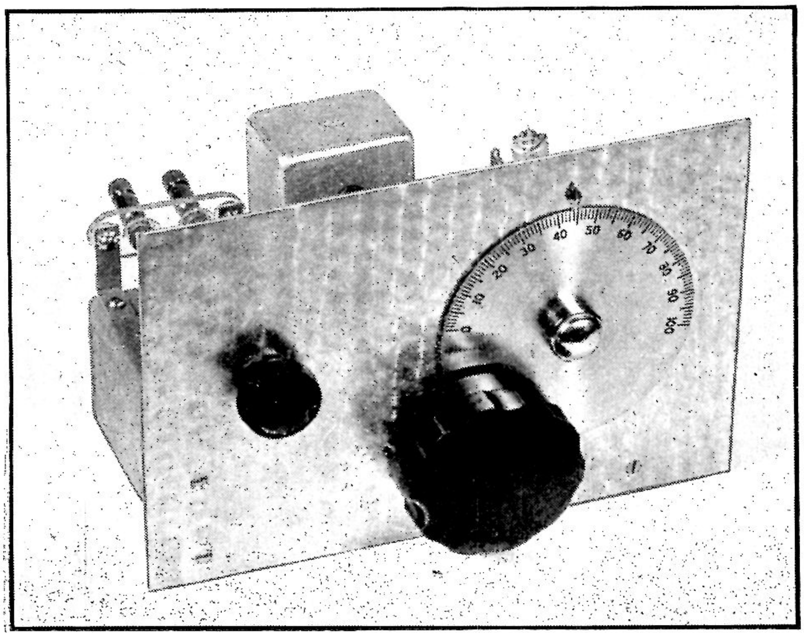


Fig. 12-16 — **Front-panel view of the simple 2-meter converter.**

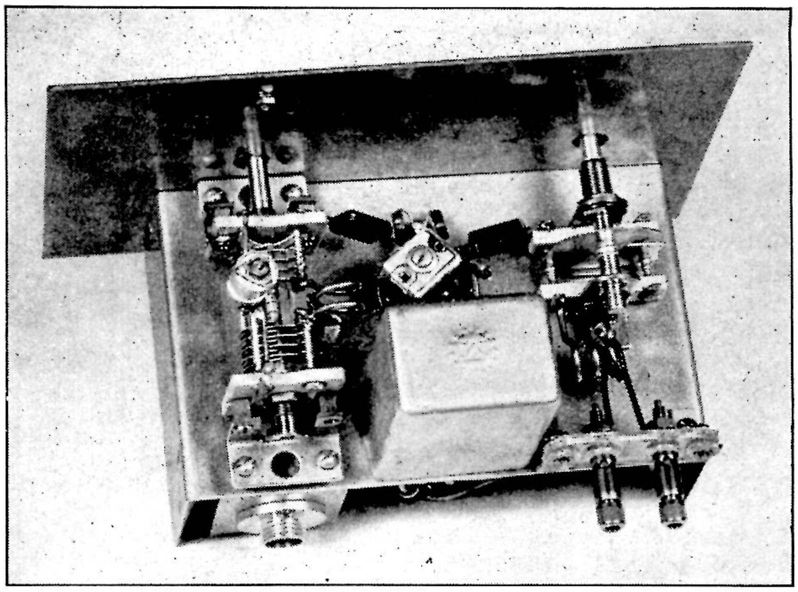


Fig. 12-17 — Top view of the simple 2-meter converter. At the right are the mixer tuned circuit and antenna coupling coil. Oscillator components are at the left. The shield at the rear of the chassis houses the output coupling transformer. The trimmer attached to the 7F8 socket terminals is the oscillator injection condenser, C_4.

band. The oscillator tuning condenser, C_2, a split-stator variable, was made from a Millen Type 21935, originally a 35-μμfd. single-section double-spaced midget variable. A section of the stator bars ¼ inch long is sawed out of the center of the condenser, leaving four stator and five rotor plates in each section. The three extra rotor plates, at the center of the rotor shaft, may be removed with long-nosed pliers. The condenser is mounted with the stator bars at the top, permitting the two-turn coil to be soldered directly to the sawed ends of the bars, for solid mounting. The parallel padder, C_3, is an air trimmer of new design (Silver Type 619), or a mica trimmer may be substituted, if necessary.

The vernier dial used is a National Type K, but a large knob is substituted for the small one with which the dial is equipped, giving the converter tuning a communications-receiver quality. The appearance of the converter was further dressed up by giving the panel a "watch-case" finish. This is done with a small wad of steel wool in a drill press, or it may be done by hand with somewhat more effort.

The tube is mounted with its socket above the chassis, providing short r.f. leads. The arrangement of the smaller parts should be obvious from the photographs. Oscillator injection is controlled by the mica trimmer, C_4, which is mounted directly on the oscillator-plate and mixer-grid prongs of the tube socket. Its setting is not critical; it may be left near the minimum capacitance of the condenser.

The output coupling transformer is housed in a cut-down i.f. shield can, with the mixer plate lead coming out of a hole in the side. Winding data for both 10.7- and 27.9-Mc. transformers are given. The higher frequency is recommended for use wherever possible. The fixed padder, C_7, and the by-pass condenser, C_8, are mounted inside the i.f. shield. Converter output is taken off through a coaxial cable and fitting, though the latter may be eliminated if desired.

Ordinarily the receiver with which the converter is to be used will be capable of supplying the 6.3 volts a.c. at 0.3 amp. and 150 volts d.c. at 5 ma., but a separate supply may be used if desired. If the supply voltage is much over 150 volts, a dropping resistor should be used to bring it down to about that value.

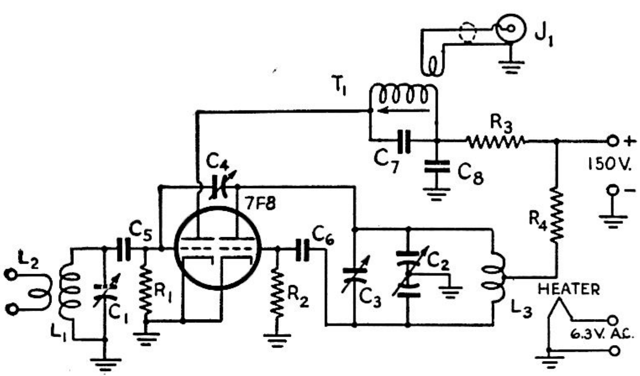


Fig. 12-18 — Schematic diagram of the 7F8 converter for 144 Mc.

C_1 — 10-μμfd. variable.
C_2 — 15-μμfd. per section, split stator. See text.
C_3 — 3–30-μμfd. air trimmer (Silver Type 619).
C_4 — 3–30-μμfd. mica trimmer.
C_5, C_6 — 47-μμfd. mica or ceramic.
C_7 — 27-μμfd. mica or ceramic.
C_8 — 470-μμfd. mica.
R_1 — 1 megohm, ½ watt.
R_2, R_3 — 10,000 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
L_1 — 3 turns No. 12, ⅜-inch i.d., ½ inch long.
L_2 — 2 turns "push-back," ⅜-inch i.d., inserted in cold end of L_1.
L_3 — 2 turns No. 12, ½-inch i.d., spaced ¼ inch, center-tapped.
J_1 — Coaxial jack (Jones S-201).
T_1 — 29.7-Mc. i.f.: 9 turns No. 22 d.s.c. wire, spaced one diameter, on National XR-50 form (slug-tuned). Coupling winding: 2 turns No. 22 d.s.c. wire, interwound in cold end of main winding.
— 10.7-Mc. i.f.: 22 turns No. 22 enameled wire, close-wound on National XR-50 form (slug-tuned). Coupling winding: 3 turns No. 22 enameled wire wound over cold end of main winding. Insulate between two windings with polystyrene or other insulating tape.

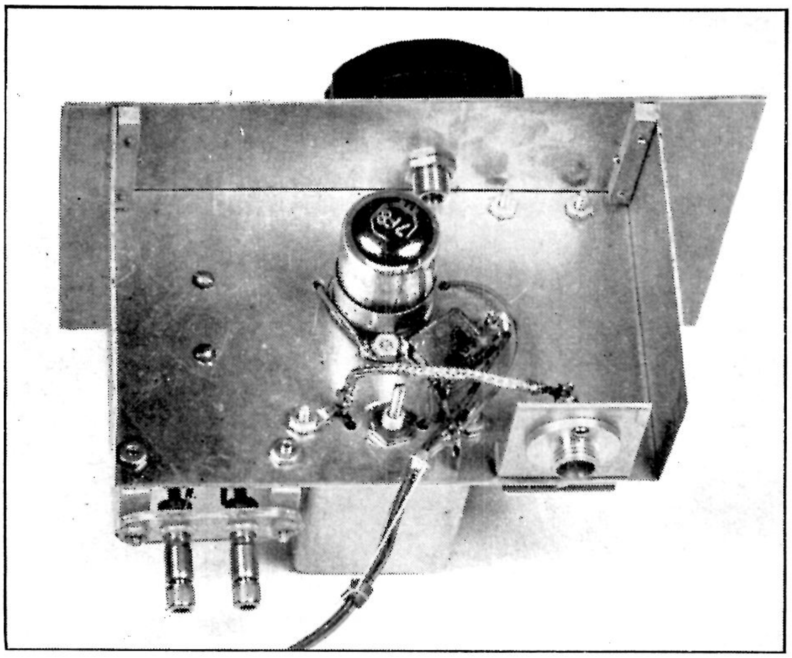


Fig. 12-19 — Bottom view of the converter shows the 7F8 tube mounted below the chassis. The i.f. core adjusting screw is in front of the tube. At the right, on a bracket, is the coaxial fitting for i.f. output.

The converter is built on a folded aluminum chassis made from a sheet 4 × 10 inches in size, 2 inches being folded over on either end. The panel is 5 × 8 inches. The coaxial fitting and the antenna terminals are mounted on small aluminum brackets. The panel is fastened to the chassis by means of ¼-inch-square rods, but small angle brackets would serve equally well.

Adjustments

Adjustment and testing of the converter are simple enough, if a calibrated signal generator is available. Lacking this, harmonics of a VFO, or even the radiation from a receiver oscillator, may be used. A superregenerative receiver, the tuning range of which is known, may also be used as a signal generator. First, the i.f. output transformer should be tuned to 27.9 or 10.7 Mc. by means of its adjustable core. The exact frequency employed is, of course, unimportant, as the coaxial cable between the converter and receiver will prevent appreciable pick-up at the i.f. frequency. If any strong signals are present, the i.f. may be shifted to any clear frequency.

Next, the tuning range of the oscillator should be checked. When the converter is to be used with a 27.9-Mc. i.f., the tuning range of the oscillator will be 116.1 to 120.1 Mc. With a 10.7-Mc. i.f. it will be 133.3 to 137.3 Mc. Either of these ranges can be reached by adjustment of the parallel air trimmer, C_3. Bandspread, with the higher i.f., will be about 80 divisions. With the low i.f. it will be somewhat less. The oscillator may be checked with a calibrated absorption-type wavemeter, or by listening to it in a calibrated receiver.

A strong signal near 146 Mc. should then be tuned in, and the mixer condenser adjusted for maximum response. As the oscillator frequency varies when the mixer tuning is changed, it will be necessary to rock the oscillator dial back and forth across the signal as the mixer tuning is adjusted. Once the proper setting of C_1 has been determined, it may be left set for the entire band. When a sensitive receiver is used as an i.f. system quite good sensitivity will be obtained, and a signal of one microvolt or less will produce a plainly-audible response.

A 144-MC. GROUNDED-GRID R.F. AMPLIFIER

The two-stage r.f. amplifier unit shown in Figs. 12-20 through 12-22 may be used with the simple converter described above, or with any other 2-meter receiver where an improvement in sensitivity and image rejection is desired. It employs two 6J4 triodes in a grounded-grid circuit. Other tubes might also be used, but the 6J4 was designed especially for this service and has internal shielding which reduces the likelihood of self-oscillation troubles. In the grounded-grid circuit, the signal input is to the cathode of the tube instead of the grid, which is connected directly to ground, as the name implies.

In the unit shown, the antenna is coupled to a tuned circuit in the cathode of the first tube, the plate circuit of which is also tuned. Coupling between stages is effected by a small capacity, C_9, which is tapped down on the plate coil, L_3. The cathode of the second stage is

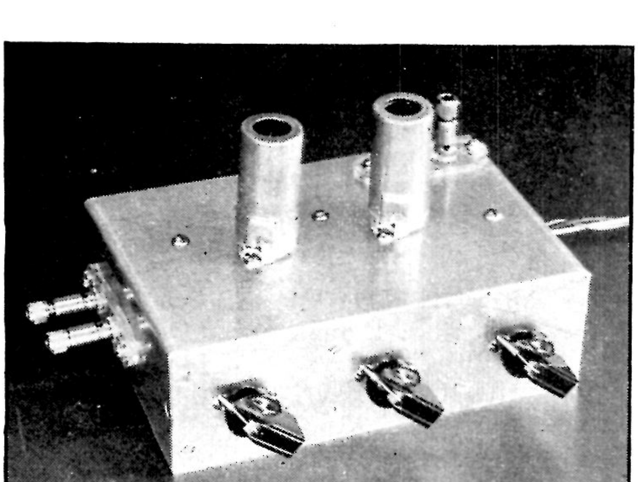


Fig. 12-20 — A two-stage grounded-grid r.f. amplifier for 144 Mc.

returned to ground through an r.f. choke and a bias resistor, R_3. Both cathodes are maintained above ground for r.f. by insertion of r.f. chokes in the heater leads. The plate circuit of the second stage is link-coupled to the mixer, the coupling line being brought out through two National FWG terminals at the back of the chassis. All three tuned circuits are provided with front-panel controls, for ease of adjustment, but only the plate tuning of the second stage will require any readjustment in tuning over the band. The other two controls may be set at 146 Mc. and no noticeable change in signal strength will be obtained if they are repeaked for a signal at either end.

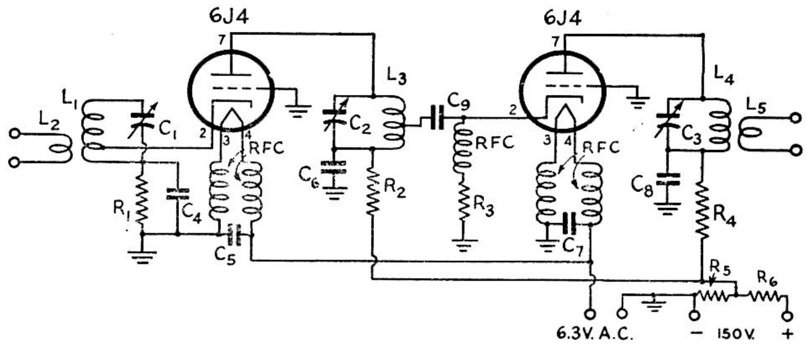


Fig. 12-21 — Schematic diagram of the 6J4 r.f. amplifier.

C_1, C_2, C_3 — 15-$\mu\mu$fd. midget variable (Millen 20015).
C_4, C_5, C_6, C_7, C_8 — 470-$\mu\mu$fd. midget mica.
C_9 — 47-$\mu\mu$fd. mica.
R_1, R_3 — 220 ohms, ½ watt.
R_2, R_4 — 470 ohms, ½ watt.
R_5, R_6 — 10,000 ohms, 10 watts.
L_1 — 4 turns No. 14, ⅜-inch inside diameter, ½ inch long, tapped 1½ turns from cold end.
L_2 — 2 turns "push-back," interwound in L_1.
L_3 — 3 turns No. 14, ⅜-inch inside diameter, ⅜ inch long, center-tapped.
L_4 — Same as L_3, but without tap.
L_5 — 2 turns "push-back," interwound in L_4.
RFC — No. 22 d.s.c. wire closewound on 1-watt carbon resistor. Winding length 1 7/32 inch.

The r.f. amplifier is mounted on a chassis similar to that used in the simple converter. It is 2 × 4 × 6 inches in size, and was bent from a 4 × 10-inch piece of aluminum. The front panel is cut to fit, being approximately 2 × 6 inches in size. Interstage shields are sheets of copper, notched to fit closely over the center of the 6J4 tube sockets, which are mounted in such a position that the cathode and plate terminals come on opposite sides of the shield. The three grid pins (1, 5, 6) are soldered directly to the shield itself, or grounded to soldering lugs under the screws with which the sockets are mounted. All leads should be as short as possible.

The amplifier unit may be operated from the same power supply as that used with the converter, but care should be taken to see that the plate voltage on the 6J4s does not exceed 150 volts. The performance of the unit does not change materially with a considerably-lower voltage, and it was found advisable to operate it from a bleeder tap (R_5 R_6) as shown in the schematic diagram, when the amplifier was used in conjunction with the 7F8 converter and a 150-volt supply.

With the constants shown, the circuits will tune near minimum capacitance. Tuning the circuits to resonance produces a different result for each circuit. The noise output drops appreciably as the first circuit hits resonance; in the second there is only a slight noise change; while in the third the noise *increases* noticeably at resonance. Best results will be obtained if each circuit is adjusted while listening to a signal. If the receiver has an S-meter, the amplifier should be tuned for maximum reading on a medium-strength signal; if no S-meter is available, the adjustments should be made with the a.v.c. off, otherwise small changes in signal level will be difficult to detect. Once C_1 and C_2 have been set near the middle of the band they require no further adjustment, and C_3 may be repeaked for maximum noise as the converter is tuned across the band.

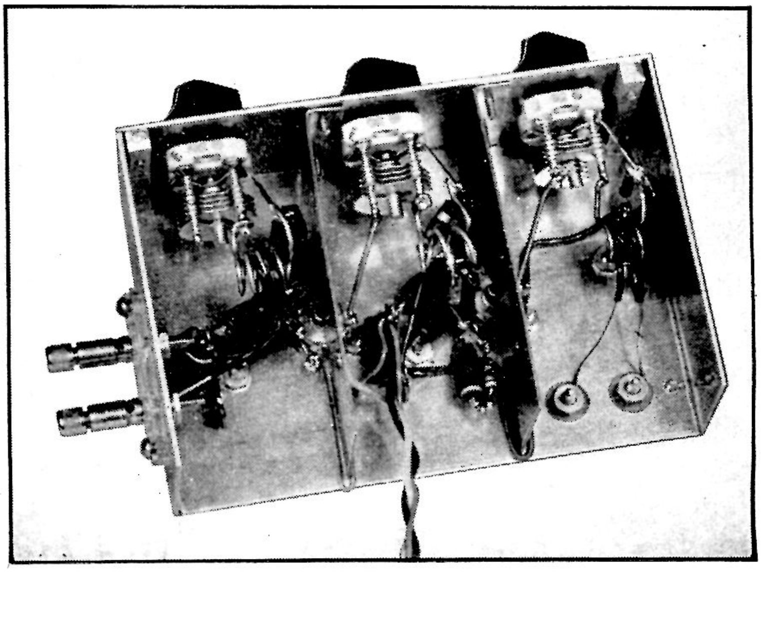


Fig. 12-22 — Bottom view of the grounded-grid r.f. amplifier. At the left are the cathode input circuit and antenna coupling. The center compartment contains the plate circuit of the first stage and cathode circuit of the second. The right-hand compartment houses the plate circuit of the second stage, and the output coupling.

Mobile Receiving Equipment

Probably the most satisfactory method of obtaining reception on 50 and 28 Mc., in cars equipped with broadcast receivers, is the use of a converter of simple design, working into the car receiver at 1600 kc. Some other arrangement must be made for 144 Mc. and higher, however, as the i.f. used for these frequencies must be higher than 2 Mc., to avoid image troubles. One solution, in areas where there is extensive use of crystal-controlled transmitters, would be a converter having an i.f. of about 30 Mc., working into a second mixer-oscillator whose output would be 1600 kc., for working into the car broadcast receiver.

Most mobile reception on frequencies from 144 Mc. up has been done with simple superregenerative units, or, in a few cases, complete receivers designed especially for 144-Mc. work. With either of these approaches it is difficult to achieve satisfactory performance on all three of the popular mobile bands, 2, 6 and 10 meters. One way of attaining this end involves the use of a superheterodyne converter (or converters) and a high-frequency i.f. (in the vicinity of 10–20 Mc.) working into a superregenerative second detector. This is particularly useful for installations where no broadcast receiver is available.

A SUPERREGENERATIVE I.F. AND AUDIO UNIT

The high sensitivity, noise rejection, and a.v.c. characteristics of the superregenerative detector make it useful in mobile operation. The chief difficulties inherent in this type of receiver, broadness of tuning and radiation of an interfering signal, can be overcome at least partially through the use of a superregenerative stage as the second detector in a superheterodyne receiver. The i.f. amplifier and audio unit shown in Figs. 12-23–12-26 was designed especially for mobile operation. Two converters, shown in Figs. 12-23 and 12-27–12-31, working with this unit, provide mobile reception on 2, 6, 10 and 11 meters. The space available in a particular make of car will influence the form factor of the units, but these are representative designs. The two converters, one for 6–11 meters and one for 2 meters, are intended for steering-post mounting, while the i.f.-audio unit is shaped to fit into either a glove or radio compartment.

Little need be said about the i.f. unit, as there are few critical factors, and mechanical layout is relatively unimportant. Only four tubes are used: a 6AG5 11-Mc. i.f. amplifier, a 6C4 superregenerative second detector, a 6C4 first audio amplifier, and a 6AK6 second audio. Note that both audio stages are transformer-coupled, this method having been used in preference to resistance coupling, as experience has shown that the former makes for smooth, quiet operation when superregenerative detectors are employed.

The input stage of the unit should be well shielded, not only to prevent oscillation, but to reduce pick-up on 11 Mc. When the unit is installed in a car this is not troublesome, but in home-station work, 11-Mc. interference can become quite severe, especially during evening hours.

The tuned circuits used in the 11-Mc. amplifier, the superregenerative detector, and as output coupling units in the two converters, are all similar. The coils are wound of No. 22 enameled wire on National XR-50 core-tuned forms, the secondary winding occupying the entire winding space. A simple way of securing the primary is to wrap a layer of Scotch Tape, sticky side *out*, around the ground end of the secondary. The primary winding will then stick as it is wound on, and holding it in place will be no problem. A small tab of tape, or household cement, will suffice.

A CONVERTER FOR 50 AND 28 MC.

The three-tube converter shown in Figs. 12-23, 12-27 and 12-28 covers the 50–54 Mc.

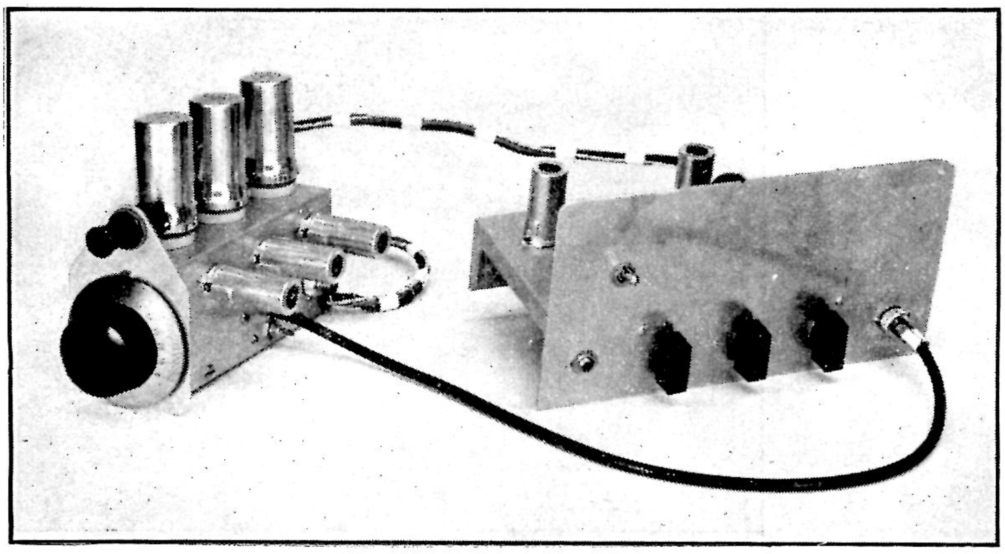


Fig. 12-23 — The three-tube converter for 6 and 10 meters connected to the 11-Mc. i.f. amplifier and audio system. The converter is mounted on the steering post, while t e i.f. unit is designed for glove-compartment mounting. The object above the converter dial is an adjustable-beam dial light.

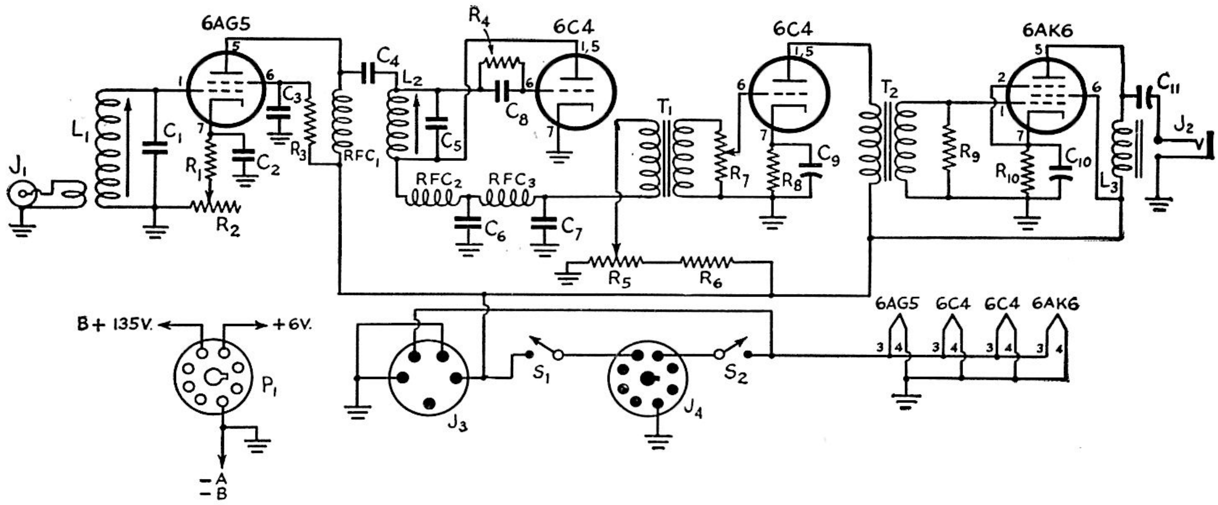


Fig. 12-24 — Wiring diagram of the i.f. unit using a superregenerative second detector and two audio stages.

C_1, C_5 — 47-μμfd. ceramic.
C_2, C_3 — 470-μμfd. midget mica.
C_4, C_8 — 100-μμfd. midget mica.
C_6, C_7 — 0.0068-μfd. mica.
C_9, C_{10} — 25-μfd. 50-volt electrolytic.
C_{11} — 0.1-μfd. 600-volt tubular.
R_1 — 270 ohms, carbon.
R_2 — 10,000-ohm potentiometer.
R_3 — 1000 ohms.
R_4 — 4.7 megohms.
R_5 — 50,000-ohm potentiometer.
R_6 — 47,000 ohms, 1 watt.
R_7 — 0.25-megohm potentiometer.
R_8 — 2200 ohms.
R_9 — 0.22 megohm.
R_{10} — 680 ohms.
All resistors ½-watt type unless otherwise indicated.
L_1, L_2 — 22 turns No. 22 enam., close-wound on National XR-50 form. Primary: 3 turns No. 22 enam. close-wound on layer Scotch Tape over ground end of L_1.
L_3 — Midget filter or audio choke.
J_1 — Coaxial socket (Jones S-201).
J_2 — 'Speaker or headphone jack.
J_3 — 5-prong plug for converter power, mounted on back of chassis.
J_4 — Octal plug, mounted on back of chassis.
P_1 — Octal socket on power cable.
RFC_1 — 2.5-mh. r.f. choke (National R-100).
RFC_2 — One "pie" from National R-100, mounted on 1-watt resistor.
RFC_3 — 80-mh. r.f. choke.
S_1 — S.p.s.t. toggle switch, bat-handle type.
S_2 — S.p.s.t. switch, mounted on R_7.
T_1, T_2 — Midget interstage audio transformers.

and 27–30 Mc. ranges by means of plug-in coils. Using the 11-Mc. intermediate frequency, it is possible to cover the two bands with a common oscillator coil, the oscillator running on the low side of the signal frequency for 50–54 Mc. and on the high side for 27–30 Mc. It is thus merely necessary to change the mixer and r.f. coils when changing bands. Three tubes are used: a 6AK5 r.f. amplifier, a 6AK5 mixer, and a 6C4 oscillator.

The converter layout, shown in Fig. 12-27, makes some sacrifices in accessibility for the sake of compactness; however, by planning the construction carefully, the builder should have no trouble in assembling or adjusting the converter. Parts are mounted on an "L"-shaped aluminum chassis, with a cover of the same general shape, making a case that is 2 inches wide, 3 inches high, and 6½ inches long.

Octal sockets for the plug-in coils (Millen

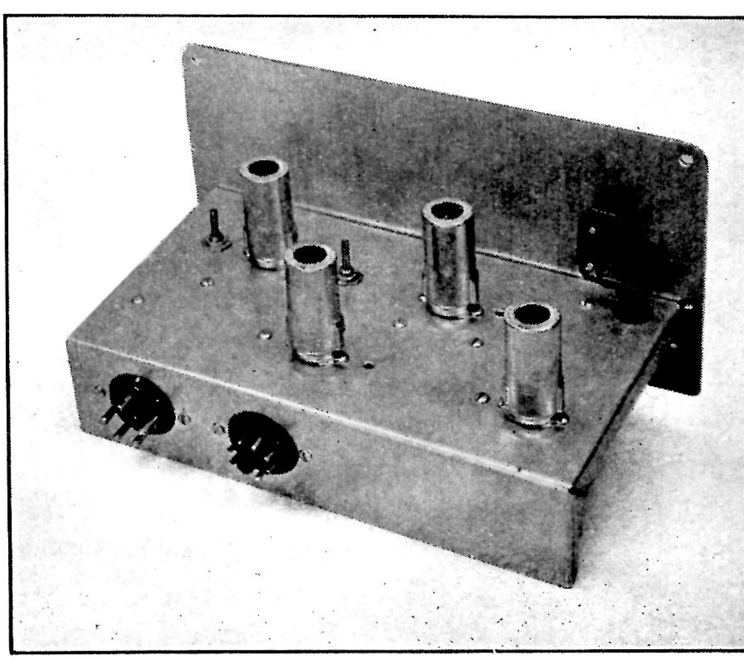


Fig. 12-25 — Rear view of the 11-Mc.-i.f./audio unit. The tubes nearest the panel are the i.f. amplifier, left, and the superregenerative detector. The octal plug on the back of the chassis is for the power cable, while the 5-prong plug connects through another cable to the converter. The toggle switch is the B+ stand-by switch.

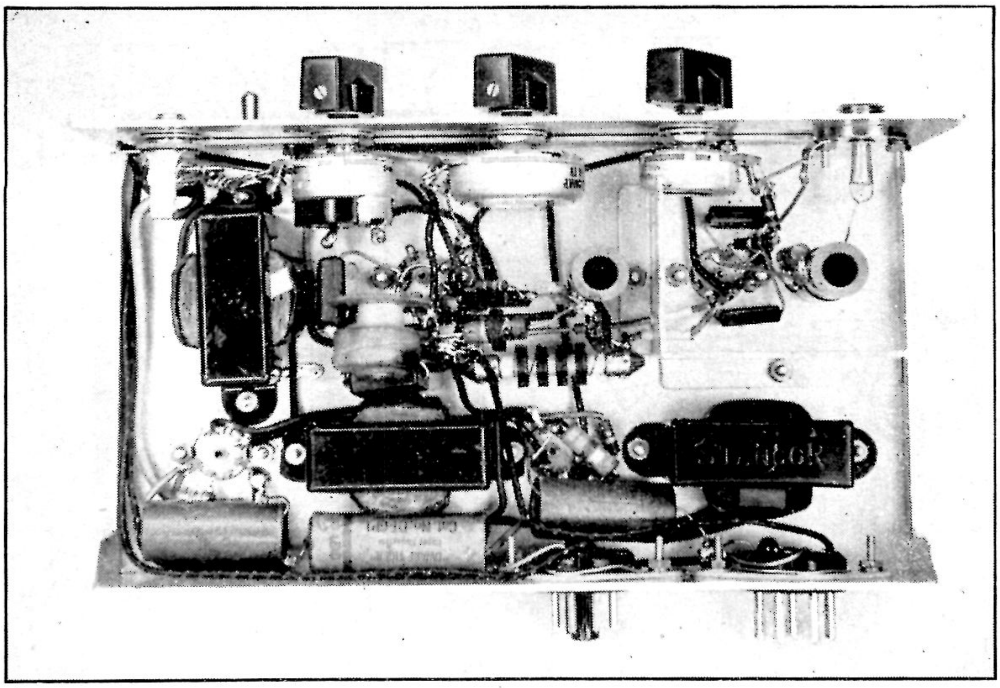


Fig. 12-26 — Bottom view of the i.f.-audio unit, showing arrangement of parts. At the upper right, in a partially-shielded compartment, are the parts comprising the i.f.-amplifier input circuit. In the center are the detector socket and associated parts. At the left and rear are the audio components.

74001 shielded core-tuned forms) are mounted along the top edge, with the corresponding tube sockets projecting from the right side. The oscillator compartment is at the front, nearest the dial — a "must" when flexible couplings are used for ganging. The middle compartment houses the mixer-stage components, including the core-tuned i.f. output coupling transformer. Coupling between the oscillator and mixer is obtained by means of a piece of "push-back" wire which is soldered to the oscillator tuned circuit and then wrapped around the r.f.-plate or mixer-grid lead. The coupling should be set at the lowest value that will provide maximum signal strength. At the back is the r.f. section, which is provided with a coaxial input jack for antenna connection.

As this converter may be used with conventional i.f. systems, provision was made for incorporating a.v.c. Instead of grounding the grid returns from the r.f. and mixer tubes, these returns are brought out, through resistors R_1 and R_5, to a separate pin on the power-cable socket. The corresponding pin in the i.f. unit is connected to ground.

The oscillator circuit is high C, for maximum stability, the capacitance other than that of the variable condenser being supplied by a ceramic padder, consisting of 20-$\mu\mu$fd. and 27-$\mu\mu$fd. units in parallel with the tuning condenser. Adjustable padders are used in the

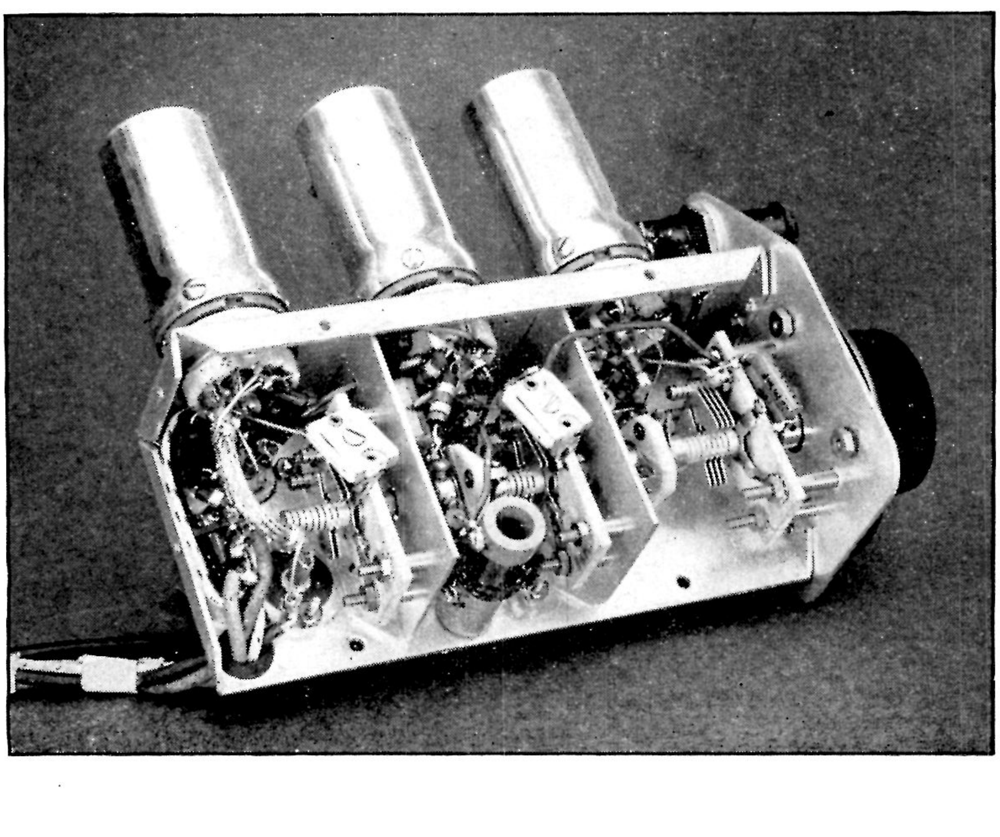


Fig. 12-27 — Interior view of the 28- and 50-Mc. converter, with cover removed. The mica trimmers are adjusted through small holes in the chassis cover. The oscillator compartment is at the front (right), the mixer in the middle, and the r.f. amplifier at the left.

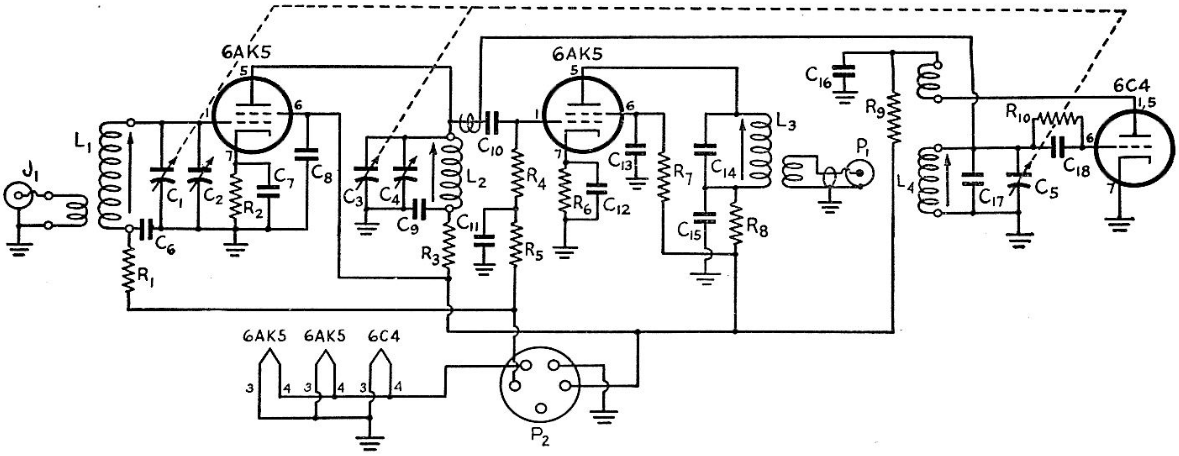


Fig. 12-28 — Schematic diagram of the mobile converter for 27 to 54 Mc.

C_1, C_3 — R.f. and mixer tuning condensers (National UM-15 reduced to 2 stator and 2 rotor plates).
C_2, C_4 — 3–30-μμfd. mica trimmer.
C_5 — Oscillator tuning condenser (National UM-35 reduced to 4 stator and 4 rotor plates).
C_6, C_7, C_8, C_9, C_{11}, C_{12}, C_{13}, C_{15}, C_{16} — 470-μμfd. midget mica.
C_{10}, C_{18} — 100-μμfd. mica.
C_{14} — 47-μμfd. ceramic.
C_{17} — 47-μμfd. ceramic (20 and 27 μμfd. in parallel).
R_1, R_5 — 0.22 megohm.
R_2, R_3, R_8, R_9 — 270 ohms, carbon.
R_4, R_7 — 1.0 megohm.
R_6 — 6800 ohms.
R_{10} — 47,000 ohms.
(All resistors ½-watt rating.)

L_1 — R.f. coil. 28 Mc.: 10 turns No. 22 enam., ¾ inch long. Primary: 2 turns No. 28 d.s.c. interwound in cold end of L_1. 50 Mc.: 5 turns No. 22 enam., ⅜ inch long. Primary similar to 28-Mc. coil.
L_2 — Mixer coil. 28 Mc.: 9 turns No. 22 e., ¾ inch long. 50 Mc.: 4 turns No. 22 e., ⅜ inch long.
L_3 — I.f. output transformer. 22 turns No. 22 enam., close-wound on National XR-50 form. Coupling winding: 2 turns No. 20 "push-back," wound at cold end of L_3.
L_4 — Oscillator coil. 2¼ turns No. 22 enam., 9/16 inch long. Feed-back winding: 2 turns No. 28 d.s.c. interwound between turns of L_4.
J_1 — Coaxial socket (Jones S-201).
P_1 — Coaxial plug (Jones P-201).
P_2 — 5-prong socket on power cable.

mixer and r.f. circuits to facilitate tracking. These are mica trimmers, but the coil inductance is adjusted so that the trimmers tune nearly wide open, and small changes in plate spacing have a negligible effect on the capacitance. Tracking is made easy by the adjustable-inductance feature of the coil forms.

In putting the converter into operation it is best to start by establishing the tuning range of the oscillator, which may be checked with an absorption wavemeter or monitored by a receiver that is capable of tuning from 37 to 43 Mc. It is useful to have the receiver capable of tuning in the high end of the old FM band, so the oscillator may be made to hit 37 Mc. or so at the low-frequency end of its range. If the inductance of the coil is properly adjusted, 43 Mc. (oscillator frequency) will come at the high end. This gives a spread of about 70 divisions for the 50-Mc. band, and about 50 divisions for 27 to 30 Mc. If more spread is desired for the 10-meter band, a separate oscillator coil for that band may be made, and additional padder capacitance built into the r.f. and mixer coils for 10 meters.

◆

Once the oscillator is tuning the desired range, the mixer should be put into operation. For test purposes, a temporary primary may be wound on the mixer coil, using two of the spare pins on the coil and socket for bringing out the leads thereto. From here on, a signal generator which tunes the desired frequency ranges is useful, but it is not absolutely necessary. A signal from a VFO, or the harmonics of several crystals, can be made to serve the same purpose. The signal from the oscillator in a communications receiver can be used also.

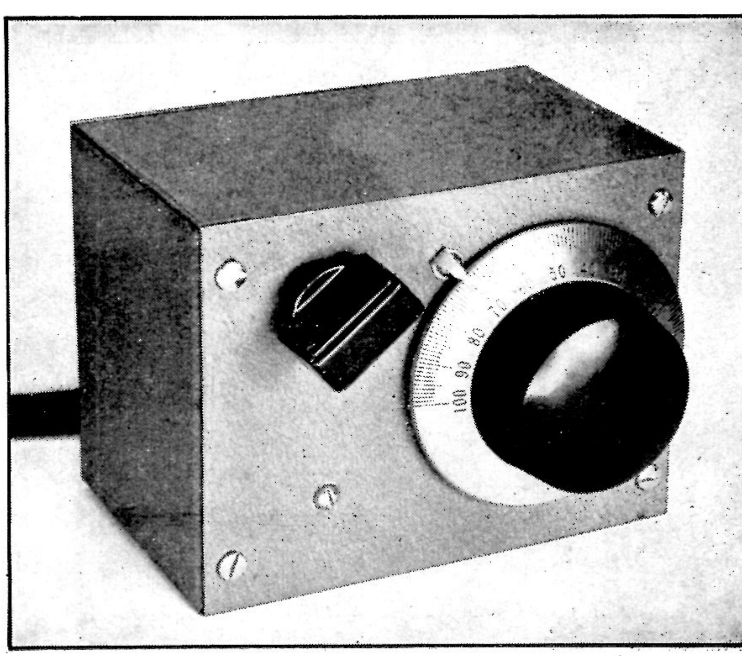


Fig. 12-29 — Front view of the 144-Mc. converter. The entire unit is contained in a standard 3 × 4 × 5-inch case.

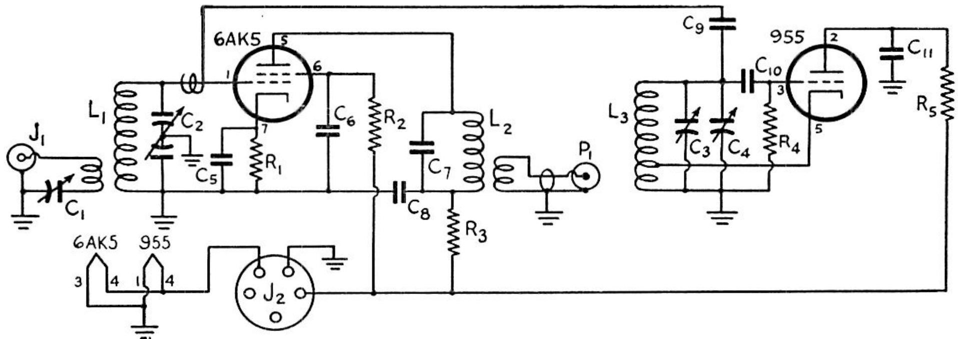


Fig. 12-30 — Schematic diagram of the 144-Mc. converter with 11-Mc. output.

C_1 — 3–30-μμfd. mica trimmer.
C_2 — Cardwell "butterfly" condenser, 1 rotor plate with 1 stator plate on each side. See text.
C_3 — 25-μμfd. trimmer with screwdriver adjustment (Millen 26025).
C_4 — Oscillator tuning condenser (Millen 20015 reduced to 1 stator and 1 rotor plate).
C_5, C_6, C_8, C_{11} — 470-μμfd. mica midget.
C_7 — 47-μμfd. ceramic.
C_9 — 4.7-μμfd. ceramic.
C_{10} — 100-μμfd. mica midget.
R_1 — 10,000 ohms.
R_2 — 1.0 megohm.
R_3 — 270 ohms.
R_4 — 22,000 ohms.
R_5 — 10,000 ohms.
All resistors ½-watt carbon.
L_1 — 3 turns No. 12 tinned, ⅜ inch long, ⅜-inch inside diameter. Primary: 2 turns No. 20 "push-back" interwound at cold end of L_1.
L_2 — 22 turns No. 22 enam., close-wound on National XR-50 form. Coupling winding: 3 turns No. 22 enam. wound on layer of Scotch Tape over cold end of L_2.
L_3 — 3 turns No. 12 tinned, ½ inch long, ¼-inch inside diameter, tapped 1 turn from cold end.
J_1 — Coaxial socket (Jones S-201).
J_2 — 5-prong socket on power cable.
P_1 — Coaxial plug (Jones P-201).

The signal source should be fed into the converter by direct connection to the temporary primary or by means of a pick-up antenna, and the output of the converter fed into a communications receiver tuned to 11 Mc. If the converter is working there will be an appreciable increase in receiver noise as the plate voltage is applied to the mixer, and this will increase as the mixer grid and plate circuits are resonated.

Tracking is accomplished in the usual way, except that no squeezing of turns is required for inductance adjustment. With a signal near the high end of the band, adjust the trimmer, C_4, for maximum signal or noise. Tune to near the low end, and recheck the setting of C_4. If the trimmer capacitance has to be increased, the coil inductance is low; if the capacitance has to be decreased the inductance is too high. Adjust the inductance by moving the core (moving the core inward increases the inductance) and repeat the trimmer-setting process until the band can be tuned without any readjustment of C_4. When the mixer is functioning properly the same procedure should be followed with the r.f. coil. It is well to note the performance of the mixer alone, as this will serve to determine whether the r.f. stage is performing as it should. There should be a noticeable increase in sensitivity when the r.f. stage is added, but if the mixer is functioning correctly it should be possible to get quite good performance with the mixer alone.

It is well to make all converter adjustments with a communications receiver serving as the i.f., as it is difficult to observe minor changes when the superregenerative detector is used, because of its strong a.v.c. characteristics.

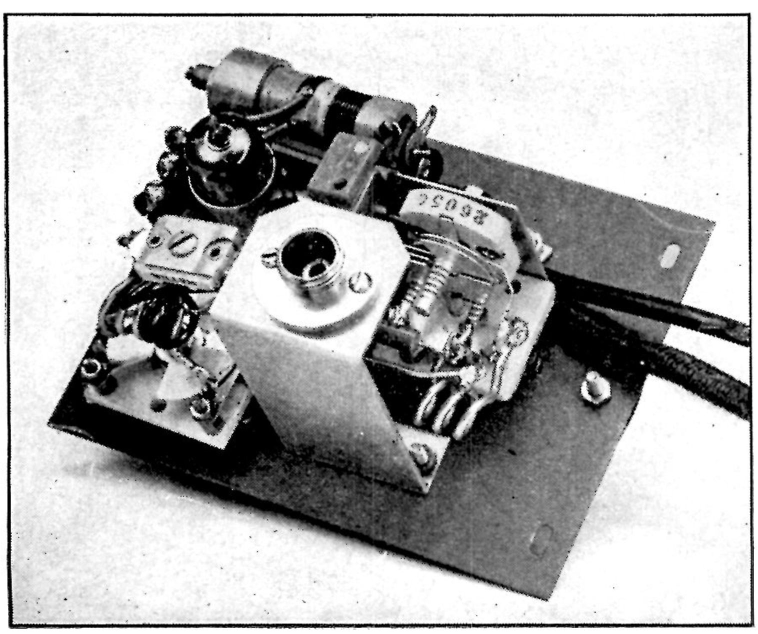


Fig. 12-31 — Back view of the 2-meter converter. Two similar condensers mounted at right angles comprise the tuning assembly for the oscillator stage.

The i.f. system should be peaked at 11 Mc. with a signal generator, and then the converter connected to it for an over-all check. The performance using the superregenerative i.f. unit will be somewhat lower than that of the converter-receiver combination, but ordinarily it should be possible to copy any signal on the mobile set-up that is solidly readable when the communications receiver is used for an i.f. system.

A MOBILE CONVERTER FOR 144 MC.

The circuit of the two-tube 144-Mc. converter, shown in Figs. 12-29–12-31, is similar to the lower-frequency unit, except that the r.f. stage is omitted for the sake of simplicity. Even without the r.f. stage, performance well above that of the better superregenerative receivers is obtainable. The 2-meter converter uses a 6AK5 mixer and a 955 oscillator. Because the mixer tuning is fairly broad, no attempt was made to gang the tuned circuits, and only the oscillator is tuned by the vernier dial. The mixer tuning condenser is provided with a front-panel knob, but once set for maximum signal at 146 Mc., it can be left in the same position for both ends of the band with a negligible sacrifice in sensitivity.

From the schematic diagram, Fig. 12-30, it may be seen that the circuits of the converters are somewhat similar except for the elimination of the r.f. stage and the use of a cathode-tapped coil in the oscillator circuit of the 2-meter unit. The converter was originally laid out using a 6J6 push-push mixer, but because of the difficulty of obtaining satisfactory performance with this arrangement, it was changed to the 6AK5. The "butterfly" tuning condenser used is a hangover from the 6J6 set-up — an ordinary Trim-Aire, with its stator sawed in half, would do.

All the parts are mounted on the front panel, so that the complete unit can be removed from the case intact. Sections of the folded-over edge of the case were sawed out at several points to provide space for easy removal. The oscillator and mixer assemblies are mounted on individual subpanels of folded aluminum, and most of the wiring can be done before these assemblies are fastened to the front panel. The coaxial socket for the antenna connection is mounted on a separate aluminum bracket, and projects through a hole located in the back of the case.

Injection of oscillator voltage is accomplished in a manner similar to that used in the other converter, except that a smaller capacitor must be used, otherwise the oscillator will "pull out" when the mixer circuit is tuned to resonance. A 4.7-$\mu\mu$fd. ceramic condenser is connected to the hot end of the oscillator tuned circuit, and the coupling lead is run from this condenser to the mixer grid lead. By bringing the two tuned circuits closer together, it would be unnecessary to provide any coupling other than that between the two coils.

The oscillator tuning condensers, C_3 and C_4, are similar mechanically, except that one has a shaft to which is affixed the vernier dial, and the other a screwdriver adjustment. It is important that two similar condensers be used in this arrangement, where the two are mounted at right angles, in order that the stators and rotors line up for direct connection without leads. With the condensers and coil used here, the 144-Mc. band covers about 50 divisions on the dial, permitting coverage up to 150 Mc. This is useful, as commercial signals are available in this range in many locations, and they are quite helpful in making necessary receiver adjustments and in judging the condition of the band.

To do a completely-effective job of mobile operation requires considerable attention to noise reduction. With this sort of receiver, the worst interference comes, not from the car's ignition system, but from the generator. The superregenerative detector provides effective silencing for noise pulses of short duration, such as ignition interference, but its inherent a.v.c. characteristics make it respond to a continuous noise such as the whine of the generator, to the exclusion of any weaker signal. It is for this reason that the use of "B" batteries for receiver plate supply is recommended. There is almost certain to be enough noise from any vibrator or generator plate supply to effect at least a slight reduction in the over-all sensitivity of a receiver of this type.

Modes of Operation

Several types of reception are possible through variation in the setting of the regeneration control. With the plate voltage on the detector near maximum, the loudest "shush" and widest bandwidth are obtained. This is the setting normally used for 144-Mc. reception of nonstabilized signals. Backing off the regeneration control reduces the hiss level and sharpens the response, and best all-around reception on 28 or 50 Mc. is usually obtained in this position. Further reduction of the plate voltage results in a whistle being heard as carriers are tuned in, and quite satisfactory c.w. reception is possible at this setting. From here down, the detector is operating in a condition in between superregeneration and straight regeneration for a considerable variation in the plate voltage. It goes into straight oscillation and then out of oscillation entirely as the voltage is reduced nearly to zero. Reception of modulated signals is possible when the detector is operated in a manner similar to that used with regenerative detectors, and "hissless" reception is possible at this point. Sensitivity is considerably lower, however, giving striking proof of the value of superregeneration as a means of attaining high performance with a few tubes.

A Mobile Converter for the Car Receiver

The converter shown in Figs. 12-32–12-36 is designed for use with a mobile broadcast receiver. Two sets of plug-in coils cover 6, 10 and 11 meters, the set not in use being plugged into a pair of dummy sockets in the base of the converter. Power for the unit is obtained from the car-receiver power supply, and a switching arrangement allows a standard car antenna to be used for either broadcast or amateur reception. The performance of the converter may be somewhat below that of more advanced types employing r.f. stages, but it is adequate for mobile operation, where high noise level and the low power of the transmitter are limiting factors.

Circuit Details

As may be seen from the diagram, Fig. 12-36, the converter uses a single 6BE6 pentagrid converter tube, employing electronic injection. A Colpitts oscillator is used, permitting the rotor of the tuning condenser to be grounded and doing away with the need for a cathode tap or tickler coil. The mixer section also uses a grounded-rotor condenser. The output transformer, C_5, L_4 and L_5, is made as a plug-in unit. Switch sections S_{1A} and S_{1B} transfer the antenna from the receiver input circuit to the converter, and, at the same time, connect the transformer output winding, L_5, to the receiver input. Toggle switches S_2 and S_3 are used for heater and stand-by purposes A VR-105 regulates the oscillator plate voltage, preventing oscillator frequency fluctuation which would otherwise result from the voltage variation usually encountered in mobile equipment.

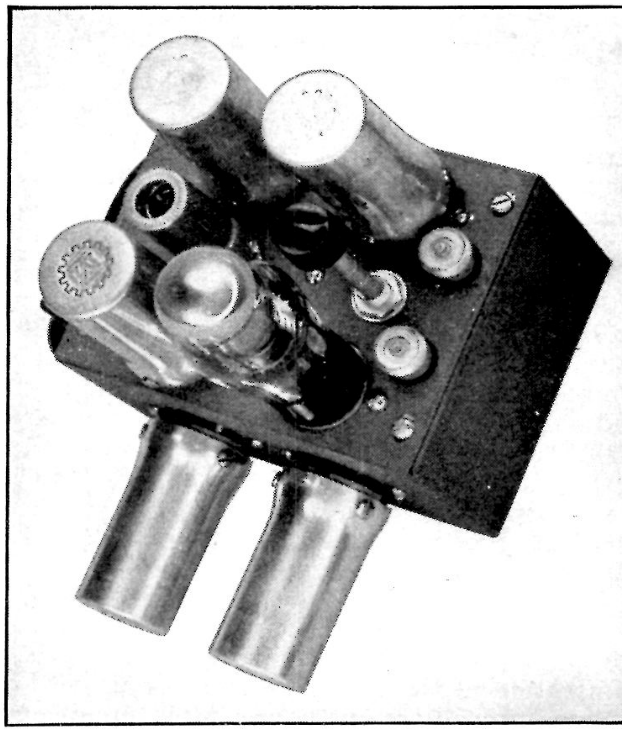


Fig. 12-33 — A side view of the mobile converter. The long shaft on the antenna change-over switch allows the switch to be operated without reaching down in among the other components.

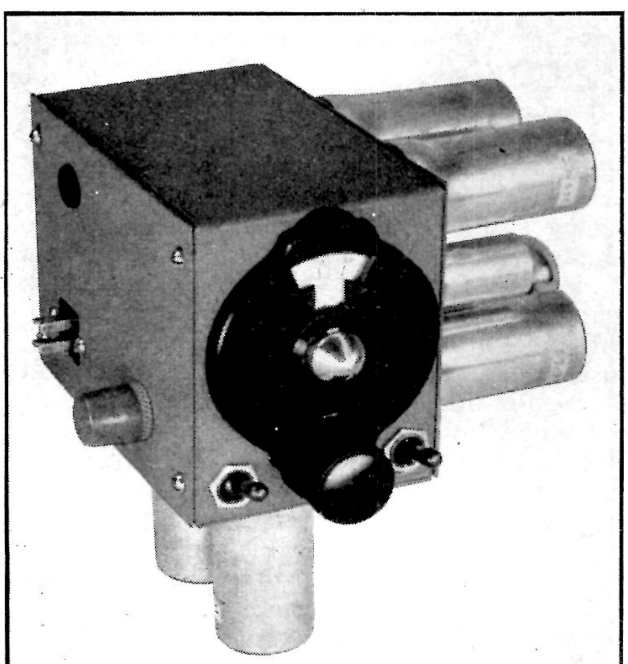


Fig. 12-32 — Front view of the mobile converter. The heater- and plate-voltage switches are to the left and right of the vernier-dial control knob. The pilot light and input connector are at the left side of the case. A hole for screwdriver adjustment of the output transformer is located above the input plug. Tube sockets, mounted in the bottom of the case, are used as holders for the extra set of coils.

Physical Layout

A side view of the converter, Fig. 12-33, shows most of the components mounted on one of the detachable plates of a 3 × 4 × 5-inch utility box. From top to bottom on the left side of the cover plate are the oscillator coil, the 6BE6 tube, and the mixer coil. The output transformer, the antenna switch, and the regulator tube are next in line, with the output jack, J_2, just to the right of the output transformer, and the input jack, J_1, directly below J_2.

The inside view of the unit, Fig. 12-35, shows the parts closely grouped around the tuning condenser, C_{1A}, C_{1B}, the mounting of which requires care in order that the control shaft can be made to line up with the vernier dial which is mounted on the small surface of the case. It is suggested that the layout drawing, Fig. 12-34, be followed as closely as possible. The condenser, mounted after other components have been fastened in place, is equipped with the brackets that are supplied with the Trim-Aire type of condenser. These brackets are attached to the outer surfaces of the ceramic end plates, with the mounting lips

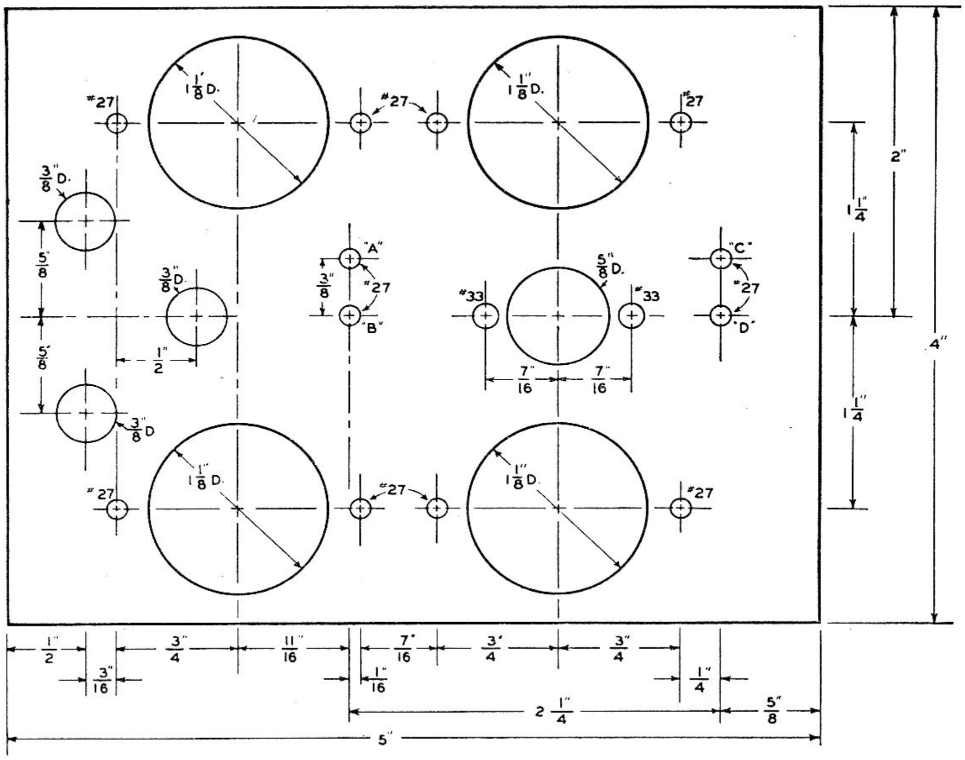


Fig. 12-34 — A mechanical drawing of the subchassis layout.

facing toward the back of the condenser. Between the brackets and the panel are metal pillars $1\frac{1}{16}$ inch long. The position of the condenser is such that the stator terminals are just above the tube- and coil-socket prongs.

It should be possible to follow the layout shown, as there is no excessive crowding of parts. Looking at Fig. 12-35, the oscillator components are at the left of the tuning condenser, with the mixer parts at the right. The output transformer is at the lower left and the regulator tube socket at the right, with the antenna switch in between. The band-set condensers are soldered directly to the coil-socket prongs. All four sockets have extra prongs, which may be used as tie-points for other components.

The main part of the utility box needs some modification before the dial, switches and spare-coil sockets are mounted. It is necessary to fold back the two ½-inch flanges which are at the top and bottom edges of the right side of the case, and the flange at the rear must be filed out in two places to provide clearance for jacks J_1 and J_2.

Details of the coils are given in Fig. 12-36. The oscillator and mixer coils employ Millen No. 74002 shielded plug-in forms, and the output transformer uses the No. 74001 permeability-tuned form of similar construction. These were selected because of the protection their design affords against the rough handling to which such equipment is subjected. The output transformer is designed to resonate near 1500 kc., and is core-adjusted to this frequency.

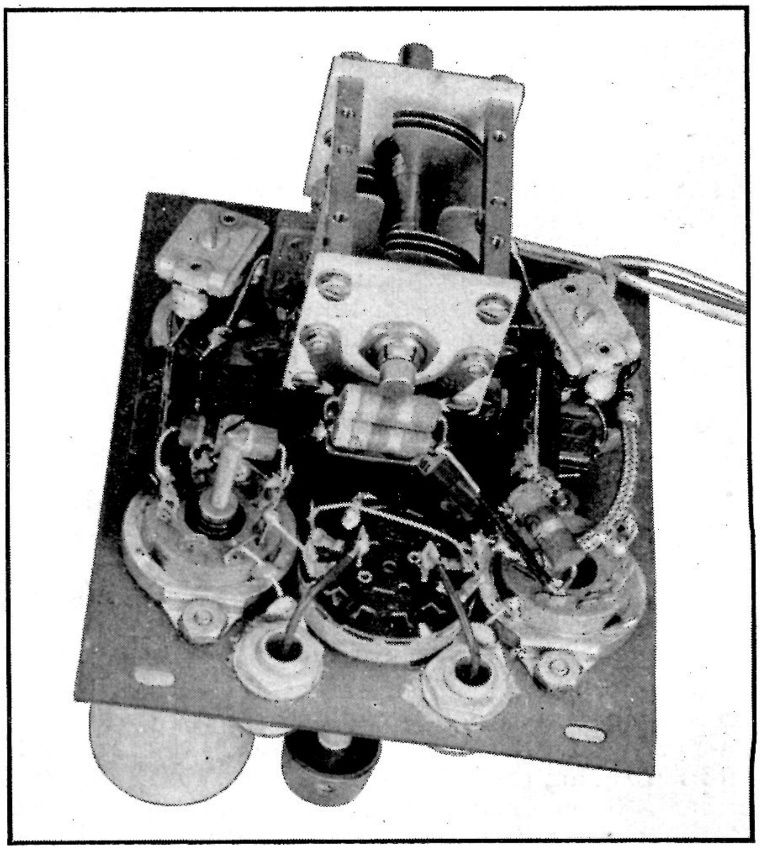


Fig. 12-35 — Inside view of the 6-, 10- and 11-meter mobile converter. Practically all of the construction and wiring can be completed before the subchassis is attached to the case.

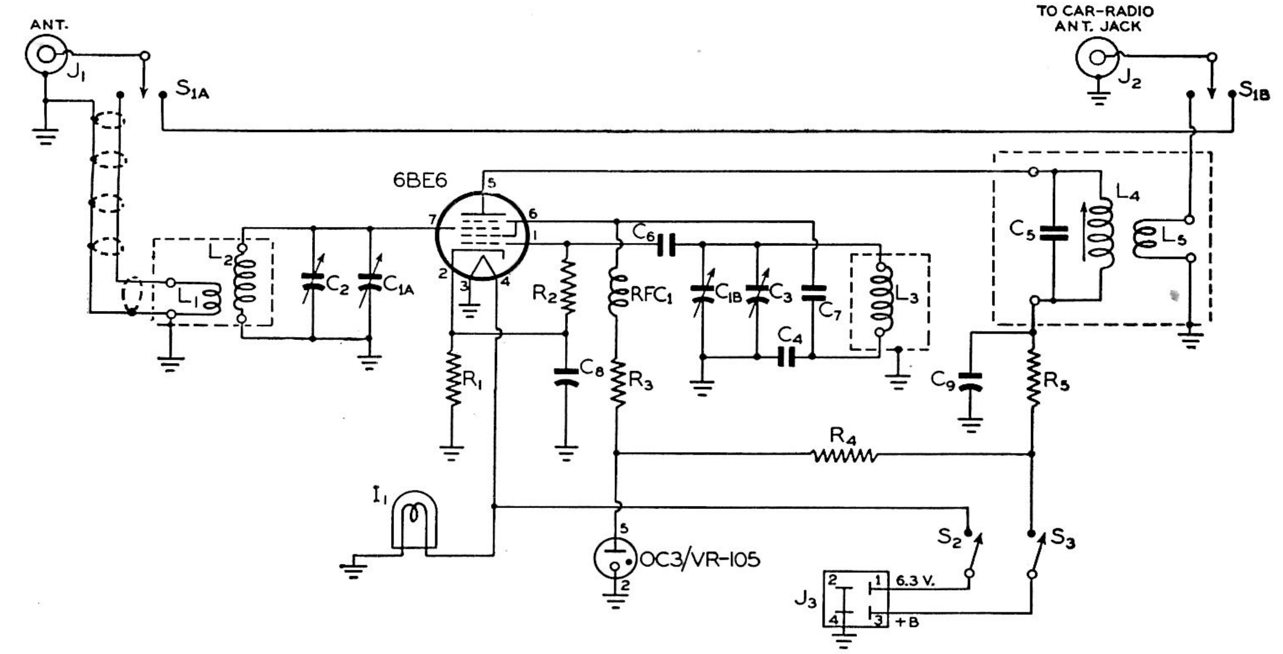


Fig. 12-36 — Wiring diagram of the mobile converter for 6, 10 and 11 meters.

C_{1A}, C_{1B} — Split-stator condenser, 15 μμfd. per section (Cardwell ER-15-AD).
C_2, C_3 — 3–30-μμfd. ceramic trimmer.
C_4, C_5 — 100-μμfd. midget mica.
C_6, C_7 — 0.0022-μfd. mica.
C_8, C_9 — 0.01-μfd. paper tubular.
R_1 — 150 ohms, ½ watt.
R_2 — 22,000 ohms, ½ watt.
R_3 — 680 ohms, 1 watt.
R_4 — 2200 ohms, 2 watts.
R_5 — 1000 ohms, 1 watt.
I_1 — 6.3-volt pilot lamp.
J_1, J_2 — Coaxial-cable jack.
J_3 — 4-prong male plug.
S_{1A}, S_{1B} — 2-pole 2-circuit selector switch.
S_2, S_3 — S.p.s.t. toggle switches.
L_1 — 50 Mc.: 2⅓ turns No. 18 d.c.c. wire, interwound in cold end of L_2.
— 28 Mc.: 3¼ turns No. 18 d.c.c. wire at cold end of L_2.
L_2, L_3 — 50 Mc.: 4½ turns No. 18 enameled wire, spaced one diameter.
— 28 Mc.: 11½ turns No. 18 enameled wire, 9/16 inch long.
L_4 — No. 36 d.s.c. wire, close-wound, one inch long.
L_5 — 15 turns No. 36 d.s.c. wire, close-wound over B-plus end of L_4. Note that condenser C_5 is mounted inside the coil shield.

Alignment and Installation

The converter may be lined up using an a.c. power supply and a communications receiver. A good signal generator simplifies this task, but if none is available the job may be done as follows: Turn on the receiver, set the volume control at maximum, and with the receiver dial at 1500 kc. adjust the output transformer on the converter for greatest noise. With the 10-meter coils in place and the tuning condenser set at about half capacitance, set the oscillator trimmer near maximum capacitance. With the aid of a signal, which may be noise from ignition or an electric razor, adjust the mixer trimmer for maximum response. Final adjustments can be made using amateur signals, resetting the oscillator trimmer as may be required to bring the band at the desired dial settings. The procedure outlined should be followed for the 6-meter range as well. Bandspread will be approximately 60 divisions for the 50-Mc. band and 85 for 27–29.7 Mc.

The exact anode voltage for the 6BE6 will not be known until the converter is actually tested in the car installation. The tube is designed to operate with 250 volts on the output plate and 100 volts on the oscillator plate. With a 150-ohm cathode resistor the cathode current is approximately 10 ma. It is important that the regulator tube be allowed to function normally at all times, and in cases where the receiver voltage is abnormally low it may be necessary to change the value of the limiting resistor, R_4. The proper value will permit the tube to regulate (as indicated by a constant glow) over the complete voltage range, as the input voltage varies with the condition of the car battery and the charging rate of the car generator.

In modifying the car receiver it is necessary to remove it from its case and install a power plug for carrying the heater and plate voltages and the ground lead to the converter. The heater lead should be connected directly to one of the receiver tube sockets, to take advantage of any 6-volt line filtering which may be incorporated in the receiver. Similarly, it is wise to tap for plate power at a point where the r.f. circuits are most heavily decoupled. This may result in a somewhat lower plate voltage than the maximum obtainable, but the extra filtering aids in eliminating hum and vibrator hash.

Antenna and power cables should be made up at this time. One coaxial cable must reach from the car antenna to the converter and the second from the converter to the receiver antenna connection. The 3-wire shielded power cable will connect between the receiver output plug and the 4-prong plug at the converter end.

In putting the converter into operation in the car it may be necessary to readjust the output transformer slightly, and retune the r.f. trimmer in the broadcast receiver slightly to compensate for loading effect of the converter. It may also be found that, although the broadcast receiver may be quiet in its operation, there may be considerable noise from ignition and the car generator when the converter is used. Even with the best available filters and suppressors the ignition noise may still be excessive, in which case the best solution to the problem is the installation of some form of noise silencer in the car receiver. Suitable noise silencers are completely described in Chapter Five.

Wide-Band FM Reception

Wide-band FM may be used in the 50-Mc. band above 52.5 Mc., and elsewhere throughout the v.h.f. and higher bands. It represents an excellent means of obtaining high-quality noise-free reception at these frequencies, where the width of the passband required is not a problem. It shares with narrow-band FM an almost complete freedom from the broadcast-interference problems that plague the urban amateur who uses amplitude modulation. A receiver designed for wide-band FM is also useful in work with stations using modulated-oscillator type transmitters, on 144 Mc. and higher. Standard FM broadcast receivers may be used in conjunction with suitable converters for wide-band FM reception on any frequency. This technique is applicable up through the microwave range, and several of the workers in the amateur microwave bands have used wide-band FM detection in their pioneering efforts in this field. Receiving techniques for narrow-band FM are detailed in Chapter Five.

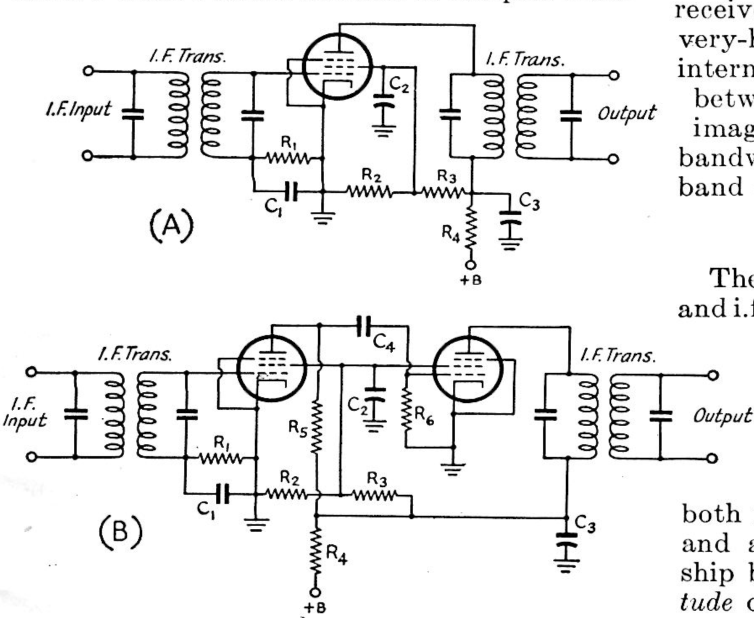


Fig. 12-37 — **FM limiter circuits. A, single-tube plate-saturation limiter; B, cascade limiter. Typical values are:**

	Circuit A	*Circuit B*
C_1 —	100 $\mu\mu$fd.	100 $\mu\mu$fd.
C_2, C_3 —	0.1 μfd.	0.1 μfd.
C_4 —		220 $\mu\mu$fd.
R_1 —	0.1 megohm.	47,000 ohms.
R_2 —	2200 ohms.	2200 ohms.
R_3 —	47,000 ohms.	47,000 ohms.
R_4 —	0–50,000 ohms.	0–50,000 ohms.
R_5 —		3900 ohms.
R_6 —		0.22 megohm.

Plate-supply voltage is 250 in both circuits.

FM RECEIVERS

A frequency-modulation receiver differs in circuit design from one designed for amplitude modulation chiefly in the arrangement used for detecting the signal. Detectors for amplitude-modulated signals do not respond to frequency modulation. It is also necessary, for full realization of the noise-reducing benefits of the FM system, that the signal applied to the detector be completely free from amplitude modulation. In practice, this is attained by preventing the signal from rising above a given amplitude by means of a limiter. Since the weakest signal must be amplitude-limited, high gain must be provided ahead of the limiter; the superheterodyne type of circuit almost invariably is used to provide the necessary gain.

The r.f. and i.f. stages in a superheterodyne for FM reception are practically identical in circuit arrangement with those in an AM receiver. Since the use of FM is confined to the very-high frequencies (above 29 Mc.) a high intermediate frequency is employed, usually between 4 and 5 Mc. This not only reduces image response but also provides the greater bandwidth necessary to accommodate wide-band frequency-modulated signals.

Receiver Requirements

The primary requirements are sufficient r.f. and i.f. gain to "saturate" the limiter even with a weak signal, sufficient bandwidth to accommodate the full frequency deviation either side of the carrier frequency without undue attenuation at the edges of the band, a limiter circuit that functions properly on both rapid and slow variations in amplitude, and a detector that gives a linear relationship between frequency deviation and *amplitude* output. The audio circuits are the same as in other receivers, except that in communications-type receivers it is desirable to cut off the upper audio range by a low-pass filter because higher-frequency noise components have the greatest amplitude in an FM receiver

The Limiter

Limiter circuits generally are of the plate-saturation type, where low plate and screen voltage are used to limit the plate-current flow at high signal amplitudes. Fig. 12-37A is a typical circuit. The tube is self-biased by a grid leak, R_1, and condenser, C_1. R_2, R_3 and R_4 form a voltage divider which puts the desired voltages on the screen and plate. The lower the voltages the lower the signal level at which limiting occurs, but the r.f. output voltage of the limiter also is lower. C_2 and C_3 are the plate and screen by-pass condensers, of conventional value for the intermediate frequency used. The time constant of R_1C_1 determines the behavior of the limiter with respect to rapid and slow amplitude variations. For best operation on impulse noise the time constant should be small, but a too-small time constant limits the range of signal strengths the limiter can handle without departing from the constant-output condition. A larger time constant is better in this respect but is not so effective for rapid variations. Compromise constants are shown in Fig. 12-37.

The cascade limiter, Fig. 12-37B, overcomes this by making the time constant in the first grid circuit suitable for effective operation on impulse noise, and that in the second grid (C_4R_6) optimum for a wide range of input signal strengths. This results, in addition, in more constant output over a very wide range of input signal amplitudes because the voltage at

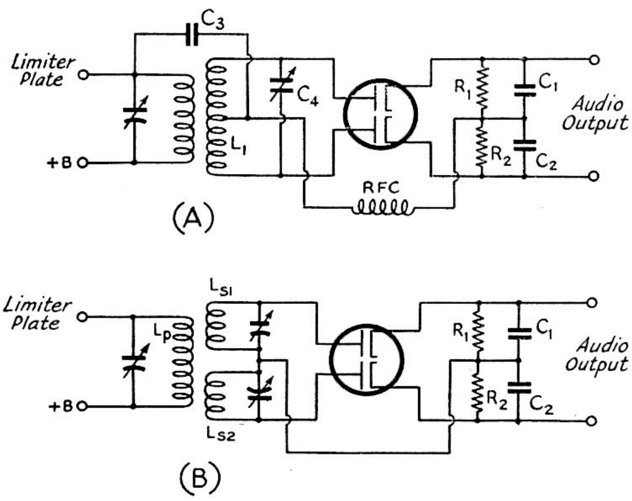


Fig. 12-38 — **FM discriminator circuits. In both circuits typical values for C_1 and C_2 are 100 $\mu\mu$fd. each; R_1 and R_2, 0.1 megohm each. C_3 in A is approximately 50 $\mu\mu$fd., depending upon the intermediate frequency; *RFC* should be of a type designed for the i.f. in use (2.5 millihenrys is satisfactory for intermediate frequencies of 4 to 5 Mc.).**

the grid of the second stage already is partially amplitude-limited. Resistance coupling ($R_5C_4R_6$) is used for simplicity and to prevent unwanted regeneration, additional gain at this point being unnecessary.

The rectified voltage developed across R_1 in either circuit may be applied to the i.f. amplifier for a.v.c.

Discriminator Circuits and Operation

The FM detector commonly is called a *discriminator*, because of its ability to discriminate between frequency deviations above and below the carrier frequency.

A rectifier connected to an ordinary tuned circuit adjusted so that the signal frequency falls on one side of the response curve constitutes an elementary discriminator, because the rectifier output will vary with a change in the carrier frequency. If two such circuits are used with a balanced rectifier, one tuned above and the other below the signal frequency, amplitude variations are balanced out and the combined rectified current is proportional to the frequency deviation.

The circuit most widely used is the "series" or center-tuned discriminator shown in Fig. 12-38A. A special i.f. coupling transformer is used between the limiter and detector. Its secondary, L_1, is center-tapped and is connected back to the plate side of the primary circuit, which otherwise is conventional. C_4 is the tuning condenser. The load circuits of the two diode rectifiers ($R_1C_1R_2C_2$) are connected in series; constants are the same as in ordinary diode detector circuits. Audio output is taken from across the two load resistances.

The primary and secondary circuits are both adjusted to resonance in the center of the i.f. passband. The voltage applied to the rectifiers consists of two components, that induced in the secondary by the inductive coupling and that fed to the center of the secondary through C_3. The phase relations between the two are such that at resonance the rectified load currents are equal in amplitude but flow in opposite directions through R_1 and R_2, hence the net voltage across the terminals marked "audio output" is zero. When the carrier deviates from resonance the induced secondary current either lags or leads, depending upon whether the deviation is to the high- or low-frequency side, and this phase shift causes the induced current to combine with that fed through C_3 in such a way that one diode gets more voltage than the other when the frequency is below resonance, while the second diode gets the larger voltage when the frequency is higher than resonance. The voltage appearing across the output terminals is the difference between the two diode voltages. Thus a characteristic like that of Fig. 12-39 results, where the net rectified output voltage has opposite polarity for frequencies on either side of resonance, and up to a certain point becomes greater in amplitude as the frequency deviation is greater. The straight-line portion of the curve is the useful detector characteristic. The separation between the peaks that mark the ends of the linear portion of the curve depends upon the Qs of the primary and secondary circuits and the degree of coupling. The separation becomes greater with low Qs and close coupling. The circuit ordinarily is designed so that the peaks fall just outside the limits of the passband, thus utilizing most of the straight portion of the curve. Since the audio output is proportional to the change in d.c. voltage with deviation, it is advantageous for maximum output to keep the frequency separation between peaks down to the minimum value necessary for a linear characteristic.

A second type of discriminator is shown in Fig. 12-38B. Two secondary circuits are used, one tuned above the center frequency of the i.f. passband and the other below. They are coupled equally to the primary, which is tuned to the center frequency. As the carrier frequency deviates the voltages induced in the secondaries will change in amplitude, the larger voltage appearing across the secondary being nearer resonance with the instantaneous frequency. The detection characteristic is similar to that of the center-tuned discriminator. The peak separation is determined by the Qs of the circuits, the coefficient of coupling, and the tuning of the secondaries. High Qs and loose coupling are required for close peak separation.

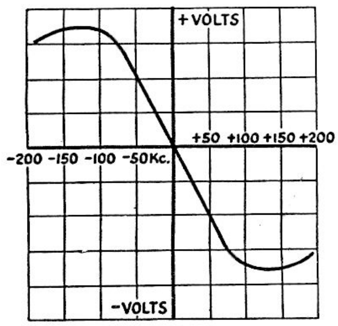


Fig. 12-39 — Characteristic of a typical FM detector. The vertical axis represents the voltage developed across the load resistor as the frequency varies from the exact resonance frequency. This detector would handle FM signals up to a bandwidth of 150 kc. over the linear portion of the curve.

A simple self-quenched superregenerative receiver may be used as a frequency detector if it is tuned so that the carrier frequency falls along the slope of the resonance curve. Two such detectors, off-tuned on either side of the carrier, may be used in push-pull. An alternative arrangement employing a superregenerative stage as a first i.f. amplifier at 75 Mc., following a converter unit, provides high gain and linear response with relatively few stages.

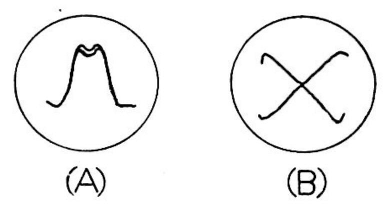
FM-Receiver Alignment

Alignment of FM receivers up to the limiter is similar to other superheterodynes. For output measurement, a 0–1 milliammeter or 0–500 microammeter should be connected in series with the limiter grid resistor (R_1 in Fig. 12-37A) at the grounded end; or, if the voltage drop across R_1 is used for a.v.c. and the receiver is provided with a tuning meter, the tuning meter may be used as an output meter. An accurately-calibrated signal generator or test oscillator is desirable, since the i.f. should be aligned to be as symmetrical as possible; that is, the output reading should be the same for any two test-oscillator settings the same number of kilocycles above or below resonance. It is not necessary to have uniform response over the whole band to be received, although the output at the edges of the band (limit of deviation of the transmitted signals) should not be less than 25 per cent of the voltage at resonance. In communications work, a bandwidth of 30 kc. or less (15 kc. or less deviation) is commonly used. Output readings should be taken with the oscillator set at intervals of a few kilocycles either side of resonance up to the band limits.

After the i.f. (and front-end) alignment, the limiter operation should be checked. This can be done by temporarily disconnecting C_3, if the discriminator circuit of Fig. 12-38A is used, disconnecting R_1 and C_1 on the cathode side, and inserting the milliammeter or microammeter in series with R_2 at the low end. This converts the discriminator into an ordinary diode rectifier. Varying the signal-generator frequency over the channel, with the discriminator transformer adjusted to resonance, should show no change in output (at the bandwidths used for communications purposes) as indicated by the rectified current read by the meter. At this point various plate and screen voltages can be tried on the limiter tube or tubes, to determine the set of conditions that gives maximum output with adequate limiting (no change in rectified current).

When the limiter has been checked the discriminator connections can be restored, leaving the meter connected in series with R_2. Provision should be made for reversing the connections to the meter terminals, to take care of the reversal in polarity of the net rectified current. Set the signal generator to the center frequency of the band and adjust the discriminator-transformer trimmer condensers to resonance, which will be indicated by zero rectified current. Then set the test oscillator at the deviation limit on one side of the center frequency, and note the meter reading. Reverse the meter terminals and set the test oscillator at the deviation limit on the other side. The two readings should be the same. If they are not, they can be made so by a slight adjustment of the primary trimmer. This will necessitate rechecking the response at resonance to make sure it is still zero. Generally, the secondary trimmer will chiefly affect the zero-response frequency, while the primary trimmer will have most effect on the symmetry of the discriminator peaks. A detector curve having satisfactory linearity can be obtained by cut-and-try adjustment of both trimmers.

Fig. 12-40 — Oscilloscope patterns in FM i.f. alignment. A — I.f.-amplifier response. B — Over-all characteristic through the FM detector.



A visual curve tracer is particularly advantageous in aligning the wide-band i.f. amplifiers of FM receivers. The i.f. is first aligned with the discriminator circuit converted into an AM diode detector, as described above, the pattern appearing as in Fig. 12-40A. The over-all characteristic, including the FM detector, is shown in Fig. 12-40B.

V.H.F. Transmitters

Beginning with the v.h.f. region, frequency assignments are no longer in direct harmonic relationship. This fact, coupled with the necessity for extreme care in selection and arrangement of components for low circuit capacitance and minimum lead inductance, makes it highly desirable to construct separate r.f. equipment for v.h.f. work, rather than attempt to adapt for v.h.f. use a transmitter designed for the lower frequencies.

Transmitter stability requirements for 50 Mc. are the same as for the lower-frequency bands, and, by careful attention to component placement, a rig may be made to serve well on 50, 28, and even 14 Mc., but incorporation of 50 Mc. and higher in the usual "all-band" transmitter is not generally feasible.

At 144 Mc. and higher, no restrictions are imposed on transmitter stability, except that the whole emission must be kept within the band limits. This permits the use of modulated-oscillator transmitters, and many of the stations now working on 144 Mc. and above still employ this simple type of gear. By proper choice of tubes and circuits, crystal control is applicable to 144 Mc. however, and the greatly-increased occupancy of the band in metropolitan areas makes stabilization of at least the higher-powered stations almost mandatory, if the full possibilities of the band are to be realized. Crystal control, or its equivalent, may even be employed on 235 and 420 Mc., but the use of these frequencies has not reached the point where stabilization is particularly important.

Above 51 Mc., and higher throughout the v.h.f. and u.h.f. regions, frequency modulation as well as amplitude modulation is permitted by the amateur regulations. The 200-watt transmitter for 50 and 144 Mc. described in this chapter makes provision for the use of FM, and any crystal-controlled transmitter can be adapted for FM by using a frequency-modulated oscillator to replace the crystal, in the manner described in Chapter Nine.

At 420 Mc. and higher, most standard transmitting tubes cannot be used with any degree of success. Instead, special tubes designed for these frequencies must be employed. Such tubes have extremely-close electrode spacing, to reduce transit-time effects, and are constructed with leads having virtually no inductance. Several more-or-less-conventional tubes are now available which will operate with fair efficiency up to above 500 Mc., and the disk-seal or "lighthouse" variety will function up to about 3000 Mc.

Above about 2000 Mc. the most useful types of tubes are the klystron and magnetron. These are essentially one-band devices, the frequency-determining circuits being an integral part of the tube itself. Tuning over a small frequency range, such as an amateur band, is possible, usually by warping the cavity employed, but the tubes are not independent of frequency in the conventional sense.

A 60-Watt AM–FM 50-Mc. Transmitter or Exciter

The transmitter shown in Figs. 13-1–13-3, inclusive, has an output of approximately 40 watts in the 50-Mc. band and is so designed that either frequency or amplitude modulation may be used. Aside from power supplies, no auxiliary apparatus is needed for FM transmission, since the primary frequency control is a variable-frequency oscillator and a reactance modulator is included in the unit. For amplitude modulation, a modulator having an audio power output of about 30 watts is required.

As an alternative to electron-coupled VFO control, provision also is made for crystal control, using a Tri-tet oscillator. As shown in the circuit diagram, Fig. 13-2, the crystal oscillator and e.c. oscillator have a common plate circuit, the frequency being doubled in this circuit in both cases. The oscillators are followed by a 6V6 doubler, and this in turn drives the final amplifier, an 815.

The tuned circuits are designed to cover a little more than the range required for the 50-Mc. band so that the transmitter as shown can be used to drive a power frequency multiplier tripling into the 144-Mc. band. The VFO grid circuit tunes from 12 to 13.5 Mc., the range from 12.5 to 13.5 Mc. being used for the 50-Mc. band, and the range from 12 to 12.35 Mc. being available for the 144-Mc. band. When crystal control is to be used, frequencies within the appropriate ranges should be selected, since the oscillator portion of the Tri-tet circuit works over the same frequency range as the grid circuit of the VFO. Appropriate crystals in the 6-, 8-, or 12-Mc. ranges may be used, as the 6AG7 crystal oscillator will operate effectively as a quadrupler or tripler, as well as a doubler.

The common oscillator plate circuit tunes from 24 to 27 Mc., with the 6V6 doubling to 48 to 54 Mc. Either oscillator may be selected by means of a switch, $S_{1A\text{-}B\text{-}C}$, which closes the cathode circuit of the desired oscillator. To prevent any possibility of accidental frequency modulation when amplitude modulation is being used, a three-position switch is employed, giving a front-panel selection of crystal or VFO control (for AM or c.w.) and VFO control with FM.

Stability under changes in supply voltage is attained by supplying the VFO screen from a VR-150. This holds the screen voltage at 150 when the plate potential is varied from 150 to 600 volts. The cathode current to the oscillator, measured in J_2, remains practically constant when the plate voltage is varied over this wide range, and the total frequency shift is only a few hundred cycles. With variations in plate voltage which would result from even the most severe line-voltage fluctuations, the frequency shift in the oscillator is only a few cycles.

The transmitter is built on a $10 \times 17 \times 3$-inch chassis, with all components except tubes, crystal and the final-stage output circuit mounted below the deck. Viewing the unit from the top front, the microphone transformer and 6SA7 reactance modulator are at the right front, with the VR-150 at the rear, adjacent to the antenna-coupling assembly. The crystal, crystal oscillator and VFO are grouped near the middle of the chassis, with the doubler and final tubes at the left.

The front panel is a standard $8\frac{3}{4} \times 19$-inch crackle-finished Masonite unit. The VFO tuning dial is centrally placed, with the oscillator and doubler tuning condensers at the left and the AM/FM switch and deviation control at the right. The final plate tuning knob is above the VFO dial, at the left, and the swinging-link adjustment is at the right. Jacks, from left to right, are J_4, J_3, J_2 and J_1.

The two wires protruding through the chassis close to the 815 are neutralizing "condensers," labeled C_{N1} and C_{N2} on the schematic diagram. They consist of two pieces of No. 14 enameled wire soldered to the grid prongs of the 815 socket, crossed under the chassis, and brought through the chassis and held in position by two small Isolantite feed-through bushings (Millen 32150).

Adjustment Procedure

Adjustment is simple and straightforward. The tuning range of the VFO should be checked first. This may be done with only the two oscillator tubes in place and the AM/FM switch in the VFO position. The oscillator plate condenser should be tuned for maximum r.f. indication in a neon bulb adjacent to L_2, and the frequency checked in a receiver having a fairly-accurate calibration for the region around 12, 24, or 48 Mc.

The size of the VFO grid coil, L_1, is extremely critical, and if some pruning of this coil is to be avoided it would be advisable to make the 50-$\mu\mu$fd. section of C_{10} an adjustable padder condenser, such as a Hammarlund APC-50, which can then be adjusted until 12 Mc. appears at about 90 on the VFO vernier dial. The high-frequency limit, 13.5 Mc., should then come at approximately 10, giving a spread of about 18 divisions for the 144-Mc. band and 54 divisions for the 50-Mc. band. Without such a variable condenser, the number of turns on L_1 must be adjusted by cut-and-try until the proper tuning range is secured. In either case, the final adjustment of band coverage should be made with the 6SA7 reactance modulator in its socket so that its plate-to-ground capacitance will be across the tuned circuit.

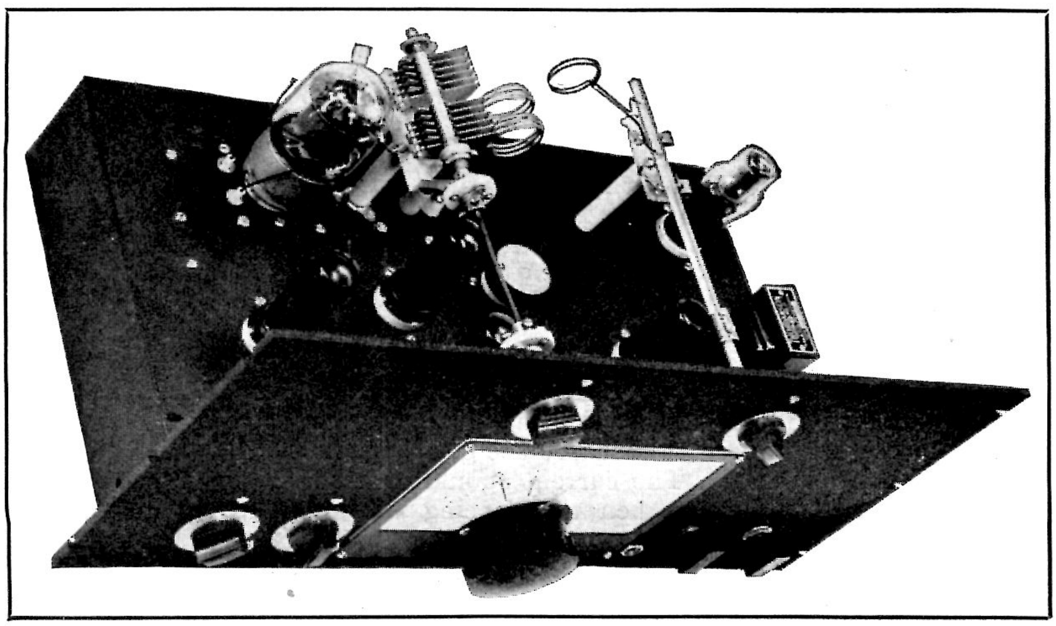


Fig. 13-1 — Front view of the 50-Mc. AM/FM transmitter. The r.f. section of the unit occupies the left-hand portion of the chassis. The VR-150, 6SA7 reactance modulator, and microphone transformer are at the right. Note the neutralizing-capacitance wires at the left of the 815.

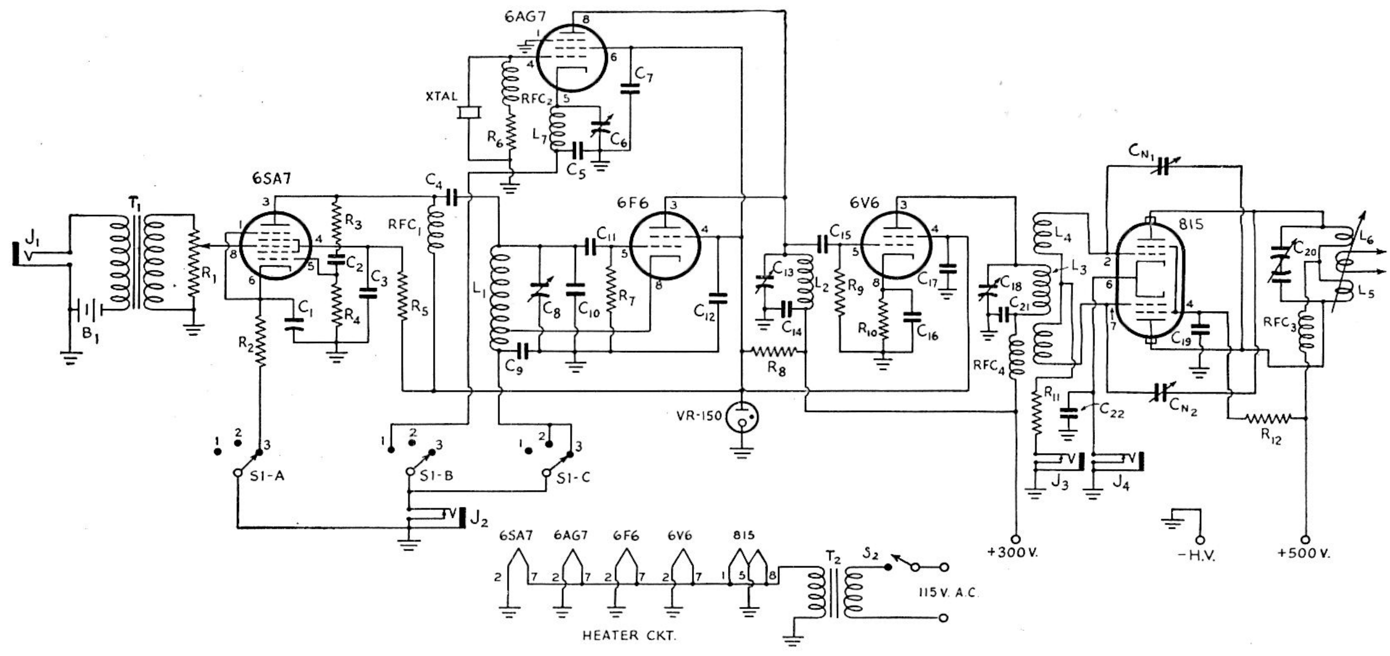


Fig. 13-2 — Wiring diagram of a 50-Mc. AM/FM transmitter.

C_1 — 0.01-μfd. 400-volt paper tubular.
C_2 — 0.001-μfd. mica.
C_3 — 8-μfd. 450-volt electrolytic and 0.0047-μfd. mica in parallel.
C_4, C_{19} — 470-$\mu\mu$fd. mica.
C_5, C_7, C_9, C_{12}, C_{14}, C_{16}, C_{17}, C_{21}, C_{22} — 0.0022-μfd. mica.
C_6 — 100-$\mu\mu$fd. midget variable, screwdriver adjustment (Hammarlund APC-100).
C_8 — 50-$\mu\mu$fd. variable, "straight-line-frequency" type (Hammarlund MC-50-M).
C_{10} — 100-$\mu\mu$fd. and 50-$\mu\mu$fd. in parallel (Sickles Silver-cap). See text.
C_{11} — 100-$\mu\mu$fd. mica.
C_{13}, C_{18} — 50-$\mu\mu$fd. variable (Hammarlund MC-50-S).
C_{15} — 47-$\mu\mu$fd. mica.
C_{20} — 35 $\mu\mu$fd. per section, split stator (Hammarlund MCD-35-MX).
C_{N1}, C_{N2} — Neutralizing capacitors. See text.
R_1 — 0.5-megohm volume control, switch type.
R_2 — 680 ohms, ½ watt.
R_3 — 47,000 ohms, ½ watt.
R_4, R_6 — 0.22 megohm, ½ watt.
R_5 — 4700 ohms, ½ watt.
R_7, R_9 — 0.1 megohm, ½ watt.
R_8 — 5000 ohms, 5 watts.
R_{10} — 220 ohms, 1 watt.
R_{11} — 15,000 ohms, 1 watt.
R_{12} — 15,000 ohms, 5 watts.
L_1 — 8 turns No. 18 tinned, ¾-inch diameter, 1-inch length on National PRF-2 form. Tapped 2 t. from ground end.
L_2 — 10 turns No. 14 e., ½-inch diameter, spaced one diameter, air-wound.
L_3 — 4 turns No. 14 e., ½-inch diameter, spaced one diameter, air-wound.
L_4 — 5 turns each section, No. 14 e., ½-inch diameter. Adjust spacing for best coupling. See text.
L_5 — 3 turns each section, No. 12, tinned, 1⅛-inch diameter, spaced one diameter.
L_6 — 2 turns No. 14 e., 1-inch diameter, swinging link. See photos and text.
L_7 — 35 turns No. 24 d.c.c., close-wound on 9/16-inch diameter form (National PRE-3).
B_1 — Microphone battery (Burgess).
J_1 — Open-circuit jack.
J_2, J_3, J_4 — Closed-circuit jack.
RFC_1, RFC_2, RFC_4 — 2.5-mh. r.f. choke (National R-100).
RFC_3 — 2.5-mh. r.f. choke, end-mounting (National R-100-U).
$S_{1A\text{-}B\text{-}C}$ — 3-position 3-contact rotary switch (Mallory).
S_2 — Switch on deviation control, R_1.
T_1 — Single-button microphone transformer (Thordarson T-83A78).
T_2 — 6.3-volt 4-amp. filament transformer.

Operation of the crystal oscillator may next be checked. With a 100-ma. meter connected through J_2, and the AM/FM switch in the "crystal" position, adjust the crystal-oscillator cathode tuning, C_6, until the current dips sharply, indicating oscillation. This control should be set at the point that gives the lowest cathode current consistent with easy crystal starting. Cathode current should be similar for both oscillators — about 20 ma.

The doubler stage may next be tested by installing the 6V6 and 815 tubes, leaving the plate power off the 815. A meter having a 10-ma. range should be used to measure the grid current in the 815, at J_3. The current should come up to about 6 ma. when the spacing between L_3 and L_4 is optimum, though this is more than is actually needed for satisfactory operation of the 815.

Next the position of the neutralizing wires can be adjusted. The 815 plate tuning condenser, C_{20}, should be rotated slowly, meanwhile watching the grid current for any variation. The position of the neutralizing wires should be adjusted until there is no sign of fluctuation in grid current as the tuning condenser is rotated. A length of wire extending about one inch above the metal ring on the 815, at a position about ⅛ inch from the glass envelope, should be sufficient. If this should be inadequate, small tabs of copper or brass can be soldered to the ends of the wires to afford additional capacitance to the tube plates. The neutralizing capacitance is necessary in order to ensure completely-stable operation.

After neutralization, power may be applied to the 815 plates, while noting the cathode current as indicated on a 200-ma. meter plugged

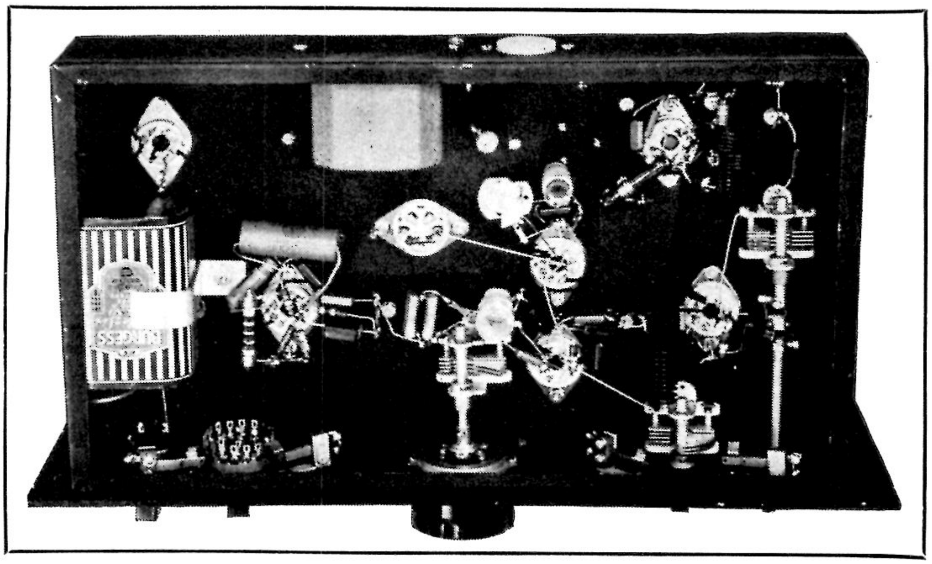


Fig. 13-3 — Under-chassis view of the 50-Mc. AM/FM transmitter. At the lower center are the VFO grid coil and associated components. Over these are the crystal and cathode circuit for the 6AG7 crystal oscillator. At the upper right are the inductively-coupled doubler plate coil and final grid coil. The coil and condenser at the lower right comprise the plate circuit which is common to both oscillators. The doubler plate tuning condenser is at the extreme right.

into J_4. The dip at resonance should bring the current to about 50 ma. with no load. A 40-watt lamp connected across the swinging-link terminals should then give a full-brilliancy indication when the link is adjusted for maximum coupling. This is with 500 volts applied, which should be used only after it has been determined that everything is functioning properly. If trouble is encountered, further tests should be made with reduced voltage to avoid damaging the tube.

When the transmitter is put on the air, the full 500 volts at 150 ma. may be used for FM or c.w. operation. For plate modulation, the voltage should be reduced to about 400 for maximum tube life, even though the tube plates may show no color at the higher plate voltage.

For frequency modulation, the 6SA7 reactance modulator provides the simplest possible means of obtaining the desired swing in frequency. It may be operated with a single-button microphone plugged into J_1, or the modulator may be driven from a speech amplifier and crystal or dynamic microphone. The output of the speech amplifier should then be connected across potentiometer R_1, and T_1 may be omitted. In either case, potentiometer R_1 serves as a deviation control, the frequency swing being adjusted to suit the receiver at the station being worked.

For 144-Mc. work, or for operation above 52.5 Mc. in the 6-meter band, wide-band FM may be used if desired, in which case the setting of R_1 may be anything up to full-on. The transmitter may also be used for narrow-band FM on any frequency above 51 Mc. Considerable care should then be used in adjusting the deviation, to be certain that the swing is within the prescribed limits. The procedure outlined in Chapter Nine may be followed for checking the deviation in NFM operation.

In addition to the filament transformer, T_2, indicated in the circuit diagram, the transmitter requires two plate power supplies. One, for the 815, should have an output of 400 to 500 volts at 175 ma.; the other, for the remaining tubes, should deliver 300 volts at approximately 100 milliamperes.

If more power is desired, an 829 may be substituted for the 815. In this case an input of 100 watts or more may be run, the output being as high as 85 watts at maximum operating conditions.

200-Watt Driver-Amplifier for 50 and 144 Mc.

A companion medium-power driver-amplifier for the 815 rig just described is shown in Figs. 13-4 to 13-7. The amplifier uses a pair of 24G triodes in push-pull, while the driver, a frequency tripler used for 144-Mc. operation only, is a single 829-B. If operation on 144 Mc. is not contemplated, all to the left of the final grid coil, L_3 in Fig. 13-6, may be omitted.

Looking at the front-panel view, Fig. 13-4, the two large dials are the plate tuning controls for both stages. The small dial at the left controls the swinging link, the center one is the grid tuning control for the final, and the one at the far right is the tripler grid tuning.

The rear view shows the general placement of parts. At the left rear is a jack-bar, containing terminals *AA* and *BB*, into which the link from the exciter is plugged to furnish excitation for the tripler (terminals *AA*) or final stage (terminals *BB*). The tripler grid-circuit components, L_1C_1, are adjacent to the jack-bar. C_1 is mounted with its shaft parallel to the panel, in order to permit short leads, and is tuned by means of a flexible shaft. The 829-B and its plate-circuit components are mounted in such position as to permit inductive coupling between the plate coil of the tripler and the grid coil of the final, when the transmitter is used on 144 Mc. For 50-Mc. work the 829-B is inoperative, and the 50-Mc. grid coil (the same

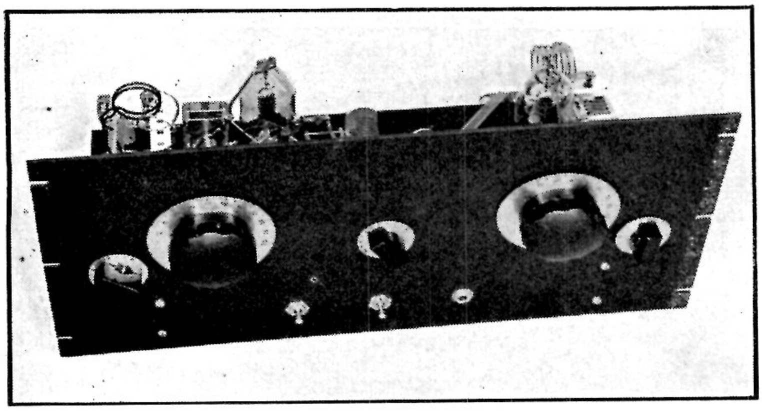


Fig. 13-4 — Front view of the 200-watt driver-amplifier for 50 and 144 Mc. The two large dials are the plate tuning controls. The small dial at the left adjusts the position of the output coupling link, the center dial is the grid tuning control for the final, and the third small dial is the tripler grid-tuning control. Across the lower center are the filament switches and grid-current meter jack.

coil is used in both tripler and final) just clears the plate coil of the 829-B. Care must be used in parts placement to work this out correctly.

Between the grid tuning condenser, C_3, and the 24G tubes are the two neutralizing condensers. These are triple-spaced midgets, mounted back-to-back, with coupled shafts. The stator plates had to be filed out to reduce the minimum capacitance to the small value needed to neutralize the 24Gs. The final tank condenser, C_5, and the jack-bar for the final plate coil are positioned for the short leads that are required if satisfactory performance on 144 Mc. is to be attained. R.f. leads in the final stage are made with ¼-inch silver ribbon, which is appreciably better than braid at these frequencies. Thin copper strip may also be used. All connections in the plate circuit of the 24Gs should be made with bolts and nuts, as the tank circuit will heat sufficiently during 144-Mc. operation to melt soldered connections and increase losses. Plate connections to the 829-B are made by means of small Fahnestock clips. The tripler plate coil is supplied with two of these clips (Millen 36021), so that it may be removed separately from the condenser leads, in replacing the tube.

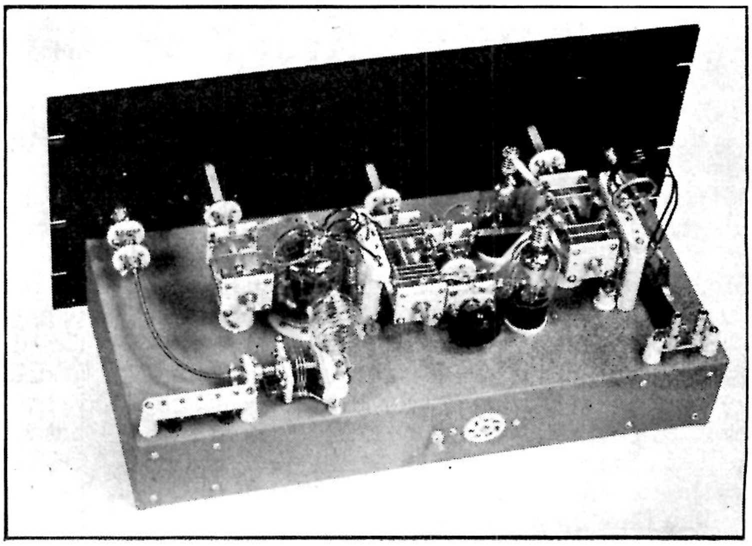
Operation on 50 Mc.

When the amplifier is to be used on 50 Mc. the switch, S_1, is left open, so that the heater of the tripler will not be energized when S_2 is closed. The link from the exciter is plugged into terminals *BB* in the jack-bar, which is a Millen 41205 coil socket. The output of the exciter is thus connected to the link terminals of the final grid-coil socket, a National XB-16.

The final stage should be tested on 50 Mc. before attempting 144-Mc. operation. With the proper coils inserted at L_3 and L_4, and with power on the exciter but no plate voltage on the final, rotate C_3 for maximum grid current. Set the neutralizing condensers at maximum capacitance and rotate C_5. If the plate circuit is capable of being resonated there will be a kick in the grid current as the circuit passes through resonance. The neutralizing condensers should be rotated, a small amount at a time, until the kick in the grid current disappears. This will probably occur close to the minimum-capacitance setting of C_{4a} and C_{4b}.

Power may now be applied to the plate circuit. It is advisable to make initial tuning adjustments at low voltage, preferably 750 volts or less. If everything is in order, the plate current will drop to about 20 ma. at resonance, at this voltage. A load of some sort should now be connected across the output terminals and the operation tested at increasingly higher voltages. The final tubes should be capable of an input of 250 watts or more on 50 Mc. without exceeding their normal plate dissipation of 50 watts for the pair, indicated by a bright orange color.

Fig. 13-5 — Rear view of the driver-amplifier for 50 and 144 Mc., with coils in place for 144-Mc. operation. At the lower left are the link socket and the grid circuit of the tripler stage. The plate coil of the tripler is inductively-coupled to the final grid coil, a single turn. All parts are grouped as closely as possible, for efficient performance on 144 Mc. R.f. leads in the final stage are made of ¼-inch silver ribbon.



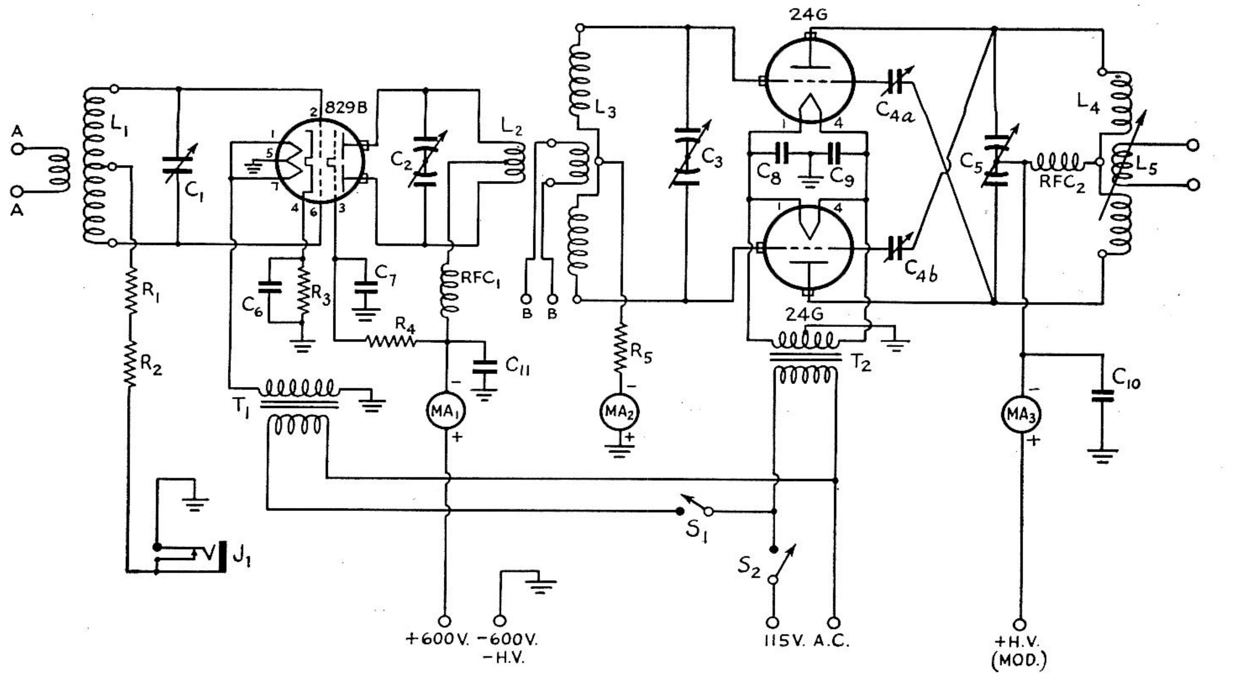


Fig. 13-6 — Schematic diagram of the 200-watt driver-amplifier.

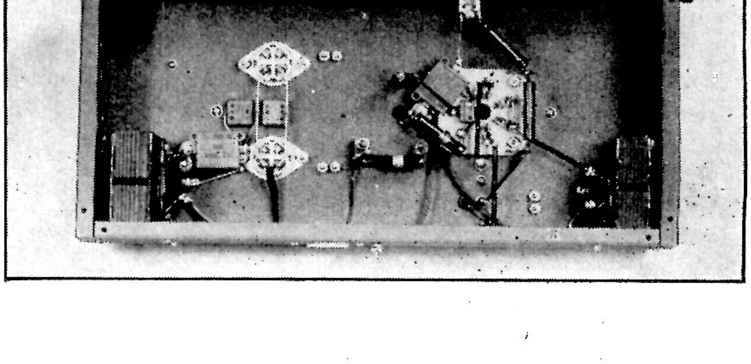
C_1 — 15-$\mu\mu$fd. variable (Cardwell ZR-15-AS).
C_2 — 10-$\mu\mu$fd.-per-section split stator (Cardwell ER-10-AD).
C_3 — 15-$\mu\mu$fd.-per-section split stator (Cardwell ET-15-AD).
C_{4a}, C_{4b} — 2-plate triple-spaced midget variable (Cardwell ZS-4-SS with stator plates filed to reduce minimum capacitance).
C_5 — 4-$\mu\mu$fd.-per-section split stator (Cardwell ES-4-SD).
C_6, C_7 — 470-$\mu\mu$fd. midget mica.
C_8, C_9 — 0.001-μfd. mica.
C_{10} — 500-$\mu\mu$fd. 2500-volt mica.
C_{11} — 500-$\mu\mu$fd. 1000-volt mica.
R_1 — 4700-ohm 2-watt carbon.
R_2 — 50,000 ohms, 10 watts.
R_3 — 250 ohms, 10 watts.
R_4 — 15,000 ohms, 10 watts.
R_5 — 3000 ohms, 10 watts.
L_1 — 4 turns No. 18, 1¼-inch diameter, 1 inch long, 3-turn center link (National AR-16, 10-C, with 2 turns removed from each end).
L_2 — 2 turns No. 12 enameled, ⅞-inch diameter, spaced ¼ inch.
L_3 — 50 Mc.: Use L_1. 144 Mc.: Center-tapped "U" made from No. 12 enameled wire, ⅝ inch high, ¾ inch wide. See rear-view photograph. Sockets for L_1 and L_3 are National XB-16.
L_4 — 50 Mc.: 4 turns each side of center-tap, spaced diameter of wire, on Millen No. 40205 base.
— 144 Mc.: 2 turns ⅛-inch brass rod, spaced to fit into socket terminals, 1¼-inch inside diameter. May be silver-plated for best results.
L_5 — Swinging link: 3 turns No. 12 enameled wire, 1⅜-inch diam. Mount for adjustment on polystyrene rod; see rear-view photograph.
J_1 — Closed-circuit jack.
MA_1 — 0-200 d.c. milliammeter.
MA_2 — 0-100 d.c. milliammeter.
MA_3 — 0-300 d.c. milliammeter.
RFC_1, RFC_2 — Ohmite Z-0.
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — 6.3 volts, 2.5 amp.
T_2 — 6.3 volts, 6 amp.

Checking on 144 Mc.

For operation on 144 Mc., the switch, S_1, should be closed, energizing the heater of the 829-B tripler. The link from the exciter should be plugged into terminals *AA* on the jack-bar, so that drive is applied to the tripler grid circuit. The 144-Mc. coils should be inserted at L_3 and L_4. It will be noted that these coils have no bases. The grid coil (merely a center-tapped "U") is made of No. 12 wire, which fits snugly in the coil-socket terminals. It should be bent slightly toward the 829-B plate coil, L_2, so that it fits between the turns at a position that provides optimum inductive coupling. The final plate coil is made of ⅛-inch brass rod, which provides a tight fit in the jack-bar. The coil may be silver-plated, if desired.

The grid coil for the tripler (L_1) may be the same coil as is used for 50-Mc. operation at L_3. With this coil in place and the output of the exciter on 48 Mc., the operation of the tripler may be checked, using a voltage of around 400

Fig. 13-7 — Bottom view of the 50-144-Mc. driver-amplifier.



on the plates of the 829-B initially. The grid circuit of the final should then be tuned to resonance as indicated by maximum grid current. The plate voltage on the tripler may be raised to 600 volts, if necessary, to assure adequate grid drive for the final stage. Typical operating conditions are as follows: tripler plate voltage — 600; plate current — 125 ma.; tripler grid current (read in J_1) — 10 ma. or more; final grid current — 35 to 50 ma.

Neutralization of the final should be rechecked, as the position of C_{4a} and C_{4b} may be slightly different from the 50-Mc. setting. Power may then be applied to the final stage, using low voltages at first. The final stage should not be operated without load, except with low plate voltages, as tank-circuit losses will cause excessive heating otherwise. The dip in plate current at resonance will be less than at 50 Mc., and minimum current at 1000 volts will probably not drop below about 65 ma. The maximum recommended plate potential for 144-Mc. operation is about 1250 volts, though tests have been made on this amplifier at voltages as high as 1700.

A safe check on the operation of the tubes is to adjust the plate voltage (with no excitation applied) until an input of 50 watts is being run to the pair. Note the color of the plates at this input — a bright orange. If this color is not exceeded in normal operation, one may be sure that the tubes are being operated within safe limits. An input of 200 watts can be handled safely on 144 Mc. if the stage is running properly.

The transmitter may be used for c.w. work on either band by keying the cathode of the driving stage. If the 829-B tripler cathode is to be keyed a jack will have to be added. Provision should be made to add fixed bias in series between the final grid meter, MA_2, and ground. A 45-volt "B" battery will suffice to hold the 24G plate current to a safe value when the excitation is removed.

The three meters shown in the schematic diagram, but not in the photographs, are mounted on a separate meter panel. Another useful addition, not shown, is a 3500-ohm 10-watt potentiometer, connected in series with R_5, so that the final-stage bias can be varied to suit different operating conditions. This is particularly useful if the transmitter is to be used on c.w. and FM, as well as AM voice operation.

A 60-Watt Transmitter for 50, 28 and 14 Mc.

The transmitter-exciter shown in Figs. 13-8, 13-9 and 13-10 is a three-stage unit designed for use in the 50-, 28- and 14-Mc. bands. It employs a 6V6GT Tri-tet oscillator, a 6V6GT frequency multiplier, and an 815 operating as a straight amplifier without neutralization. It is capable of an input of 75 watts when being operated as an exciter or c.w. transmitter, but the power should be reduced to 60 watts input if the amplifier is modulated. Plug-in coils are employed for simplicity and flexibility.

Circuit Features

The Tri-tet oscillator has a fixed-frequency cathode circuit which resonates at approximately 21.5 Mc. With the cathode circuit so tuned, it is possible to employ a wide variety of crystal frequencies. For operation on 14 Mc., 3.5-Mc. crystals may be used, with the oscillator plate circuit tuned to 7 Mc., doubling in the second 6V6GT to 14 Mc. Most 3.5-Mc. crystals will deliver sufficient output from the oscillator on 7 Mc. to permit operating the second stage as a quadrupler to 28 Mc. also. Crystals between 7000 and 7200 kc. can be used with the oscillator working straight-through, provided that the oscillator plate circuit is not resonated at exactly the crystal frequency. Crystals from 7000 to 7425 kc. are used for operation of the amplifier in the 28-Mc. band, the oscillator doubling in this case.

For 50-Mc. operation crystals between 6250 and 6750 kc. are recommended. The oscillator

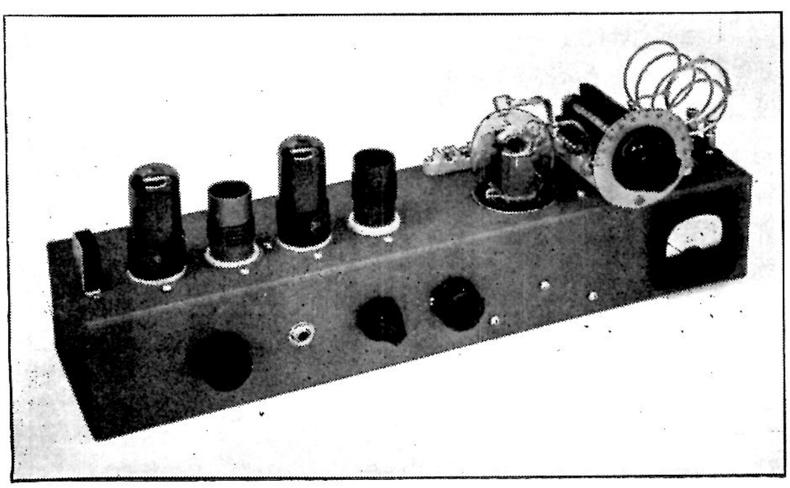


Fig. 13-8 — A front view of the 815 transmitter-exciter for 50, 28 and 14 Mc. Coils for 50-Mc. operation are in place.

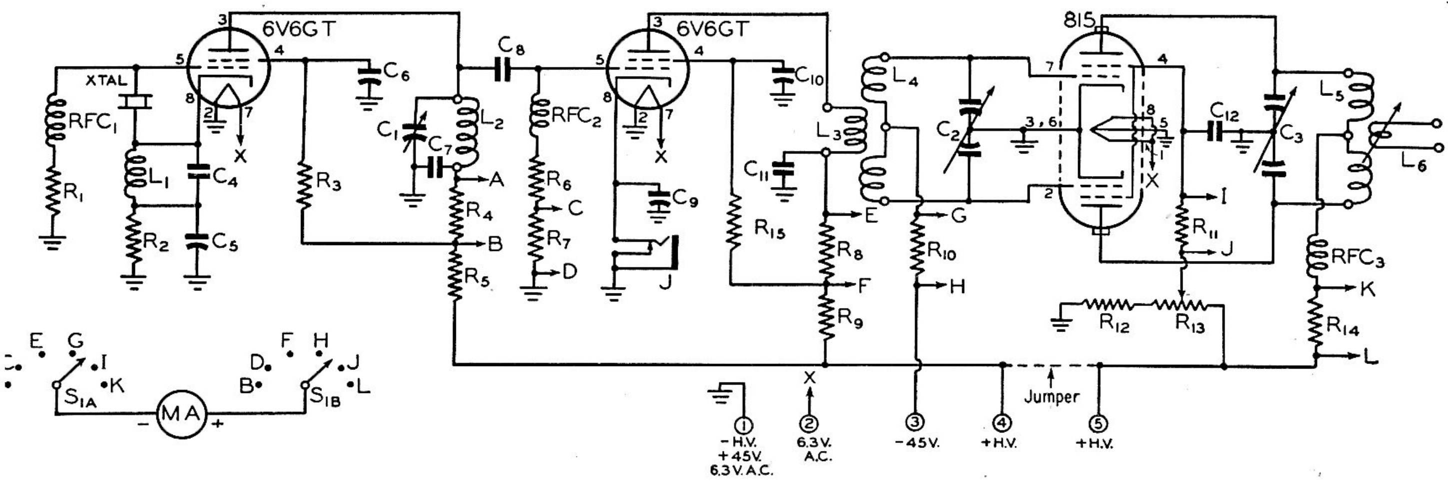


Fig. 13-9 — Circuit diagram of the 815 transmitter-exciter.

C_1 — 25-$\mu\mu$fd. variable (Cardwell ZR-25-AS).
C_2 — 50-$\mu\mu$fd.-per-section variable (Bud LC-1662).
C_3 — 35-$\mu\mu$fd.-per-section variable (Hammarlund MCD-35-SX).
C_4 — 68-$\mu\mu$fd. mica.
C_5, C_6, C_9, C_{10} — 0.01-μfd. paper.
C_7 — 0.0022-μfd. mica.
C_8 — 100-$\mu\mu$fd. mica.
C_{11}, C_{12} — 470-$\mu\mu$fd. mica.
R_1 — 0.12 meghom, ½ watt.
R_2 — 220 ohms, ½ watt.
R_3 — 15,000 ohms, 1 watt.
R_4, R_7, R_8, R_{10}, R_{11}, R_{14} — 100 ohms, ½ watt.
R_5, R_9 — 5,000 ohms, 10 watts, adjustable.
R_6 — 47,000 ohms, ½ watt.
R_{12}, R_{13} — 5,000 ohms, 10 watts, adjustable.
R_{15} — 40,000 ohms, 10 watts.
L_1 — 8 turns No. 18 enameled, close-wound, ½-inch dia.
L_2 — 7 Mc. — 25 turns No. 20 d.c.c., close-wound, 1 inch long.
14 Mc. — 12 turns No. 20 d.c.c., space-wound, 1 inch long.
28 Mc. — 5 turns No. 18 enameled, space-wound, ½ inch long.
L_3 — 14 Mc. — 9 turns No. 20 d.c.c., close-wound, ⅜ inch long.
28 Mc. — 6 turns No. 20 d.c.c., space-wound, ⅜ inch long.
50 Mc. — 3 turns No. 18 enameled, space-wound, ¼ inch long.
L_4 — 14 Mc. — 14 turns No. 20 d.c.c., close-wound, 7 turns each side of primary.
28 Mc. — 6 turns No. 20 d.c.c., spaced diam. wire, 3 turns each side of primary.
50 Mc. — 4 turns No. 18 enameled, spaced diam. wire, 2 turns each side of primary.
Above coils are wound on 1-inch diameter forms (Millen 45004 for L_2; Millen 45005 for L_3-L_4). Approximately ⅛ inch between L_3 and L_4.
L_5 — 14 Mc. — 14 turns No. 16, 1⅞-inch diameter, 2 inches long. (B & W 20-JVL.)
28 Mc. — 8 turns No. 12, 1⅞-inch diameter, 2 inches long. (B & W 10-JVL.)
50 Mc. — 4 turns No. 12, 1⅞-inch diameter, 2 inches long. (B & W 10-JVL with 2 turns removed from each end.)
Above coils are wound in two sections with half the total number of turns each side of center. A ½-inch space is left at the center to permit the use of a swinging link (L_6). The Barker & Williamson coils are mounted on five-prong bases of the type which plug into tube sockets.
J — Closed-circuit jack.
MA — 0-50 milliammeter.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke.
S_{1A}, S_{1B} — 2-circuit 6-position selector switch (Mallory 3226J).

is then operated as a quadrupler and the second stage as a doubler. The oscillator may also be operated as a tripler, using crystals between 8334 and 9000 kc., or as a doubler with 12.5–13.5-Mc. crystals. The complete unit may also be used as a driver for a 144-Mc. tripler stage by using the above crystal types, except that the ranges would then be 6000–6166, 8000–8222, or 12,000–12,333 kc.

Cathode bias is used on the oscillator stage, maintaining the plate current at a reasonable figure. The common-power-supply voltage is lowered to 300 volts by the dropping resistor, R_5, and the screen voltage of 250 is obtained from resistor R_3. The by-pass condenser, C_7, is inserted between the cold end of the plate coil, L_2, and ground, in order to permit grounding of the tuning-condenser rotor. Capacitive coupling, through C_8, is used between the oscillator and the multiplier stage.

The multiplier stage has a self-resonant plate circuit which is inductively-coupled to the grid circuit of the final amplifier. This arrangement was found to give better balance of excitation to the final-stage grids than did the use of a self-resonant grid circuit in the final stage and a tuned plate circuit for the 6V6GT. Resistors R_9 and R_{15} drop the screen and plate voltages to the proper values. Grid bias is developed across R_6. The cathode by-pass condenser, C_9, is required because of the lengthening of the cathode lead by the insertion of the keying jack, *J*. No neutralization is required in this stage, since it is operated as a frequency multiplier at all times.

Bias for the final amplifier is obtained from a 45-volt "B" battery, permitting the preceding stage to be keyed for c.w. operation. The voltage-divider network, R_{12} and R_{13}, can be adjusted to provide the proper screen voltage for either type of operation. The plate tuning-condenser rotor is connected directly to ground. Power output is taken from the plate circuit by means of the adjustable link, L_6, which is part of the plug-in plate-coil assembly.

A switching system is provided for measuring all the necessary currents with one 50-ma. meter, connecting it across shunt resistors R_4, R_7, R_8, R_{10}, R_{11} and R_{14}. The range of the meter is extended to 300 ma. by means of R_{14},

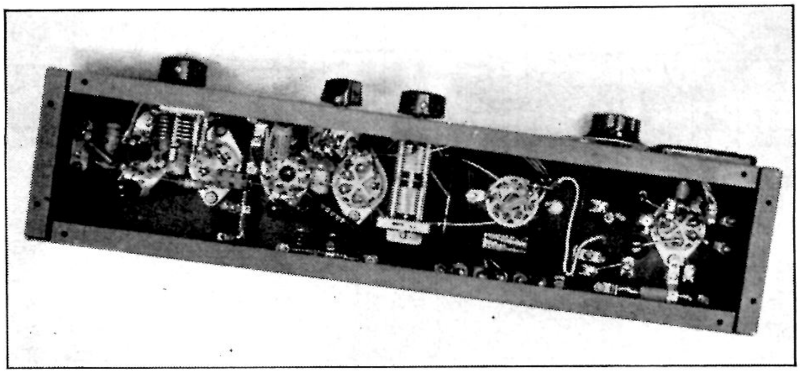


Fig. 13-10 — A bottom-inside view of the 815 transmitter.

which consists of about 31 inches of No. 30 insulated wire scramble-wound on a 100-ohm resistor. The length of wire required for this shunt may vary with different types of meters.

Mechanical Details

The front-view photograph, Fig. 13-8, shows the arrangement of the principal components. The chassis measures 3 × 4 × 17 inches. It is suggested that this layout be followed closely, particularly as regards the arrangement of parts in the 815 plate and grid circuits, since this layout permits operation of the 815 without neutralization. Components, along the top of the chassis, from left to right, are the crystal, oscillator tube, oscillator plate coil, multiplier tube, multiplier plate and final grid coil, final-amplifier tube, and amplifier tank circuit. A 5-connector terminal strip may be seen just to the rear of the 815. Across the front wall are the oscillator tuning knob, the keying jack, meter switch, multiplier tuning knob, and meter.

The 815 socket is mounted on a subchassis which can be seen in the bottom view, Fig. 13-10. This unit, Millen No. 80009, also includes the tube shield, which was cut off in this case so that, with the shield just flush with the chassis, the tube socket is 1½ inches below. Arrangement of other components is apparent from the bottom view. Resistors R_{12} and R_{13} are at the right end near the coil socket. The amplifier plate choke, RFC_3, is on a stand-off adjacent to the coil socket, and the meter shunt, R_{14}, is connected between the choke and one end of R_{13}. The dropping resistor, R_9, is parallel to the rear wall of the chassis, and R_5 is at right angles to it, both below, and to the left of, the sockets for the 6V6GT and its associated plug-in coil. RFC_1 and RFC_2 are mounted on stand-off insulators at the rear of the tube sockets. The cathode coil, L_1, is self-supporting, and is mounted directly on the tube-socket terminals. All other small parts are grouped closely adjacent to their respective tube sockets.

Coil information is given in full detail under Fig. 13-9. The number of turns specified should be satisfactory for the frequencies of operation referred to, though some adjustment of the coupling between windings of L_3 and L_4 may be required. Since variation in spacing of these windings affects the tuning range of the L/C combination to a noticeable degree, the setting of C_2 should be checked as the coupling adjustment is being carried on.

Testing Procedure

The power supply for the transmitter should be capable of delivering 275 to 300 ma. at 400 or 500 volts. The filament transformer should supply 6.3 volts at 2.5 amperes or more. It is advisable to test the oscillator and multiplier stages before applying plate and screen voltage to the 815. The jumper between Terminals 4 and 5 on the terminal strip should be left off during this operation. With plate voltage applied to the oscillator and doubler, the oscillator plate circuit should be tuned to resonance as indicated by maximum grid current in the next stage. The amplifier grid circuit should then be tuned for maximum grid current. With the coil specifications given it is unlikely that the circuits will tune to an incorrect harmonic; nevertheless, it is wise to check with a calibrated absorption-type wavemeter to be sure that such is not the case. Dropping resistors R_5 and R_9 should be set at their full value of 5000 ohms during this operation, final setting of the adjustments being made after the power supply is loaded by the entire transmitter. Grid current to the final stage should be about 15 ma. for all bands at this point. This may be adjusted by changing the turns on L_3, or by detuning C_2, if the grid current is excessive. Detuning of C_2 is recommended as the more satisfactory of the two methods.

The final amplifier may now be put into operation. The screen voltage should be tapped in between the screen divider, R_{12} and R_{13}, and the jumper connected between Terminals 4 and 5. With plate voltage and grid excitation applied, the off-resonance plate current should be about 250 ma., dropping to about 25 ma. when the plate circuit is tuned to resonance. A dummy load such as a lamp should be connected across the output terminals, and the coupling adjusted to bring the plate current up to 150 ma. at resonance.

The oscillator and multiplier plate voltages and the amplifier screen voltage should now be adjusted to 300 and 200 volts, respectively, by adjusting the taps on R_5, R_9 and R_{13}. It is probable that the amplifier plate current will change appreciably at this point, so it will be desirable to readjust the coupling so that the current is again 150 ma., and readjust the resistor taps as required to secure the correct voltages on the various tube elements. Adjustment for 'phone operation is similar, except that the amplifier screen and plate voltages should be 175 and 400 volts, respectively.

With all voltages set for the proper values, the currents will run about as follows: oscillator and doubler plates — 35 ma.; doubler grid — 1 to 3 ma.; 815 grid — 5 to 6 ma.; 815 screen — 17 ma.; 815 plate — 150 ma. The 815 screen current should drop to 15 ma. for 'phone operation. A grid current of 4 to 6 ma. in the 815 is adequate. Though more grid current is usually available, an increase beyond 6 ma. does not improve the output. In c.w. operation, it will be noticed that the 815 plate current does not drop completely off when the excitation is removed. This is no cause for concern, so long as the plate and screen dissipation are held to recommended levels.

The amplifier plate coils are provided with links designed for working into a low-impedance line. The amplifier may thus be used for direct feed in the case of antenna systems having matched lines, or it may work into a line feeding an antenna-coupling unit, in case tuned-feeder antenna systems are employed.

A 3-Stage Stabilized Transmitter

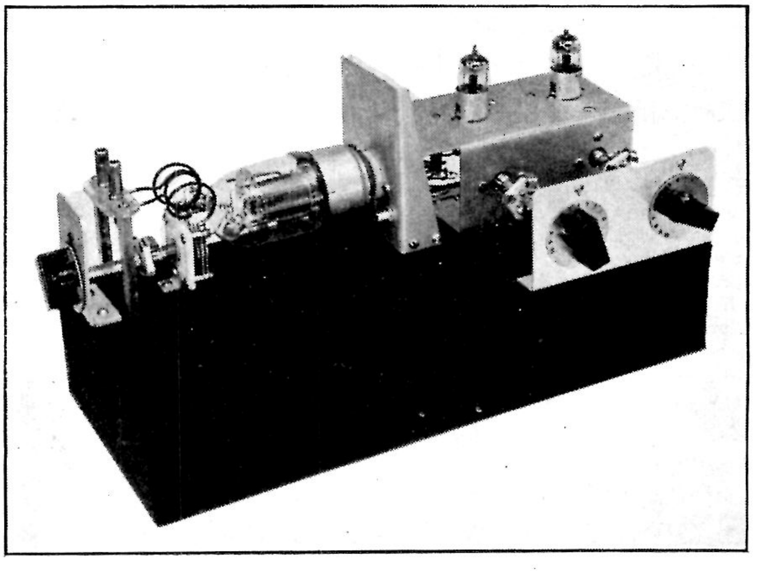
A three-stage transmitter in which frequency-modulation effects are quite small is shown in Figs. 13-11–13-14, inclusive. It includes a 6C4 oscillator, 6C4 neutralized buffer amplifier, and 815 final amplifier, as shown in the circuit diagram, Fig. 13-12. The oscillator and buffer are built as a unit on a "U"-shaped piece of aluminum 6½ inches long on top, 2⅜ inches high, and 2⅞ inches deep on the top. The 815 is mounted on a vertical aluminum piece measuring 4¼ inches high and 3 inches wide, reinforced by bending side lips as shown in the photographs. The two sections are assembled on a 6 × 14 × 3-inch chassis.

The oscillator circuit is "high-C," using a butterfly-type condenser with three circular rotor plates, two butterfly rotors, and four split-stator plates. This condenser has a high minimum capacitance and a small tuning range.

The construction of the buffer amplifier is quite similar to that of the oscillator. The buffer tuning condenser consists of a rotor having three butterfly plates and two stators each having two 90-degree plates. The grid circuit of the buffer is self-resonant, the tuning being adjusted by squeezing the turns of the grid coil L_2 together, or prying them apart. The buffer neutralizing condenser, C_7, mounted directly between the grid of the 6C4 and the lower set of stator plates of C_8, is a 3-30-$\mu\mu$fd. trimmer with the movable plate removed and a washer soldered under the head of the adjusting screw. The washer, by replacing the movable plate, reduces the capacitance of the condenser to a value suitable for neutralizing the 6C4. This capacitor may be conveniently adjusted through the open end of the chassis. Its location is clearly shown in Fig. 13-13.

The grid coil of the final amplifier also is resonant with the input capacitance of the 815, just as the buffer grid circuit is self-resonant. For best operation, the 815 requires neutralization at this frequency. The neutralizing "condensers," C_9 and C_{10} in the circuit diagram, are simply pieces of No. 14 wire extending from the grid of one section of the 815 to the vicinity of the plate of the other section. The wires are crossed at the bottom of the tube socket and go through Millen 32150 bushings in the metal partition. The screen and filament by-pass condensers are mounted so that the leads between the socket prongs and the nearest ground point are as short as possible. This wiring should be done before mounting the partition.

Fig. 13-11 — A three-stage transmitter using a 6C4 master oscillator, 6C4 buffer amplifier, and 815 final amplifier for stabilized transmission in the 144-Mc. band. The oscillator and buffer are built as a unit on the folded-aluminum chassis at the right. The transmitter develops a carrier output of about 40 watts.



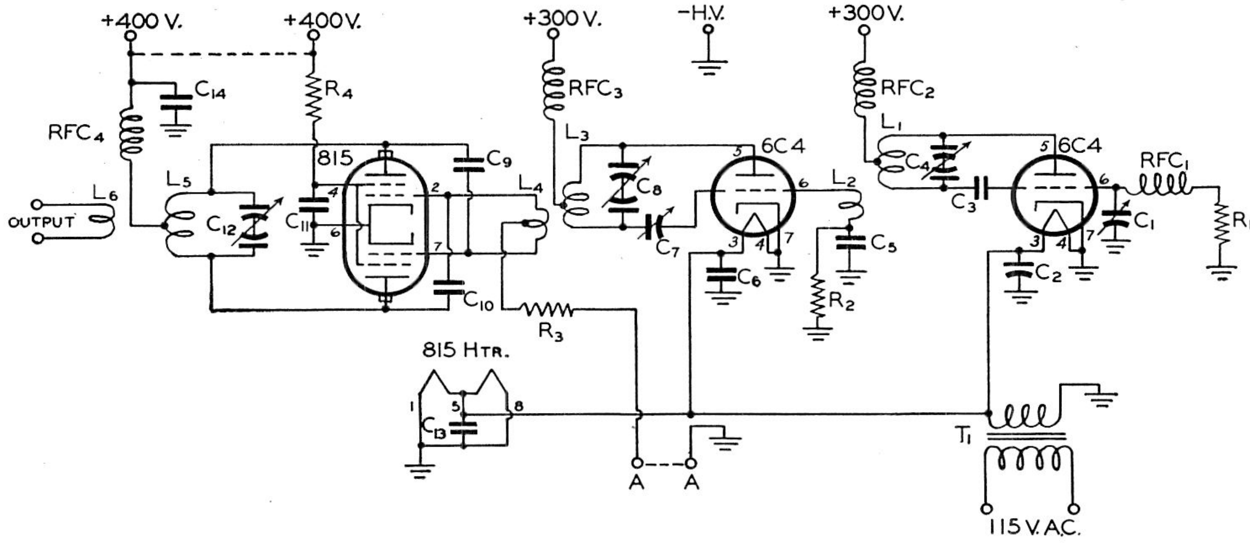


Fig. 13-12 — Circuit diagram of the stabilized 144-Mc. transmitter.

C_1 — 3–30-μμfd. trimmer.
C_2, C_6, C_{11}, C_{13} — 470-μμfd. midget mica.
C_3, C_5 — 47-μμfd. midget mica.
C_4 — Oscillator tuning; Cardwell ER-14-BF/SL.
C_7 — Neutralizing capacitor; see text.
C_8 — Buffer tuning; Cardwell ER-6-BF/S.
C_9, C_{10} — Amplifier neutralizing; see text.
C_{12} — Amplifier tuning; Cardwell ER-6-BF/S.
C_{14} — 100 μμfd., 2500 volts.
R_1, R_2 — 22,000 ohms, ½ watt.
R_3 — 15,000 ohms, 1 watt.
R_4 — 15,000 ohms, 10 watts.
T_1 — 6.3-volt 2-amp. filament transformer.
L_1 — 2 turns No. 12 bare wire; inside diameter 9/16 inch, length 1 inch; plate-supply tap at center.
L_2 — 2 turns No. 14, inside diameter ½ inch; turns spaced wire diameter.
L_3 — 4 turns No. 14, inside diameter ¾ inch, length 1 inch; plate-supply tap at center.
L_4 — 2 turns No. 14, inside diameter ½ inch; turns spaced diameter of wire; tapped at center.
L_5 — 2 turns No. 12, inside diameter 1 inch, length 1 inch; plate-supply tap at center.
L_6 — 2 turns No. 12, inside diameter ¾ inch.
RFC_1, RFC_2, RFC_3, RFC_4 — 1-inch winding No. 24 d.s.c. on ¼-inch diam. polystyrene rod.

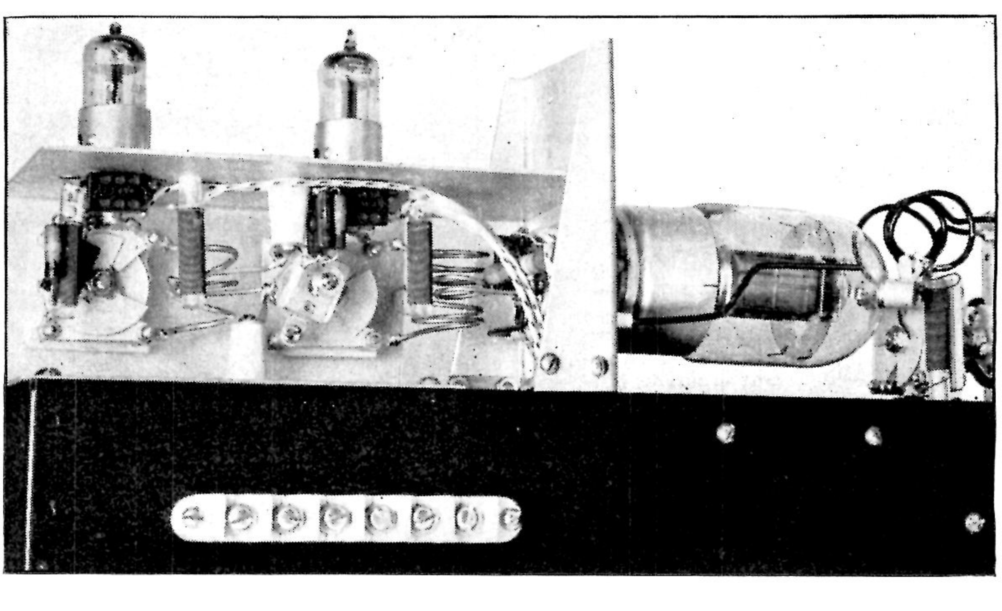
The amplifier plate tank circuit uses a condenser of the same construction as that used in the buffer tank. It is mounted as closely as possible to the plate caps on the 815, and to preserve circuit symmetry the condenser is tuned from the left-hand edge of the chassis. If the transmitter is to be equipped with a regular panel this condenser may be operated by a right-angle drive from the front.

The output terminals are a standard binding-post assembly on polystyrene, mounted on metal posts 2⅜ inches high to bring the coupling coil in proper relation to the amplifier plate tank coil, L_5. Antenna coupling is adjusted by bending L_6 toward or away from L_5.

The plate by-pass condenser and screen-dropping resistor are mounted underneath the chassis, as shown in Fig. 13-14, together with the filament transformer. Separate power-supply terminals are provided for the oscillator plate, buffer plate, amplifier grid (terminals *AA*), amplifier screen, and amplifier plate so that the currents can be measured separately. An external 0–200 d.c. milliammeter will serve in making all adjustments. However, if a meter of lower range is available, it may be used profitably in the low-current circuits.

In putting the transmitter into operation,

Fig. 13-13 — A rear view of the three-stage 144-Mc. transmitter. The oscillator is at the left, with the buffer amplifier in the center. The 815 final is at the right.



the first step is to adjust the frequency range of the oscillator, using Lecher wires or a calibrated absorption-type wavemeter. This should be done after C_1 has been adjusted for maximum output. Then, using loose coupling between the buffer grid coil, L_2, and the oscillator tank coil, L_1 (the coupling may be adjusted by bending L_2 away from L_1 on its mounting lugs), adjust L_2 by changing the turn spacing until the grid circuit is resonant. Resonance will be indicated by maximum oscillator plate current; it can also be checked by measuring the voltage across the buffer grid leak, R_2, with a high-resistance voltmeter. The maximum voltmeter reading (about 40 volts) indicates resonance. The buffer should next be neutralized by varying the capacitance of C_7 until there is no change in the voltage across R_2 when the buffer tank condenser, C_8, is tuned through resonance. The point of correct neutralization also can be determined by coupling a sensitive absorption wavemeter such as is described in the chapter on Measuring Equipment to the buffer plate coil, and adjusting C_7 for minimum reading. With this method, care must be used to avoid coupling between the wavemeter and the oscillator; link coupling between L_3 and the wavemeter, with the latter far enough away so that it does not give a reading from the oscillator alone, should be used. Another method of checking neutralization is to adjust the turn spacing of the amplifier grid coil, L_4, to resonance and measure the 815 grid current (with no plate or screen voltage on the tube) and adjust C_7 for zero grid current.

After the buffer is neutralized, plate voltage may be applied and C_8 adjusted to resonance, as indicated by minimum plate current. If the coupling to the final amplifier is quite loose, the minimum plate current should be approximately 17 ma. The amplifier grid coil may next be resonated (by adjusting the spacing between turns) and the coupling increased until the maximum grid current is secured. The grid current should be 4 milliamperes or more and the buffer plate current should rise to about 28 ma.

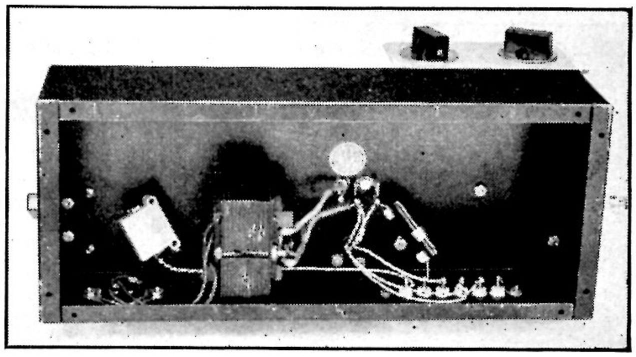


Fig. 13-14 — Underneath the chassis of the 144-Mc. MOPA transmitter. The filament transformer, amplifier plate by-pass condenser, and screen-dropping resistor are mounted here.

Neutralization of the 815 is the next step. The process is similar to that outlined for the 815 transmitter described at the beginning of the chapter. Plate and screen voltage may then be applied. With no load on the amplifier the plate current should dip to approximately 65 ma. at resonance. Loading the amplifier to a plate current of 150 ma. should not cause the grid current to drop below about 3.5 ma. A 40-watt lamp used as a dummy load should light to practically normal brightness at this input, using a plate-supply voltage of 400.

For greatest stability, the coupling between the oscillator and buffer should be as loose as possible. It is better to obtain the rated 815 grid current of 3 milliamperes by using tight coupling between the buffer and amplifier and loose coupling between the oscillator and buffer than vice versa. With normal operation the oscillator plate current should be approximately 25 ma. and the buffer plate current 28 ma., at 300 volts.

A modulator for the transmitter should have an audio output of 35 watts, using a coupling transformer designed to work into a 2500-ohm load. This transmitter is described in greater detail in *QST* for April, 1946.

144-Mc. Double Beam-Tetrode Power Amplifier

An amplifier set-up suitable for use with double beam-tetrode tubes is shown in Figs. 13-15, 13-16 and 13-17. The tube in the photographs is an 829, but an 815 or 832 can be used in the same layout. The only change that might be required would be in the inductances of the grid and plate coils, L_2 and L_3; these may have to be made slightly smaller or larger in diameter to compensate for the differences in input and output capacitances in the various types. The physical arrangement of the components is similar to that used for the 815 amplifier incorporated in the three-stage transmitter described above. When an 829 is used, the amplifier is well suited for use as an outboard unit with war-surplus transmitters such as the SCR-522.

The amplifier is built on an aluminum chassis formed by bending the long edges of a 5 × 10-inch piece of aluminum to form vertical lips ¾ inch high, so that the top-of-chassis dimensions are 3½ by 10 inches. The tube socket is mounted on a vertical aluminum partition measuring 3½ inches high by 3¼ inches wide on the flat face, with the sides bent as shown in the photographs to provide bracing. The partition is mounted to the chassis by right-angle brackets fastened to the sides. The socket is mounted with the cathode connection at the top, the cathode prong being directly grounded to the nearest mounting screw for the socket. The heater by-pass condenser, C_6, is mounted directly over the center of the tube socket, extending between the

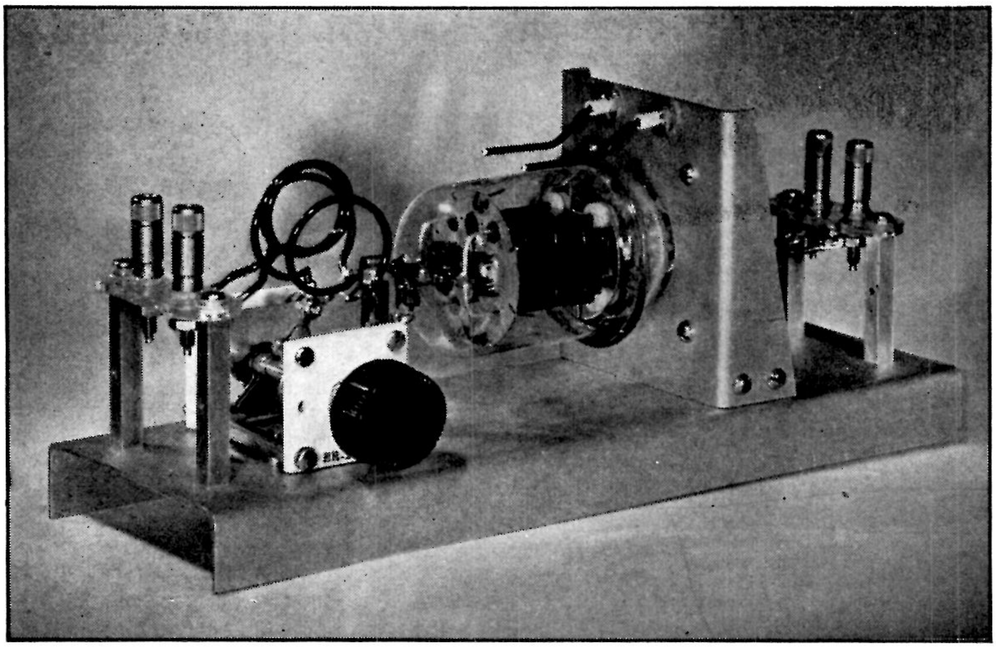


Fig. 13-15 — A 144-Mc. amplifier using a double beam tetrode. This type of construction is suitable for the 815 and 832 as well as the 829 shown. The vertical partition provides support for the tube as well as shielding between the input and output circuits. Note the neutralizing "condensers" formed by the wires near the tube plates.

paralleled heater prongs at the bottom and the cathode prong at the top. The screen by-pass is connected with short leads between the screen prong and the nearest socket screw.

The grid coil, L_2, is supported by the grid prongs on the socket. The two turns of the coil are spaced about one-half inch to allow room for the input coupling coil L_1 to be inserted between them. The coupling is adjusted by bending L_1 into or out of L_2. The grid tuning condenser, C_1, is mounted between the socket prongs; although the condenser has mica insulation it is used essentially as an air-dielectric condenser since the movable plate does not actually contact the mica at any setting inside the band. The coupling link is soldered to lugs under binding posts on a National FWG strip, the strip being mounted on metal pillars 1½ inches high to bring the link to the same height as the grid coil.

Although the shielding between the input and output circuits of the tube is sufficiently good so that the circuit will not self-oscillate, tuning of the plate circuit will react on the grid circuit to some extent because the grid-plate capacitance, while small, is not zero. To eliminate this reaction it is necessary to neutralize the tube. The neutralizing "condensers" are lengths of No. 12 wire soldered to the grid prongs on the socket. The wires are crossed over the socket and then go through small ceramic feed-throughs at the top of the vertical shield, projecting over the tube plates on the other side as shown in Fig. 13-15.

Connections between the plate tank condenser, C_7, and the tube plate terminals are made by means of small Fahnestock clips soldered to short lengths of flexible wire. The tank coil, L_3, is mounted on the same condenser terminals to which the plate clips make connection. The output link, L_4, is mounted similarly to the grid link except that the posts are 1⅞ inches high. The plate choke, RFC_1, is mounted vertically on the chassis midway between the plate prongs of the tube, the mounting means being a short machine screw threaded into the end of the polystyrene rod. The "cold" lead of the choke is by-passed by C_5 underneath the chassis.

Supply connections are made through a 5-post strip on the rear edge of the chassis. The dotted lines between connections in Fig. 13-16 indicate that these connections are normally short-circuited; leads are brought out so that the grid and screen currents can be measured separately.

In adjusting the amplifier, the plate and

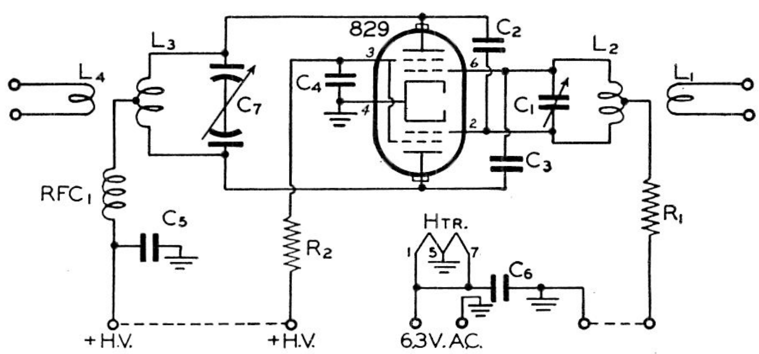


Fig. 13-16 — Circuit of the 829 amplifier for 144 Mc.

C_1 — 3–30-μμfd. ceramic trimmer.
C_2, C_3 — Neutralizing condensers; see text.
C_4 — 500-μμfd. mica, 1000 volts.
C_5 — 500-μμfd. mica, 2500 volts.
C_6 — 470-μμfd. mica.
C_7 — Split stator, 15 μμfd. per section (Cardwell ER-15-AD).
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.
L_1 — 2 turns No. 12, diameter ½ inch.
L_2 — 2 turns No. 12, diameter ½ inch, length ½ inch.
L_3 — 2 turns No. 12, diameter 1⅛ inches, length 1 inch.
L_4 — 2 turns No. 12, diameter 1 inch.
RFC_1 — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.

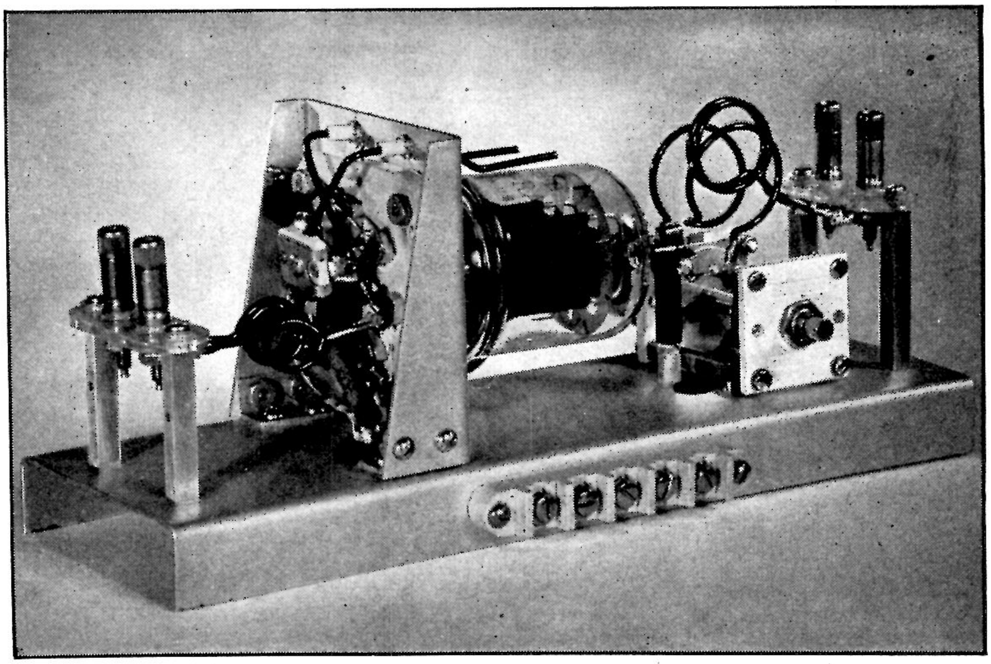


Fig. 13-17 — Another view of the 144-Mc. amplifier. The neutralizing wires are crossed over the socket before going through the feed-through insulators. The input circuit is designed for link coupling to the driver stage.

screen voltages should be left off and the d.c. grid circuit closed through a milliammeter of 0–25 or 0–50 range. The driver should be coupled to the amplifier input circuit through a link (Amphenol Twin-Lead is suitable, because of its constant impedance and low r.f. losses). Use loose coupling between L_1 and L_2 at first, and adjust C_1 to make the grid circuit resonate at the driver frequency, as indicated by maximum grid current. The coupling between L_1 and L_2 may then be increased to make the grid current slightly higher than the rated load value for the tube used — approximately 12 ma. for the 829. If the driver is an oscillator, the coupling between L_1-L_2 should be as loose as possible with proper grid current.

After neutralization, the procedure for which has been given in connection with other similar amplifiers, plate and screen voltage may be applied. If possible, the plate voltage should be low at first trial so there will be no danger of overloading the tube. Adjust C_7 to resonance, as indicated by minimum plate current (this should be measured independently of the screen); with the 829, the minimum plate current should be in the neighborhood of 80 milliamperes with 400 volts on the plate and no load on the circuit. A dummy load such as a 60-watt lamp should light to something near full brilliance when the coupling between L_3 and L_4 is adjusted to make the tube draw a plate current of 200 ma. When the loading is set, the grid current should be checked to make sure it is up to the rating for the tube. If it has decreased, the coupling between L_1 and L_2 should be increased to bring it back to normal.

Power-supply and modulator requirements will depend upon the particular tube used. For the 829, the plate supply should have an output voltage of 400 to 500 with a current capacity of 250 milliamperes. With a 400-volt supply the modulator power required is 50 watts, with an output transformer designed to work into a 1600-ohm load; with a 500-volt supply slightly over 60 watts of audio power is needed, the load being 2000 ohms.

A Low-Cost 2-Meter Transmitter

Until very recently, most 144-Mc. stations employed simple transmitters of the modulated-oscillator type. Since the superregenerative receiver was also widely used, the instability of the transmitters was not a matter of great importance; but with the rapid swing to stabilized transmitters and selective receivers now in evidence, most of the modulated-oscillator signals are no longer readable. It is, however, still possible, by careful design and proper operation, to use the simple and economical oscillator rig and yet radiate a signal that can be copied on all but the most selective receivers. Such a transmitter is shown in Figs. 13-18, 13-21 and 13-22.

Oscillator Ills and Their Treatment

There are two principal faults in most simple 2-meter transmitters. Many use filament tubes with a.c. applied to the filaments, causing severe hum modulation. Others, through poor design, have insufficient feed-back (as evidenced by low grid current) so that they are unable to sustain strong oscillation under load. Lack of sufficient excitation also renders them incapable of maintaining oscillation at low plate voltages, causing them to go out of oscillation over a considerable portion of the modulation cycle. Such oscillators suffer from extreme frequency modulation, making their signals unreadable on all but the very broadest

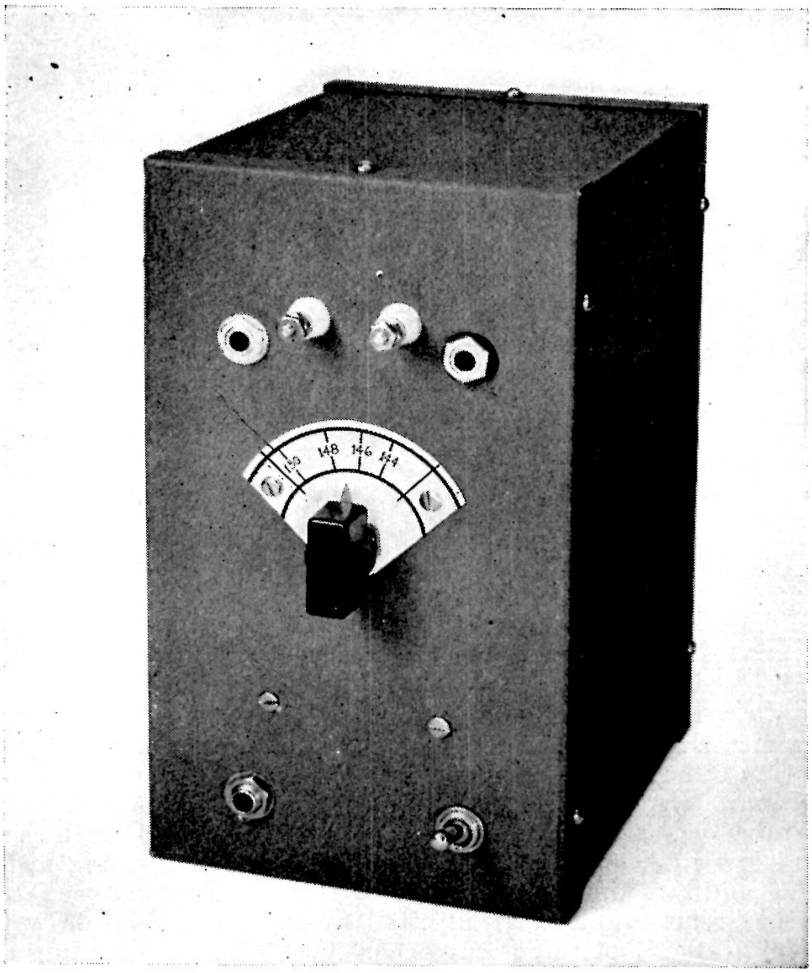


Fig. 13-18 — Front view of the simple 144-Mc. transmitter. The jacks at each side of the antenna terminals are for insertion of a meter in the oscillator grid (left) and plate (right) circuits. The microphone jack is at the lower left and the on-off switch is at the right. The calibration scale is drawn with India ink on heavy white paper.

♦

receivers, and even on these the quality is poor indeed.

No simple transmitter can hope to overcome these faults entirely, but they are materially reduced in the rig described herewith. A.c. hum modulation is reduced through the use of indirectly-heated tubes; and stability is improved through the use of a high-*C* push-pull oscillator, employing the familiar "unity-coupled" circuit. This arrangement, wherein the grid coil is fed through the inside of a plate tank made of copper tubing, provides adequate excitation. Stability over wide ranges of plate voltage is quite good, and the degree of frequency modulation is not too severe if the modulation is held to 75 per cent or less. It is laid out so that it is stable mechanically, reducing possible frequency changes from vibration.

Mechanical Details

The transmitter is designed for use with a plate supply of 250 to 300 volts, making it useful for mobile or low-powered home-station operation. It employs a pair of 2C22 tubes (also known as 7193s) as oscillators, a 6V6GT modulator, and a 6C4 as a speech amplifier and source of microphone voltage. It is housed in a standard 5 × 6 × 8-inch utility cabinet, the back and front of which are removable. The schematic diagram is shown in Fig. 13-19.

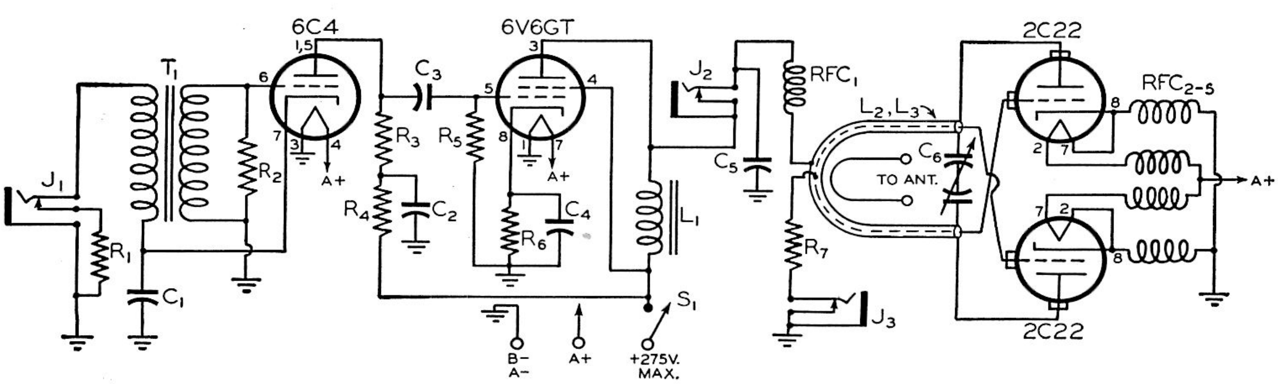


Fig. 13-19 — Schematic diagram of the simple 144-Mc. transmitter.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3, C_5 — 0.01-μfd. 600-volt paper.
C_6 — "Butterfly" variable (Cardwell ER-14-BF/S modified; see text).
R_1 — 470 ohms, 1 watt.
R_2 — 0.33 megohm, ½ watt.
R_3, R_4 — 5000 ohms, 5 watts.
R_5 — 0.47 megohm, ½ watt.
R_6 — 680 ohms, 1 watt.
R_7 — 10,000 ohms, 1 watt.
L_1 — Midget filter choke.
L_2, L_3 — Unity-coupled grid and plate coils. See text and Fig. 13-20.
J_1, J_2, J_3 — Closed-circuit jack.
RFC_1 — No. 28 d.s.c. wire, close-wound on 1-watt resistor, ¼-inch diam., ⅝ inch long.
RFC_2, RFC_3, RFC_4, RFC_5 — 20 turns No. 20 d.s.c. wire close-wound on ¼-inch polystyrene rod.
S_1 — S.p.s.t. toggle switch.
T_1 — Single-button microphone transformer (UTC "Ouncer" — surplus).

The plate tank "coil" is made of $\frac{3}{16}$-inch copper tubing, bent into a "U" which is two inches long overall. The ends of the "U" are made into spade lugs, as shown in Fig. 13-20, the slotted ends providing a small range of inductance adjustment. The lug ends are fastened directly to two of the stator terminals of the butterfly-type tank condenser, C_6. Part of the "U" is cut out at the curved end, to provide an opening for the center-tap of the

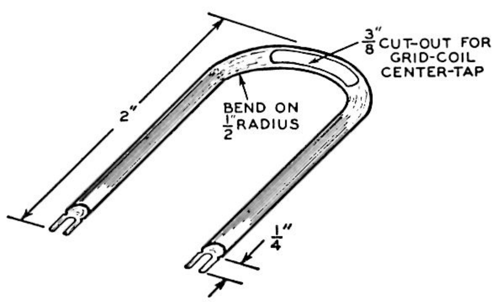


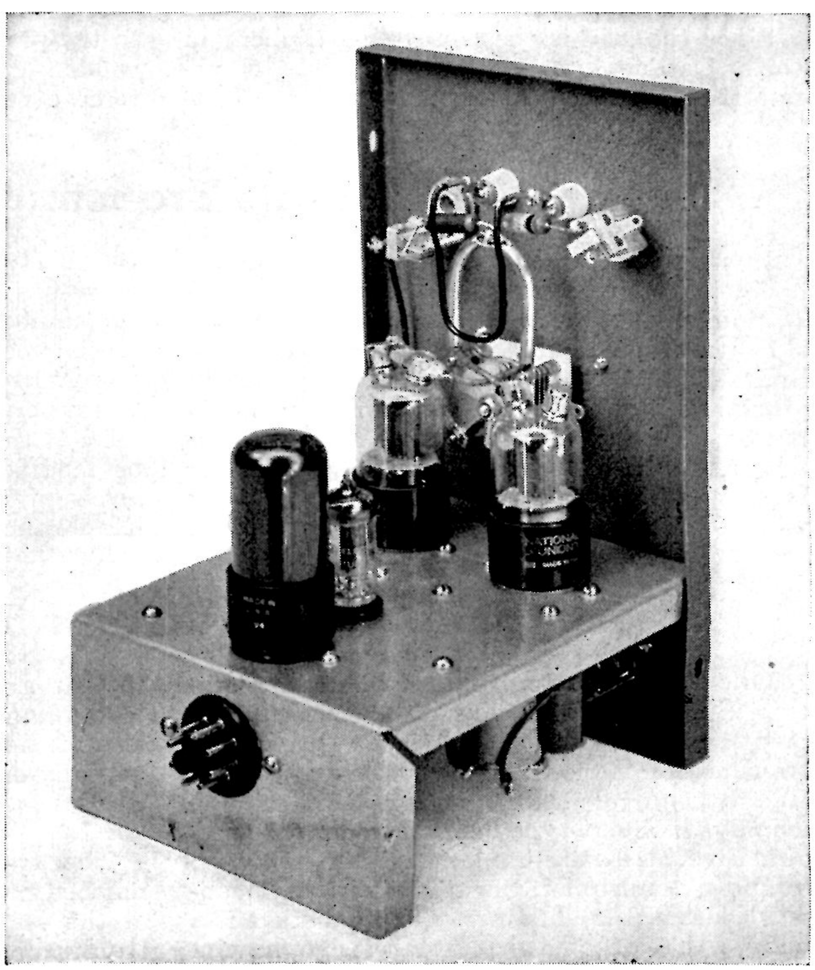
Fig. 13-20 — Detail drawing of the oscillator plate inductance. It is made from $\frac{3}{16}$-inch copper tubing, bent into a "U" shape. Ends of the "U" are formed into spade lugs, the slots in which provide a means of slight inductance adjustment. It is mounted directly on the stator terminals of the tuning condenser.

grid coil. An easy way to make the grid coil is to cut two pieces of flexible insulated wire, about four inches long, and feed them into the "U" through the center opening. The protruding tap, made by twisting the ends of the wires together, should be coated with household cement after the grid resistor has been soldered to it. Note that the grid leads are transposed. The 2C22s will not oscillate if these are improperly connected. The plate leads may be made of $\frac{1}{4}$-inch copper braid, or copper or silver ribbon is even better, if available. If braid is used, it may be made solid at the end by flowing solder over the last half inch, after which it may be drilled, to pass the stator terminal screw.

Provision is made for reading both grid and plate current to the oscillator, two meter jacks being mounted on either side of the plate tank. Their terminals make convenient mounting places for R_7 and RFC_1. Note that the jacks are connected so that the meter leads need not be reversed when changing from one jack to the other. The plate-meter jack must, of course, be insulated from the metal panel.

No battery is required for microphone current, this being obtained by running the cathode current of the 6C4 speech amplifier through the microphone transformer. The 6C4 cathode is by-passed with a large electrolytic condenser, and the plate is decoupled and by-passed to reduce hum. Since the 6C4 stage is used principally as a source of microphone current, resistance coupling to the 6V6GT modulator gives adequate drive. No gain control is included, as the full output of the modulator is insufficient for overmodulation.

Fig 13-21 — Back view of the 2-meter transmitter, showing the symmetrical arrangement of components. Note that the "U"-shaped tank inductance is mounted directly on the stator terminals of the butterfly tuning condenser.



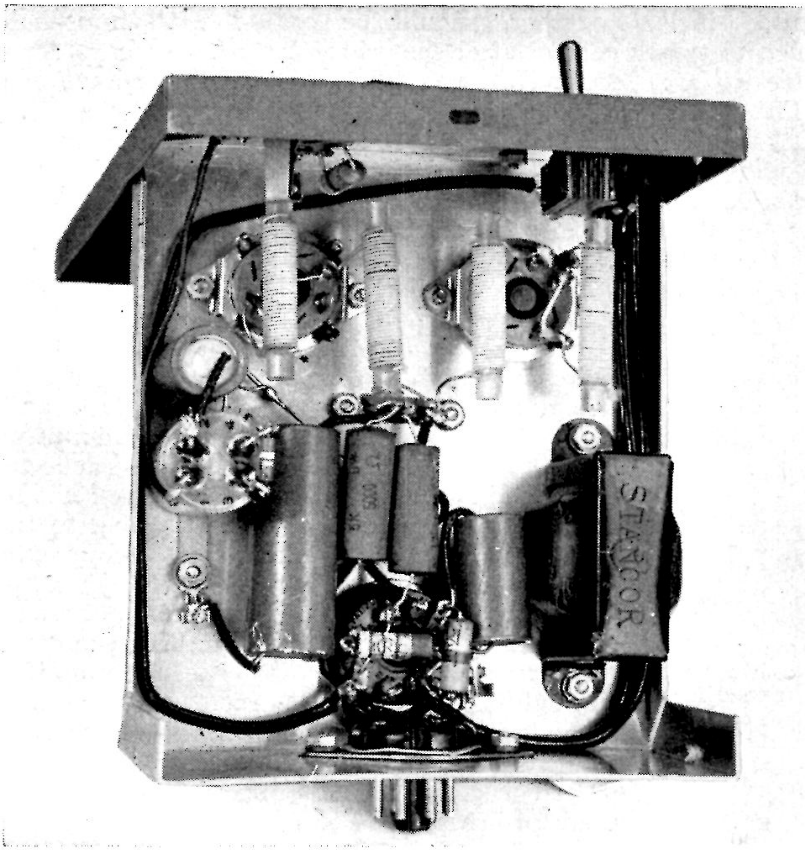


Fig. 13-22 — Under-chassis view shows the four heater chokes and audio components. The small round object, left center, is the microphone transformer, a surplus midget unit. The audio choke is at the right.

Testing

Since the grid is the controlling element in the operation of any Class C stage, it is important that the grid current be observed in adjusting the oscillator. The plate current may be almost meaningless, as an indication of the proper functioning of such a stage, but the grid current shows plainly if the oscillator is functioning correctly. If the grid current and bias are normal for the tubes used, the plate current can be ignored, except to see that the input is not excessive. Grid current in this oscillator should run about 8 ma. when a plate voltage of 275 or so is used and the oscillator is loaded by a lamp or antenna. The "U"-shaped antenna-coupling loop should be adjusted until the grid current is approximately this value. The plate current will be about 60 ma. with 275 volts on the plates.

The transmitter frequency should be checked with Lecher wires, or by listening to the signal in a calibrated receiver. In either case there should be a load across the antenna terminals, as the frequency may be appreciably different between loaded and unloaded operation.

The rough calibration scale shown was first roughed on a white card using pencil, and afterward drawn over in India ink. The calibration card is glued to the panel, and further held in place by the condenser mounting nut and two small machine screws.

Mobile Transmitters

In most respects, gear designed for mobile operation is similar to that used for home-station service, except for the additional considerations imposed by space and current-drain limitations and the need to withstand constant vibration. Though there are various types of power supplies capable of delivering more power, the most satisfactory arrangement for most mobile installations is the generator or vibrator supply that furnishes 300 volts at 100 to 150 ma. This power is within the capability of the average family-car battery, making unnecessary the separate batteries and special generators usually needed when higher-powered systems are employed. The transmitters described below are designed for 300-volt service, though in several instances they may be modified readily for use at higher power levels, if the car battery and generator will handle the extra load imposed.

A Mobile Transmitter for 50 and 28 Mc.

Low over-all battery drain in mobile operation is best obtained through the use of filament-type tubes which are lighted only during transmission periods. The mobile unit for 6, 10 and 11 meters, shown in Figs. 13-23–13-27, employs filament-type beam tetrodes throughout. Five 2E30s are used, as crystal oscillator, frequency multiplier, Class A driver, and push-pull Class AB modulators. The final stage is a 2E25, a tube of somewhat larger design, having its plate connection at the top of the envelope. Total filament current is only 4.3 amperes, and there is no drain whatever when the rig is not actually on the air.

Mechanical Details

The transmitter is housed in a crackle-finished cabinet which may be mounted in back of the seat in coupé-type vehicles or in the trunk compartment of sedans.

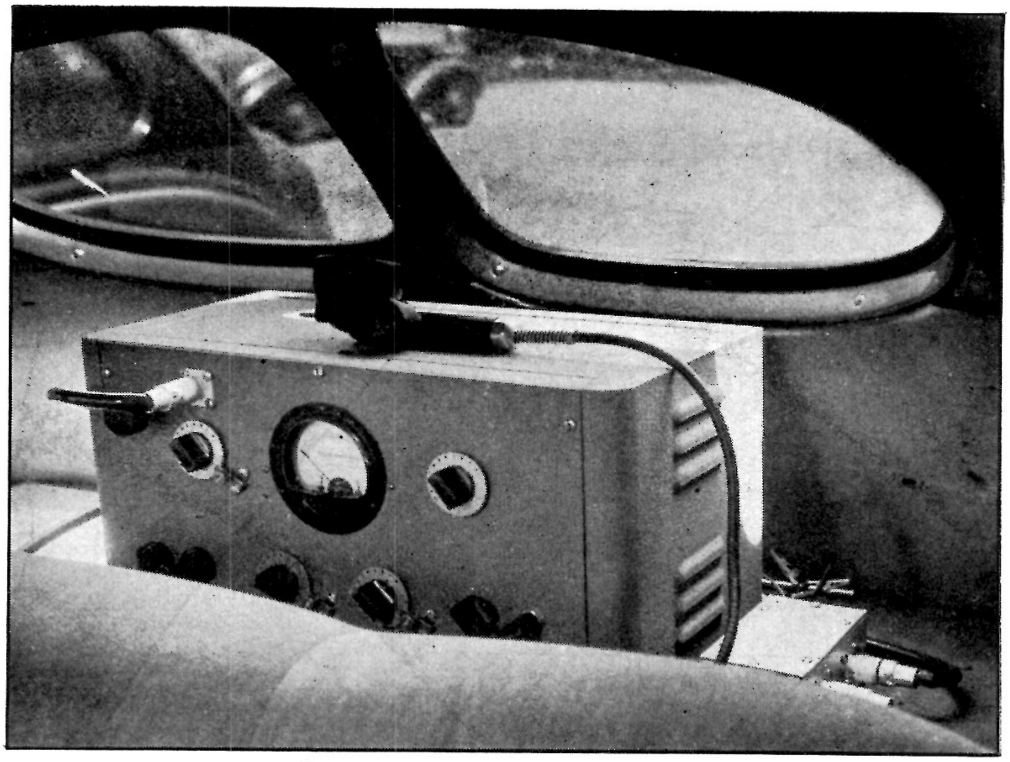


Fig. 13-23 — A typical installation of the 6- and 10-meter mobile transmitter. The small aluminum box at the right of the unit houses the antenna change-over relay. The genemotor and its starting relay are mounted under the hood, adjacent to the car battery. Operation of the transmitter is controlled entirely by the push-to-talk switch on the hand microphone.

Special attention is paid to ruggedness of construction, all leads being made as short and direct as possible. Small components are supported with terminal strips at each end where possible, and tuning controls are equipped with dial locks (National ODL). The meter (a Marion 0–10-ma. sealed unit) is back-of-panel mounted, with a sheet of Lucite serving as a protecting window. This method of mounting the meter, about ½ inch in back of the panel, also provides a convenient method for illuminating the meter face. Dial lights are mounted at either side of the meter, as shown in Figs. 13-24 and 13-26.

By using 100-μμfd. variable condensers for C_2 and C_3, the range of the oscillator and multiplier plate circuits is extended, so that it is unnecessary to change these coils in changing bands. Only the crystal and the final plate coil, L_5, need be changed. Complete push-to-talk operation is made possible through the use of two relays. Ry_1 starts the genemotor and applies the filament voltage to the transmitter. Ry_2 transfers the antenna from receiver to transmitter. Both are controlled by the switch on the microphone, which may be any single-button type that has a control switch. The Army T-17-B, now currently available as government surplus, is shown with the rig.

The Circuit

The crystal oscillator is a Tri-tet, modified for filament-type tubes. Interwound coils are inserted in the filament leads, and one of these is tuned. The setting of this adjustment is not critical and may be left near maximum setting, for both 7- and 8.4-Mc. crystals. The oscillator doubles in its plate circuit for both bands.

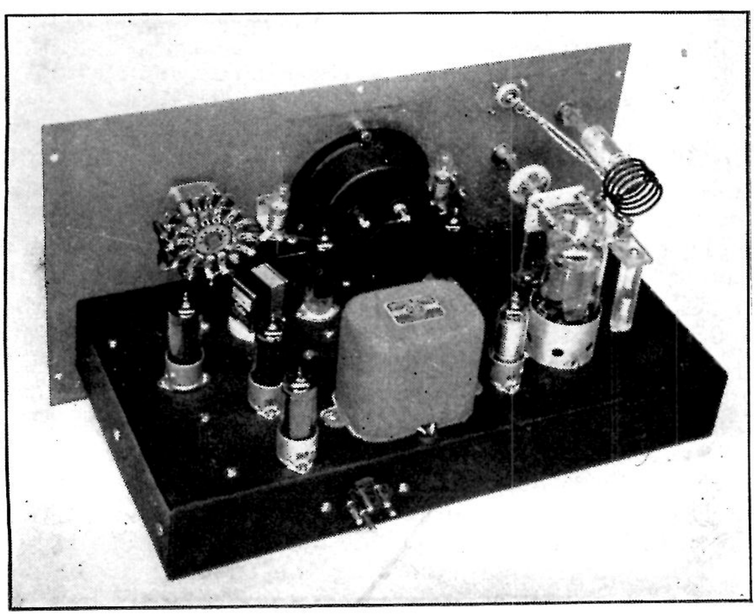


Fig. 13-24 — The plate circuit of the final stage of the mobile transmitter is the only r.f. circuit above the chassis. The three tubes at the left are the driver and audio stages, with the oscillator and multiplier tubes directly in back of the meter. The tube to the right of the modulation transformer is the 0A2 voltage regulator. Chassis size is 7 × 13 × 2 inches.

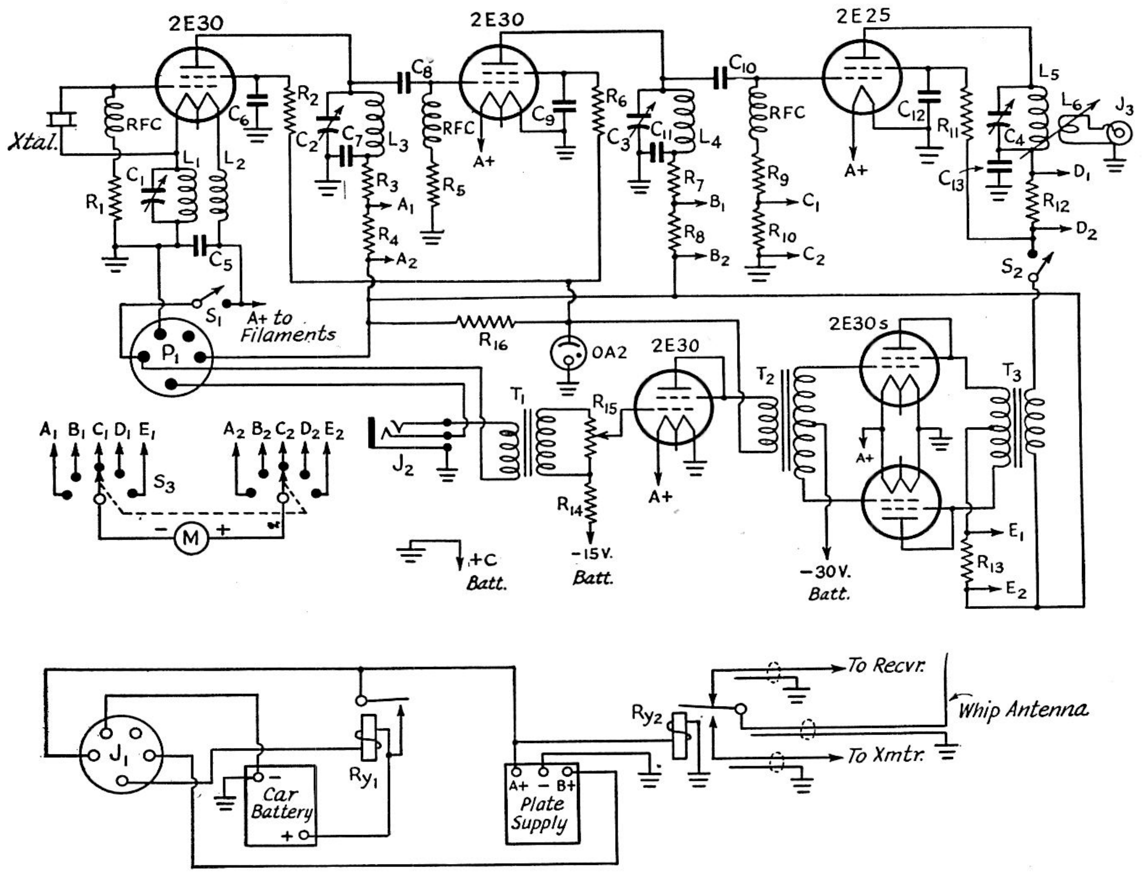


Fig. 13-25 — Wiring diagram of the mobile rig for 6 and 10 meters.

C_1 — 100-$\mu\mu$fd. midget, screwdriver-adjustment type (Hammarlund APC-100).
C_2, C_3 — 100-$\mu\mu$fd. midget, shaft type (Hammarlund HF-100).
C_4 — 15 $\mu\mu$fd., double spaced (Hammarlund HFA-15-E).
C_5 — 0.001-μfd. mica.
C_6, C_7, C_9, C_{11}, C_{12}, C_{13} — 470-$\mu\mu$fd. midget mica.
C_8, C_{10} — 100-$\mu\mu$fd. midget mica.
R_1 — 82,000 ohms, 1 watt.
R_2, R_6 — 1000 ohms, ½ watt.
R_3, R_7, R_{10} — 100 ohms, ½ watt.
R_4, R_8, R_{12}, R_{13} — Special shunts. (See text.)
R_5 — 150,000 ohms, 1 watt.
R_9 — 33,000 ohms, 1 watt.
R_{11}, R_{16} — 5000 ohms, 10 watts.
R_{14} — 10,000 ohms, ½ watt.
R_{15} — 0.5-megohm potentiometer.
L_1, L_2 — 7 turns each, No. 20 d.c.c., 9/16 inch long on 1-inch dia. form, windings interwound.
L_3 — 10 turns No. 12 enam., close-wound on 1-inch diam. form.
L_4 — 6 turns No. 12 enam., ¾ inch long, ½-inch inside diam., self-supporting.
L_5 — 28 Mc.: 10 turns No. 12 enam., 1½ inches long, 1-inch inside diam., self-supporting.
50 Mc.: 5 turns No. 12 enam., 1 inch long, 1-inch inside diam., self-supporting.
L_6 — 3 turns on ½-inch polystyrene rod — see text and detail photo.
J_1 — Socket on power cable, 5 prong.
J_2 — Double-button microphone jack. If T-17-B microphone is used, a special jack designed for this microphone must be obtained.
J_3 — Coaxial fitting (Amphenol 83-1R. Matching plug is 83-1SPN).
M — 0–10-ma. sealed unit (Marion).
P_1 — Power plug on transmitter chassis.
RFC — 2.5-mh. r.f. choke, National R-100.
R_{y1}, R_{y2} — See text.
S_1, S_2 — S.p.s.t. snap switch.
S_3 — 2-section 5-position wafer-type switch.
T_1 — Single-button microphone transformer.
T_2 — Driver transformer (Stancor A-4752).
T_3 — Modulation transformer (UTC S-18).

The stage following the oscillator is operated as a doubler for 27- and 28-Mc. work, and as a tripler for 50 Mc. The 2E30 is an effective frequency multiplier, and there is adequate excitation for the final in either case. Screen voltage on the exciter stages is stabilized with a miniature voltage-regulator tube, an 0A2. With a screen voltage of 150, the plate input to both 2E30s is held to about 6 watts per tube.

The final stage uses a 2E25, whose top-cap plate connection permits the mounting of the plate circuit above the chassis, well isolated from the other tuned circuits. A small shield, cut from an old-style tube shield to a length of about one inch, comes up to the bottom of the 2E25 plate assembly. These precautions are sufficient to provide completely-stable operation without neutralization.

The antenna coupling coil, L_6, is wound on a short length of polystyrene rod ½ inch in diameter, into which is inserted a ¼-inch rod of the same material. This shaft projects through the front panel, where a shaft-locking panel bushing (Bud PB-532 bushing, Millen 10061 shaft lock) holds it in the desired position. Coupling is adjusted by pushing or pulling the knob affixed to the shaft, following which the bushing may be tightened for permanent setting. The bushing may also be set finger-tight, allowing the coupling to be ad-

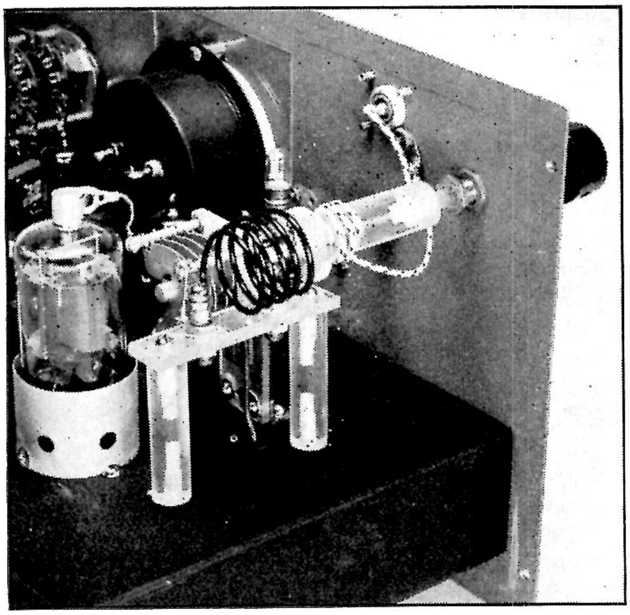


Fig. 13-26 — Detail photo of the 2E25 final stage, showing method of coupling to the antenna. The coupling coil, wound on a polystyrene rod, is adjustable from the front panel. The plate coil is mounted by means of G.R. plugs.

justed, yet holding it with sufficient firmness to prevent its being jarred out of position.

Three 2E30s are used for the modulator, one as a Class A driver and two in push-pull as Class AB modulators. All three are triode-connected. Bias is supplied by a 30-volt hearing-aid battery, which can be tapped at 15 volts by opening up the cardboard case and soldering on a lead at the point where the two 15-volt sections are joined together. This lead is brought out to the unused terminal on the battery socket and plug.

Metering of all circuits is provided by a 10-ma. meter, a 2-section 5-position switch, and a set of shunts. The shunts are made from small 100-ohm resistors, on which is wound about 7 feet of No. 30 enameled wire. The shunts should be wound with an excess of wire, the length of which may be reduced until the multiplication of the meter scale is just right. The resistor R_{10} in the final grid circuit is left without a shunt, giving direct reading on the 10-ma. scale for measuring the final grid current.

Testing

Except for the speech stages, the unit may be tested using 6.3 volts a.c. on the filaments and an a.c. power supply. A storage battery must be used for filament supply when the speech equipment is to be tested, as a.c. on the filaments will produce excessive hum. Initial testing should be carried on with about 200 volts on the tube plates. When operation has been found to be satisfactory, this may be raised to 300.

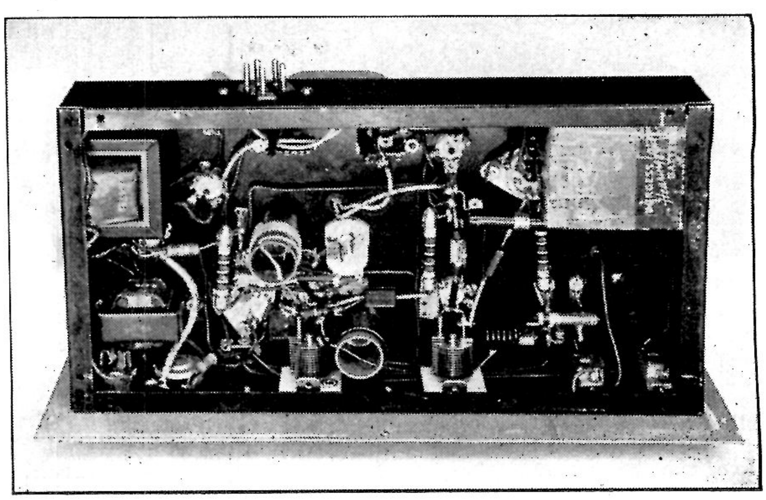
To place the unit in operation, set S_1 to the "on" position, leaving S_2 "off." With the meter switch in position *A*, apply plate voltage and note meter reading, which is the oscillator plate current. This will be about 20 ma., dipping slightly at resonance as C_2 is adjusted. Next switch to position *B* and adjust C_3. The dip here may not be as pronounced as in the oscillator, and the final grid current, position *C*, 10-ma. scale, is the best indication of resonance in the preceding adjustments. This reading should be about 4 ma., dropping to 3 ma. under load. With S_2 turned on, the final plate current, position *D*, should drop to below 10 ma. at resonance, and coupling of the antenna should raise it to 50 to 60 ma. Modulator plate current will be about 20 ma., rising to 60 ma. or more on audio peaks. No metering position is provided for the Class A driver current, but this should be approximately 10 ma.

With the coil and condenser values given, it is impossible to get output from the final stage on a wrong frequency, but excitation to the final may be obtained on incorrect harmonics; hence it is advisable to check the frequency of each stage with a calibrated absorption-type wavemeter.

Installation

For maximum convenience, the same antenna should be used for both transmission and reception. Antenna change-over is handled with a conventional 6-volt antenna relay which is mounted in a small box made up for the purpose from folded sheet aluminum. Amphenol coaxial fittings, mounted on the sides

Fig. 13-27 — Bottom view of the mobile rig. At the left center are the interwound coil and tuning condenser which are part of the oscillator filament circuit. Audio components are at the left, with oscillator and multiplier plate circuits near the front panel.



of the relay box as close to the relay contacts as possible, provide for connection to the transmitter, the receiver, and the antenna by means of coaxial line. The relay case is grounded and only the inner conductor of the coaxial line is switched.

A headlight relay for genemotor starting may be purchased from any auto-accessory store, and this and the genemotor should be mounted as close to the car battery as possible, in order to minimize voltage drop. Battery wiring and filament cables should be as heavy wire as possible, with No. 12 as the minimum for the genemotor leads.

For actual mobile operation, the quarter-wave telescoping "whip" antenna, operating as a Marconi in the manner shown in Fig. 14-11, is convenient. Much greater range in stationary operation from high locations may be had with half-wave radiators or multielement arrays, either of which may be arranged for easy on-the-spot assembly. An example of such a portable array for 50 Mc. is shown in Fig. 14-12, Chapter Fourteen.

Mobile Transmitter for 144-Mc.

A crystal-controlled transmitter designed for mobile operation on 2 meters is shown in Figs. 13-28 through 13-31. It includes a modulator and will handle 15 watts input when used with a 300-volt power supply.

The circuit diagram of the transmitter, Fig. 13-29, shows a Type 6V6GT tube used in a Tri-tet oscillator, with a 24-Mc. crystal. The oscillator has a fixed-tuned cathode circuit and a self-resonant plate tank coil that tunes to 48 Mc. A series-dropping resistor, R_2, maintains the screen voltage at the proper level, when a 300-volt supply is used. The oscillator, like the other stages of the transmitter, includes a jack wired in the cathode lead for metering purposes.

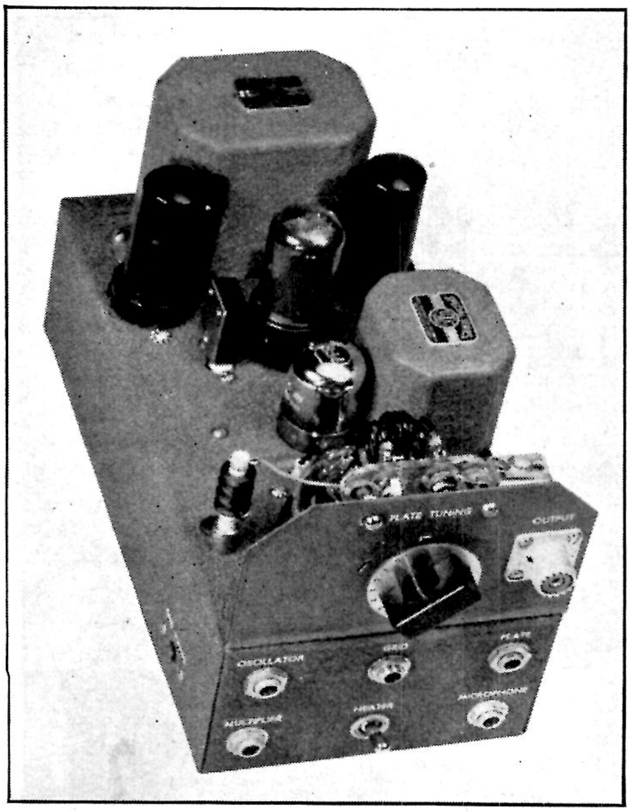


Fig. 13-28 — **Crystal-controlled transmitter for 144-Mc. mobile use. The amplifier plate tuning condenser and the antenna jack are mounted on the front panel along with the pick-up link assembly and the antenna trimmer condenser. The microphone jack, the metering jacks, and the filament on-off switch are mounted on the front wall of the chassis. A hole to permit screwdriver adjustment of the amplifier grid condenser is located at the left side of the chassis.**

A 7F8 dual triode serves as a push-pull tripler to 144 Mc. The tripler grid coil, L_3, is tuned by the trimmer condenser, C_1, and its plate coil, L_4, is self-resonant. Cathode bias, developed across R_4, prevents excessive plate current flow during off-resonant adjustment of either the oscillator or the tripler circuits.

The final stage employs an 832A dual tetrode as a neutralized amplifier on 144 Mc. Grid bias is developed by the flow of grid current through resistor R_5. If desired, protective bias can be used by decreasing the value of R_5 and by connecting a battery in series with the grid resistor.

A small amount of neutralization was required to assure completely-stable operation. The neutralizing condensers, C_{11} and C_{12} in the circuit diagram, are pieces of No. 12 wire extending from the grid of one section of the 832A to the vicinity of the plate of the other section. The wires are crossed at the bottom of the tube socket and go through Millen 32150 bushings mounted in the chassis between the 7F8 and the 832A sockets. It is possible that use of a shielded tube socket would eliminate the tendency toward oscillation in the 832A.

A series-tuned antenna circuit, consisting of C_4 and L_7, is intended for use with any of the low-impedance antenna feed systems commonly used for mobile work. The amount of loading is adjusted by varying the position of the pick-up link, L_7.

The modulator employs a pair of 6V6 tubes working Class AB. A speech-amplifier stage is not required as long as a single-button carbon microphone is used. Voltage for the microphone is taken from the junction of the two cathode-biasing resistors, R_7 and R_8, thus eliminating the need for a microphone battery.

The microphone and modulation transformers used are both large and expensive for the job at hand and were used only because they happened to be available. The microphone transformer can be any single-button-microphone-to-push-pull-grids transformer and the modulation transformer need not be rated at more than 10 watts. It should be capable of matching a pair of 6V6 tubes to an r.f. load of

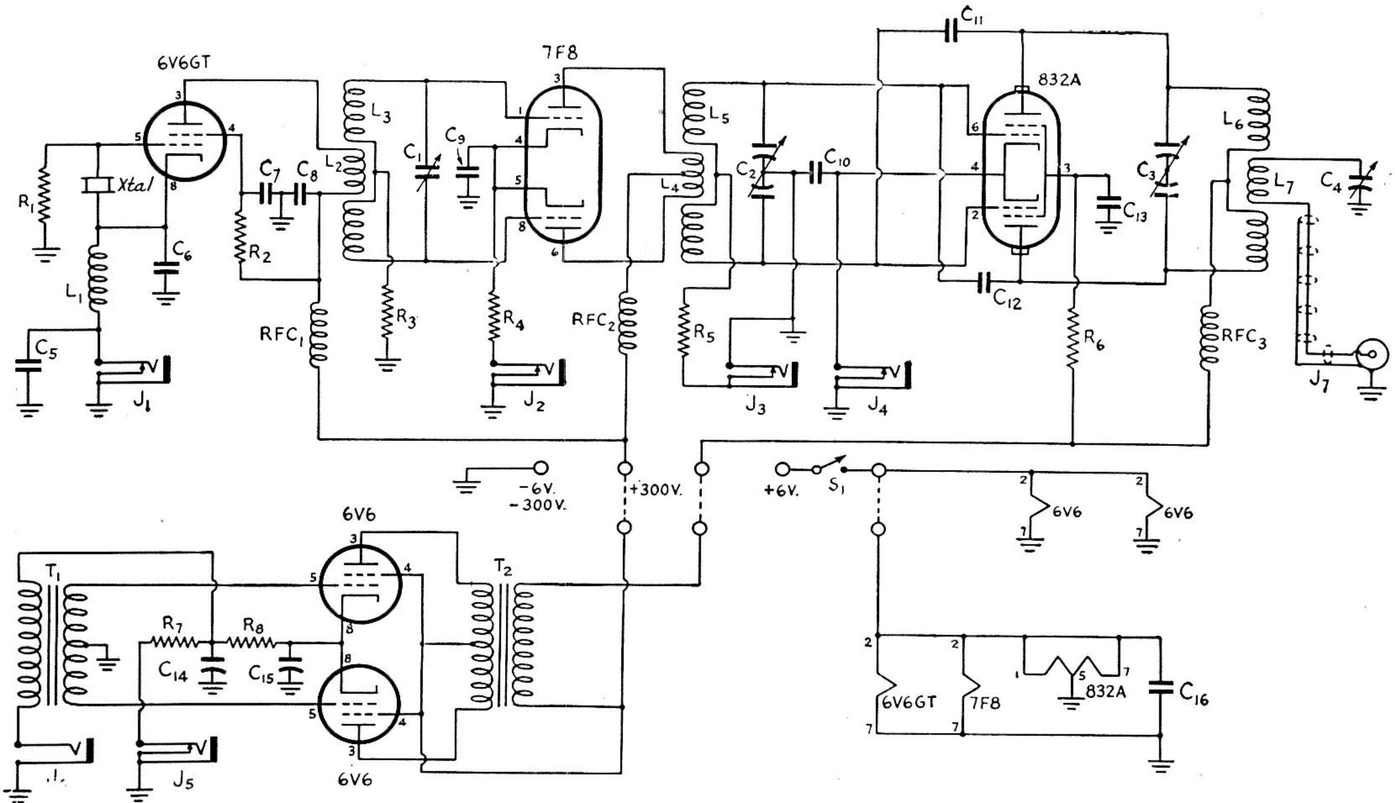


Fig. 13-29 — Circuit diagram of the mobile transmitter for 144 Mc.

C1, C4 — 3-30-μμfd. mica trimmer.
C2 — 15-μμfd.-per-section split stator (Bud LC-1660).
C3 — "Butterfly" condenser, 6 μμfd. per section (Cardwell ER-6-BF/S).
C5, C7, C8 — 0.0047-μfd. mica.
C6 — 100-μμfd. midget mica.
C9, C10, C13, C16 — 470-μμfd. midget mica.
C11, C12 — Neutralizing wires. (See text.)
C14, C15 — 10-μfd. 25-volt electrolytic.
R1 — 0.1 megohm, ½ watt.
R2 — 47,000 ohms, ½ watt.
R3 — 33,000 ohms, ½ watt.
R4 — 470 ohms, ½ watt.
R5 — 22,000 ohms, ½ watt.
R6 — 25,000 ohms, 10 watts.
R7 — 100 ohms, 1 watt.
R8 — 150 ohms, 1 watt.
L1 — 3 turns No. 18 enam., close-wound, ½-inch diam.
L2 — 4 turns No. 18 enam., ⅜ inch long.
L3 — 10 turns No. 18 enam.; coil wound in two sections with 5 turns each side of *L*2, each section ⅜ inch long. A ½ inch is left between windings. Form for *L*2*L*3 is a Millen 30003 Quartz-Q stand-off insulator, ¾-inch diam.
L4 — 3 turns No. 18 enam., ½ inch long, 9/16-inch diameter.
L5 — 2 turns No. 18 enam., interwound with turns of *L*4. *L*4 and *L*5 are wound on a National PRE-3 coil form.
L6 — 4 turns No. 12 enam., ½-inch i.d., wound in two sections with 2 turns each side of center-tap and a ½-inch space at the center, turns spaced wire diameter.
L7 — 3 turns No. 12 enam., ½-inch diam., turns spaced wire diameter.
J1-J5 — Closed-circuit jack.
J6 — Open-circuit jack.
J7 — Coaxial-cable connector.
RFC1, RFC2 — 300-μh. r.f. choke (Millen 34300).
RFC3 — 2.5-mh. r.f. choke (Millen 34102).
S1 — S.p.s.t. toggle.
T1 — Single-button microphone transformer (UTC S-7).
T2 — Modulation transformer (UTC S-19).

5000 to 7000 ohms, depending upon the input at which the 832A is operated.

It will be noticed that the modulator input and output connections are not wired directly into the transmitter proper and that separate terminal blocks are used for the r.f. and audio sections. When the modulator is used with the 144-Mc. transmitter, jumper connections are made between the two sets of terminals as shown by the circuit diagram. Removal of the jumpers will allow the modulator to be used with another transmitter, which might very well be the 50-Mc. rig shown in Figs. 13-32 to 13-35. *S*1, the filament switch, will open and close the modulator- and r.f.-section filament circuits when the jumper connections are made, and will operate with the modulator alone when the jumpers are removed.

The photographs of the transmitter show how the parts are mounted on a metal chassis measuring 3 × 5 × 10 inches. The front panel measures 3 × 5 inches and has a ½-inch lip for fastening to the chassis. The construction of the antenna assembly and the method of mounting the components on the panel are identical to the 50-Mc. transmitter. A recommended system of mounting the 832A tube socket is also detailed in the text referring to the 50-Mc. unit.

No special care need be given to the wiring of the audio circuit, but the r.f. leads should be kept as short as possible. The use of four tie-point strips will simplify the mounting and wiring of parts. A single tie-point is mounted to the rear of the oscillator tube socket and is used as the junction of *R*7, *R*8, *C*14 and the primary lead of the microphone transformer. A double tie-point strip is mounted to the right

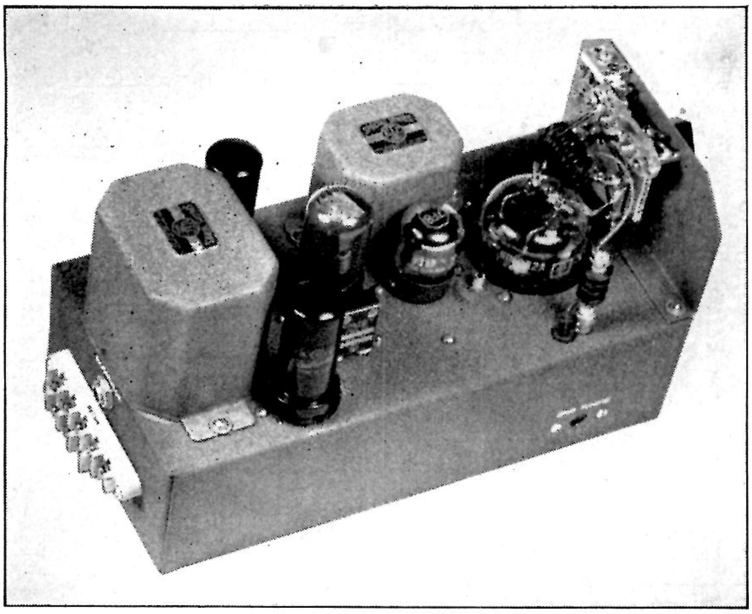


Fig. 13-30 — A top view of the mobile transmitter. The modulation transformer and the modulator tubes are at the left end of the chassis. The 6V6GT crystal oscillator, the 7F8 frequency multiplier, and the 832A amplifier are in line along the center of the chassis. The microphone transformer is in back of the 7F8. J_5, the modulator metering jack, is mounted on the rear wall of the chassis just above the terminal strips.

of the crystal socket (as seen in Fig. 13-31); one lug is used as the connecting point for the positive high-voltage lead and the bottom ends of RFC_1 and RFC_2; the bottom of L_1 and the top ends of C_5 and J_1 are connected to the second terminal. The cathode end of L_1 is connected to the cathode side of the crystal socket. The third tie-point strip is mounted on the 832A tube socket and serves as the connecting point between R_4 and J_2; the bottom end of R_6 connects to the high-voltage lead at the second lug. The fourth strip (single lug) is mounted on the frame of C_2 and the leads between R_5 and J_3 join at this point.

The construction of the driver-stage coils is not difficult if the coil forms are properly prepared in advance. A study of Fig. 13-31 will show how the windings are placed on the forms, and the lengths of the windings are given in the parts list. The forms should be marked and drilled to accommodate the windings with the holes for the ends of the windings passing directly through the forms. L_3 should be wound in two sections with the inside ends being soldered together after the winding of L_2 has been completed. The center-taps for L_4 and L_5 are made by cleaning and twisting the wire at the center of each winding. Condenser C_1 is soldered across the grid ends of L_3 before the coil is connected to the tube socket.

Adjustment and Testing

When testing the transmitter, it is advisable to start with the high voltage applied to the first two stages only. With a 100-ma. meter plugged in J_1 the oscillator cathode current at resonance should be approximately 30 ma. A low-range milliammeter should now be plugged in J_3 and the final grid circuit should be brought into resonance by adjustment of C_2. Proper operation of the tripler stage will be indicated by a cathode current of approximately 20 ma. and a final-amplifier grid current of 2.5 to 3 ma. The tripler grid condenser, C_1, should be retuned after the amplifier grid circuit has been peaked, to assure maximum overall operating efficiency.

The amplifier should be tested for neutralizing requirements after adequate grid drive has been obtained. If a well-shielded tube socket has been used, it is possible that the amplifier grid current will not be affected by tuning the 832A plate circuit through resonance. However, if the grid current does kick down as the plate circuit is tuned, it will be necessary to add the neutralizing wires referred to in the text and parts list as C_{11} and C_{12}. After installation these wires should be adjusted until no kick in grid current is seen as the 832-A plate circuit is tuned through resonance.

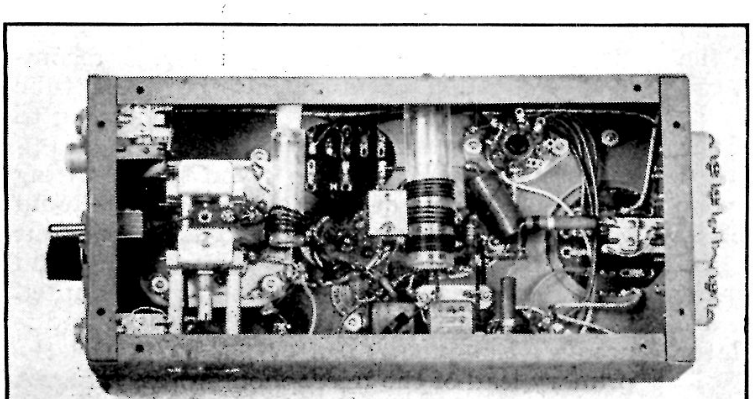


Fig. 13-31 — Bottom view of the 144-Mc. transmitter. The coil forms for L_2L_3 and L_4L_5 are mounted on the side wall of the chassis; the form for L_4L_5 is mounted on a small stand-off insulator so that the windings can be brought out to the center line of the chassis. C_1, the grid condenser for the frequency multiplier, is soldered across the grid ends of L_3. The amplifier grid tuning condenser, C_2, is mounted on metal pillars having a length of 1⅝ inches.

Plate and screen voltage can now be applied to the 832A and the plate circuit tuned to resonance, as indicated by a dip in the cathode current to 40 ma. or less. Then a dummy load (a 15-watt light bulb will do) is connected to the antenna jack and the loading adjusted by varying the position of L_7 and the capacitance of C_4, to cause a cathode current of 60 to 70 ma. Approximately 10 ma. of the total cathode current will be drawn by the screen of the 832A and this value should be subtracted from the cathode current in determining the plate input. Amplifier grid-current should be 1.5 to 2 ma. under load.

The jumper connections can now be made between the r.f. and modulator terminal blocks and, with power applied to the entire transmitter, the modulator cathode current should be 75 ma.; 85 ma. with modulation.

The cathode meter reading will decrease slightly when the microphone is plugged into the circuit. This is caused by the parallel current path that exists when the microphone circuit is completed.

A Mobile Low-Power 50-Mc. Transmitter

The transmitter shown in Figs. 13-32 through 13-35 is designed for mobile operation with a power input of 12 to 20 watts, as a companion unit to the 144-Mc. rig described above. However, the rating of the amplifier tube used is such that the unit can be used at higher levels, delivering an output of approximately 20 watts if a 500-volt supply is available.

As may be seen from the circuit diagram, in Fig. 13-33, a Tri-tet oscillator-doubler, employing a 6V6GT tube and a 25-Mc. crystal, is used to drive an 832A amplifier. The oscillator requires no manual adjustment, once set, as the cathode circuit is fixed-tuned and the plate circuit has a self-resonant coil. The screen voltage for the 6V6 is reduced to the proper value by the dropping resistor, R_2.

The push-pull amplifier employs split-stator input and output circuits. The grid circuit is inductively-coupled to the oscillator plate coil, and grid bias is developed across resistor R_3. Usually, for the sake of convenience, it is desirable to employ self-bias during mobile operation and the amplifier, as shown, is set up for this type of operation. The grid leak should be reduced in value, and a battery or bias supply should be connected between the upper end of the grid-metering jack, J_1, and ground, if the unit is to be operated at maximum rated power input. Because of the isolation afforded by the placement of the grid-circuit components below the chassis and the plate circuit above, the amplifier has no tendency toward self-oscillation. This may not be true if the parts layout differs from that shown. Screen voltage for the 832A is maintained at the proper level for both low and maximum power input to the amplifier, by the dropping resistor, R_4. A jack, J_2, provides for metering of the cathode current of the tube. If the transmitter is to be keyed for c.w. operation, the key can be plugged into this cathode jack.

An antenna tuning assembly, intended for operation with coaxial feed systems normally used in mobile installations, is included as part of the transmitter circuit. The degree of antenna loading can be regulated by adjustment of the coupling between the plate coil, L_4, and the output link, L_5. A short length of coaxial cable completes the circuit between the antenna tuning components and the output jack.

The photographs show how a metal box measuring 3 × 4 × 5 inches serves as the chassis for the transmitter. The bottom plate of the box is removed and used as a panel, and is held in place by the screws and nuts that hold the top cover and the box together. In Figs. 13-32 and 13-35 the condenser, C_2, and the antenna jack may be seen mounted on the panel. Metal pillars, ¼ inch long, are used to space the condenser away from the panel. A National FWB polystyrene insulator is used as a mounting support for the antenna coil, L_5, and the insulator is mounted on ¾-inch metal posts. The antenna tuning condenser, C_3, is supported by its own mounting tabs, and is connected between one end of the pick-up link and ground.

The rear and bottom views of the transmitter show how the rest of the components are laid out on the top plate of the metal box.

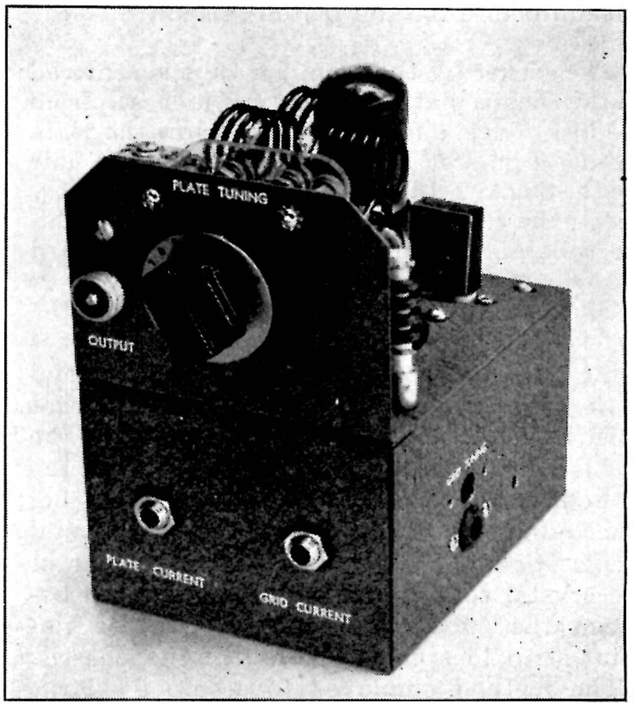


Fig. 13-32 — A compact mobile transmitter or exciter unit for 50 Mc.

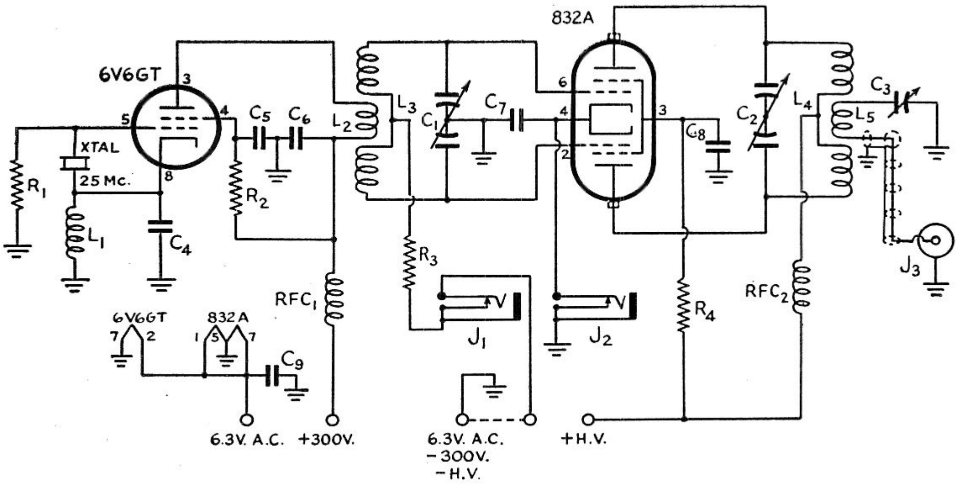


Fig. 13-33 — Circuit diagram of the mobile transmitter for 6 meters.

C_1 — 15-μμfd. per section (Bud LC-1660).
C_2 — "Butterfly" condenser, 15 μμfd. total (Cardwell ER-15-BF/S).
C_3 — 3–30-μμfd. mica trimmer.
C_4 — 100-μμfd. midget mica.
C_5, C_6 — 0.0047-μfd. mica.
C_7, C_9 — 470-μμfd. midget mica.
C_8 — 0.001-μfd. mica.
R_1 — 0.12 meghom, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 22,000 ohms, ½ watt.
R_4 — 25,000 ohms, 10 watts.
L_1 — 3 turns No. 18 enameled wire, close-wound, ½-inch diam.
L_2 — 5 turns.
L_3 — 9 turns, 4½ each side of center, with a ⅞-inch space between sections.
L_4 — 10 turns, 5 each side of center, with a ¾-inch space between sections.
L_5 — 3 turns. L_2 through L_5 have an inside diameter of ¾ inch; No. 12 enameled wire, turns spaced wire diameter.
J_1, J_2 — Midget closed-circuit jack.
J_3 — Coaxial-cable connector.
RFC_1 — 10-μh. r.f. choke (Millen 34300).
RFC_2 — 2.5-mh. r.f. choke (Millen 34102).

This plate should be removed from the box while the construction and wiring are being carried on. All of the wiring, with the exception of the d.c. leads to the metering jacks and the input terminals, can be completed in convenient fashion before the top plate is attached to the metal box.

The socket for the amplifier tube is centered on the chassis plate at a point 2⅜ inches in from the front edge, and is mounted below the plate on metal pillars ⅝ inch long. A clearance hole for the 832A, 2¼ inches in diameter, is directly above the tube socket. Sockets for the oscillator tube and the crystal are mounted toward the rear of the chassis.

The oscillator coil, L_2, is mounted on the 6V6 socket; the spare pin, No. 6, of the socket being used as the tie-point for the cold end of the plate coil and the other connections that must be made at this part of the circuit. The oscillator cathode coil is mounted between the cathode pin of the 6V6 and a soldering lug placed under the mounting screw of the crystal socket. C_5 and C_6 can be seen to the rear of the crystal socket, and RFC_1 is mounted between the tube socket and a bakelite tie-point strip located at the left of the chassis.

The method employed to assure good r.f. grounding of the amplifier components is visible in Fig. 13-35. Soldering lugs are placed beneath the mounting nuts of the 832A socket, and these lugs are joined together with a No. 12 lead which, in turn, is carried on to the common ground point for the oscillator circuit. The filament, cathode, and screen by-pass condensers for the amplifier are all returned to the common ground. These three condensers, C_7, C_8 and C_9, all rest on the 832A tube socket.

The amplifier grid coil, L_3, is self-supporting, with the ends connected to the grid pins of the 832A socket. The tuning condenser, C_1, is actually supported on metal pillars at the right-hand side of the metal box, but the condenser can be wired in place if the operation is carried out in the proper order. First, mount the chassis plate on the box and locate the proper place for the condenser. Next, determine the length of the leads to connect the condenser to the tube socket, and then remove the chassis from the case. The condenser may now be wired into the circuit, and the

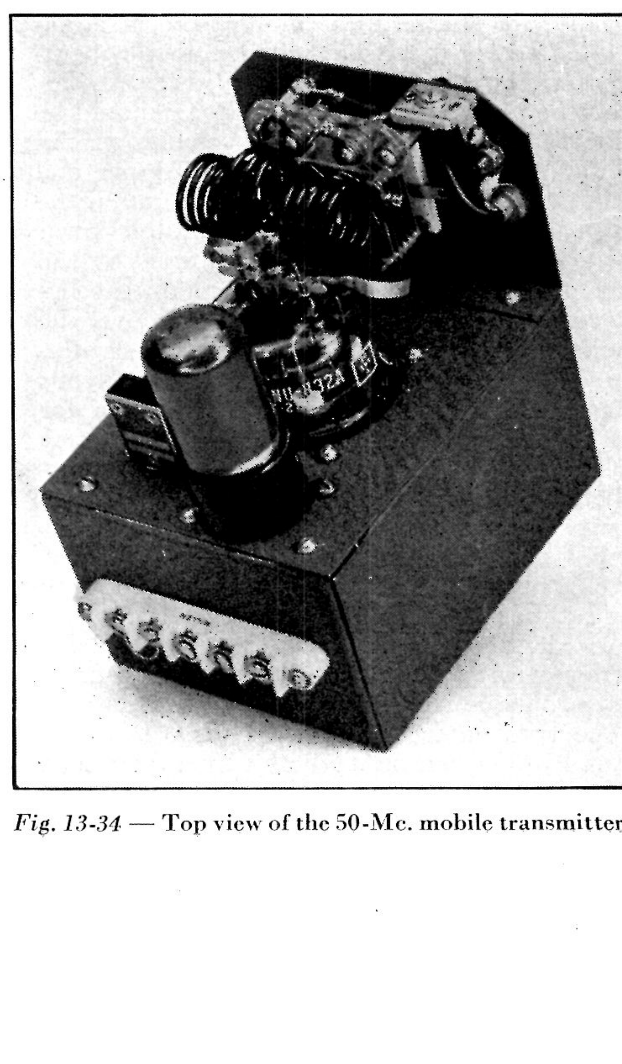


Fig. 13-34 — Top view of the 50-Mc. mobile transmitter.

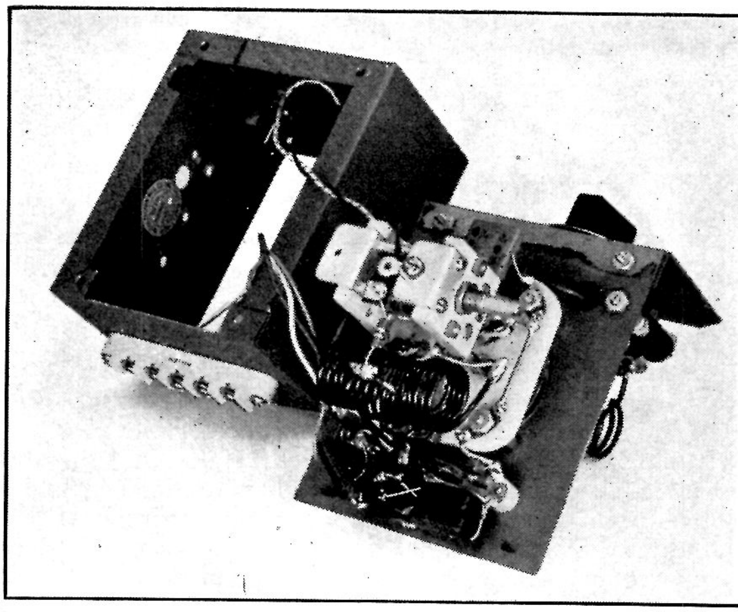


Fig. 13-35 — Bottom view of the mobile transmitter, showing all major components attached to the top plate.

rigid mounting of C_1, by means of metal posts 1¼ inches long, can be done during the final assembly of the unit.

The grid leak, R_3, is connected between the center-tap of L_3 and a tie-point strip that is mounted on the condenser frame. RFC_2 is mounted toward the front of the chassis, and the grommet-fitted hole to the left of the choke (Fig. 13-35) carries the lead between the plate-voltage terminal and the choke.

The metering jacks and the power terminal strip may now be mounted on the front and rear walls of the metal box. Holes to permit mounting and adjustment of C_1 should also be drilled at this time. Portions of top flanges of the metal case must be cut away in order to provide clearance for the oscillator section and the mounting nut for the amplifier plate choke. After the case, chassis and panel have been fastened together, the wiring of the amplifier plate circuit may be completed.

Test Procedure

A power supply capable of delivering 300 volts at 100 ma. and 6.3 volts at 2 amp. may be used for testing the transmitter. The high voltage should not be applied to the 832A plates until the oscillator has been checked. For initial tests the input voltage can be reduced to approximately 150 volts while the circuits are checked for resonance and proper operation. Squeezing or spreading the turns of the coils should bring the circuits into resonance, as indicated by maximum grid current to the 832A. If the oscillator is performing correctly, the 6V6 plate current will be between 1 and 4 ma., rising to the latter value when the circuits are tuned to the second harmonic of the crystal by adjustment of C_1. The grid current should fall to zero, and the plate current of the oscillator tube should rise considerably when the crystal is removed from the socket.

The amplifier plate and screen voltage can be applied at this point. The unloaded cathode current of the amplifier should be about 15 ma., rising to a maximum of 75 or 80 ma. under load, which may be a 15-watt light bulb connected to the antenna jack. C_3 should be adjusted along with the coupling between L_4 and L_5 until maximum output is obtained. The correct degree of loading has been obtained when the plate current at resonance is 10 to 15 ma. below the off-resonance value. The plate tuning condenser, C_2, should be reset each time that a loading adjustment is made.

A final check of voltages and currents should show the following: oscillator and amplifier plate, 300 volts; oscillator screen, 200 volts; amplifier screen, 150 volts; amplifier bias (read at the grid-coil center-tap with a high-resistance voltmeter), 65 volts, negative.

The oscillator plate current should be 28 to 30 ma. and amplifier grid current should be about 3 ma. Under load, the amplifier cathode current should be approximately 60 ma. with 8 or 10 ma. of this amount being drawn by the 832A screen.

Modulation can be supplied by the audio system used in the 2-meter rig shown in Fig. 13-28, or a similar unit may be added, if only 50-Mc. operation is desired.

Transceivers

The transceiver is a combination transmitter-receiver in which, by suitable switching of d.c. and audio circuits, the same tube and r.f. circuit functions either as a modulated transmitting oscillator or as a superregenerative detector. This makes for extreme compactness and light weight, making the transceiver popular for hand-carried portable equipment. It is a compromise with respect to other features, however. The transceiver can be a source of serious interference, and its efficiency is not equal to that of other types of gear wherein separate tubes and circuits are used for transmission and reception.

As a matter of good amateur practice the use of transceivers should be confined to very low-power operation — as in "walkie-talkie" or "handie-talkie" equipment — in the 144-Mc. band, and to experimental low-power operation in the higher-frequency bands. The use of transceivers should be avoided entirely for regular operation on the 144-Mc. band.

V.H.F. Antennas

While the basic principles of antenna operation are essentially the same for all frequencies, certain factors peculiar to v.h.f. work call for changes in antenna technique for the frequencies above 50 megacycles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in nearly every location, and experimentation with various types of arrays is an important part of the program of most progressive amateurs. The importance of high-gain antennas in v.h.f. work cannot be overemphasized. A good antenna system is often the sole difference between routine operation and outstanding success in this field. By no other means can so large a return be obtained from a small investment as results from the erection of a good directional array.

Design Factors

Beginning with the 50-Mc. band, the frequency range over which antenna arrays should operate effectively is often wider in percentage than that required of lower-frequency systems; thus greater attention must be paid to designing arrays for maximum frequency response, possibly to the extent of sacrificing other factors such as high front-to-back ratio.

As the frequency of operation is increased, losses in the transmission line rise sharply; hence it becomes more important that the line be matched to the antenna system correctly. Because any v.h.f. transmission line is long, in terms of wavelength, it is often more effective to use a high-gain array at relatively low height, rather than to employ a low-gain system at great height above ground, particularly if the antenna location is not completely shielded by heavy foliage, buildings, or other obstructions in the *immediate* vicinity.

This concept is in direct contrast to early notions of what was most desirable in a v.h.f. antenna system. An appreciable clearance above surrounding terrain is desirable, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained by many v.h.f. workers, especially on 50 and 144 Mc., with antennas not more than 25 to 40 feet above ground. DX can be worked on 50 Mc. with arrays as low as a half-wave above the ground level.

Polarization

Practically all the early work on frequencies above 30 Mc. was done with vertical antennas, probably because of the somewhat stronger field in the immediate vicinity of a vertical system. When v.h.f. work was confined to almost pure line-of-sight distances, the vertical dipole produced a stronger signal at the edge of the working range than did the same antenna turned over to a horizontal position. With the advent of high-gain antennas and extended operating ranges, horizontal systems began to assume importance in v.h.f. work, especially in parts of the country where a considerable degree of activity had not already been established with verticals.

Numerous tests have shown that there is very little difference in the effective working range with either polarization, *if* the most effective element arrangements are used and the same polarization is employed at both ends of the path. Vertical polarization still has its adherents among 50-Mc. enthusiasts and much fine work has been done with vertical antennas, but an effective horizontal array is somewhat easier to build and rotate. Simple 2-, 3- or 4-element horizontal arrays have proven extremely effective in 50-Mc. work, and the postwar era has seen an increase in the use of such arrays which has amounted to standardization on horizontal polarization.

The picture is somewhat different when one goes to 144 Mc. and higher. At these frequencies, the most effective vertical systems (those having two or more half-wave elements, vertically stacked) are more easily erected than on 50 Mc. Important, in considering the polarization question, is the existence of countless 144-Mc. mobile stations, whose antenna systems must, of necessity, be vertical. While horizontal polarization will undoubtedly find increased favor at 144 Mc. and higher, particularly for point-to-point work in rural areas, it is probable that vertical polarization will continue in use for some years to come, particularly in areas where activity has been established with vertical systems. Under certain conditions, notably a station directly in the shadow of a hill, there may be a considerable degree of polarization shift, but ordinarily it may be assumed that best results in 144-Mc. work will be obtained by matching the polarization of the stations one desires to contact.

Impedance Matching

Because line losses tend to be much higher in v.h.f. antenna systems, it becomes increasingly important that feedlines be made as nearly "flat" as possible. Transmission lines commonly used in v.h.f. work include the open-wire line of 500 to 600 ohms impedance, usually spaced about two inches; the polyethylene-insulated flexible lines, available in impedances of 300, 150, 100, and 72 ohms; and coaxial lines of 50 to 90 ohms impedance. These may be matched to dipole or multielement antennas by any of several arrangements detailed below.

The "J"

Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave matching section, as shown at A, Fig. 14-1. The spacing between the two sides of the matching section should be two inches or less, and the point of attachment of the feedline will depend on the impedance of the line used. The feeder should be slid along the matching section until the point is found that gives the best operation. The bottom of the matching section may be grounded for lightning protection. A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 14-1. The "J" is also useful in mobile applications.

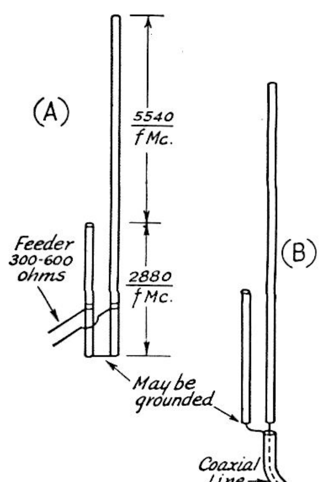


Fig. 14-1 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

The Delta or "Y"-Match

Probably the simplest arrangement for feeding a dipole or parasitic array is the familiar delta, or "Y"-match, in which the feeder system is fanned out and attached to the radiator at a point where the impedance along the element is the same as that of the line used. Information on figuring the dimensions of the delta may be found in Chapter Ten. Chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multielement array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The "Q" Section

An effective arrangement for matching an open-wire line to a dipole, or to the driven element in a 2- or 3-element array having wide (0.25 wavelength or greater) spacing, is the "Q" section (Chapter Ten). This consists of a quarter-wave line, usually of ½-inch or larger tubing, the spacing of which is determined by the impedance at the center of the array. The parallel-pipe "Q" section is not practical for matching multielement arrays to lines of lower impedances than about 600 ohms, nor can it be used effectively with close-spaced parasitic arrays. The impedance of the "Q" section required in these cases is lower than can be obtained with parallel sections of tubing of practical dimensions. A quarter-wave section of coaxial or other low-impedance line is a commonly-used means of matching a line of 300 to 600 ohms impedance to the low center impedance of a 3- or 4-element array. The length of such a line will depend on the velocity of propagation (propagation factor) of the line used. The propagation factors of all the commonly-used lines are given in table form in Chapter Ten.

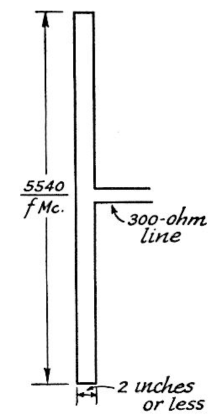


Fig. 14-2 — Details of the folded dipole.

In some installations it may be more convenient to use a line of greater length than a single quarter wave for matching purposes, in which case any odd multiple of a quarter wavelength may be used. The exact length required may be determined experimentally by shorting one end of the line and coupling it to a source of r.f., and trimming the line length until maximum loading is obtained at the center frequency of the operating range.

The "T"-Match

The principal disadvantages of the delta system can be overcome through the use of the arrangement shown in Figs. 14-5 and 14-13, commonly called the "T"-match. It has the advantage of providing a means of adjustment (by sliding the clips along the parallel conductors), yet the radiation from the matching arrangement is lower than with the delta, and its rigid construction is more suitable for rotatable arrays. It may be used with coaxial lines of any impedance, or with the various other forms of transmission lines up to 300 ohms. The position of the clips should, of course, be adjusted for maximum loading and minimum standing-wave ratio, the latter being most important as an indication of proper setting. The "T" system is particularly well suited for use in all-metal "plumbing" arrays.

The Folded Dipole

Probably the most effective means of matching various lines to the wide range of antenna impedances encountered in v.h.f. antenna work is the folded dipole, shown in its simplest form in Fig. 14-2. When all portions of the dipole are of the same conductor size, the impedance at the feed-point is equal to the *square* of the number of elements in the folded dipole times the normal center impedance which would be present if only a conventional split half-wave radiator were used. Thus, the simple folded dipole of Fig. 14-2 has a feed-point impedance of 4 × 72, or approximately 288 ohms. It may be fed with the popular 300-ohm line without appreciable mismatch. If a three-wire dipole were used, the step-up in impedance would be *nine* times. Note that this step-up occurs *only* if all portions of the folded dipole are the same conductor size.

The impedance at the feed-point of a folded dipole may also be raised by making the fed portion of the dipole smaller than the parallel section. Thus, in the 50-Mc. array shown in Fig. 14-4 the relatively low center impedance of a 4-element array is raised to a point where it may be fed directly with 300-ohm line by making the fed portion of the dipole of ¼-inch tubing, and the parallel section of 1-inch. A 3-element array of similar dimensions could be matched by substituting ¾-inch tubing in the unbroken section. Conductor ratios and spacings may be obtained from Fig. 10-80, Chapter Ten.

Antenna Systems for 50 Mc.

Since the same basic principles apply to all antennas regardless of frequency, little discussion is given here of the various simple dipoles that may be used when nondirectional systems are desired. Details of such antennas may be found in Chapter Ten, and the only modification necessary for adaptation to use on 50 Mc. or higher is the reduction in length necessary for increased conductor diameter at these frequencies.

A Simple 2-Element Array

A simple but effective array which requires no matching arrangement is shown in Fig. 14-3. Its design takes into account the drop in center impedance of a half-wave radiator when a parasitic element is placed a quarter wavelength away. A director element is shown, as the drop in impedance using a slightly-shortened parasitic element is just about right to provide a good match to a 50-ohm coaxial line. The element lengths are not extremely critical in such a simple system, and the figures presented may be used with satisfactory results.

A 4-Element Array

The importance of broad frequency response in any antenna designed for v.h.f. work cannot be overlooked. The disadvantage of all parasitic systems is that they tend to tune quite sharply, and thus are often effective over only a small portion of a given band. One way in which the response of a system can be broadened out is to increase the spacing between the parasitic elements to somewhat more than the 0.1 or 0.15 wavelength normally considered to provide optimum front-to-back ratio. Some broadening may also be obtained by making the directors slightly shorter and the reflector slightly longer than the optimum value. The folded dipole is useful as the radiator in such an array, as its over-all frequency response is somewhat broader than other types of driven elements.

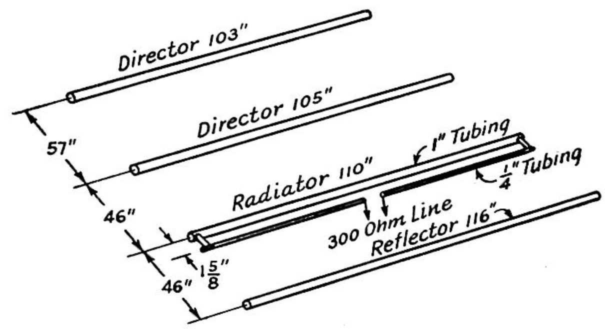


Fig. 14-4 — **Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.**

A 4-element array for 50 Mc. having an effective operating range of about 2 Mc. is shown in Figs. 14-4 and 14-5. It employs a folded dipole having nonuniform conductor size. Reflector and first director are spaced 0.2 wavelength from the driven element, while the forward director is spaced 0.25 wavelength. The spacing and element lengths given were derived experimentally, and are those that give optimum forward gain at the expense of some front-to-back ratio. As the latter quality is not of great value in 50-Mc. work, it can be neglected entirely in the tuning procedure for such an array.

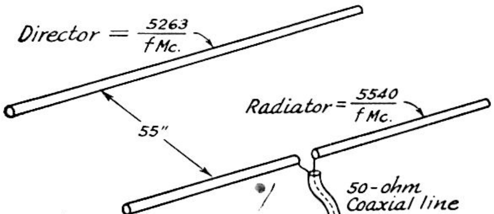


Fig. 14-3 — **A simple 2-element array for 50 Mc. No matching devices are needed with this arrangement.**

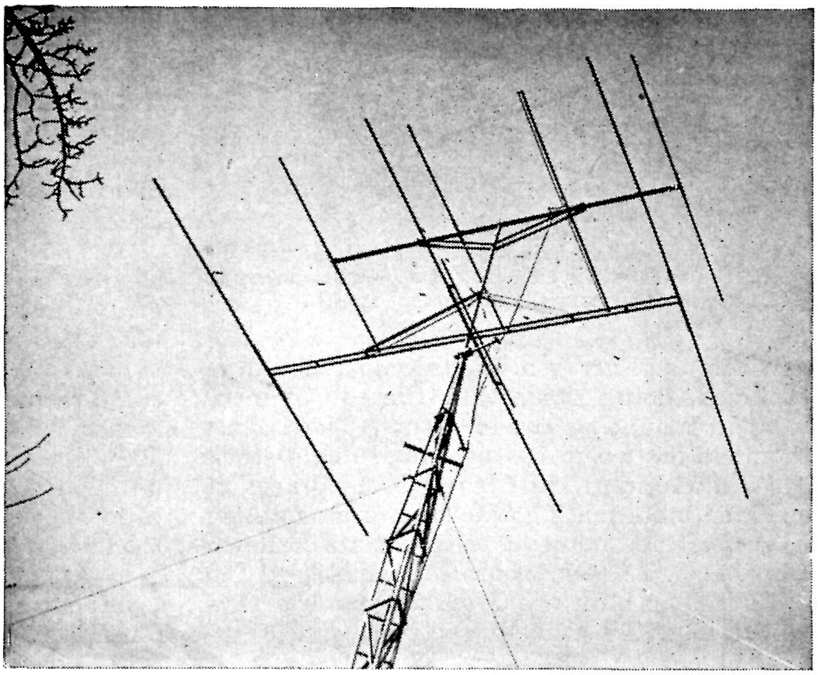


Fig. 14-5 — An example of stacking two arrays for different bands on the same support. The top section is a 4-element array for 50 Mc.; the lower a 3-element system for 28 Mc. All-metal construction is employed.

The dimensions given are for peak performance at 50.5 Mc. For other frequencies, the length of the folded dipole in inches should be figured according to the formula

$$L = \frac{5540}{f_{Mc.}}$$

The reflector will be 5 per cent longer, the first director 5 per cent shorter, and the second director 6 per cent shorter than the driven element. A broadening of the response may be obtained, at a slight sacrifice in forward gain, by adding to the reflector length and subtracting from the director lengths. For those interested in experimenting with element lengths, slotted extensions may be inserted in the ends of the various elements, other than the dipole, as shown in Fig. 14-7. A 3-element array may be built, using the same general dimensions, except that the unbroken section of the folded dipole, in this case, should have a ¾-inch diameter element in place of the 1-inch tubing used in the 4-element array.

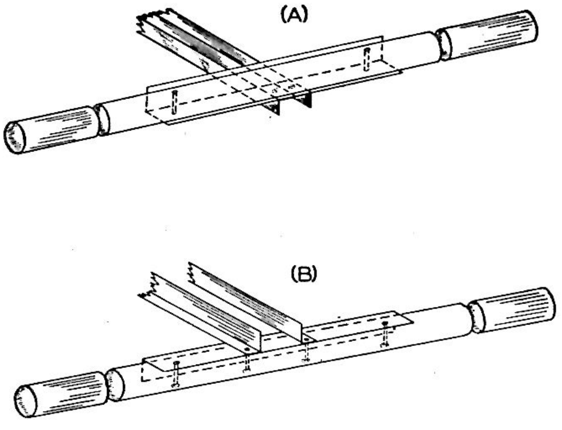


Fig. 14-6 — Detail sketches showing method of mounting elements in the dual array for 28 and 50 Mc. A — The 50-Mc. boom is comprised of two pieces of angle stock mounted edge-to-edge to form a channel. The elements are fastened to the boom by means of a cradle, also of angle stock. B — In the 10-meter array, the two portions of the boom are separated, and mount on either side of the vertical support. The elements and their supporting crossarms are attached to the lower surface of the boom.

Stacked Antennas

Excellent results in long-distance work are being obtained by 50-Mc. stations using various more-complex directional arrays than the ones described above. The most important factor in such work is the attainment of the lowest-possible radiation angle, and this purpose is well served by stacking of elements, in either vertical or horizontal systems. The use of two parasitic arrays, one a half-wavelength above the other, fed in phase, provides a gain of 3 db. or more over that of a single array. The system shown (for 144 Mc.) in Fig. 14-8 is excellent for either vertical or horizontal polarization, as is the "H" array, using four half-wave elements, with or without parasitic elements.

Stacked Arrays for Two Bands

As many 50-Mc. enthusiasts also operate on 28 Mc., it is often desirable to stack arrays for the two bands on a common tower and rotating device. Such a dual array, combining a 4-element system for 50 Mc. with a 3-element array for 28 Mc., is shown in Fig. 14-5.

If space limitations make it absolutely necessary, the two arrays may be mounted with but a few inches separating them, but experience has shown that some effectiveness is sacrificed, particularly in the array for the higher frequency. A separation of at least three feet is recommended as the minimum for avoiding harmful interaction. In the example shown the separation is six feet, at which distance each array performs equally as well as it would if mounted alone.

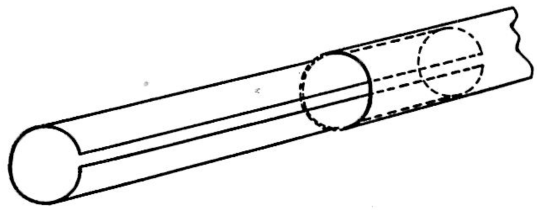


Fig. 14-7 — Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.

In this dual array all-metal construction is employed, doing away with the use of insulators in mounting the elements. The booms are made of two pieces of 1-inch angle stock (24ST aluminum), with supporting braces of the same material. The method of assembling the booms and mounting the elements is shown in Fig. 14-6. The booms are 150 inches and 160 inches in length for the 6- and 10-meter arrays respectively. To prevent swaying of the 10-meter elements, they are braced with guy wires, which are broken up with small insulators. These sway-brace wires are attached to the elements at approximately the midpoint between the boom and the outer end, and are brought up to the vertical support at the point of attachment of the horizontal fore-and-aft braces.

The 50-Mc. portion of the array is similar in element length and spacing to the 4-element array already described. The element spacing for the 10-meter array is 0.2 wavelength, or 80 inches. The driven element is 198 inches long, the director 188 inches, and the reflector 208 inches. It is fed by means of a "T"-match and a 300-ohm line. These dimensions give quite uniform performance and low standing-wave ratio over the range from 28 to 29.1 Mc., and the array will take power and show appreciable gain over a half-wave from 27.2 to 29.7 Mc. Complete details of this dual array will be found in *QST* for July, 1947.

Antenna Systems for 144 Mc.

The antennas already described may, of course, be used for 144 as well as 50 Mc., but since they are designed for maximum effectiveness in a horizontal position, other designs may be used more effectively where vertical polarization is desired. With either polarization, the stacking of elements vertically lowers the radiation angle and extends the operating range. The smaller size of 144-Mc. arrays makes such stacking of elements a much simpler procedure than on 50 Mc. Another advantage of the array employing elements fed in phase is that its frequency response is likely to be less critical than an array that achieves the same gain with but one driven element and parasitic directors and reflectors. Thus the element lengths, even in such complex systems as the 16-element array shown in Fig. 14-9, are not at all critical.

Plane reflectors are usable at 144 Mc., their size at this frequency being within reason. An interesting possibility in connection with this type of reflector is its use with two different sets of driven elements, one on each side of the reflecting screen. A set of elements arranged for vertical polarization may be used on one side, and a set of horizontally-polarized elements on the other, or the plane reflector may be made to serve on two different bands by a similar arrangement of elements for two frequencies, on opposite sides of the reflector. The screen need not be a solid sheet of metal, or even a close-mesh screen. A set of wires or rods arranged in back of the driven elements will work almost equally well. The dimensions of the reflector are not critical. For maximum effectiveness, the plane reflector should extend at least one-quarter wavelength beyond the area occupied by the elements, but reflecting curtains no larger than the space occupied by the reflectors shown in Fig. 14-9 have been used with good results.

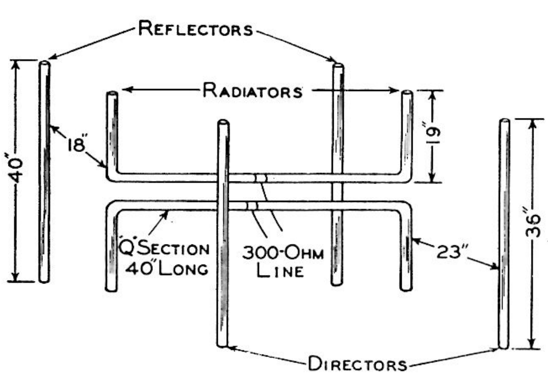


Fig. 14-8 — A double-"Q" array for 144 Mc. The horizontal portion of the half-wave "H" acts as a "Q" section, matching the antenna impedance to the 300-ohm line attached at the center of the array. This array works well in either vertical or horizontal positions.

A 6-Element Array

In designing directional arrays having more than one driven element it is advisable to arrange for feeding the array at a central point. A simple 6-element array of high performance, incorporating this principle, is shown in Fig. 14-8. All the elements may be made of soft-copper tubing, ¼ inch in diameter. The driven elements are comprised of two pieces which are bent into two "U"-shaped sections and arranged in the form of a half-wave "H." The horizontal portion of the "H" is then a double quarter-wave "Q" section, matching the impedance of the two radiators to that of the feedline. With the wide spacing used, the position of the parasitic elements is not particularly critical, except as it affects the impedance of the system, and the spacing of the elements may be varied to provide the best

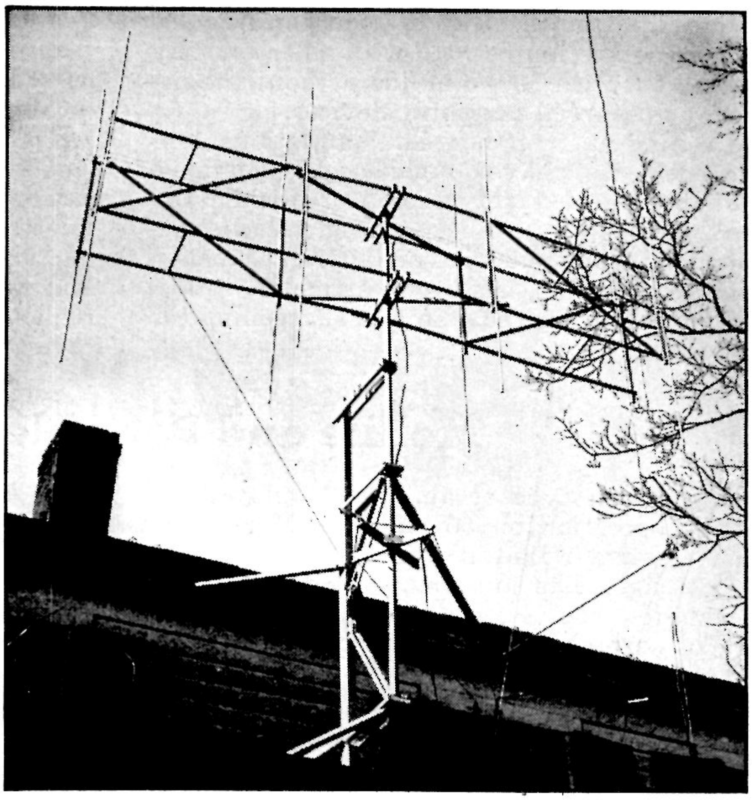


Fig. 14-9 — A 16-element array for 144 Mc., showing supporting structure and "rotating mechanism." Sash cord wrapped three times around the crisscross pulley permits 360-degree rotation.

match. The spacing of the horizontal section may be varied for the same purpose. With the dimensions given, a spacing of one inch between centers is about right for feeding with a 300-ohm line. The radiation pattern of this array is similar in both horizontal and vertical planes; thus it will work with equal effectiveness in either position, provided the polarization is the same as that of the stations to be contacted.

A 16-Element Array

By using a curtain of eight half-wave elements, arranged as shown in Figs. 14-9 and 14-10, backed up by eight reflectors, a degree of performance can be obtained which is truly outstanding. A gain of as much as 15 db. can be realized with such an arrangement, effecting an improvement in operating range which could never be obtained by any other means. Such an array is neither difficult nor expensive to construct, and its performance will more than repay the builder for the trouble involved in its construction.

The cumbersome nature of the structure required to support such an array would make its construction out of the question for a lower frequency, but for 144 Mc. the outside dimensions are only 1½ × 7 × 10 feet, and the supporting frame can be made quite light.

The center pole (a 1½-inch rug pole 10 feet long) turns in three bearings which are mounted on braced arms extending out about two feet from a "two by three," which is braced in a vertical position. An improvised pulley made of two pieces of 1 × 2-inch "furring" notched in the ends and fastened crisscross fashion near the bottom of the center pole serves as a "rotating mechanism." Sash cord wrapped three times around this "pulley" and run over to the window on small pulleys allows the beam to be rotated more than 360 degrees before reversal is required. To keep the array from twisting in high winds light sash cords are attached near each end of the supporting structure. These cords are brought through the window near the rotating ropes and are pulled up tight and fastened when the antenna is not in use.

The elements are of 7⁄16-inch soft-aluminum tubing for light weight. To stiffen the structure, and to help maintain alignment, inserts were turned down from ½-inch polystyrene rod to fit tightly into the elements at the point where the crossover or phasing wires are connected. Similar inserts are used for the reflector elements also. The interconnecting phasing sections are of No. 16 wire, spaced about 1½ inches. The feedline, connected at the center of the system, is Amphenol 21-056 Twin-Lead, 300 ohms impedance. The impedance at the center of the array is about right for direct connection of the 300-ohm line, without the necessity for a matching section of any kind. It is probably somewhat lower than 300 ohms, actually, and if a perfect match is desired, a "Q" section may be used. The performance is not greatly affected by such a change, however, as the standing-wave ratio is relatively low with the connection as shown.

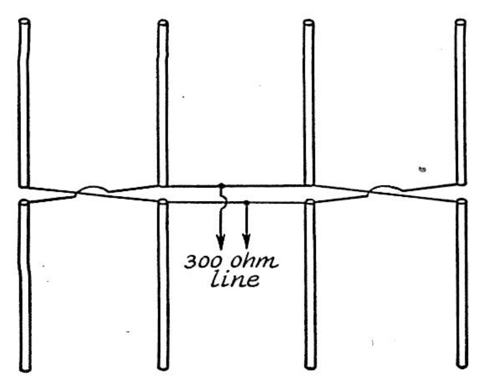


Fig. 14-10 — Schematic of the radiating portion of the 16-element 144-Mc. array. Reflectors are omitted for clarity. Radiators are 38 inches long, reflectors 40.5 inches. Crossover or phasing sections are also 40.5 inches long. Reflectors are mounted 17 inches in back of each radiator.

The center section of the array may be used without the outside 8 elements, if space is lim-