CHAPTER SEVEN

Transmitter Theory

The general function of a transmitter is to generate a signal of a desired frequency and to modulate this signal in accordance with the intelligence to be transmitted. The radio frequency energy from the transmitter is most commonly carried by a *transmission line* to a radiating system or *antenna* from whence the intelligence-carrying energy is radiated into space. Transmission lines and antennas will be treated in the chapter devoted to *Antennas;* the theory of operation of the various divisions of the transmitter proper will be discussed in the following pages.

The usual transmitter will contain the following general divisions: an oscillator, either crystal or self-controlled; one or more frequency multiplying stages; one or more radio-frequency amplifying stages and a system for either keying or modulating by voice the output of the final amplifier stage. However, a transmitter need not necessarily have all the stages mentioned above, and, in fact, may be merely an oscillator whose output is controlled by a telegraph key.

Oscillators

As was mentioned earlier, in the chapter devoted to the theory of vacuum tubes, the amplifying properties of a tube having three or more elements give it the ability to generate an alternating current of a frequency determined by the components associated with it. A vacuum tube operated in such a circuit is called an oscillator, and its function is essentially to convert a source of direct current into radio frequency alternating current of a predetermined frequency. Oscillators for controlling the frequency of conventional radio transmitters can be divided into two general classifications: self-controlled and crystal-controlled.

There are a great many types of self-controlled oscillators, each of which is best suited to a particular application. They again can further be subdivided into the classifications of: negative-grid oscillators, electron-orbit oscillators, and negative-resistance oscillators.

Negative-Grid Oscillators. A negative-grid oscillator is essentially a vacuum-tube amplifier with a sufficient portion of the output energy coupled back into the input circuit to sustain oscillation. They are called negative-grid oscillators because, in contrast to certain other oscillator circuits, the grid is biased a considerable amount negative with respect to the cathode. It is this classification of oscillator which finds most common application in low- and medium-frequency transmitter control circuits. The various common types of negative-grid oscillators are diagrammed in figure 1.

The Hartley. Figure 1 (A) illustrates the oscillator circuit which finds the most general application at the present time; this circuit is commonly called the Hartley. The operation of this oscillator will be described as an index to the operation of all negative-grid oscillators; the only real difference between the various circuits is the manner in which energy for excitation is coupled from the plate to the grid circuit.

Operation of the Hartley Oscillator. When the plate voltage is applied to the plus and minus terminals of the Hartley oscillator shown at (A), the sudden flow of plate current accompanying the application of plate voltage will cause an electro-magnetic field to be set up in the vicinity of the coil. The building-up of this field will cause an instantaneous potential drop to appear from turn-to-turn along the coil. Due to the inductive coupling between the portion of the coil in which the plate current is flowing and the grid portion, a potential will be induced in the grid portion. Since the cathode tap is between the grid and plate ends of the coil, the induced grid voltage acts in such a manner as to increase further the plate current to the tube. This action will continue for a small period of time determined by the inductance

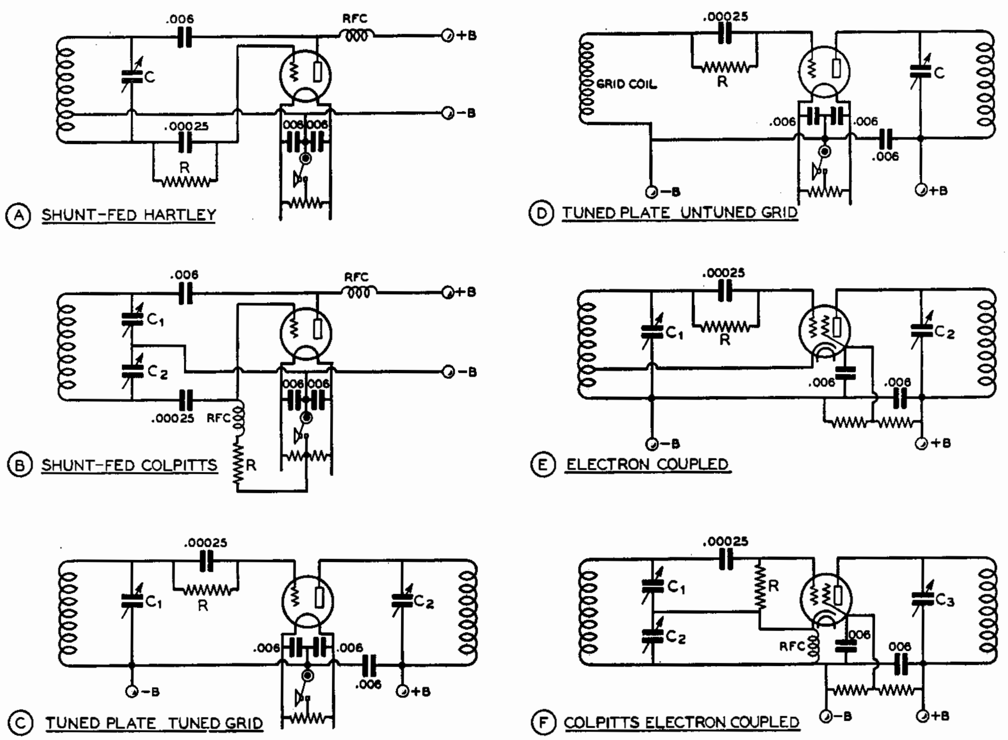


Figure 1.
COMMON TYPES OF SELF-EXCITED OSCILLATORS.

and capacity of the tuned circuit, until the "flywheel" effect of the tuned circuit causes this action to come to a maximum and then to reverse itself. The plate current then decreases, the magnetic field around the coil also decreasing, until a minimum is reached when the action starts again in the original direction and at a greater amplitude than before. The amplitude of these oscillations, the frequency of which is determined by the coil-condenser circuit, will increase in a very small period of time to a limit determined by the plate voltage or the cathode emission of the oscillator tube.

The Colpitts. Figure 1 (B) shows a version of the colpitts oscillator. It can be seen that this is essentially the same circuit as the Hartley except that a pair of capacitances in series are employed to determine the cathode tap, instead of actually using a tap on the tank coil. Also, the net capacity of these two condensers comprises the tank capacity of the tuned circuit.

The T.P.T.G. The tuned-plate tuned-grid or t.p.t.g. oscillator illustrated at (C) has a tank circuit in both the plate and grid circuits. The feedback of energy from the plate to the grid circuits is accomplished by the plate-to-grid interelectrode capacity within the tube.

For best operation of the Hartley and Colpitts oscillators the voltage from grid to cathode, determined by the tap on the coil or the setting of the two condensers, should be from one-third to one-fifth that appearing between plate and cathode. In the t.p.t.g. oscillator the grid circuit should be tuned to a frequency slightly lower than that of the plate circuit for best operation. The frequency of oscillation is determined primarily by the constants of the plate circuit, and therefore a broadly resonant or aperiodic coil may be substituted for the grid tank to form the *T.N.T.* oscillator shown at (D).

Electron-Coupled Oscillators. In any of the three oscillator circuits just described it is possible to take energy from the oscillator circuit by coupling an external load to the tank circuit. Since the tank circuit determines the frequency of oscillation of the tube, any variations in the conditions of the external circuit will be coupled back into the frequency determining portion of the oscillator. These variations will result in frequency instability.

Two oscillators in which the frequency determining portion of the oscillator is coupled to the load circuit only by an electron stream are illustrated in (E) and (F) of figure 1. When it is considered that the screen of the tube acts as the plate to the oscillator circuit, the plate acting as a coupler to the load, then the similarity between the cathode-grid-screen circuit of these oscillators and the cathode-grid-plate circuits of the corresponding prototype.

The advantage of the electron-coupled oscillator over conventional types is in the greater stability with respect to load and voltage variations that can be obtained. Load variations have very little effect on the frequency of operation of the e.c.o., since the only coupling between the oscillating circuit and the load is through the electron stream flowing through the other elements to the plate. The plate is electrostatically shielded from the oscillating portion by the screen, which is at ground potential with respect to r.f.

The stability of the electron-coupled type of oscillator with respect to variations in supply voltages comes from an entirely different source. It is a peculiarity of such an oscillator that the frequency will shift in one direction with an increase in screen voltage while an increase in plate voltage will cause it to shift in the other direction. By a proper proportioning of the resistors that comprise the voltage divider supplying screen voltage, it is possible to make the frequency of the oscillator substantially independent of supply voltage variations; the tendency of an increase in screen voltage to make the frequency shift in one direction is counterbalanced by the effect of the increase in plate voltage to make the frequency shift in the other direction.

V. F. O. Transmitter Controls. During the last year or two there has been an increasing tendency to break away from the standard crystal oscillator as the only means of controlling the frequency of a transmitter because of the necessarily limited flexibility of such an oscillator. The new tendency has been toward the use of highly stabilized *variable-frequency oscillators* as transmitter controls in amateur equipment. These oscillators are nothing more than certain types of self-excited oscillators in which adequate precautions have been taken to insure that they shall be as stable as possible with respect to load and supply voltage variations.

Due to the better inherent stability of the electron-coupled type of oscillator, a number of the recent designs for *v.f.o.'s* (as the variable-frequency oscillators for transmitter frequency control are called) have used this type of oscillator. However, one disadvantage of the electron-coupled oscillator is that the cathode and heater are not at the same r.f. potential. This gives rise to difficulties due to heater-cathode leakage, heater-cathode capacity variation with changes in temperature, and coupling of stray r.f. energy from the heater into the cathode circuit.

As a consequence of this disadvantage of the electron-coupled oscillator, another group of the recent designs for v.f.o.'s (variable-frequency oscillators) have used grounded-cathode oscillator circuits of the modified Hartley type. A v.f.o. of this design is shown in the chapter *Exciters and Low-Powered Transmitters.* Since the cathode of an oscillator of this type is at ground potential, it is impossible for r.f. energy from an external source to be coupled into the oscillating circuit from the heater circuit. However, the use of any type of oscillator as a transmitter control means that it must be carefully constructed, both from the electrical and from the mechanical standpoint.

Other Oscillator Circuits

Electron-Orbit Oscillators. Of the other oscillator circuits the negative-resistance and electron-orbit types are the most common of the self-excited class. Electron-orbit oscillators are used only for extremely high-frequency work (above 300 Mc.) and depend for their operation upon the fact that an electron takes a finite time to pass from one element to another inside a vacuum tube. The Gill-Morrell, Barkhausen-Kurtz, and Kozanowski oscillators are examples of this type and are described in the *Ultra-High Frequency Transmitters* chapter. Another special type of u.h.f. oscillator is the *magnetron,* which is also described in the u.h.f. chapter. This type employs a filament surrounded by a split plate to which are connected rods comprising a linear tank circuit. The tube is operated in a strong magnetic field; hence the name, magnetron.

Negative Resistance Oscillators. The other common type is the negative-resistance oscillator, which is used when unusually high frequency stability is desired, as in a frequency meter. The dynatron of a few years ago and the transitron of more recent fame are examples of oscillator circuits which make use of the negative resistance characteristic between different elements in a multi-grid tube. In the dynatron the negative resistance is a consequence of secondary emission of electrons from the plate of the tube. By a proper proportioning of the electrode voltages an increase of screen voltage will cause a decrease in screen current; from this comes the term, *negative resistance.* A similar effect in the *transitron* is produced by coupling the screen to the suppressor; the negative resistance in this case is due to interelectrode coupling rather than to secondary emission.

The Franklin Oscillator. Another circuit which makes use of two cascaded tubes to obtain the negative-resistance effect is the Franklin oscillator illustrated in figure 2. The tubes may be either a pair of triodes, tetrodes, or pentodes, a dual triode, or a combination of a triode and a multi-grid tube. The chief advantages of this oscillator circuit are that only very loose coupling between the two tubes and the tank circuit, LC, is required, and that the frequency determining tank only has two terminals and one side of the circuit is grounded. Condensers C_1 and C_2 need be only one or two $\mu\mu$fd. for satisfactory operation of the oscillator; this means that tube capacity and input resistance variations will have only an extremely small effect on the frequency of oscillation.

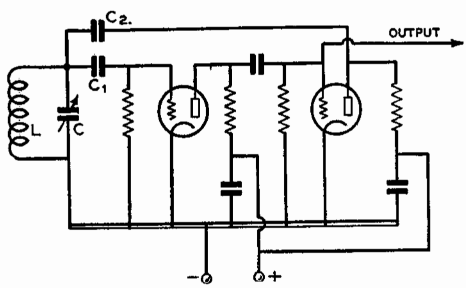


Figure 2.
THE FRANKLIN OSCILLATOR CIRCUIT.
In this oscillator a separate phase-inverter tube is used to feed a portion of the output back into the input circuit in the proper phase to sustain oscillation.

Crystal Controlled Oscillators

When it is desired to hold the frequency of a transmitter very closely to a certain definite value or to keep it within an assigned frequency tolerance, reliance is very commonly placed upon the *piezo-electric* properties of a plate cut from a natural crystal of quartz. Quartz crystals are very widely employed by amateurs and commercial services as frequency controls; hence some of the important characteristics of piezo-electric minerals will be mentioned before entering into a discussion of the oscillators that make use of these characteristics for frequency control.

Quartz Crystals. Quartz and tourmaline are naturally occurring crystals having a structure such that when plates are cut in certain definite relationships to the crystallographic axes, these plates will show the piezo-electric effect—the plates will be deformed in the influence of an electric field, and, conversely, when such a plate is compressed or deformed in any way a potential difference will appear upon its opposite sides.

When such a quartz plate is placed in a circuit with a vacuum-tube amplifier having the output circuit coupled back into the input, and a tuned circuit in series with the plate of the amplifier tube, the circuit will self-oscillate at a frequency primarily determined by the frequency of mechanical resonance of the quartz plate. The frequency of mechanical resonance or frequency of oscillation of a quartz plate is dependent upon its physical dimensions and upon a constant determined by the crystallographic (or optical) cut of the plate. The stability of the frequency of oscillation of a crystal controlled oscillator is dependent upon the Q of the quartz plate (determined by the optical cut, the accuracy of grinding, and the method of mounting) and upon the coefficient of temperature drift which is determined primarily by the optical cut of the plate.

Crystal Cuts. The face of an X cut or Y cut crystal is made parallel to the Z axis in figure 3. Special cut crystals, known as AT cut, V cut, LD2, HF2, etc., are cut with the face of the crystal at an angle with respect to the Z axis, rather than being parallel to it. The purpose of the special cuts is to increase the power handling ability of the plates in some cases, but especially to reduce their temperature coefficient. AT, V, B5 and LD2 cut crystals have temperature coefficients approaching zero, and they should be used in radio transmitters in which accurate frequency control is essential. These crystals eliminate the need of a crystal oven for amateur work. A constant operating temperature is still required for many commercial applications, but the oven temperature need not be kept within as close limits as for an X or Y cut plate.

Spurious Peaks. Crystals that oscillate at more than one frequency are commonly known as crystals with multiple peaks. The dual vibrational tendency is more pronounced with

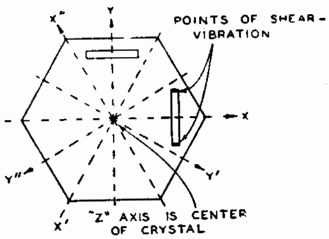


Figure 3.
SECTION THROUGH A QUARTZ CRYSTAL SHOWING THE AXES OF THE RAW QUARTZ.

Y cuts, but to a certain degree is exhibited by many X cuts. The use of a well designed, space-wound, low C tank coil in an oscillator will tend to discourage the crystal from oscillating at two frequencies, and in addition will increase the output. Experiments have shown that the frequency stability is not improved by large tank capacities, which only tend to augment the double frequency phenomenon.

Twin frequencies appear in several ways: sometimes the crystal will have two frequencies several hundred cycles apart, and will oscillate on both frequencies at the same time to produce an acoustically audible beat note. Other crystals will suddenly jump frequency as the tank tuning condenser is varied past a certain setting. Operation with the tank condenser adjusted near the point where the frequency shifts is very unstable, the crystal sometimes going into oscillation on one frequency and sometimes on the other as the plate voltage is cut on and off. Still other crystals will jump frequency only when the temperature is varied over a certain range. And some plates will jump frequency with a change in either tank tuning or temperature.

Edge-of-Band Operation. When operating close to the edge of the band, it is advisable to make sure that the crystal will respond to but one frequency in the holder and oscillator in which it is functioning; any crystal with two peaks can jump frequency slightly without giving any indication of the change in the meter readings of the transmitter. If the transmitter frequency is such that operation takes place on the edge of the band at all times, under all conditions of room temperature, some form of temperature control will be required for the crystal unless it is of the zero drift type.

When working close to the edges of the 14 or 28 Mc. band, it is essential that the crystal temperature be kept at a fairly constant value; the frequency shift in kilocycles per degree increases in direct proportion to the operating frequency, regardless of whether the fundamental or harmonic is used. When a crystal shifts its frequency by two kilocycles, its second harmonic has shifted 4 kilocycles. Amateurs not operating on the edge of the band generally need not concern themselves about frequency drift due to changes in room temperature.

If a pentode or beam tetrode tube having a plate potential of approximately 300 volts is used for the crystal oscillator, the temperature of the crystal, regardless of cut, should not increase enough to cause any noticeable drift even at 14 megacycles. When a crystal oscillator is keyed on 3.5 or 1.7 megacycles, the frequency drift is not of any consequence, even with much higher values of plate input, because of the keying and of the fact that the drift is not multiplied as it would be with harmonic operation of a final amplifier.

The Crystal Holder. Crystal holders have a large effect on the frequency; for example, the frequency of an 80-meter crystal can vary as much as 3 kilocycles in different holders. In fact, crystals can be purchased in variable gap holders which enable the operator to vary the frequency by varying the air gap. From 20 to 50 kc. shift can be obtained at 14 Mc. with the newer types of variable gap crystals.

High-Frequency Crystals. Forty-meter crystals can be treated much the same as 80-meter crystals, provided they are purchased in a dust-proof holder from a reliable manufacturer. However, it is a good idea with 40-meter crystals to make sure that the crystal current is not excessive, as it will run higher in a given oscillator circuit than when a lower frequency crystal is used in the same circuit at the same voltage. A low loss, low C tank circuit and a pentode or beam type oscillator tube are desirable.

Third-Harmonic Crystals (14 and 28 Mc.). Twenty- and 10-meter crystals, especially the latter, require more care in regard to circuit details, components and physical layout. These crystals are *not* of the zero drift type, as such crystals would be too thin to be of practical use. A special thick cut operated on a harmonic (almost always the third) is used to give the crystal sufficient mechanical ruggedness. Crystals of this cut have a drift of approximately 40-45 cycles/Mc/deg. C. This means that such crystals must be run at very low power levels not only to avoid fracture, but to prevent excessive drift. However, their use permits considerable simplification of a u.h.f. transmitter.

A type 41 tube, running at 275 volts on the plate and 100 volts on the screen, makes a good oscillator tube for a 20-meter crystal.

Bias should be obtained from a 500-ohm cathode resistor rather than from a grid leak. Very light loading, preferably with inductive coupling, is required. The tank coil should be low loss, preferably air-supported or wound on a ceramic form.

Medium high μ triodes with high transconductance and low input and output capacities make excellent 10-meter crystal oscillators. The types RK34, 6J5G and 955 are the most satisfactory oscillators, the 6J5G giving the greatest output besides being the least expensive.

Contrary to general practice with pentode crystal oscillators, the plate tank circuit should *not* be too low C; a moderate amount of tuning capacity should be used in a 10-meter triode crystal oscillator. The plate voltage on the oscillator tube should not be allowed to exceed 200 volts. About 2 watts output is obtainable from the 10-meter oscillator tank at this plate voltage. The tank coil can consist of 8 turns of no. 12 wire, air-wound to a ¾-inch diameter and spaced the diameter of the wire. Bias should be obtained from a 200-ohm cathode resistor (by-passed) and no grid leak. Connecting leads should be short and components small physically.

Both 10- and 20-meter crystal oscillators should be followed, where practicable, by a tube of high power gain, such as the 807. This reduces the number of tubes required in a high power stationary u.h.f. transmitter.

A 10-meter crystal oscillator with a 6J5G, driving a 6V6G doubler using a 150,000-ohm grid leak, makes an excellent 5-meter mobile transmitter. The latter tube can be either plate or plate-and-screen modulated. The modulation is better, especially when doubling, if both plate and screen are modulated.

Crystal Oscillator Circuits

Crystal oscillators can be divided into three classifications: (1) low power circuits, which require several additional buffer stages to drive medium or high power final amplifiers; (2) high power crystal oscillators, which minimize the number of buffer stages in a transmitter; (3) harmonic crystal oscillators, which operate on more than one harmonically related band from one quartz crystal.

Low power crystal oscillators are often required in transmitter design where extremely accurate frequency control is needed. The crystal oscillator tube is operated at low plate potential, such as 200 volts, with the result that oscillation is relatively weak. This means that there will be less heating effect in the quartz plate; the frequency drift, due to changes in temperature, is therefore minimized.

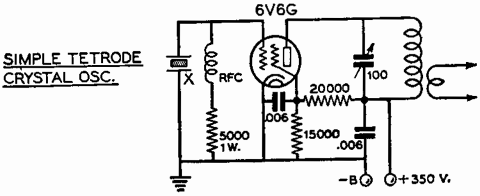


Figure 4.
TYPICAL CRYSTAL OSCILLATOR CIRCUIT.
This circuit has been found to be the most satisfactory for the frequency control of a multi-stage transmitter. A 6L6, or for that matter any pentode or power beam tetrode may be used with comparable success.

Mere operation of a quartz crystal oscillator tube at relatively low plate voltage does not necessarily mean a low degree of frequency drift; a type of crystal oscillator tube must be used which has high power sensitivity, high μ and low feedback (interelectrode) capacity. The amount of feedback determines the value of r.f. current flowing through the quartz plate and thus determines the amplitude of the physical vibration of the quartz plate. Any tube which requires only a very small amount of grid excitation voltage and has low grid-to-plate capacity can be used to supply relatively high-power output in a crystal oscillator without heating of the quartz plate.

High-power crystal oscillators are those which operate with as high a plate voltage as can be used with only moderate heating of the quartz crystal. Many transmitters, such as those used for amateur work, do not require as high a degree of frequency stability as do radiotelephone transmitters used for commercial services. The relatively high output from such crystal oscillators usually means the elimination of one or two buffer-amplifier stages. This simplifies the transmitter and may result in more trouble-free operation. There are a great many types of tubes suitable for high-power crystal oscillators, some of which are also used in high-stability low-power crystal oscillators by merely reducing the electrode voltages.

The crystal oscillator circuit in figure 4 is the standard oscillator circuit and uses either a pentode or beam tetrode tube. It operates on one frequency only, and the plate circuit is tuned to a frequency somewhat higher than that of the quartz crystal.

The actual power output of a crystal oscillator, such as shown in figure 4, is from one to fifteen watts, depending upon the values of plate and screen voltage. The use of *AT-cut*

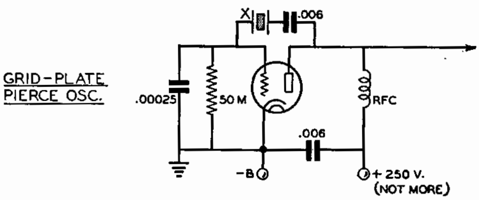


Figure 5.
TYPICAL PIERCE OSCILLATOR CIRCUIT.
No tank circuit is required with this type of crystal oscillator circuit. However, the crystal current is quite high for the amount of output voltage obtained.

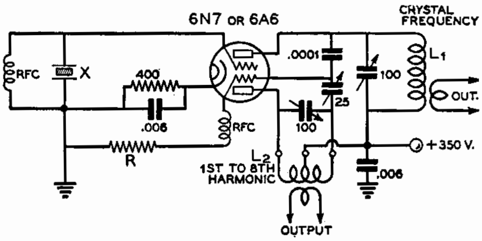


Figure 7.
REGENERATIVE DUAL-TRIODE OSCILLATOR.
By using one section of a dual triode as a crystal oscillator and the other section as a regenerative frequency multiplier, output on frequencies as high as the eighth harmonic of the crystal frequency may be obtained.

or low temperature coefficient quartz plates allows higher values of output to be obtained without exceeding the safe r.f. crystal current ratings or encountering frequency drift. *X-cut* and *Y-cut* crystals, especially the latter, must be operated with comparatively low crystal current because they not only will not stand as much r.f. crystal current, but also have a higher temperature coefficient.

Pierce Crystal Oscillator. One of the earliest crystal oscillator circuits recently enjoyed a revival in popularity. This is the Pierce oscillator, in which the crystal is connected directly from plate to grid of the oscillator tube, the crystal taking the place of the tuned tank circuit in an ultra-audion oscillator. Just as in the ultra-audion, the amount of feedback depends upon the grid to cathode capacity. Thus, it is only necessary to connect from grid to cathode a fixed condenser permitting the proper amount of feedback for the tube and frequency band used. The capacity is not at all critical, and ordinarily it is not necessary to change the capacity even when changing bands.

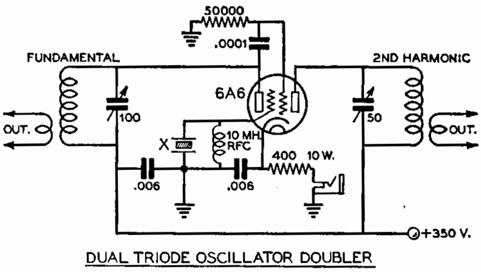


Figure 6.
TYPICAL TWIN-TRIODE OSCILLATOR-DOUBLER CIRCUIT.
Any dual triode of the 7F7, 6N7, 6F8G, 6A6, 53 class may be used in this simple circuit to obtain output on either the crystal frequency or its second harmonic.

The chief advantage of the oscillator is that it requires no tuned circuits. The chief disadvantage is that the maximum obtainable output is low, due to the fact that not over 200-250 volts can be used safely. Also, it works well only with 160- and 80-meter crystals, though many 40-meter crystals will work satisfactorily if the constants are chosen for maximum performance on 40 meters.

The oscillator may be fed plate voltage either through an r.f. choke or a resistor of high enough resistance that it doesn't act as a low impedance path for the r.f. energy. A considerably higher power output can be obtained with an r.f. choke in the plate circuit as compared to the use of a resistor in this position. However, since the plate voltage required on succeeding stages is invariably greater than that used on the Pierce crystal oscillator, the use of a resistor as the plate load is to be recommended. A popular version of the Pierce crystal oscillator circuit is shown in figure 5.

Dual-Triode Oscillator-Doubler Circuits. The types 6N7, 6A6, and 53 twin-triode tubes are popular for circuits where one triode acts as a crystal oscillator which drives the other triode as a frequency doubler; one tube, therefore, serves a dual purpose, supplying approximately 5 watts output on either the fundamental frequency or the second harmonic of the quartz crystal. Two applications of the twin-triode tube in a crystal oscillator circuit are shown in figures 6 and 7.

Figure 6 is a circuit which can be used with quartz crystals cut for 160-, 80-, 40- or 20-meter operation. The circuit shown in figure 7 can be made regenerative in the frequency-multiplier section in order to use the second triode as a tripler or quadrupler. By reducing the capacity of the feedback condenser to a low enough value, the second triode can be neutralized for use as a buffer stage. A suitable condenser for this purpose is a small mica-insulated trimmer condenser having a capacity range of from 3-to-30$\mu\mu$fds.

The resistor R shown in figure 7 should be from 30,000 to 50,000 ohms in value, and generally the r.f. choke shown in series with this resistor can be omitted.

Harmonic Oscillator Circuits. Harmonic oscillator circuits can be generally defined as those crystal oscillator arrangements which use a single tube and which allow power output to be obtained on harmonics of the crystal frequency. While these oscillator circuits have the advantage that one or more tubes are eliminated from the lineup, and sometimes that a tuned circuit is eliminated, they all have the disadvantage that they are difficult to adjust properly and they all have a tendency toward excessive crystal current when improperly tuned up. Five of the best known and most satisfactory of these oscillator circuits have been grouped together in figure 8.

The Tritet Crystal Oscillator. Any of the common pentode, tetrode, or screen-grid tubes may be used in the tritet crystal oscillator as shown in figure 8A. There are really two active circuits in this oscillator arrangement: the grid-cathode-screen circuit which acts as a triode crystal oscillator, and the cathode-grid-plate circuit which acts as an r.f. amplifier or frequency multiplier with its output circuit shielded from the oscillator portion. The tetrode or pentode plate circuit is *electron coupled* to the oscillator circuit. The plate circuit is generally tuned to the second harmonic and outputs of from 5 to 15 watts can be obtained without damage to the quartz crystal. This circuit is an improvement over the older forms of tritet in which a grid leak was used in place of the grid r.f. choke, and in which no cathode resistor and by-pass condenser were included. The improved circuit (figure 8A) decreases the crystal current as much as 50 per cent, and thereby protects the crystal against fracture. The cathode circuit is high C and is tuned to a frequency which is 40 to 50 per cent higher than that of the crystal. If an 802 or 807 is substituted for the 6L6 tube, the plate circuit can be tuned to the fundamental frequency of the crystal without making it necessary to short-circuit the cathode tuned circuit. A further reduction in r.f. crystal current may be obtained by connecting a 140-$\mu\mu$fd. variable condenser between the bottom of the crystal and the top of the cathode tank coil L_2. This condenser should be set to the smallest value of capacity which will permit steady oscillation and full output.

Regenerative Oscillator Circuits. Figures 8B and 8C show two versions of a regenerative crystal oscillator circuit which requires only one tank circuit and which is capable of giving power output on harmonics of the crystal

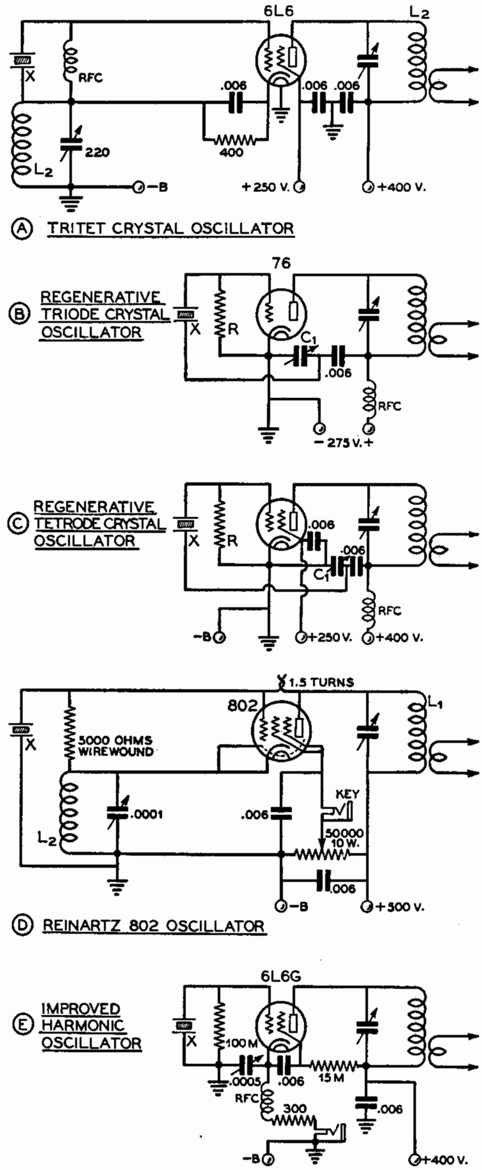


Figure 8.
REGENERATIVE OSCILLATOR CIRCUITS.
Full details of the operation of these oscillator circuits and a comparison between them is given in the text.

frequency. Figure 8B shows the circuit for use with a triode tube such as the 76, 6C5, 6J5, and 7A4, given in the order of their efficacy. 8C shows the same circuit adapted for use with a pentode or beam tetrode such as the 42, 6F6, 6V6, 6L6, or 7C5.

Triodes such as the 7A4, 6J5-GT, and the 76 will deliver as much as 2 or 3 watts with an r.f. crystal current of between 10 and 60 ma. for crystals from 160 to 10 meters. The triode circuit is excellent to drive a 6L6G buffer-doubler and the screen supply voltage for the 6L6G tube may be applied to the 76 plate circuit. This type of circuit is the only one which works with all crystals, 10, 20, 40, 80 and 160 meters, whether they are extremely active, such as a good X cut, or relatively inactive such as most high-frequency crystals. The triode will furnish from 1 to 2 watts at twice crystal frequency when used with 160-, 80- or 40-meter crystals by tuning the plate circuit to the second harmonic.

In figure 8B the cathode condenser, C_1, usually is left at some setting of from 40 to 50 $\mu\mu$fd. for 40-, 80- and 160-meter crystals.

A 6F6 or 42 works very well in the figure 8C circuit with a C_1 value of .0001 μfd. if heavily loaded. Eight to 12 watts output can be obtained easily from 160 to 20 meters and about 5 watts on 10 meters. A 6L6G tube requires a higher value of C_1, about .0004 μfd. unless heavily loaded.

Reinartz Crystal Oscillator. The Reinartz 802 crystal oscillator has a fix-tuned cathode circuit which is resonated to approximately *one-half* the crystal frequency. For example, with an 80-meter crystal the cathode circuit is tuned to 160 meters, the plate circuit to 80 meters. Either an 802 or a 6F6 tube can be used in a Reinartz crystal oscillator circuit. The output will be from 5 to 25 watts, depending upon the values of plate and screen voltages. The 6F6 is used as a high-μ triode in this same type of circuit, whereas the 802 is used as a pentode oscillator with additional control grid-to-plate capacity feedback. The circuit is shown in figure 8D.

The crystal r.f. current is quite low in this circuit, in comparison with the output power which can be obtained. The cathode circuit is tuned to half the frequency of the crystal, and the reactive effect produces regeneration at the harmonic frequency. This increases the operating efficiency of the tube without danger of uncontrollable oscillation at frequencies other than that of the crystal.

Improved Harmonic Oscillator. Figure 8E shows an improved version of a harmonic oscillator arrangement which has been suggested by Jones. It is quite similar to previous arrangements in regard to the general hookup but in this arrangement the screen is by-passed back to the cathode (which is hot to r.f.) rather than to ground. This is said to increase the stability of the oscillator and to increase the efficiency of the arrangement when operating on harmonics of the crystal frequency.

Push-Pull Crystal Oscillators. Figure 9 shows a simple crystal oscillator arrangement which makes use of one of the common dual-triode tubes as a push-pull oscillator. The type 6A6, 6N7, and 7F7 dual triodes make good push-pull crystal oscillators.

Outputs of from 5 to 10 watts can be obtained from this circuit without exceeding the ratings of the usual X-cut crystals. The crystal current for a push-pull oscillator is but little higher than for a single triode of the same type, and twice the output can be obtained.

Some push-pull oscillators will not oscillate on 160 meters, the feedback being insufficient in the push-pull connection to sustain oscillation under load.

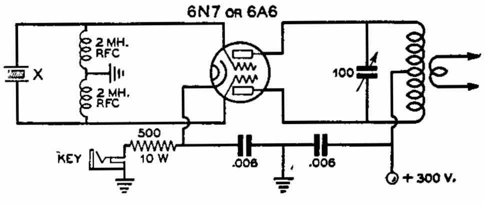


Figure 9.
PUSH-PULL CRYSTAL OSCILLATOR.
Any of the dual-triode tubes make a quite satisfactory push-pull crystal oscillator for feeding a push-pull r.f. amplifier, a push-push doubler, or merely to obtain somewhat greater output than from a single-ended oscillator.

Tuning the Crystal Oscillator

In nearly every practical transmitter circuit there will be some means for determining proper tuning of the crystal oscillator stage. Perhaps the most satisfactory of these tuning indicators is the grid milliammeter of the following stage. Maximum meter reading indicates maximum output from the crystal oscillator. Other indicators are: (1) A small neon bulb held near the plate end of the oscillator tuned circuit; maximum glow of the bulb indicates maximum oscillator output. (2) A flashlight bulb or a pilot light bulb, connected in series with a turn of wire fastened to a long piece of wood dowel (to protect the operator) can be coupled to the oscillator coil for indicating r.f. output. Maximum brilliancy of the lamp denotes maximum output from the oscillator.

Oscillator-Doubler Circuits. The type 6N7 or 6A6 oscillator-doubler circuit is adjusted by tuning the oscillator section for maximum output, and the doubler section for greatest dip in cathode or plate current. The crystal plate section should generally be tuned until the circuit approaches the point where oscillation is about to cease; this is towards the higher-capacity setting of the oscillator plate tuning condenser and operation in this manner provides most output in proportion to r.f. crystal current and frequency drift.

Harmonic Oscillators. Harmonic crystal oscillators are always tuned for maximum output and minimum plate, or cathode current. The regeneration or feedback condenser is adjusted or chosen to provide a good plate current dip when the plate circuit is tuned to the second harmonic of the crystal oscillator. Too much regeneration will cause the tube to oscillate for all settings of the plate tank condenser, without any sharp dip at the harmonic frequency of the crystal. Insufficient regeneration will result in low second harmonic output.

A plate potential of 400 volts is generally considered a safe upper limit for a type 6L6 oscillator tube. The screen-grid voltage affects the degree of regeneration and harmonic output; this voltage should generally range between 250 and 275 volts. The cathode current will run between 50 and 60 milliamperes for fundamental frequency operation, and 60 to 75 milliamperes for harmonic operation, at these plate and screen voltages. The crystal r.f. current normally runs between 25 and 75 milliamperes in this type of oscillator, depending on the frequency and plate voltage used.

Radio-Frequency Amplifiers

Since the output of the oscillator stage in a transmitter (whether it be self-controlled or crystal controlled) must be kept down to a fairly low level to maintain stability and to maintain a factor of safety from fracture of the crystal when one is used, the low power output of the oscillator is brought up to the desired power level by means of radio-frequency amplifiers. The two classes of amplifiers that find widest application in amateur transmitters are the class B and class C types.

The Class B Amplifier. Class B amplifiers are used in a radio-telegraph transmitter when maximum power gain is desired in a particular stage. A class B amplifier operates with cutoff bias and a comparatively small amount of excitation. Power gains of 20 to 200 or so are obtainable in a well-designed class B amplifier. The plate efficiency of a class B c.w. amplifier will run around 65 per cent.

The Class B Linear. Another type of class B amplifier is the class B linear stage as employed in radiophone work. This type of amplifier is used to increase the level of a modulated carrier wave and depends for its operation upon the linear relation between excitation voltage and output voltage. Or, to state the fact in another manner, the power output of a class B linear stage varies linearly with the square of the excitation voltage. The class B linear amplifier is operated with cutoff bias and a small value of excitation, the actual value of exciting power being such that the power output under carrier conditions is one fourth of the peak power capabilities of the stage. Class B linears are very widely employed in broadcast and commercial installations, but are comparatively uncommon in amateur application, since tubes with high plate dissipation are required for moderate output. Another reason for their unpopularity among amateurs is that the power limitation upon amateurs is placed upon power *input* to the final stage and not upon power *output*. The approximately 33 per cent efficiency of the class B linear makes the power capability of a transmitter with a linear amplifier in the final stage less than half that of a high-level modulated transmitter whose maximum efficiency may be as high as 75 or 80 per cent. This assumes, of course, that the maximum legal input of one kilowatt is being employed in each case.

The Class C Amplifier. Class C amplifiers are very widely employed in all types of transmitters. A good power gain may be obtained (values of gain from 3 to 20 are common) and the plate circuit efficiency may be, under certain conditions, as high as 85 per cent. Class C amplifiers operate with considerably more than cutoff bias and ordinarily with a rather large amount of excitation as compared to a class B amplifier. The bias for a normal class C amplifier is such that plate current on the stage flows for approximately 120° of the 360° excitation cycle. Class C amplifiers are used in transmitters where a fairly large amount of excitation power is available, where good plate circuit efficiency is desired, and when the stage is to be plate modulated.

Class C Plate Modulation. The characteristic of a class C amplifier which makes it linear with respect to changes in plate voltage is that which allows such an amplifier to be *plate modulated* for radiotelephony. Through the use of higher bias than is required for a c.w. class C amplifier and greater excitation, the linearity of such an amplifier may be extended from zero plate voltage to twice the normal value. The output power of a class C amplifier adjusted for plate modulation varies with the square of the plate voltage. Since this is the same condition that would take place if a resistor equal to the voltage on the amplifier divided by its plate current were substituted for the amplifier, it is said the stage presents a resistive load to the modulator.

Class C Grid Modulation. If the grid current to a class C amplifier is reduced to a low value and the plate loading is increased to the point where the plate dissipation approaches the rated value, such an amplifier may be grid modulated for radiotelephony. If the plate voltage is raised to quite a high value and the stage is adjusted carefully, efficiencies as high as 42 to 45 per cent with good modulation capability and comparatively low distortion may be obtained. This type of operation is termed class C grid modulation and is coming into increasing favor among amateur radiotelephone operators.

Grid Excitation. A sufficient amount of grid excitation must be available for class B or class C service. The excitation for a plate-modulated class C stage must be sufficient to drive a normal value of d.c. grid current through a grid bias supply of about 2½ times cutoff. The bias voltage preferably should be obtained from a combination of grid leak and fixed C-bias supply. Cutoff bias can be calculated by dividing the amplification factor of the tube into the d.c. plate voltage. This is the value normally used for class B amplifiers (fixed bias, no grid leak). Class C amplifiers use from 1½ to 5 times this value, depending upon the available grid drive, or excitation, and the desired plate efficiency. Less grid excitation is needed for c.w. operation, and the values of fixed bias (if greater than cutoff) may be reduced, or the value of the grid leak resistor can be lowered until normal d.c. grid current flows. This value should be between 75% and 100% of the value listed under tube characteristics.

The values of grid excitation listed for each type of tube may be reduced by as much as 50% if only moderate power output and plate efficiency are desired. When consulting the tube tables, it is well to remember that the power lost in the tuned circuits must be taken into consideration when calculating the available grid drive. At very high frequencies, the r.f. circuit losses may even exceed the power required for grid drive unless low loss tank circuits are used.

Readjustments in the tuning of the oscillator, buffer or doubler circuits, will result in greater grid drive to the final amplifier. The actual grid driving power is proportional to the d.c. voltage developed across the grid leak (or bias supply) multiplied by the d.c. grid current.

Link coupling between stages, particularly to the final amplifier grid circuit, normally will provide more grid drive than can be obtained from other coupling systems. The number of turns in the coupling link and the location of the turns on the coil can be varied with respect to the tuned circuits to obtain the greatest grid drive for allowable values of buffer or doubler plate current. Slight readjustments sometimes can be made after plate voltage has been applied.

Excessive grid current will damage the tubes by overheating the grid structure; beyond a certain point of grid drive no increase in power output can be obtained for a given plate voltage.

Neutralization of R. F. Amplifiers

The plate-to-grid feedback capacity of triodes makes it necessary that they be neutralized for operation as r.f. amplifiers at frequencies above about 500 kc. Those screen-grid tubes, pentodes, and beam tetrodes which have a plate-to-grid capacity of a small fraction of one micro-microfarad may ordinarily be operated as an amplifier without neutralization.

Neutralizing Circuits. The object of a neutralization circuit for an r.f amplifier is, of course, to cancel or "neutralize" the capacitive feedback of energy from plate to grid. There are two general methods by which this energy feedback may be eliminated: the first, and the most common method, is through the use of a capacity bridge, and the second method is through the use of a parallel reactance of equal and opposite polarity to the grid-to-plate capacity to nullify the effect of this capacity.

Until recently, the capacity-bridge method of neutralization was divided into two systems, grid neutralization and plate neutralization. It has always been known that the use of grid neutralization caused an amplifier to be either regenerative or degenerative, but it was not until quite recently that Doherty showed the reason for the unsatisfactory performance of grid neutralization. Hence, only plate neutralization (the capacity bridge system), and coil neutralization (the opposite reactance system) will be considered as satisfactory methods for neutralizing a single-ended r.f. amplifier stage.

Tapped-Coil Plate Neutralization. As was mentioned under *Neutralizing Circuits,* there are two general types of neutralizing circuits for a single-ended amplifier, the bridge and opposite-reactance methods. The following paragraphs will describe first the variations upon the bridge method. Figure 10A shows a circuit for the neutralization of a single-ended triode r.f. amplifier by means of a tapped coil in the plate circuit. This circuit is satisfactory for frequencies below about 7 Mc. with ordinary tubes but a considerable amount of regeneration will be found when this circuit is used on frequencies above 7 Mc. Some regeneration can be tolerated in an amplifier for c.w. use, but for phone operation either of the split-stator circuits described in the next two paragraphs should be used.

Split-Stator Plate Neutralization. Figure 10B shows the neutralization circuit which is most widely used in single-ended r.f. stages. The use of a split-stator plate condenser makes the electrical balance of the circuit substantially independent of the mutual coupling within the coil and also makes the balance independent of the place where the coil is tapped. With conventional tubes this circuit will allow one neutralization adjustment to be made on, say, 14 Mc., and this adjustment will hold for all lower frequency bands.

Capacity-Balanced Split-Stator Plate Neutralization. Figure 10C shows an alternative circuit for split-stator neutralization of a single-ended amplifier stage which, with low-capacity tubes, can be made to remain in adjustment on all bands from 56 Mc. on down in frequency. The additional balancing condenser CB serves merely as an adjustment to keep the capacity-to-ground exactly the same from each side of the balanced plate tank circuit. This condenser can be either a small adjustable one of the type commonly used for neutralization, or the relative capacity to ground of the two sides of the circuit can be proportioned so that there is a balance. In determining the balance of the circuit, it must be remembered that the plate-to-filament capacity of the power amplifier tube is the main item to cause the unbalance. If the other capacities

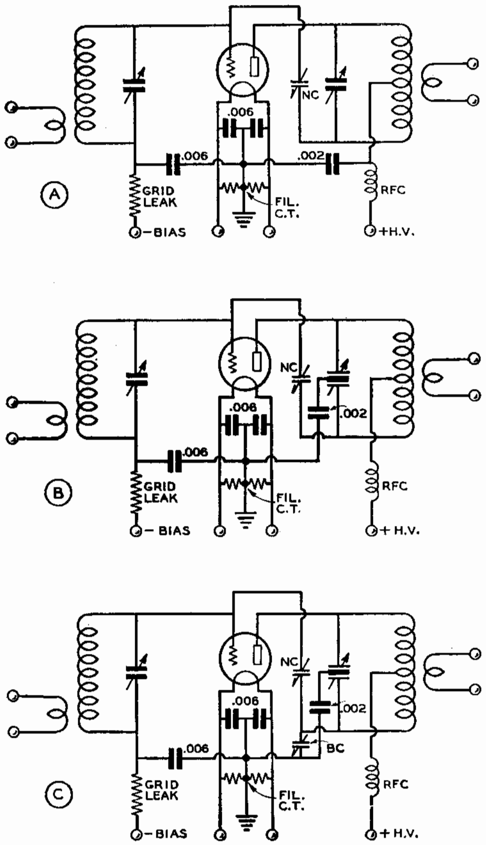


Figure 10.
PLATE NEUTRALIZING CIRCUITS FOR A SINGLE-ENDED AMPLIFIER.

(A) shows a neutralizing circuit employing a split coil plate tank which is suitable under ordinary conditions for operation at frequencies as high as 7 Mc. (B) shows conventional split-stator plate neutralization. (C) shows split-stator plate neutralization with the addition of a balancing condenser BC which compensates for the plate-to-ground capacity of the amplifier tube and thus keeps the output tank circuit balanced to ground, improving neutralization on the higher frequencies.

of the circuit are perfectly balanced with respect to ground, the capacity of the condenser CB should be approximately equal to the plate-to-ground capacity of the tube being neutralized. However, it is often just as convenient to unbalance the circuit capacities to ground until the additional capacity on the neutralizing side of the circuit is about equal to that on the plate side. At the point where the plate-to-ground capacity is exactly

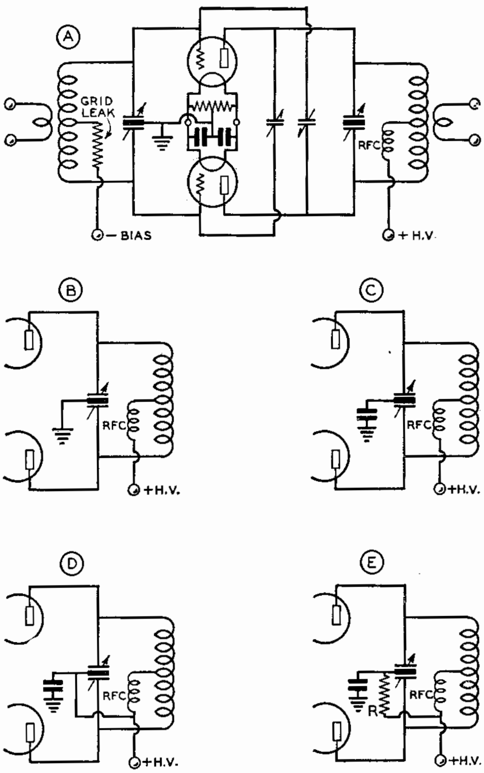


Figure 11.
PUSH-PULL AMPLIFIER NEUTRALIZATION.

(A) shows the basic circuit for a neutralized push-pull r.f. amplifier. In this circuit the nodal point for the stage is determined by the grounded rotor on the grid tuning condenser and the rotor of the plate tank condenser is allowed to float. (B), (C), (D), and (E) show alternative arrangements for returning the rotor of the plate tank condenser to ground when this grounding is deemed necessary. Discussion of the various circuits is given in the text.

balanced the amplifier will neutralize perfectly (at least as nearly perfect as a push-pull amplifier) and will stay neutralized on all bands for which the amplifier tubes are satisfactory.

Push-Pull Neutralization. Two tubes can be connected for *push-pull* operation so as to obtain twice as much output as that of a single tube. A push-pull amplifier, such as that shown in figure 11A, also has an advantage in that the circuit can more easily be balanced than a single-tube r.f. amplifier. The various interelectrode capacities and the neutralizing condensers are connected in such a manner that those on one side of the tuned circuits are exactly equal to those on the opposite side. For this reason, push-pull r.f. amplifiers can be more easily neutralized in very-high-frequency transmitters; also, they usually remain in perfect neutralization when tuning the amplifier to different bands.

The circuit shown in figure 11A is perhaps the most commonly used arrangement for a push-pull r.f. amplifier stage. The rotor of the grid condenser is grounded and the rotor of the plate tank condenser is allowed to float. Under certain conditions the circuit of 11B may be used (when the plate tank condenser has a much larger voltage rating than the maximum possible peak output of the power tubes) with the rotor of the grid condenser grounded or not, as desired. It is also possible to use a single-section grid condenser with a tapped coil (un-bypassed) for low-frequency operation with this circuit arrangement.

Figure 11C shows an alternative arrangement for the return of the rotor of the plate tank condenser which is best for use with a c.w. amplifier stage. The by-pass condenser from the rotor to ground can be any capacity from .01 μfd. down to .0005 μfd. and even down to .0001 μfd. for a u.h.f. amplifier. For phone use it is best to have some sort of a coupling arrangement to make the rotor of the tuning condenser follow plate voltage fluctuations. As long as the rotor of the tuning condenser is at the same d.c. potential as the stators there will be a much reduced chance of breakdown on modulation peaks.

Figures 11D and 11E show two arrangements which tend to keep the rotor of the condenser as nearly as possible at the same d.c. potential as the stators. In figure 11D the rotor of the condenser, and the ungrounded side of the by-pass condenser, is merely connected to the plate supply side of the r.f. choke. This is an excellent arrangement for use with moderate plate voltages but has the disadvantage that considerable stress is placed on the mica by-pass condenser, and should this condenser break down the plate supply would be shorted. Figure 11E shows an alternative arrangement which has the advantage that, should the mica by-pass condenser short out, only the resistor R will be destroyed. For a mica by-pass capacity of .001 μfd. and a maximum 100 per cent modulation frequency of 3000 cycles, a 25,000-ohm resistor will be satisfactory for R.

Shunt Neutralization. The feedback of energy from grid to plate that would cause oscillation or serious regeneration in an unneutralized r.f. amplifier is a result of the grid-to-plate capacity of the amplifier tube. A neutralization circuit is merely an electrical arrangement for nullifying the effect of this capacity. All the previous neutralization circuits have made use of a bridge circuit for balancing out the grid-to-plate energy feedback by an amount of energy equal and opposite in phase from a balanced tuned circuit.

Another method of eliminating the feedback effect of this capacity, and hence of neutralizing the amplifier stage, is shown in figure 12. The grid-to-plate capacity in the triode amplifier tube acts as a capacitive reactance coupling energy back from the plate to the grid circuit. If we parallel this capacity with an inductance having the same value of reactance (but having the opposite sign, of course) at the frequency upon which the amplifier is operating, the reactance of one will cancel the reactance of the other and we will have a high-impedance tuned circuit from grid to plate on the triode tube.

This neutralization circuit works very beautifully and can be used on ultra-high frequencies where other neutralization circuits are unsatisfactory. This is true because the lead length in the neutralization circuit is practically negligible. The circuit can also be used with push-pull r.f. amplifiers; in this case each tube is neutralized separately although both neutralizing condensers are set to the same capacity.

The big advantage of this arrangement is that it allows the use of single-ended tank circuits with a single-ended amplifier.

However, the circuit has one serious disadvantage for amateur work in which the frequency of operation is changed frequently: the neutralization holds for one frequency—that frequency where the grid-to-plate capacity is resonant with the external neutralization coil. But by the use of plug-in coils and the trimmer condenser C in parallel with the grid-to-plate capacity, it is possible to shift the band of operation and to trim to any frequency within the band. This trimmer condenser, if used, must be insulated for somewhat more voltage than the tank condenser. The .0001-μfd. condenser in series with the neutralizing circuit is merely a blocking condenser to isolate the plate voltage from the grid circuit. The coil L will have to have a very large number of turns for the band in operation in order to be resonant with the usually rather small grid-to-plate capacity. But since, in all ordinary cases with tubes operating on frequencies for which they were designed, the L/C ratio of the tuned circuit will be very high, the coil can use comparatively small wire although it must be wound on air or very low-loss dielectric and must be insulated for the sum of the plate r.f. voltage and the grid r.f. voltage.

Figure 13 shows an alternative arrangement for the neutralizing circuit in which the variable trimmer condenser is in series with the neutralizing coil instead of in parallel

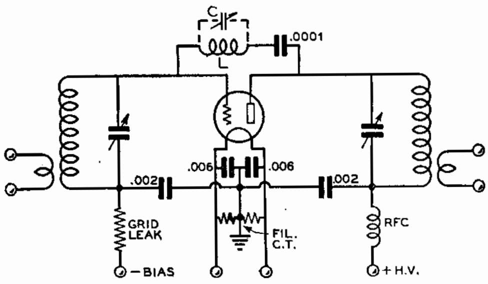
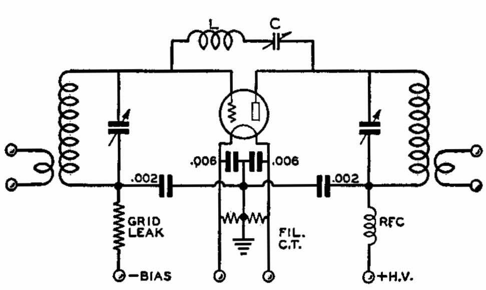


Figure 12.
SHUNT OR "COIL" NEUTRALIZATION.
This neutralization circuit makes use of a coil connected from grid to plate (with a blocking condenser in series with it) which resonates with the grid-to-plate capacity to the operating frequency. The impedance from plate to grid is thus made very high, feedback is stopped, and the amplifier is neutralized for this frequency of operation. When the frequency of operation is changed, the trimmer condenser C changes the resonant frequency of this circuit to the new operation frequency.

Figure 13.
ALTERNATIVE SHUNT NEUTRALIZATION CIRCUIT.
In this circuit the trimmer condenser for varying the frequency of resonance of the circuit is placed in series with the neutralizing coil, thus replacing the blocking condenser and reducing the necessary voltage rating for the trimmer condenser, although increasing the capacity required.



with it. This system also allows the stage to be trimmed to neutralization on any frequency in the band of operation. This condenser can have a capacity of 35 to 100 $\mu\mu$fd. and need not have nearly as much voltage insulation as the trimmer condenser shown in figure 12. A plate spacing of .070" will be ample for any plate voltage ordinarily used by the amateur.

Neutralizing Procedure

The r.f. amplifier is neutralized to prevent self-oscillation or regeneration. A neon bulb, a flashlight lamp and a loop of wire, or an r.f. galvanometer can be used as a *null indicator* for neutralizing low-power stages. Plate voltage is disconnected from the r.f. amplifier stage while it is being neutralized. Normal grid drive then is applied to the r.f. stage, the neutralizing indicator is coupled to the plate coil and the plate tuning condenser is tuned to resonance. The neutralizing condenser (or condensers) then can be adjusted until *minimum* r.f. is indicated for resonant settings of both grid and plate tuning condensers. Both neutralizing condensers are adjusted simultaneously and to approximately the same value of capacity when a push-pull stage is being neutralized.

A final check for neutralization should be made with a d.c. milliammeter connected in the grid leak or grid-bias circuit. There will be no movement of the meter reading as the plate circuit is tuned through resonance (without plate voltage being applied) when the stage is completely neutralized. The milliammeter check is more accurate than any other means for indicating complete neutralization and it also is suitable for neutralizing the stages of a high-power transmitter.

Push-pull circuits usually can be more completely neutralized than single-ended circuits when operating at very high frequencies. In the intermediate range of from 3 to 15 megacycles, single-ended circuits will give satisfactory results. Single-ended operation in the 3-to-15 megacycle range is most stable with split-stator tuning condensers.

Neutralizing Problems. When a stage cannot be completely neutralized, the difficulty can be traced to one or more of the following causes: (1) The filament leads may not be by-passed to the common ground bus connection of that particular stage. (2) The ground lead from the rotor connection of the split-stator tuning condenser to filament may be too long. (3) The neutralizing condensers may be in a field of excessive r.f. from one of the tuning coils. (4) Electromagnetic coupling may exist between grid and plate coils, or between plate and preceding buffer or oscillator circuits. (5) Insufficient shielding or spacing between stages, or between grid and plate circuits in compact transmitters may prevent neutralization or give false indications of neutralizing adjustments. (6) If shielding is placed too close to plate circuit coils, neutralization will not be secured because of induced currents in the shields. (7) Parasitic oscillations may take place when plate voltage is applied. The cure for the latter is mainly a matter of cut and try—rearrange the parts, change the length of grid or plate or neutralizing leads, insert an ultra-high-frequency r.f. choke in the grid lead or leads, or eliminate the grid r.f. chokes which may be the cause of a low-frequency parasitic (in conjunction with plate r.f. chokes).

Plate Circuit Tuning. When the amplifier is completely neutralized, reduced plate voltage should be applied before any load is coupled to the amplifier. This reduction in plate voltage should be at least 50% of normal value because the plate current will rise to excessive values when the plate tuning condenser is not adjusted to the point of resonance. The latter is indicated by the greatest dip in reading of the d.c. plate current milliammeter; the r.f. voltage across the plate circuit is greatest at this point. With no load, the r.f. voltage may be several times as high as when operating under conditions of full load; this may result in condenser flashover if normal d.c. voltage is applied. The no-load plate current at resonance should dip to 10% or 20% of normal value. If the plate circuit losses are excessive, or if *parasitic oscillations* are taking place, the no-load plate current will be higher.

Loading. The load (antenna or succeeding r.f. stage) then can be coupled to the amplifier under test. The coupling can be increased until the plate current at resonance (greatest dip in plate current meter reading) approaches the normal values for which the tube is rated. The value at reduced plate voltage should be proportionately less in order to prevent excessive plate current load when normal plate voltage is applied. Full plate voltage should not be applied to an amplifier unless the r.f. load also is connected; otherwise the condensers will arc or flash over, thereby causing an abnormally high plate current which may damage the tube. The tuned circuit impedance is lowered when the amplifier is loaded, as are the r.f. voltages across the plate condenser.

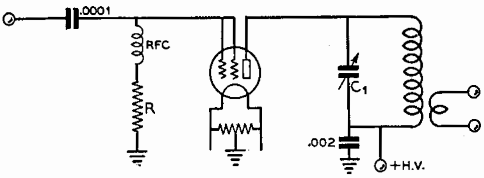


Figure 14.
CONVENTIONAL FREQUENCY DOUBLER CIRCUIT.
A high-μ, dual-grid triode, or a pentode or beam tetrode with the grid and screen paralleled makes an excellent frequency doubler. In addition, all of these types of tubes have the advantage that when the excitation is removed their plate current will fall to a very low value. The plate circuit is tuned to twice the excitation frequency.

Grid Excitation. Excessive grid excitation is just as injurious to a vacuum tube as abnormal plate current or low filament voltage. Too much grid driving power will overheat the grid wires in the tube, and will cause a release of gas in certain types of tubes. An excess of grid drive will not appreciably increase the power output and increases the efficiency only slightly after a certain point is reached. The grid current in the tube should not exceed the values listed in the *Tube Tables,* and care also should be exercised to have the bias voltage low enough to prevent flashover in the stem of the vacuum tube.

Grid excitation usually refers to the actual r.f. power input to the grid circuit of the vacuum tube, part of which is used to drive the tube, and part of which is lost in the C-bias supply. There is no way to avoid wasting a portion of the excitation power in the bias supply.

Frequency Multipliers

Quartz crystals are not ordinarily used for direct control of the output of high-frequency transmitters. *Frequency multipliers* are needed to multiply the frequency to the desired value. These multipliers operate on exact multiples of the crystal frequency; a 3.6-megacycle crystal oscillator can be made to control the output of the transmitter on 7.2 or 14.4 megacycles, or even on 28.8 megacycles, by means of one or more frequency multipliers. When used at twice frequency, as they most usually are, they are often termed *frequency doublers.* A simple doubler circuit is shown in figure 14. It consists of a vacuum tube with its plate circuit tuned to *twice* the frequency of the grid driving circuit. This doubler can be excited from a crystal oscillator, or connected to another doubler or buffer amplifier stage.

Doubling is best accomplished by operating the tube with extremely high grid bias in order to make the output plate current rich in harmonics. The grid circuit is driven approximately to the normal value of d.c. grid current through the r.f. choke and grid leak resistor, shown in figure 14. The resistance value generally is from two to five times as high as that used with the same tube for simple amplification. For the same value of grid current the grid bias is several times as high.

Neutralization is seldom absolutely necessary in a doubler circuit, since the plate is

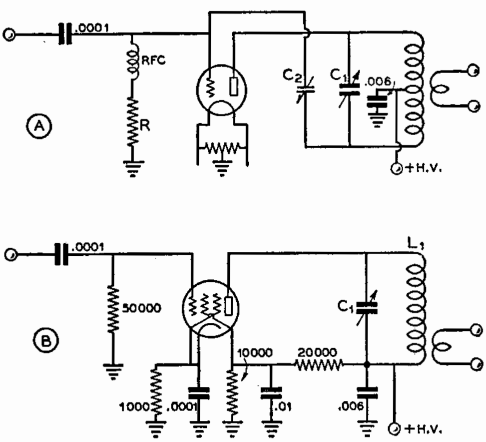


Figure 15.
REGENERATIVE DOUBLER CIRCUITS.
(A) shows a circuit which may be used either as a neutralized buffer stage or, when the capacity of C_2 is increased beyond the "neutralized" setting, as a regenerative doubler. (B) shows a frequency multiplier circuit with cathode regeneration which will give quite good results as a doubler, and very good results, compared to other multiplier circuits, as a frequency quadrupler.

tuned to twice the frequency of the grid circuit. The feedback from the doubler plate circuit to the grid circuit is at *twice* the frequency of the grid driving circuit to which the coupling condenser (figure 14) is connected. The impedance of this external tuned grid driving circuit is very low at the doubling frequency and thus there is no tendency for self-excited oscillation when ordinary triode tubes are used. At very high frequencies however, this impedance may be great enough to cause

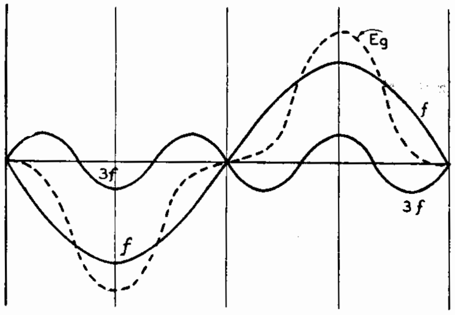


Figure 16.
PEAKED WAVEFORM OBTAINED BY ADDITION OF FUNDAMENTAL AND THIRD-HARMONIC ENERGY IN PROPER PHASE.

When fundamental frequency (f) energy and third-harmonic (3f) energy are added in the proper phase the result is a peaked waveform as shown by E_g. This peaked waveform, when used as excitation for a frequency doubler stage, gives considerably higher plate efficiency than when sine-wave excitation voltage is applied to the grid of the tube.

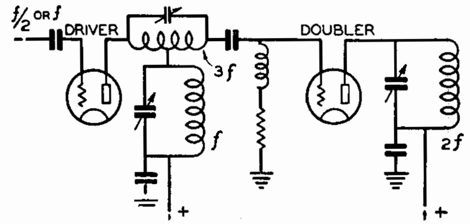


Figure 17.
CIRCUIT FOR COMBINING FUNDAMENTAL AND THIRD-HARMONIC ENERGY IN PROPER PHASE FOR PEAKED WAVEFORM.

The small third-harmonic tank circuit connected as shown adds the fundamental and third harmonic in the proper phase relation for producing a peaked excitation waveform on the grid of the doubler stage.

regeneration, or even oscillation, at the tuned output frequency of the doubler.

A doubler can either be neutralized or made more regenerative by adjusting C_2 in the circuit shown in figure 15.

When condenser C_2 is of the proper value to neutralize the plate-to-grid capacity of the tube, the plate circuit can be tuned to twice the frequency (or to the same frequency) as that of the source of grid drive; the tube can be operated either as a neutralized amplifier or doubler. The capacity of C_2 can be increased so that the doubler will become *regenerative,* if the r.f. impedance of the external grid driving circuit is high enough at the output frequency of the stage.

Frequency doublers require bias of several times cutoff; high-μ tubes therefore are desirable for this type of service. Tubes which have amplification factors from 20 to 200 are suitable for doubler circuits. Tetrodes and pentodes usually have high amplification factors. Low-μ triodes, having amplification constants of from 3 to 10, are not applicable for doubler service because in some cases the grid voltage must be as high as the plate voltage for efficient doubling action. The necessary d.c. grid voltage for high-μ tubes can be obtained more easily from average driver stages in conventional exciters.

Angle of Flow in Frequency Multipliers. The angle of plate current flow in a frequency multiplier is a very important factor in determining the efficiency. As the angle of flow is decreased for a given value of grid current, the efficiency increases. To reduce the angle of flow, higher grid bias is required so that the grid excitation voltage will exceed the cutoff value for a shorter portion of the exciting-voltage cycle. Frequency doublers of all types should have an angle of flow of 90 degrees or less, triplers 60 degrees or less and quadruplers 45 degrees or less.

Normally, a smaller angle of flow requires quite high bias and excitation. However, by altering the shape of the exciting voltage from its usual sine wave shape at the exciting frequency, it is possible to decrease the angle of flow and thus increase the efficiency without resorting to increases in the excitation voltage and bias.

The angle of flow may be decreased by adding some properly phased third harmonic voltage to the excitation. The result of adding the third harmonic voltage to the fundamental is shown graphically in figure 16. As shown by the dotted curve, E_g, when the fundamental and third harmonic voltages are added in the proper phase the result is a grid excitation voltage having a peaked wave form, exactly what is required for high-efficiency frequency multiplying. The method by which the third harmonic is added is shown in figure 17. A small, center-tapped tank circuit tuned to three times the driver frequency is placed between the driver plate and the coupling condenser to the frequency-multiplier stage. The center tap of this coil is connected to the "hot" end of the driver plate tank, which remains tuned to the fundamental frequency. The third-harmonic tank circuit can be tuned accurately to frequency by coupling to it a small, low-current dial lamp in a loop of wire and tuning for maximum brilliancy. An absorption wavemeter may be coupled to the third-harmonic tank after it has been tuned to make sure that it is on the correct harmonic. The tuning of this circuit is not critical; one setting will serve to cover an amateur band.

Push-Push Doublers. Two tubes can be connected with the grids in push-pull, and the plates in parallel, for operation in a so-called *push-push doubler,* as shown in figure 18.

This doubler circuit will deliver twice as much output as a single-tube circuit; it has proven popular in amateur transmitters because of its operating ease. In previous doubler circuits, capacitive coupling was shown. Link coupling to the tuned circuit in a preceding stage is shown in figure 18. This coupling arrangement simplifies the push-pull connection of the two grid circuits.

The circuit C_2-L_2 is tuned to the same frequency as that of the preceding tuned circuit, and the doubler plate circuit C_1-L_1 is tuned to *twice* the frequency. The grid circuit should be tuned by means of a split-stator condenser, connected as shown in figure 18, rather than by means of the single-section tuning condenser and by-passed center-tapped coil arrangement. The latter would provide a relatively high impedance at the doubling frequency. The push-push doubler then would be highly regenerative, and in most cases it would break into self-oscillation. The split-stator tuning circuit, because it has a capacitive reactance, provides a very low impedance at the doubling frequency, so that there is very little regenerative action; the circuit, therefore, is quite stable if the grid tank is not made too low C.

Some multigrid crystal oscillators are designed so that frequency doubling can be accomplished directly in the oscillator tube circuit by connecting the various grids in push-pull (2 tubes) and the output plates in parallel.

The push-push circuit makes a very efficient doubling arrangement because each grid is being excited on a positive half of the exciting voltage and, since the grids are in push-pull, this means that plate current flows to one or the other of the parallel plates twice during every cycle of the exciting voltage. Thus the current pulses in the plate circuit occur at twice the exciting-voltage frequency, resulting in extremely efficient doubling action. The push-push doubler may also be used as a quadrupler by tuning the plate circuit to the fourth harmonic of the grid-excitation frequency. As with a single ended doubler, short-pulse excitation is required for good efficiency.

Tank Circuit Capacities

Tuning capacity values for class C amplifiers are an important consideration to anyone building a radio transmitter. The best value of capacity can be determined closely by charts or formulas for any frequency of operation. The ratio of C to L, capacitance to inductance, depends upon the operating plate voltage and current, and upon the type of circuit. Proper choice of capacity-to-inductance ratio for resonance at any given frequency is important in obtaining low harmonic output and also low distortion in the case of a modulated class C amplifier.

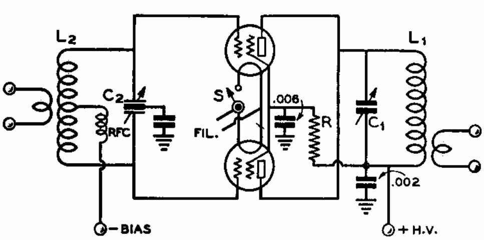


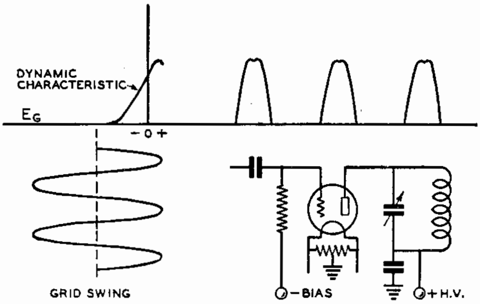
Figure 18.

PUSH-PUSH DOUBLER CIRCUIT.

In this type of doubler the grids are connected in push-pull and the plates are connected in parallel. A pair of triodes, a dual triode, or a pair of pentodes or tetrodes may be used. In the diagram shown, the heater of one of the tubes may be opened and the other tube operated as a neutralized amplifier, the other tube acting as the neutralizing condenser.

A class C amplifier produces a very distorted plate current wave form in the form of pulses as shown in figure 19. The LC circuit is tuned to resonance and its purpose is to smooth out these pulses into a sine wave of radio-frequency output, since any wave form distortion of the carrier frequency is illegal, causing harmonic interference in higher-frequency channels. A class A radio-frequency amplifier would produce a sine wave output. However, the a.c. plate current would be flowing during the full 360° of each r.f. cycle, resulting in excessive plate loss in the tube for any reasonable value of output. The class C amplifier has a.c. plate current flowing during only a fraction of each cycle, allowing the plate to cool off during the remainder of each cycle. If the plate current is zero for 2/3 of each cycle, the *angle of plate current flow* is said to be 120°, since current is flowing during 1/3 of 360°. The tube in a class C amplifier could have several times as much power input for a given plate loss as when used in a class A amplifier.

The tuned circuit must have a good fly-wheel effect in order to furnish a sine-wave output to the antenna when it is receiving energy in the form of very distorted pulses such as shown in figure 19. The LC circuit fills in power over the complete r.f. cycle, providing the LC ratio is correct. The fly-wheel effect is generally defined as the ratio of radio-frequency volt-amperes to actual power output, or VA/W. This is equivalent to Q and should not be much less than 4π, or 12.5, for a class-C amplifier. At this value of VA/W or Q, one-half of the stored energy in the LC circuit is absorbed by the antenna. If a lower value of Q is used, the storage power is insufficient to produce a sine (undistorted) wave output to the antenna and power will be wasted in radiation of harmonics.



CLASS C AMPLIFIER PLATE CURRENT WAVEFORM

Figure 19.

Too high a value of VA/W or Q will result in excessive circulating r.f. current loss in the LC circuit and lowered output to the antenna. In high-fidelity radiophone transmitters, too high a Q will cause attenuation of the higher sideband frequencies and consequent loss of the higher audio frequencies. Too low a Q has its disadvantages also; so most transmitters are operated with LC circuit values of between 10 and 25. A value of 20 seems to be high enough for modulated class C amplifiers; about 10 to 12 is enough for c.w. transmitters. With values of Q less than about 10, the maximum r.f. output will not occur at the point of minimum plate current in the amplifier tuning adjustment.

Harmonic Radiation vs. Q. Opinions vary as to the correct value of Q, but a careful analysis of the whole problem seems to indicate that a value of 12 is suitable for most amateur phone or c.w. transmitters. A value of 15 to 20 will result in less harmonic radiation at the expense of a little additional heat power loss in the tank or LC circuit. The charts shown have been calculated for an operating value of Q = 12.

The curves shown in figure 20 indicate the sharp increase in harmonic output into the antenna circuit for low values of Q. The curve for the second harmonic rises nearly vertically for Q values of less than 10. The third harmonic does not become seriously large for values of Q less than 4 or 5. These curves show that push-pull amplifiers may be operated at lower values of Q if necessary, since the second harmonic is cancelled to a large extent *if there is no capacitive or unbalanced coupling* between the tank circuit and the antenna feeder system.

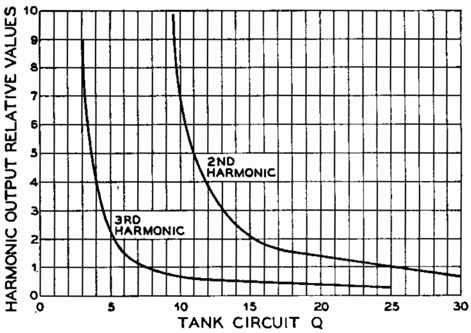
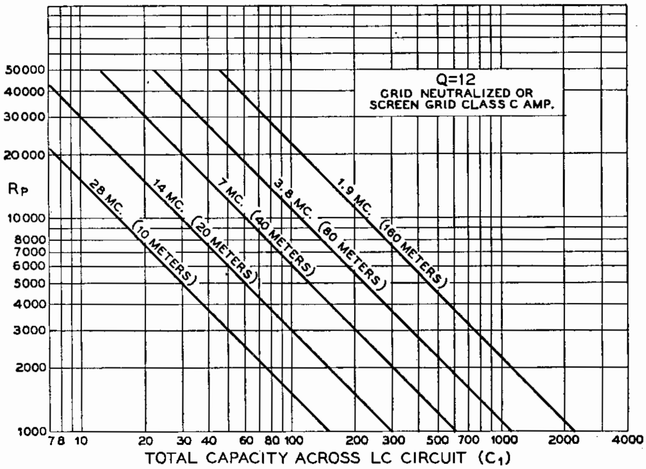
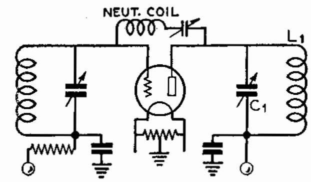


Figure 20.
SECOND AND THIRD HARMONIC OUTPUT PLOTTED AGAINST TANK CIRCUIT Q.

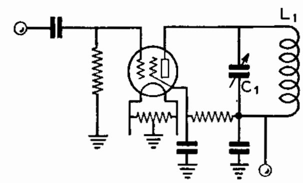
Effect of Loading on Q. The Q of a circuit depends upon the resistance in series with the capacitance and inductance. This series resistance is very low for a low-loss coil not loaded by an antenna circuit. The value of Q may be from 100 to 200 under these conditions. Coupling an antenna circuit has the effect of increasing the series resistance, though in this case the power is



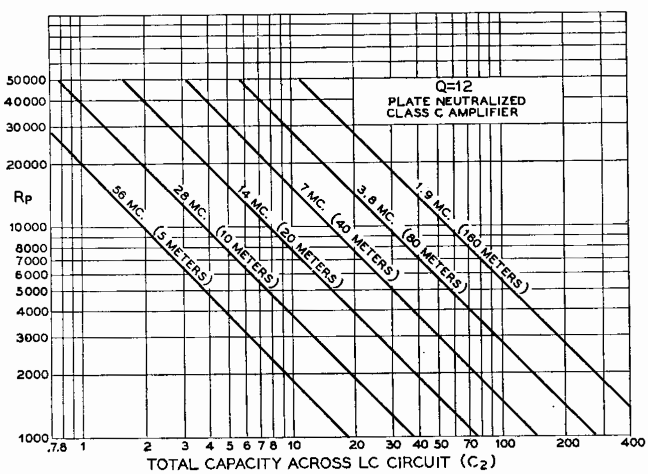


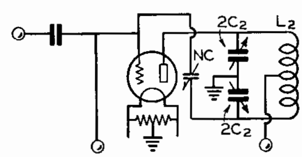
Ⓐ COIL NEUTRALIZED AMPLIFIER

Figure 21.



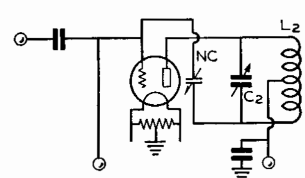
Ⓑ SCREEN GRID AMPLIFIER





Ⓐ

Figure 22.



Ⓑ PLATE NEUTRALIZED AMPLIFIERS

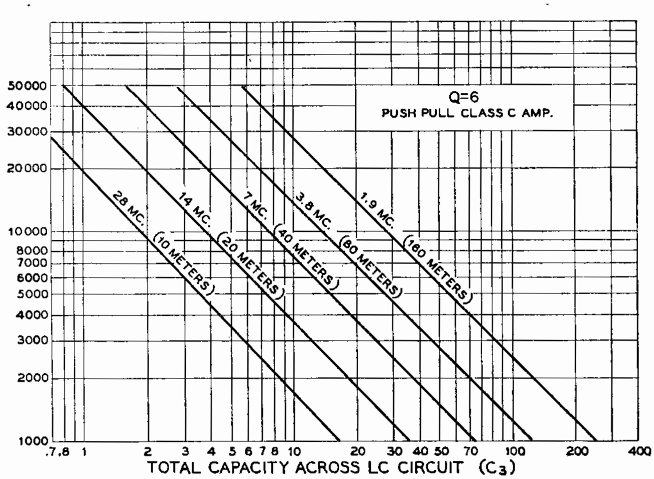
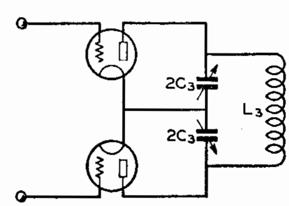


Figure 23.



PUSH PULL AMPLIFIER

consumed as useful radiation by the antenna. Mathematically, the antenna increases the value of R in the expression $Q = \omega L/R$ where L is the coil inductance and ω is the term $2\pi f$, f being in cycles per second.

The antenna coupling can be varied to obtain any value of Q from 3 to values as high as 100 or 200. However, the value of $Q = 12$ (or $Q = 20$ if desired) will not be obtained at normal values of d.c. plate current in the class C amplifier tube unless the C-to-L ratio in the tank circuit is correct for that frequency of operation.

The values of C_1, C_2 and C_3 shown in figures 21, 22 and 23 are for the total capacity across the inductance. This includes the tube interelectrode capacities, distributed coil capacity, wiring capacities and tuning condenser capacity. If a split-stator condenser is used, the effective capacity is equal to half of the value of each section since the two sections are in series across the tuned circuit. The total stray capacities range from approximately 2 up to 30 $\mu\mu$fd. and largely depend upon the type of tube or tubes used in the class C amplifier.

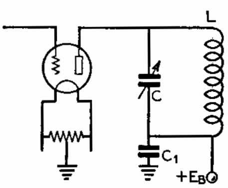


Figure 24.

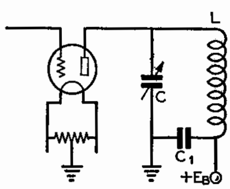


Figure 25.

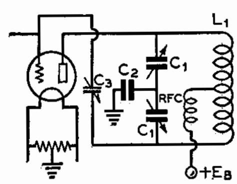


Figure 26.

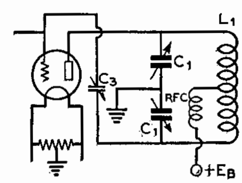


Figure 27.

In the push-pull circuit of figure 23, each tube works on a portion of each half cycle so less storage or flywheel effect is needed and a value of $Q = 6$ may be used instead of $Q = 12$.

The values of R_p are easily calculated by dividing the d.c. plate supply voltage by the total d.c. plate current (expressed in amperes). Correct values of total tuning capacity are shown in the charts for the different amateur bands. The shunt stray capacity can be estimated closely enough for all practical purposes. The coil inductance should then be chosen which will produce resonance at the desired frequency with the total calculated tuning capacity.

The capacities shown are the minimum recommended values and they should be increased 50% to 100% for modulated class C amplifiers where economically feasible. The values shown in the charts are sufficient for c.w. operation of class C amplifiers. It is again emphasized that these values are *total capacities* across the tank circuit, and should not be considered as the capacity *per section* for a *split-stator* condenser. If a split-stator condenser is to be used, the *per section* capacity should be *twice* that indicated by the charts.

Tuning Condenser Air Gap

Plate-Spacing Requirements for Various Circuits and Plate Voltages. In determining condenser air gaps the peak r.f. voltage impressed across the condenser is the important item, since the experimental and practical curves of air gap versus peak volts as published by the Allen D. Cardwell Mfg. Corp. may be applied to any condenser with polished plates having rounded edges. Typical peak breakdown voltages for corresponding air gaps are listed in the table. These values can be used in any circuit. The problem is to find the peak r.f. voltage in each case and this can be done quite easily.

The r.f. voltage in the plate circuit of a class C amplifier tube varies from nearly zero to twice the d.c. plate voltage. If the d.c. voltage is being 100 per cent modulated by an audio voltage, the r.f. peaks will reach four times the d.c. voltage. These are the highest values reached in any type of loaded amplifier: a class B linear, class C grid- or plate-modulated or class C c.w. amplifier. The circuits shown in figures 25 and 27 require a tuning condenser with plate spacing which will have an r.f. peak breakdown rating at least equal to 2 times or 4 times the d.c. plate voltage for c.w. and plate-modulated amplifiers respectively.

It is possible to reduce the air gap to one-half by connecting the amplifier so that the d.c. plate voltage does not appear across the tuning condenser. This is done in figures 24 and 26. These circuits should always be used in preference to those of figures 25 and 27 since the tuning condenser is only about one-fourth as large physically for the same capacity. Consequently, it is proportionately less expensive.

The peak r.f. voltage of a plate-modulated class C amplifier varies at 100% modulation from nearly zero to four times E_b, the d.c. plate voltage, but only one-half of this voltage is applied across the tuning condensers of figures 24 and 26. For a class B linear, class C grid-modulated or c.w. amplifier, the r.f. voltage across the tube varies from nearly zero up to twice E_b. The r.f. voltage is an a.c. voltage varying from zero to a positive and then to a negative maximum over each cycle. The fixed (mica) condenser C_1 in figure 24, and C_2 in figure 26 insulates the rotor from d.c. and allows us to subtract the d.c. voltage value from the tube peak r.f. voltage value in calculating the breakdown voltage to be expected.

This gives us a simple rule to follow for a normally-loaded plate-modulated r.f. amplifier. The peak voltage across the tuning condenser C or C_1 of figures 24 and 26 respectively will be *twice the d.c. plate voltage.* If a single-section condenser is used in figure 26, with the by-pass condenser C_2 connected to the coil center tap, the plate spacing or air gap must be twice as great as that of a split-stator condenser; so there is no appreciable saving in costs for a given capacity.

In c.w. amplifiers the air gap must be great enough to withstand a peak r.f. voltage *equal to the d.c. plate voltage,* for each section C_1 of figure 26, or, C of figure 26.

These rules apply to a loaded amplifier or buffer stage. If the latter is ever operated without an r.f. load, the peak voltages may be very much greater—by as much as two or three times in ordinary LC circuits. For this reason no amplifier should be operated without load when anywhere near normal d.c. plate voltage is applied.

A factor of safety in the air-gap rating should be applied to insure freedom from r.f. flashover. This is especially true when using the circuits of figures 25 and 27; in these circuits the plate supply is shorted when a flashover occurs. Knowing the peak r.f. voltage, an air gap should be chosen which will be about 100% greater than the breakdown rating. The air gaps listed will break down at the approximate peak voltages in the table. If the circuits are of the form shown in figures 25 and 27, the peak voltages across the condensers will be nearly twice as high and twice as large an air gap is needed. The fixed condensers, usually of the mica type, shown in figures 24 and 26, must be rated to withstand the d.c. plate voltage plus any audio voltage. This condenser should be rated at a d.c. working voltage of at least *twice the d.c. plate supply in a plate modulated amplifier* and at least *equal to the d.c. supply* in any other type of r.f. amplifier.

Push-Pull Stages. The circuits of figures 26 and 27 apply without any change in calculations to push-pull amplifiers. Only one tube is supplying power to the tuned circuit at any given instant, each one driving a part of each half cycle. The different value of Q and increased power output increase the peak voltages slightly but for all practical purposes, the same calculation rules may be employed.

These rules are based on average amateur design for any form of r.f. amplifier

BREAKDOWN RATINGS OF COMMON PLATE SPACINGS

AIR-GAP IN INCHES	PEAK VOLTAGE BREAKDOWN
.030	750
.050	1500
.070	3000
.078	3500
.084	3800
.100	4150
.144	5000
.175	5700
.200	6200
.250	7200
.300	8200
.350	9250
.375	10,000
.500	12,000

Recommended Air gap (approx. 100% factor of safety) for the circuits of figures 24 and 26. Spacings should be multiplied by 1.5 for same factor of safety with circuits of figures 25 and 27.

D.C. PLATE VOLTAGE	C. W.	PLATE MOD.
400	.030	.050
600	.050	.070
750	.050	.100
1000	.070	.084
1250	.070	.144
1500	.078	.200
2000	.100	.250
2500	.175	.375
3000	.200	.500
3500	.250	.600

with a recommended factor of safety of 100% to prevent flashover in the condenser. This is sufficient for operation into normal loads at all times, providing there are no freak parasitic oscillations present. The latter sometimes cause flashover across air gaps which should ordinarily stand several times the normal peak r.f. voltages. This is especially true of low-frequency parasitics.

The actual peak voltage values of a stable, loaded r.f. amplifier are somewhat less than the calculations indicate, which gives an addictional factor of safety in the design.

Parasitic Oscillation in R.F. Amplifiers

Parasitics are undesirable oscillations either of very high or very low frequencies which occur in radio-frequency amplifiers.

They may cause additional signals (which are often rough in tone), other than normal harmonics, hash on each side of a modulated carrier, key clicks, voltage breakdown or flashover, instability or inefficiency, and shortened life or failure of the tubes. They may be damped and stop by themselves after keying or on modulation cycles, or they may be undamped and built up during ordinary unmodulated transmission, continuing if the excitation is removed. They may be at audio or radio frequency, in either type of amplifier (though only the r.f. amplifier is treated in this discussion). They may result from series or parallel resonant circuits of all types including the dynatron. Due to the neutralizing lead length or the nature of most parasitic circuits, the amplifier usually is not neutralized for the parasitic frequency.

Sometimes the fact that the plate supply is keyed obscures parasitic oscillations that might be very severe if the plate voltage were left on and only the excitation removed.

In some cases, an all-wave receiver will prove helpful in finding out if the amplifier is without spurious oscillations, but it may be necessary to check from one meter on up, to be perfectly sure. A normal harmonic is weaker than the fundamental but of good tone; a strong harmonic or a rough note at any frequency generally indicates trouble.

Low-Frequency Parasitics. One type of unwanted oscillation often occurs in shunt-fed circuits in which the grid and plate chokes resonate, coupled through the tube's inter-electrode capacity. It can also happen with series feed. This oscillation is generally at a lower frequency than the desired one and causes additional carriers to appear, spaced from twenty to a few hundred kilocycles on either side of the main wave. One cure is to change the type of feed in either the grid or plate circuit or to eliminate one choke. Another is to use much less inductance in the grid choke than in the plate choke, or to replace the grid choke by a wire-wound resistor if the grid is series fed. In a class C stage with grid-leak bias, no r.f. choke is required if the bias is series fed.

This type of parasitic may take place in push-pull circuits, in which case the tubes are effectively in parallel for the parasitic and the neutralization is not effective. The grids or plates can be connected together without affecting the undesired oscillation; this is a simple test for this type of parasitic oscillation.

Parallel Tubes. A very high frequency inter-tube oscillation often occurs when tubes are operated in parallel. Noninductive damping resistors or manufactured parasitic suppressors in the grid circuit, or short interconnecting grid leads together with small plate choke coils, very likely will prove helpful.

Tapped Inductances. When capacity coupling is used between stages, particularly when one of the stages is tapped down from the end of the coil, additional parasitic circuits are formed because of the multiple resonant effects of this complex circuit. Inductive or link coupling permits making adjustments without forming these undesired circuits. Likewise, a condenser tapped across only part of an inductance, for bandspread tuning or capacity loading, makes the situation more complex.

Multi-Element Tubes. Screen-grid, pentode, and beam tetrode tubes may help to eliminate parasitic circuits by using no neutralization, but their high gain occasionally makes parasitic oscillation easy, particularly when some form of input-output coupling exists. Furthermore, the by-pass circuit from the additional elements to the filament must be short and effective, particularly at the higher frequencies, to prevent undesired internal coupling. At the high frequencies, a variable screen by-pass condenser at some settings may improve the internal shielding without causing a new parasitic oscillation. A blocking (relaxation) effect may occur if the screen is fed through a series resistor. The screen circuit can, of course, act as the plate in a tuned-grid tuned-plate oscillation that can be detuned or damped at the control grid terminal.

Crystal Stages. Crystal oscillators are seldom suspected of parasitic oscillation troubles, but are often guilty. Ordinary as well as parasitic circuit coupling between the

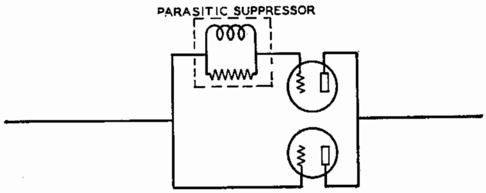


Figure 28.

Showing the use of a parasitic suppressor in series with one grid of a pair of paralleled tubes. In a push-pull amplifier which develops parasitics, the parasitic suppressor can be connected in series with the lead from the grid tank circuit to the grid of one of the tubes.

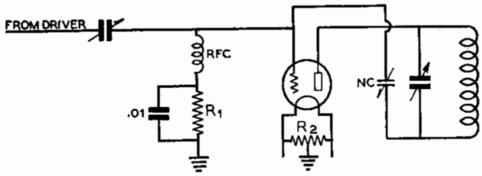


Figure 29.
GRID LEAK BIASED STAGE.
Showing how a resistor may be connected in series with the grid return lead to obtain bias due to the flow of rectified grid current through the resistor.

grid and plate circuits should be held to a minimum by separating or shielding the grid and plate leads, and by reducing the area of the loop from the grid through the crystal holder to the filament. Keeping the grid circuit short, even adding a small choke coil of a few turns in the plate lead next to the tube, will probably eliminate the possibility of high-voltage series-tuned parasitics.

Parasitic Suppressors. The most common type of parasitic is of the u.h.f. type, which fortunately can usually be dampened by inserting a parasitic suppressor of the type illustrated in figure 28 in the grid lead, or in *one* grid lead of either a push-pull or parallel tube amplifier.

Grid Bias

Radio-frequency amplifiers require some form of *grid bias* for proper operation. Practically all r.f. amplifiers operate in such a manner that plate current flows in the form of short peaked impulses which have a duration of only a fraction of an r.f. cycle. The plate current is cut off during the greater part of the r.f. cycle, which makes for high efficiency and high power output from the tubes, since there is no power being dissipated by the plates during a major portion of each r.f. cycle. The grid bias must be sufficient to cut off the plate current, and in very high efficiency class C amplifiers this bias may be several times the cutoff value. Cutoff bias, it will be recalled, is that value of grid voltage which will reduce the plate current to zero, and the method for calculating it has been indicated previously. This theoretical value of cutoff will not reduce the plate current completely to zero, due to the variable-μ tendency which is characteristic of all tubes as the cutoff point is approached. This factor, however, is of no importance in practical applications.

Class C Bias. Radiophone class C amplifiers should be operated with the grid bias adjusted to values between two and three times cutoff at normal values of d.c. grid current to permit linear operation (necessary when the stage is plate-modulated). C.w. telegraph transmitters can be operated with bias as low as cutoff, if limited excitation is available and high plate efficiency is not a factor. In a c.w. transmitter, the bias supply or resistor should be adjusted to the point which will allow normal grid current to flow for the particular amount of grid driving r.f. power available. This form of adjustment will allow more output from the under-excited r.f. amplifier than when twice cutoff, or higher bias is used with low values of grid current.

Grid-Leak Bias. A resistor can be connected in the grid circuit of an r.f. amplifier to provide grid-leak bias. This resistor R_1 in figure 29 is part of the d.c. path in the grid circuit.

The r.f. excitation is applied to the grid circuit of the tube. This causes a pulsating d.c. current to flow through the bias supply lead and any current flowing through R_1 produces a voltage drop across that resistance. The grid of the tube is positive for a short duration of each r.f. cycle, and draws electrons from the filament or cathode of the tube during that time. These electrons complete the circuit through the d.c. *grid return.* The voltage drop across the resistance in the grid return provides a *negative bias* for the grid. The r.f. chokes in figures 29, 30, 31, and 32 prevent the r.f. excitation from flowing through the bias supply, or from being short-circuited to ground. The by-pass condenser across the bias source proper is for the purpose of providing a low impedance path for the small amount of stray r.f. energy which passes through the r.f. choke.

Grid-leak bias automatically adjusts itself even with fairly wide variations of r.f. excitation. The value of grid-leak resistance should be such that normal values of grid current will flow at the maximum available amount of r.f. excitation. Grid-leak bias cannot be used for grid-modulated or linear amplifiers in which the average d.c. grid current is constantly varying with modulation.

Grid-leak bias alone provides no protection against excessive plate current in case of failure of the crystal oscillator, or failure of any other source of r.f. grid excitation. A *C-battery* or *C-bias* supply can be connected in series with the grid leak, as shown in figure 30. This additional C-bias should at least be made equal to cutoff bias. This will protect the tube in the event of failure of grid excitation.

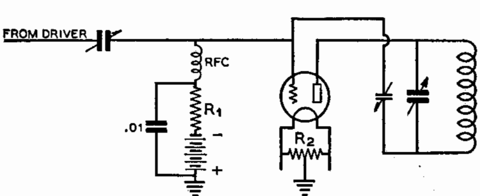


Figure 30.
GRID LEAK AND BATTERY BIAS.
A battery may be added to the grid-leak bias system of figure 30 to provide protection in case of excitation failure.

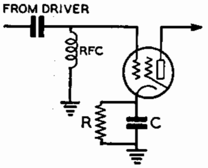
Cathode Bias. A resistor can be connected in series with the cathode or center-tapped filament lead of an amplifier to secure *automatic bias*. The plate current flows through this resistor, then back to the cathode or filament, and the voltage drop across the resistor can be applied to the grid circuit by connecting the grid bias lead to the grounded, or power supply end of the resistance R, as shown in figure 31.

The grounded (B-minus) end of the cathode resistor is negative relative to the filament by an amount equal to the voltage drop across the resistor. The value of resistance must be so chosen that the desired plate current flowing through the resistor will bias the tube for proper operation at that plate current.

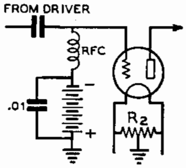
This type of bias is used more extensively in audio-frequency than in radio-frequency amplifiers. The voltage drop across the resistor must be subtracted from the total plate supply voltage when calculating the power input to the amplifier, and this loss of plate voltage in an r.f. amplifier may be excessive. A class A audio amplifier is biased only to approximately one-half cutoff, whereas an r.f. amplifier may be biased to twice cutoff, or more, and thus the plate supply voltage loss may be a large percentage of the total available voltage when using low- or medium-μ tubes.

Separate Bias Supply. C-batteries or an external C-bias supply sometimes are used for grid bias of an amplifier, as shown in figure 32.

Battery bias gives very good voltage regulation and is satisfactory for grid-modulated or linear amplifiers, which operate nearly at zero grid current. In the case of class C amplifiers which operate with high grid current, battery bias is not very satisfactory.



CATHODE BIAS
FIGURE 31



BATTERY BIAS
FIGURE 32

A resistor in the cathode lead gives cathode, or "automatic" bias as shown in figure 31. The voltage drop across the cathode resistor due to the flow of plate and grid current is applied to the grid in the form of negative bias. Figure 32 shows the use of a battery only as bias—this arrangement is suitable only for stages which do not draw over about 15 ma. of grid current.

This d.c. current has a charging effect on the dry batteries; after a few months of service the cells will become unstable, bloated and noisy.

A separate a.c. operated power supply can be used as a substitute for dry batteries. The bleeder resistance across the output of the filter can be made sufficiently low in value that the grid current of the amplifier will not appreciably change the amount of negative grid-bias voltage. This type of bias supply is used in class B audio and class B r.f. linear amplifier service where the voltage regulation in the C-bias supply is important. For a class C amplifier it is not so important, and an economical design of components in the power supply therefore can be utilized. However, in a class C application the bias voltage must be adjusted with normal grid current flowing as the grid current will raise the bias when it is flowing through the bias-supply bleeder resistance.

Interstage Coupling

Energy can be coupled from one circuit in a transmitter into another in the following ways: *capacitive coupling, inductive coupling* or *link coupling.* The latter is a special form of inductive coupling. The choice of a coupling method depends upon the purpose for which it is to be used.

Capacitive Coupling. Capacitive coupling between an amplifier or doubler circuit and a preceding driver stage is shown in figure 33.

The coupling condenser, C, isolates the d.c. plate supply from the next grid and provides a low impedance path from the r.f. energy between the tube being driven and the driver tube. This method of coupling is simple and economical for low-power amplifier or exciter stages, but has certain disadvantages. The grid leads in an amplifier should be as short as possible, but this is difficult to attain in the physical arrangement of a high-power amplifier with respect to a capacitively-coupled driver stage.

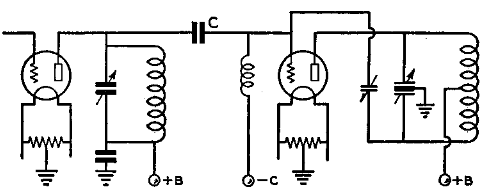


Figure 33.
CAPACITIVE INTERSTAGE COUPLING.

Disadvantages of Capacity Coupling. The r.f. choke in series with the C-bias supply lead must offer an extremely high impedance to the r.f. circuit, and this is difficult to obtain when the transmitter is operated on several harmonically related bands. Another disadvantage of capacitive coupling is the difficulty of adjusting the load on the driver stage. Impedance adjustment can be accomplished by tapping the coupling lead a part of the way down on the plate coil of the tuned stage of the driver circuit. However, when this lead is tapped part way down on the coil, a *parasitic oscillation* tendency becomes very troublesome and is difficult to eliminate. If the driver stage has sufficient power output so that an impedance mismatch can be tolerated, the condenser C in figure 33 can be connected directly to the top of the coil, and made small enough in capacity for the particular frequency of operation that not more than normal plate current is drawn by the driver stage.

The grid circuit impedance of a class C amplifier may be as low as a few hundred ohms in the case of a high-μ tube, and may range from that value up to a few thousand ohms for low-μ tubes.

Capacitive coupling places the grid-to-filament capacity of the driven tube directly across the driver tuned circuit, which reduces the LC ratio and sometimes makes the r.f. amplifier difficult to neutralize because the additional driver stage circuit capacities are connected into the grid circuit. Difficulties from this source can be partially eliminated by using a center-tapped or split-stator tank circuit in the plate of the driver stage and capacity coupling to the opposite end from the plate. This method places the plate-to-filament capacity of the driver across one half of the tank and the grid-to-filament capacity of the following stage across the other half. This type of coupling is shown in figure 34.

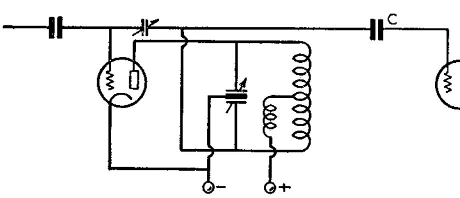


Figure 34.
BALANCED CAPACITIVE COUPLING.
This type of capacitive interstage coupling helps to equalize the capacities across the two sides of the driver tank circuit.

Capacitive coupling can be used to advantage in reducing the total number of tuned circuits in a transmitter so as to conserve space and cost. It also can be used to advantage between stages for driving tetrode or pentode amplifier or doubler stages. These tubes require relatively small amounts of grid excitation.

Inductive Coupling. The r.f. amplifier often is coupled to the antenna circuit by means of *inductive coupling,* which consists of two coils electromagnetically coupled to each other. The antenna tuned circuit can be of the series-tuned type, such as is illus-

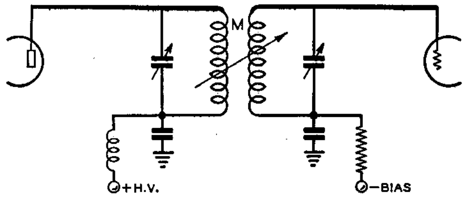


Figure 35.
INDUCTIVE INTERSTAGE COUPLING.

trated for *Marconi*-type 160-meter antennas in the chapter on *Antennas.* Parallel resonant circuits sometimes are used, as shown in figure 35, in which the antenna feeders are connected across the whole or part of the secondary circuit.

The degree of coupling is controlled by varying the mutual inductance of the two coils, which is accomplished by changing the spacing between the coils.

Inductive coupling also is used extensively for coupling r.f. amplifiers in radio receivers, and occasionally in transmitting r.f. amplifier circuits. The mechanical problems involved in adjusting the degree of coupling in a transmitter make this system of limited practical value.

Link Coupling. A special form of inductive coupling which is applied to radio transmitter circuits is known as *link coupling.* A low impedance r.f. transmission line, commonly known as a *link,* couples the two tuned circuits together. Each end of the line is terminated in one or more turns of wire, or *loops,* wound around the coils which are being coupled together. These loops should be coupled to each tuned circuit at the point of zero r.f. potential. This *nodal* point is the center of the tuned circuit in the case of plate-neutralized or push-pull amplifiers, and at the positive-B end of the tuned circuit in the case of screen grid and grid-neutralized amplifiers.

The nodal point in an antenna tuned circuit depends upon the type of feeders, and the node may be either at the center or at one end of the tuned circuit.

The nodal point in tuned grid circuits is at the C-bias or grounded end of plate-neutralized or screen-grid r.f. amplifiers, and at the center of the tuned grid coil in the case of push-pull or grid-neutralized amplifiers. The link coupling turns should be as close to the nodal point as possible. A ground connection to one side of the link is used in special cases where harmonic elimination is important, or where capacitive coupling between two circuits must be minimized.

Typical link coupled circuits are shown in figures 36 and 37.

Some of the advantages of link coupling are listed here:

(1) It eliminates coupling taps on tuned circuits.
(2) It permits the use of series power supply connections in both tuned grid and tuned plate circuits, and thereby eliminates the need of r.f. chokes.
(3) It allows separation between transmitter stages of distances up to several feet without appreciable r.f. losses.

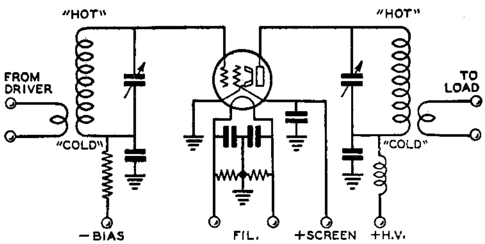
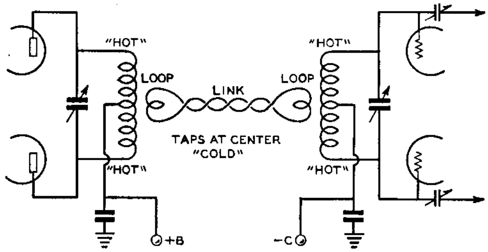


Figure 36.
LINK COUPLED CIRCUIT.
Showing link coupling into and out of a single-ended beam-tetrode amplifier stage. The coupling links should be placed at the "cold" or low-potential ends of the grid and plate coils.

Figure 37.
PUSH-PULL LINK COUPLING.
When link coupling is used between push-pull stages or between "split" tank circuits, the coupling loops are placed at the center of the coils.



(4) It reduces capacitive coupling and thereby makes neutralization more easily attainable in r.f. amplifiers.
(5) It provides semiautomatic impedance matching between plate and grid tuned circuits, with the result that greater grid swing can be obtained in comparison to capacitive coupling.
(6) It effectively reduces harmonic radiation when a final amplifier is coupled to a

tuned antenna circuit, due to the additional tuned circuit and, particularly, it eliminates capacitive coupling to the antenna.

The link coupling line and loops can be made of no. 18 or 20 gauge push back wire for coupling low-power stages. High-power circuits can be link-coupled by means of no. 8 to no. 12 rubber-covered wire, twisted low-impedance antenna-feeder wire, concentric lines or open-wire lines of no. 12 or no. 14 wire spaced ¼ to ½ inch.

The impedance of a link coupling line varies from 75 to 200 ohms, depending upon the diameter of the conductors and the spacing between them.

Radio-Frequency Chokes

Radio-frequency chokes are connected in circuits for the purpose of preventing r.f. energy from being short-circuited, or escaping into power supply circuits. They consist of inductances wound with a large number of turns, either in the form of a solenoid or universal pie-winding. These inductances are designed to have as much inductance and as little distributed or shunt capacity as possible, since the capacity by-passes r.f. energy. The unavoidable small amount of distributed capacity resonates the inductance, and this frequency normally should be lower than the frequency at which the transmitter or receiver circuit is operating. R.f. chokes for operation on several harmonically related bands must be designed carefully so that the impedance of the choke will be extremely high (several hundred thousand ohms) in each of the bands.

The r.f. choke is resonant to the harmonics of its fundamental resonant frequency; however, the *even* harmonics have a very low impedance, so that an r.f. choke designed for maximum impedance in the 80-meter amateur band would not be satisfactory for operation in the 40-meter band. The harmonic resonance points of the r.f. choke usually are made to fall between frequency bands, so that a reasonably high value of impedance is obtained on all amateur bands. The d.c. current which flows through the r.f. choke largely determines the size of wire to be used in the windings. The inductance of r.f. chokes for very short wave-lengths is much less than for chokes designed for broadcast and ordinary short-wave operation, so that the impedance will be as high as possible in the desired range of operation. A very high inductance r.f. choke has more distributed capacity than a smaller one, with the result that it will actually offer *less* impedance at very high frequencies.

Shunt and Series Feed. Direct-current grid and plate connections are made either by *series* or *parallel feed* systems. Simplified forms of each are shown in figures 38 and 39.

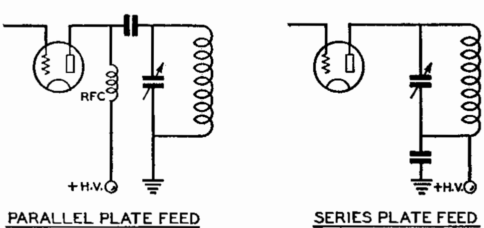


Figure 38.

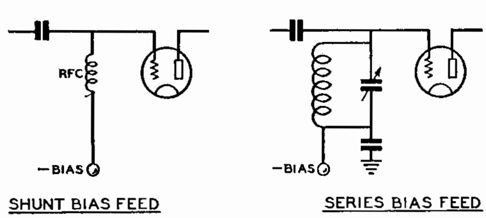


Figure 39.

Series feed can be defined as that in which the d.c. connection is made to the grid or plate circuit at a point of very low r.f. potential. Shunt feed always is made to a point of high r.f. voltage and always requires a high impedance r.f. choke or resistance in the connection to the high r.f. point to prevent loss of r.f. power.

Parallel and Push-Pull Tube Circuits

The comparative r.f. power output from parallel or push-pull operated amplifiers is the same if proper impedance matching is accomplished, if sufficient grid excitation is available in both cases, and if the frequency of measurement is considerably lower than the frequency limit of the tubes.

Parallel Operation. Operating tubes in parallel has some advantages in transmitters designed for operation on 40, 80 and 160 meters, or for broadcast band operation.

Only one neutralizing condenser is required for parallel operation, as against two for push-pull. However, on wavelengths below 40 meters, parallel tube operation is not advisable because of the unbalance in capacity across the tank circuits. Low-C types of vacuum tubes can be connected in parallel with less difficulty than the high-C types, in which the combined interelectrode capacities might be quite high in the parallel connection.

Push-Pull Operation. The push-pull connection provides a well-balanced circuit insofar as miscellaneous capacities are concerned; in addition the circuit can be neutralized more easily, especially in high-frequency amplifiers. The L/C ratio in a push-pull amplifier can be made higher than in a plate-neutralized parallel-tube operated amplifier. Push-pull amplifiers, when perfectly balanced, have less second-harmonic output than parallel or single-tube amplifiers. In actual practice, undesired capacitive coupling and circuit unbalance tend to offset the theoretical harmonic-reducing advantage of push-pull r.f. circuits.

Transmitter Keying

The carrier frequency signal from a c.w. transmitter must be broken into dots and dashes in the form of *keying* for the transmission of code characters. The carrier signal is of a constant amplitude while the key is closed, and is entirely removed when the key is open. If the change from the no-output condition to *full-output* occurs too rapidly, an undesired *key-click* effect takes place which causes interference in other signal channels. If the opposite condition of full output to no output condition occurs too rapidly, a similar effect takes place.

Excitation or Plate Voltage Keying. The two general methods of keying a c.w. transmitter are those which control either the excitation, or the plate voltage which is applied to the final amplifier. Plate voltage control can be obtained by connecting the key in the primary line circuit of the high voltage plate power supply. A slight modification of direct plate voltage control is the connection of the c.w. key or relay in the filament center-tap lead of the final amplifier. *Excitation keying* can be of several forms, such as crystal oscillator keying, buffer stage keying or blocked-grid keying.

Key Clicks. Key clicks should be eliminated in all c.w. telegraph transmitters. Their elimination is accomplished by preventing a too-rapid make-and-break of power to the antenna circuit. A gradual application of power to the antenna, and a similarly slow cessation, will eliminate key clicks. Too much lag will prevent fast keying, but fortunately key clicks can be practically eliminated without limiting the speed of manual (hand) keying. Some circuits which eliminate key clicks introduce too much time-lag and thereby add *tails* to the dots. These tails may cause the signals to be difficult to copy at high speeds.

Click Filters. Eliminating key clicks by some of the key-click filter circuits illustrated in the following text is not certain with every individual transmitter. The constants in the time-lag and spark-producing circuits depend upon the individual characteristics of the transmitter, such as the type of filter, power input and various circuit impedances. All keying systems have one or more disadvantages, so that no particular method can be recommended as an ideal one. An intelligent choice can be made by the reader for his particular transmitter requirements by carefully analyzing the various keying circuits.

Primary Keying. Key clicks (except those arising from arcing at the key, which usually do not carry beyond a few hundred feet) can be eliminated entirely by means of primary keying, in which the key is placed in the a.c. line supply to the primary of the high voltage plate supply transformer. This method of keying also has the advantage that grid leak bias can be used in the keyed stages of the transmitter. As ordinarily applied, the plate voltage to the final amplifier is controlled by the action of the key. The filter in the high voltage rectifier circuit creates a time-lag in the application and removal of the d.c. power input to the r.f. amplifier. Too much filter will introduce too great a time lag, and add tails to the dots. If a high-power stage is keyed, the variation in load on the house-lighting circuits may be sufficient to cause blinking of the lights. A heavy-duty key or keying relay is necessary for moderate or high-power transmitters to break the inductive a.c. load of the power supply. The exciting current or surge current may be several times as high as the average current drawn by the transformer which is being keyed. This will cause difficulty from sticking key contacts or burnt points on the keying relay. This effect can be minimized by proper design of the power transformer, which should have a high primary inductance and an iron core of generous size.

Lag-Less Primary Keying Circuit. An improved primary keying circuit is shown in figure 40. This circuit makes high speed key-

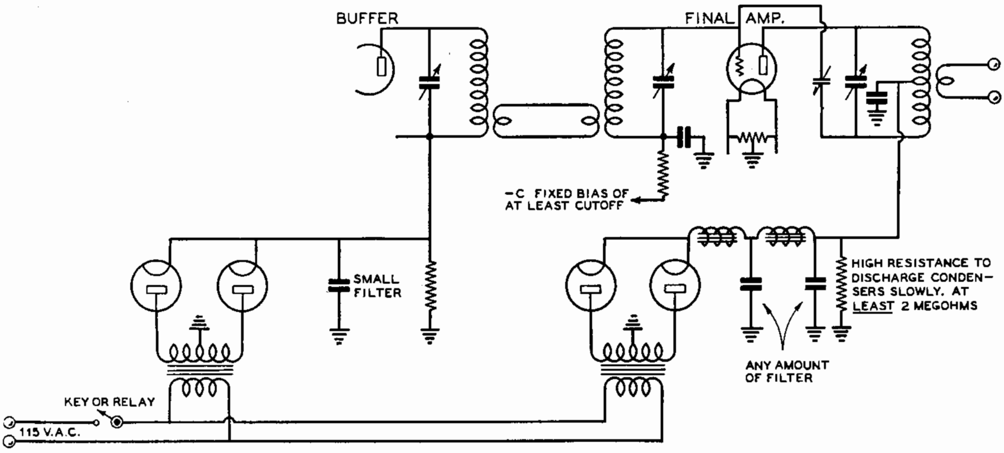


Figure 40.
IMPROVED PRIMARY KEYING WITHOUT CLICKS OR "TAILS."

ing possible, without clicks or tails, and the plate supply to the final amplifier can be very well filtered without introducing tails to the dots.

The final amplifier must have a fixed bias supply equal to more than cut off value, so that when the grid excitation from the buffer stage is removed the amplifier output will drop immediately to zero, in spite of the filter condenser's being fully charged in the final amplifier circuit. The bleeder across the final plate supply filter should have a very high resistance so that the filter condenser will hold its charge between dots and dashes. This will allow a quick application of plate voltage as soon as the grid excitation, supplied by the buffer stage, is applied to the final amplifier.

The buffer plate supply is keyed; its filter circuit consists of a single 2-μfd. filter condenser, shunted by the usual heavy-duty high-current bleeder resistor. This small filter has no appreciable time-lag, and will not add tails to the dots and dashes, but it does provide sufficient time-lag for key click elimination. The small amount of filter will not introduce a.c. hum modulation into the output of the final amplifier, because the latter is operated in class C, under saturated grid conditions. A moderate a.c. ripple in the grid excitation will not introduce hum in the output circuit under this operating condition.

Grid-Controlled Rectifiers. By the incorporation of grid-controlled rectifiers in a high-voltage power supply, one can enjoy keying that has practically all the advantages of primary keying with none of the disadvantages. The only disadvantage to this type of keying as compared to primary keying is that of the small amount of additional equipment needed and the additional expense of the special rectifiers.

Inasmuch as no power is required to block the grids, there is little sparking at the relay contacts. And because the keying is ahead of the power supply filter, the wave train or keying envelope is rounded enough that clicks and keying impacts are eliminated. In fact, it is important that no more filter be used than is required to give a good T 9 note, inasmuch as excessive filter will introduce lag and put tails on the keying. The optimum ratio and amounts of inductance and capacity in the filter will be determined by the load on the filter (plate voltage divided by plate current). With high plate voltage and low plate current (high impedance load) more inductance and less capacity should be used, and vice versa.

Of the large number of possible circuit combinations, three of the most practical are illustrated. The circuit shown in figure 41 at A is perhaps the simplest and most trouble-free, but has the disadvantage of requiring bias batteries. The relay contacts handle little power, but must be insulated from ground for the high voltage.

At B is shown the simplest method not requiring batteries. If used as shown, the bias transformer must be insulated for the full plate voltage (secondary to both primary and case). Unfortunately, b.c.l. transformers

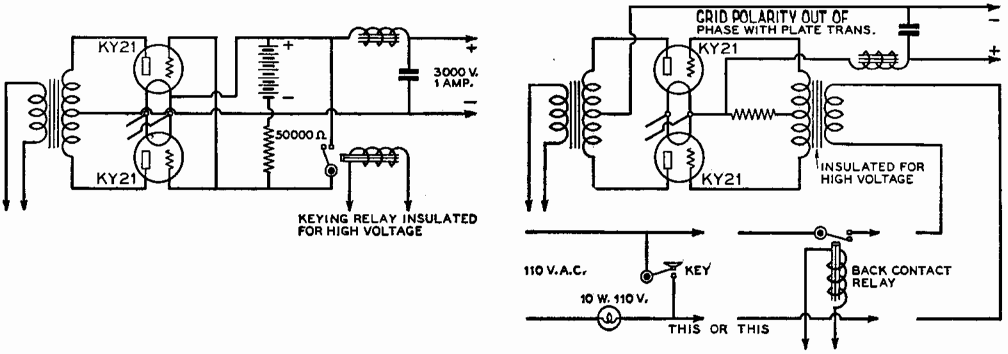


Figure 41.
GRID-CONTROLLED RECTIFIER KEYING SYSTEMS.
Two methods of using grid-controlled rectifiers for clickless, sparkless keying.

were not designed to withstand 3,000 or 4,000 volts r.m.s., either between windings or to the case. The circuit shown in figure 42 allows the use of a small broadcast-receiver type transformer for bias supply to the rectifiers. In this case the whole transformer is at the power-supply voltage above ground and it must be well insulated from metal chassis and other grounded portions of the circuit.

Blocked Grid Keying. The negative grid bias in a medium- or low-power r.f. amplifier can easily be increased in magnitude sufficiently to reduce the amplifier output to zero. The circuits shown in figures 43 and 44 represent two methods of such blocked grid keying.

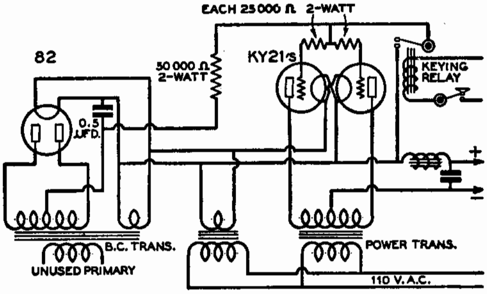


Figure 42.
ALTERNATIVE GRID-CONTROLLED RECTIFIER KEYING ARRANGEMENT.
An ordinary broadcast-receiver power transformer may be used with this circuit. The whole transformer must be well insulated from grounded parts of the circuit.

In figure 43, R_1 is the usual grid leak. Additional fixed bias is applied through a 100,000-ohm resistor R_2 to block the grid current and reduce the output to zero. As a general rule, a small 300- to 400-volt power supply with the positive side connected to ground can be used for the additional C-bias supply.

The circuit of figure 44 can be applied by connecting the key across a portion of the plate supply bleeder resistance. When the key is open, the high negative bias is applied to the grid of the tube, since the filament center tap is connected to a positive point on the bleeder resistor. Resistor R_2 is the normal bleeder; an additional resistor of from one-fourth to one-half the value of R_2 is connected in the circuit for R_1. A disadvantage of this circuit is that one side of the key may be placed at a positive potential of several hundred volts above ground, with the at-

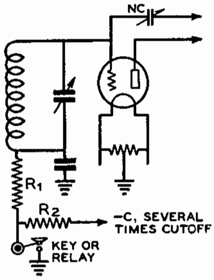


Figure 43.

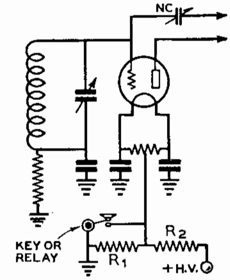


Figure 44.

ALTERNATIVE BLOCKED-GRID KEYING CIRCUITS.

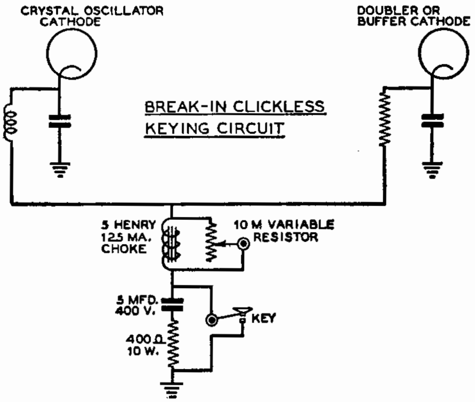


Figure 45.
BREAK-IN KEYING CIRCUIT.
This circuit arrangement may be used for break-in operation where the receiver is kept in operation on the same band as the transmitter during transmission.

tendant danger of shock to the operator. Blocked grid keying is not particularly effective for eliminating key clicks.

Oscillator Keying. A stable and quick-acting crystal oscillator may be keyed in the plate, cathode or screen-grid circuit for the purpose of minimizing key clicks and for break-in operation. This type of keying requires either fixed or cathode bias on all following r.f. stages, since the r.f. excitation is removed from all of the grid circuits. The key clicks are minimized by the presence of several tuned circuits between the antenna and crystal oscillator in a multistage transmitter. The key clicks act as sideband frequencies and are attenuated somewhat in a multistage transmitter by the resonant tuned circuits which are tuned to the carrier frequency.

If a key click filter is placed in the crystal oscillator circuit, the tone may become chirpy and tails may be added to the ends of the transmitted characters. A practical circuit for clickless keying is illustrated in figure 45, in which both the cathode of the crystal oscillator and the cathode of the next succeeding buffer or doubler circuit are connected through a key click filter.

Two tubes can be keyed very effectively with this type of circuit. The choke coil,

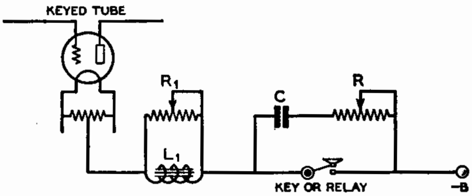


Figure 46.
CLICK FILTER CIRCUIT.
This circuit shows simple center-tap keying with an adjustable click filter to reduce interference caused by this keying method. The amount of inductance and capacity used in the filter depends upon the amount of current being keyed. Ordinarily, L_1 will be between 1 and 5 henrys, R_1 20,000 ohms, C between 0.25 and 2 μfd., and R about 2000 ohms.

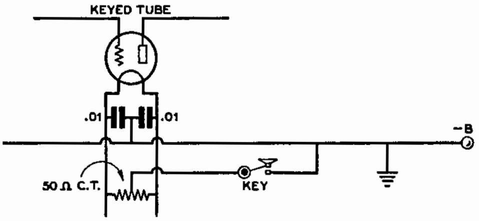


Figure 47.
ORDINARY CENTER-TAP KEYING.
The center tap of the filament transformer must not be grounded, and this transformer must not be used to supply filament voltage to any other stages.

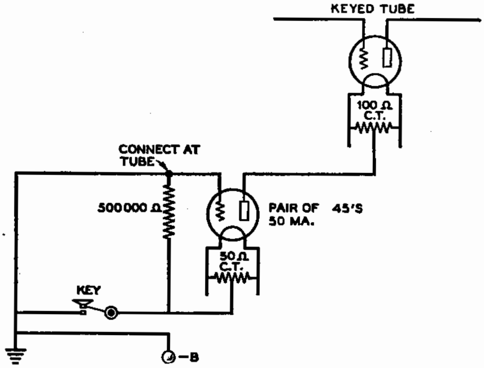


Figure 48.
VACUUM-TUBE KEYING CIRCUIT.
One of the more simple of the vacuum-tube keying circuits. Some current flows through the key in this circuit, and clicks are sometimes produced when the key is opened. Both filament transformers must be insulated from each other and from ground.

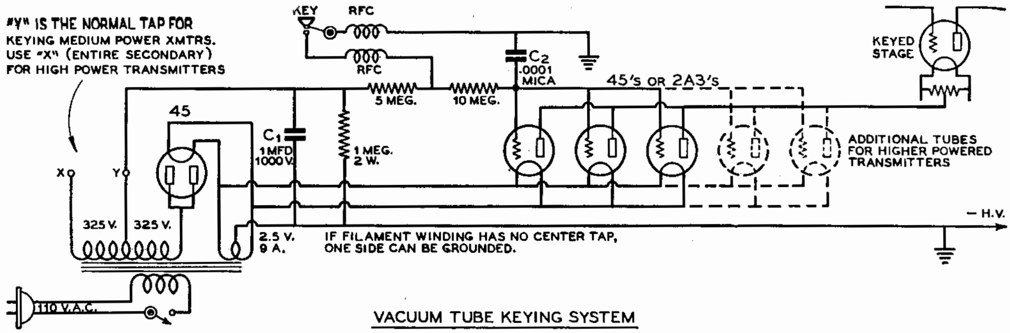


Figure 49.
VACUUM-TUBE KEYING SYSTEM.

shunted with a semi-variable resistor, provides a series inductance for slowing down the application of cathode current to the two tubes. The inductance of the choke coil can effectively be lowered to one or two henrys by shunting it with a semi-variable resistance so that the time lag will not be excessive. The 0.5-μfd. condenser and 400-ohm resistor are connected across the key contacts, as *close to the key* as possible, and these serve to absorb the spark at the telegraph key each time the circuit is opened. This effectively prevents a click at the end of each dot and dash. This same type of key click filter can be connected in the center-tap lead of a final amplifier or buffer-amplifier stage for the elimination of clicks.

Parasitics with Oscillator Keying. When keying in the crystal stage, or for that matter any stage ahead of the final amplifier, the stages following the keyed one must be absolutely stable so that parasitic or output-frequency oscillation will not occur when the excitation is rising on the beginning of each keying impulse. This type of oscillation gives rise to extremely offensive key clicks which cannot be eliminated by any type of click filter; in fact a filter designed to slow up the rate at which signal comes to full strength may only make them worse.

Center-Tap Keying. The lead from the center-tap connection to the filament of an r.f. amplifier tube can be opened and closed for keying a circuit. This opens the B-minus circuit, and at the same time opens the grid-bias return lead. For this reason the grid circuit is blocked at the same time that the plate circuit is opened, so that excessive sparking does not occur at the key contacts. Unfortunately, this method of keying applies the power too suddenly to the tube, producing a serious key click in the output circuit, which generally is coupled to the antenna. This click often can be eliminated with the key click eliminator shown in figures 45 and 46.

Vacuum Tube Keying. Center-tap keying as shown in figure 47 never should be used, because this circuit produces extremely bad key clicks. The key click filter in figure 46 always can be connected into the center-tap lead as an external unit. A more effective key click filter for the center-tap lead is made possible through the use of vacuum tubes. A simple vacuum tube keying circuit is shown in figure 48.

The keying tube is connected in series with the center-tap lead of the final r.f. amplifier. The grid of the keying tube is short-circuited to the filament when the key is closed, and the keying tube then acts as a low resistance in the center-tap lead. When the key is opened, the grid of the keying tube tends to block itself and the plate-to-filament resistance of the tube increases to a high value, which reduces the output of the r.f. amplifier approximately to zero. A more effective vacuum tube keying system is shown in figure 49.

In this system, the grids of the keying tubes are biased to a high negative potential when the key is open and to zero potential when the key is closed. The fixed bias supply to the keying tubes provides very effective keying operation. The degree of time-lag (key click elimination) can be adjusted to suit the individual operator, by varying both the capacity of the condenser which is shunted from grid to filament, and the values of the two high resistances in series with the grid and power supply leads. R.f. chokes can be connected in series with the key directly at the key terminals, to prevent the minute spark at the key contacts from causing interference in nearby broadcast receivers. These r.f. chokes are of the conventional b.c. type. There is no danger of shock to the operator when this keying circuit is used.

The small power supply for this keying circuit requires very little filter and can be of the half-wave rectifier type with a '45 tube as the rectifier. The negative voltage from this power supply only needs to be sufficient to provide cutoff bias to the type 45 keying tubes; potentials of from 100 to 300 volts are needed for this purpose. Approximately 50 milliamperes of plate current in the final amplifier should be allowed per type 45 keying tube. If the final amplifier draws 150 milliamperes, for example, three type 45 keying tubes in parallel will be required.

One disadvantage of vacuum tube keying circuits is a plate supply potential loss of approximately 100 volts, which is consumed by the keying tubes. The plate supply therefore should be designed to give an output of 100 volts more than ordinarily is needed for the r.f. amplifier. This loss of plate voltage is encountered because the plate-to-filament resistance of the type 45 tubes, at 50 milliamperes of current and zero grid potential, is approximately 2000 ohms.

Vacuum-tube keying is applicable to high-speed commercial transmitters, as well as for amateur use.